NASA CR-143813

REPORT NO. 2265. 30/459
HAC REF. NO. D2998

(NASA-CR-143813) ELECTRONICALLY STEERABLE MILLIMETER WAVE ANTENNA TECHNIQUES FOR SPACE SHUTTLE APPLICATIONS Interim Report, Sep. 1974 - Feb. 1975 (Hughes Aircraft Co.) 111 p HC $5.25 CSCL 17B G3/32

N75-29291

Unclas
33005

MARCH 1975
INTERIM REPORT FOR PERIOD SEPTEMBER 1974 - FEBRUARY 1975

ELECTRONICALLY STEERABLE MILLIMETER WAVE ANTENNA TECHNIQUES FOR SPACE SHUTTLE APPLICATIONS

W. H. KUMMER, Program Manager

Prepared for:

NASA GODDARD SPACE FLIGHT CENTER
GREENBELT, MARYLAND 20711

AEROSPACE GROUPS

HUGHES

HUGHES AIRCRAFT COMPANY
CULVER CITY, CALIFORNIA

TECHNICAL REPORT STANDARD TITLE PAGE

1. Report No.	2. Government Accession No.	3. Recipient's Catalog No.
4. Title and Subtitle Electronically Steerable Millimeter Wave Antenna Techniques for Space Shuttle Applications		5. Report Date March 1975
		6. Performing Organization Code
7. Author(s) W.H. Kummer, Program Manager		8. Performing Organization Report No. 2265.30/462
9. Performing Organization Name and Address Hughes Aircraft Company Culver City, CA 90230		10. Work Unit No.
		11. Contract or Grant No. NAS5-20726
		13. Type of Report and Period Covered Interim Report Sept.1974-Feb. 1975
12. Sponsoring Agency Name and Address NASA Goddard Space Flight Center Greenbelt, MD 20711 L. J. Ippolito, Code 951, Technical Officer		14. Sponsoring Agency Code
15. Supplementary Notes		

16. Abstract

This Interim Report describes the investigation of a large multi-function antenna aperture and related components that will perform electronic steering of one or more beams for two of the three applications envisioned: 1) communications, 2) radar, and 3) radiometry.

The array consists of a 6-meter folded antenna that fits into two pallets. The communications frequencies are 20 and 30 GHz, while the radar is to operate at 13.9 GHz.

Weight, prime power, and volumes are given parametrically.

17. Key Words (Selected by Author(s))	18. Distribution Statement
Shuttle Millimeter Waves Communication Radar	Phased Array Antennas Transmitters, Receivers Multibeam Antennas

19. Security Classif. (of this report)	20. Security Classif. (of this page)	21. No. of Pages	22. Price*
Unclassified	Unclassified		

*For sale by the Clearinghouse for Federal Scientific and Technical Information, Springfield, Virginia 22151.

PREFACE

The NASA Space Shuttle Program will provide the opportunity for a new generation of experiments in space science and technology applications. The large shuttle payload capability and the inclusion of astronaut-technologists as part of the space segment offer unique possibilities for the development of systems and experiments without the severe constraints of size, power, weight, and launch stress survival required on present space vehicles. The Millimeter-Wave Large Aperture Antenna Experiment will provide a high-gain, wideband, multi-frequency, scanning antenna system for a number of potential applications in the area of communications, propagation and radiation measurements, and high-resolution remote sensing. The systems are proposed for application to three antenna experiments: (1) a communication link in the 20- to 30-GHz region, (2) a radar system at 13.9 GHz, and (3) a set of radiometers at 10, 18, 22, 33 to 37, 55 to 60, and 94 GHz.

Although the details of the antenna designs for the three experiments are to be defined during the study, it is clear that the three experiments should share as much of the basic antenna structure and electronics as possible. Consequently, it is envisioned that the three experiments could be designed to employ an antenna subaperture common to all three. The subaperture structure would be stored in the vehicle in a folded configuration. The segments of the subaperture would contain the electronics common to all the experiments. The antenna modules would encompass the radiating aperture and the electronics unique to a system for a particular experiment. There are several problems to be considered. With a maximum of commonality, the number of electronic modules must be sufficient to provide the required set of phase shifters and down-converters for the most demanding experiment. The frequencies selected in these modules must be chosen to minimize the generation of unwanted harmonics in all systems. Since the array is large in terms of wavelength, phase shifting of the antenna modules will cause bandwidth constraints not compatible with the desired 500-MHz bandwidth. Thus, the subaperture must be divided into sections that are steerable by true-time-delay devices.

The general approach will be to convert the incoming signals to lower frequencies and perform the processing after the signal-to-noise ratio (SNR) is established. On transmit, the procedure will be reversed; either up-converter and/or solid-state amplifiers will be used.

The program is concerned with the preliminary design of the three experiments in terms of feasibility, size, weight, and cost. It is expected that several iterations will be performed in the designs to bracket the weight, size and costs within the constraints of the shuttle mission.

The most detailed designs have been applied to the communications experiment. It appears that a six-meter antenna system weighing less than 1.5 tonnes and providing 0.1° beams is feasible.

Further designs will be generated for the radar and radiometric antennas that are compatible with the designs for the communications antenna system.

ACKNOWLEDGEMENT

The personnel participating in this program include:

Mr. H. K. Blomseth
Dr. W. H. Kummer
Mr. A. F. Seaton
Dr. A. T. Villeneuve
Dr. R. J. Wagner

The typing was handled by Ms. M. Reitberger.

TABLE OF CONTENTS

TABLE OF CONTENTS (Continued)

LIST OF ILLUSTRATIONS

LIST OF ILLUSTRATIONS (Continued)

LIST OF TABLES

I. INTRODUCTION

The NASA Space Shuttle Program will provide the opportunity for a new generation of experiments in space science and technology applications. The large shuttle payload capability and the inclusion of astronaut-technologists as part of the space segment offer unique possiblities for the development of systems and experiments without the severe constraints of size, power, weight, and launch stress survival required on present space vehicles. The Millimeter-Wave Large Aperture Antenna Experiment will provide a high gain, wideband, multi-frequency, scanning antenna system for a number of potential applications in the area of communications, propagation and radiation measurements, and high-resolution remote sensing. This report presents discussions of systems for application to two antenna experiments: 1) a communication link in the 20- to 30-GHz region, and 2) a radar system at 13.9 GHz. The shuttle orbital parameters are listed in Table I-I. The experiments will be performed one at a time. The initial design parameters for the three experiments are shown in Table I-II. It will be noted that not all required constraints have been defined. For the radar system a resolution of 100 meters is desired, which indicates an antenna beamwidth on the order of 0.01 degree and an aperture size larger than 10 meters at 13.9 GHz. The designs have undergone review and changes during the course of the program as outlined below.

TABLE I-I

Orbital Parameters

Parameter	Value
Altitude	370 km
Velocity	7.68 km/sec
Round trip time	1 to 3 milliseconds

The pallets for the Space Shuttle are available in 3-meter modules, thus, a 10-meter aperture would take up from 3 to 4 pallets. Those many pallets assigned to one experiment would mean that the shuttle

MODE	COMMUNICATION LINK FREQUENCY (GHz)		RADAR FREQUENCY (GHz)		RADIOMETER FREQUENCY (GHz)					
	20	30	13.9	13.9	10	18	22	33-37	55-60	94
TRANSMITTER	20W									
RECEIVER		X								
NUMBER OF BEAMS	1,3	1,3	1	1	← 2 →					
BANDWIDTH	← 500 MHz →									
	1500 MHz (3 BEAMS)									
SCANNING	±15°	±15°	← ±15°* →							
	(CONE, 2-DIMENSIONAL)		1-DIMENSIONAL		← 1-DIMENSIONAL →					
SIDELOBES										
POLARIZATION	LINEAR	ORTHOGONAL LINEAR	LINEAR		LINEAR BOTH HORIZONTAL AND VERTICAL IN TRACK PLANE					
BEAMWIDTH	0.1°	0.1°			← 0.3° →					
RESOLUTION			100 m							

*FOR A SWATH WIDTH OF 100 KM

TABLE I-II

Initial Antenna Design Parameters

flight would have to be dedicated to this one experiment. As the study progressed, it was felt that a better compromise would be to limit this experiment to the shared use of two pallets. This would limit the basic antenna to a diameter of about 6 meters.

The extent of inclusion of the astronauts as part of the experiment has also changed during the course of this program. The original concept included the active participation of the astronaut in the deployment and reployment of the antenna system. Also included was the reconfiguration of the aperture during flight so that all three experiments would be stowed on-board and flown during one extended mission. Since it became apparent that the astronaut's time would be limited when exposed to the outdoor environment it was decided to

1) Make the experiment self-deployable,
2) To fly only one experiment (communication, radar, or radiometer) per mission,
3) Astronauts EVA participation limited to observation and emergency repair,
4) Have array stow itself on command,
5) Each experiment be completely assembled and checked-out before launch.

These changes have the following effects on the experiments:

1) Increased deployment problems since tie-down devices have to operate with minimum of human intervention,
2) Decreased need for weight on-board since stowage of two of the three experiments is not needed,
3) Reduced constraints on interchangeability, reduced need for quick disconnect plugs, modular construction, storage of components in pallets.

The antenna designs will be taken up first. A 6-meter communication array is postulated in the discussion to follow.

II. ANTENNA DESIGNS

The antenna for the millimeter wave communications experiment must produce 0.1° beams at 20 and 30 GHz* for transmission and reception, respectively. Two-dimensional electronic scanning is required over a cone of 15° radius. These factors imply a very large number of elements, each of which must be properly phased. Hence, potential solutions to the design problem include methods to reduce the number of elements used. It is desired that this experiment share as much basic hardware as possible with the other two experiments; hence, that may be a restraint on the possible solutions to be considered.

The required scan angle of $\pm 15^{\circ}$ indicates that the elements or modules of the array cannot be larger than about $1.7\,\lambda_0$ in diameter. Calculations show that at 20 GHz over 50,000 such elements would be required to fill a 6-meter diameter aperture. Since this is an unrealistically large number, methods have been considered for reducing it.

The most obvious method of reducing the number of elements is by thinning. The remaining elements must then be randomly spaced to reduce the magnitude of grating lobes that result. The average level to which these lobes can be reduced depends on the number of elements remaining after the thinning is done. An approximate expression relating the two factors is:

$$\mathrm{dB} = 10 \log N.$$

Thus, it should be possible to keep the average grating lobe down to -37 dB if we retain on the order of 5,000 elements; and -34 dB if 2,500 elements are retained. This is lower than the -30 dB assumed for the design, so should not degrade performance significantly.

---

* The beamwidth is 0.11° at 30 GHz and 0.17° at 20 GHz. For purposes of discussion the 0.1° number will be used.

Another method of reducing the number of elements is by space tapering. This is a type of thinning also, but it is applied in such a manner that the amount of thinning is a function of the distance from the center of the array to the element. Therefore, an amplitude distribution across the array can be approximated by this type of thinning; hence, the name space tapering. In general, it is not possible to eliminate a large percentage of the elements in this fashion, but it alleviates the probelm somewhat and makes it possible to excite all the remaining elements equally while achieving a tapered amplitude distribution.

Both of the above methods are used in an array computer program which has the capability of randomly placing elements in an aperture. Since the number of elements will still be large, the program has been streamlined so that it will run in the most efficient manner possible.

A. Communication Antenna

1.0 Array Patterns

The array computer program mentioned above has been modified so that it will run in the most efficient manner possible. This was accomplished by removing all aspects of the program that did not directly contribute to the desired output. Computer core space was conserved by changing all possible variable arrays into simple variables. Statement arithmetic was checked and changed when it was found that the statement could be reworked in such a way as to reduce CPU time. Even with all of these simplifications, a CPU time of approximately 3 minutes on the IBM 370 is required to compute a pattern for the communications antenna. This results primarily because of the large number of elements in the array (4,000), and the large number of points that must be computed (800 over an angle of $\pm 20^{\circ}$) to insure locating any high sidelobes that may be present.

A preliminary design has been completed for the communications antenna portion of the experiment. For this design a circular aperture was assumed because circular distributions are more efficient than square ones from the viewpoint of generating uniformly low sidelobes with a given number of elements. The array was thinned by a factor of 85% by increasing the nominal distance between elements by 2.6 times the normal inter-element spacing. The locations of the elements were then perturbed in a random fashion about their nominal locations to suppress the grating lobes that would otherwise be generated.

The 6-meter diameter aperture is large enough to generate a beam with less than the desired 0.1°, 3 dB beamwidth at 30 GHz, if a uniform distribution is used. At 20 GHz the beamwidth will be greater than 0.1°; but, since this is the transmit beam it is not felt that the 0.1° resolution requirement is quite as important as it is at the receive frequency.

It is desirable to use a tapered distribution that will yield low side-lobes while keeping the theoretical beamwidth at or less than the 0.1° value if possible. A low sidelobe preliminary design allows future designs to trade sidelobe performance for simplified fabrication and assembly methods. For instance, it may be desirable to build the array in modular form and to keep the number of different types of modules to a minimum. Such a procedure means that the tapered distribution, as well as the completely random placement of elements, must be approximated in a step-wise fashion. This would tend to raise the sidelobes in certain directions.

For the preliminary design a Taylor distribution for circular apertures was chosen for which the first several sidelobes should be 30 dB down (Hansen, 1960). The remaining sidelobes then drop off and should be more than 30 dB down. In the interests of economy and simplicity this tapered distribution was implemented by the process known as space tapering. Whereas in conventional antenna design a desired tapered distribution is achieved by reducing the excitation to the elements in the outlying regions of the array, in space tapering all of the elements are excited equally, but complete elements are dropped from the array in those outer regions. In order to reduce the probability of creating high sidelobes that might result from dropping out elements in some systematic fashion, the space tapering was done on a random basis. This was accomplished by drawing a random number from the computer (range = 0 to 1.0) for each element in the non-tapered array and comparing that number with the normalized excitation that the element should have in a conventionally tapered array. If the random number was greater than the desired excitation, the element was dropped; otherwise, it was retained with an excitation of unity. This process results in practically no elements being dropped near the center of the array and about 60% of them being dropped near the edge. The overall result is that a little over 46% of the elements are dropped so that the total number of elements in a 6-meter diameter aperture drops from 7,303 to 3,894.

Computed patterns for this preliminary design array are given in Figures II-1 and II-2. The beam is shown scanned off to the maximum scan angle of 15°. The sidelobes are seen to be over 30 dB down in most places, and the average sidelobe level is on the order of 35 dB down. Grating lobes in a uniformly spaced version of this array would tend to lie at -8.7° and -32.4° in θ. The one at -8.7° is apparently completely suppressed by the random spacing in the current design of the array. In an earlier design the grating lobes were not completely suppressed, retaining a magnitude of about 20 dB down. In that design the amount of offset allowed to the elements from their nominal locations was limited to an amount that eliminated the possiblity of any element ever physically interfering with its neighbors. Since the nominal locations were on a triangular grid, and the displacement area for each element was circular, this left an area near the center of the equilateral triangle formed by any three adjacent nominal locations that could not be reached by the element under any circumstances. Thus "holes" were systematically left in the aperture with no probability of finding an excited element there. In order to reduce the grating lobes produced by these holes, it was necessary to increase the radius of the offset circle so that the holes were much reduced in size. This, however, allows an overlap of the offset circles for adjacent element so that now there is the possiblity that finitely sized (approx. 2.5 cm in diameter) elements may interfere physically with each other.

To obtain an idea of the distribution of the elements, and the amount of physical interference of adjacent elements, the locations of the elements were plotted on a rectangular grid representing the 6-meter aperture. Each element was plotted as a tiny circle that approximated the size of an actual element. This plot is shown on Figure II-3 and clearly shows the space tapering that results from dropping out elements to approximate the Taylor 30 dB distribution. It also shows several instances of interference between adjacent elements. In a final design these interferences would be eliminated by small adjustments of the locations of the elements involved.

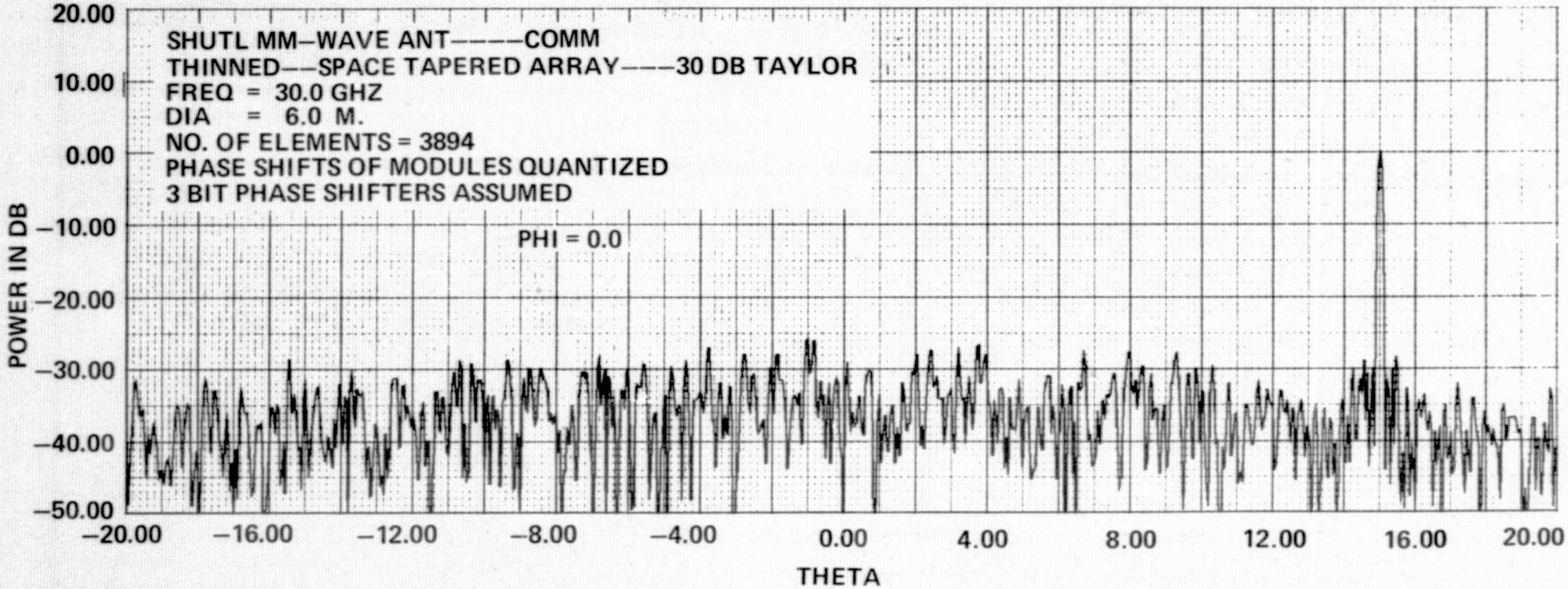


FIGURE II-1

Communications Receive Array

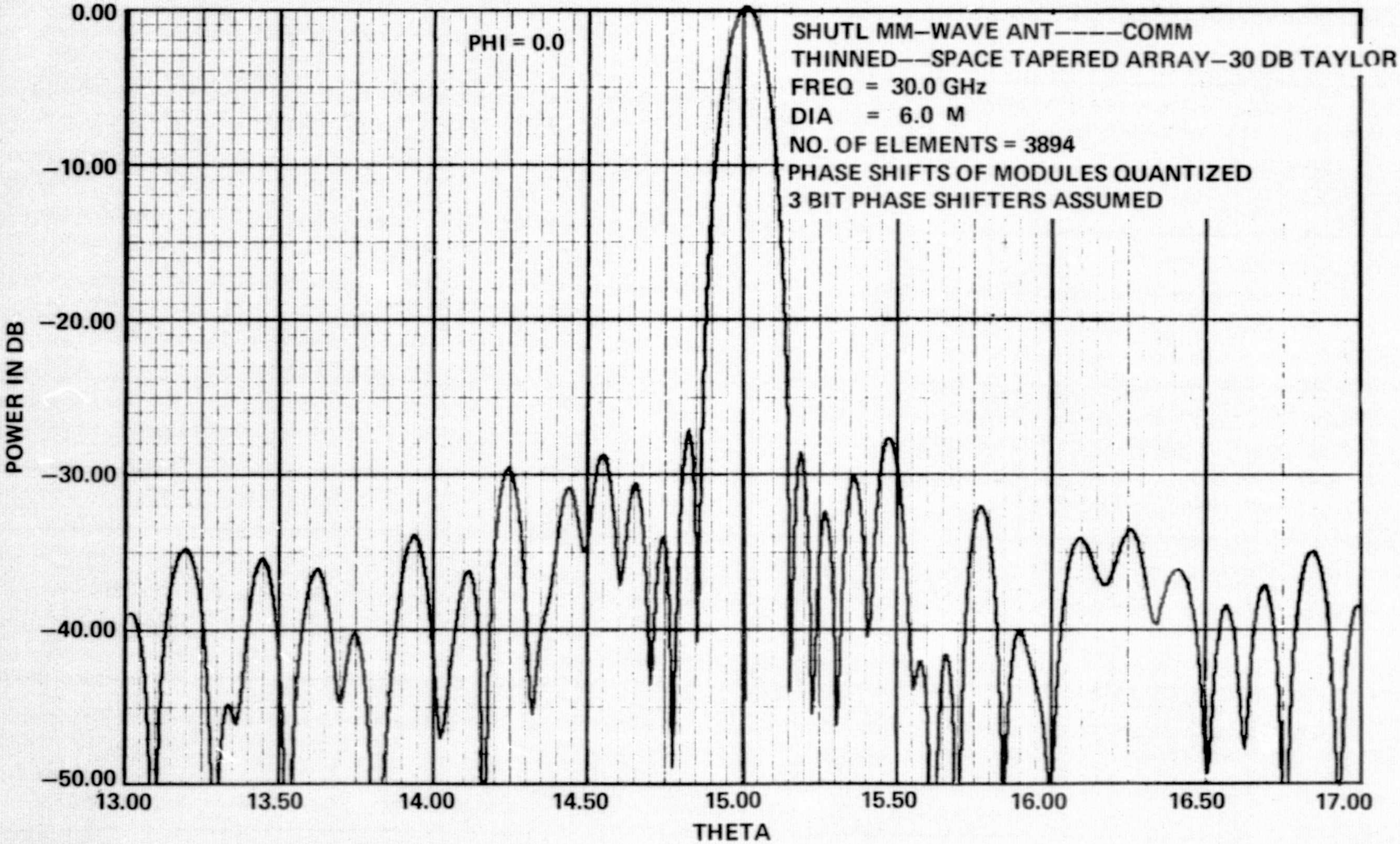


FIGURE II-2

Expanded Portion of Pattern of Figure II-1

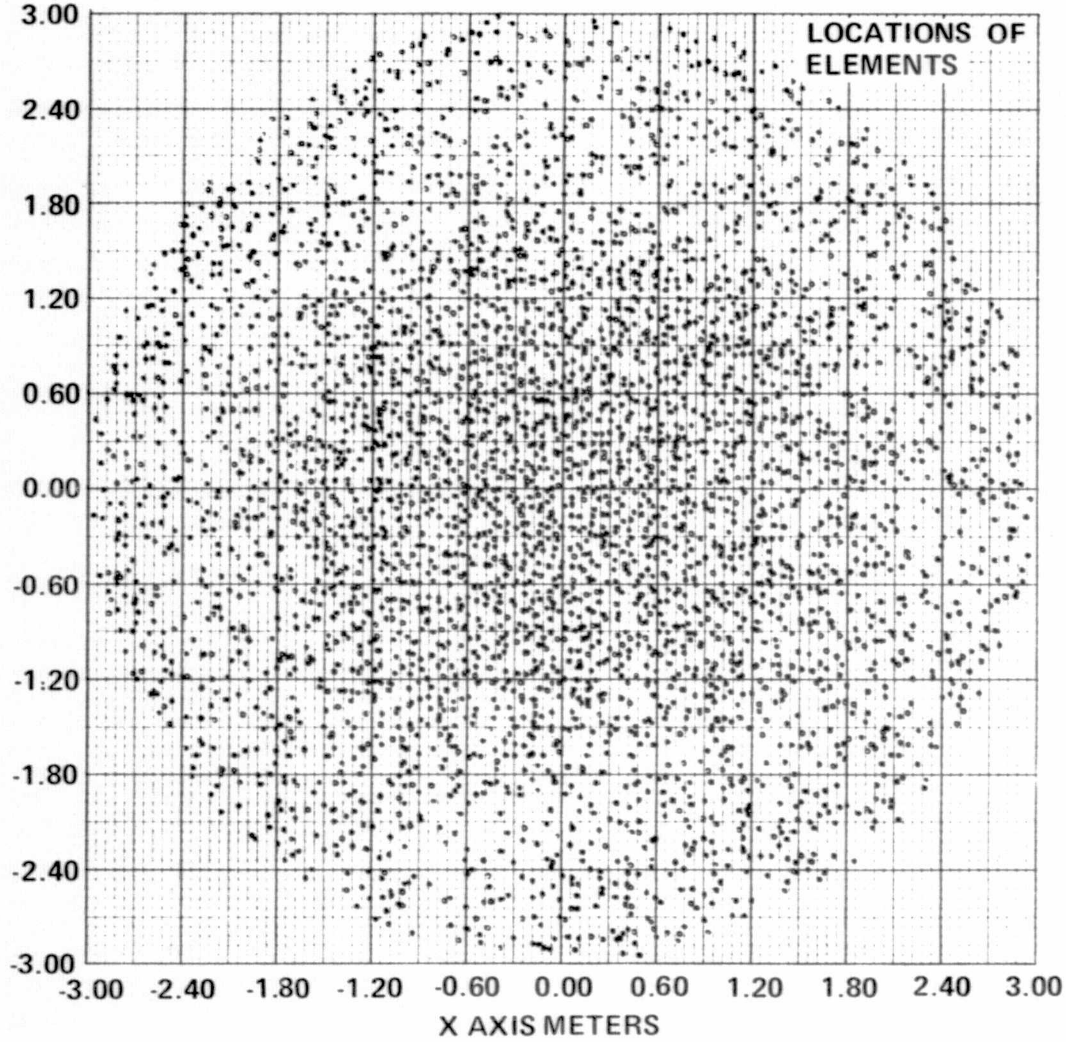


FIGURE II-3

Location of the Elements

As suggested above, some form of modular approach will probably be used in a final design. The elements within each module would be randomly placed, and, since there is a much smaller number of elements involved, the computer program could easily be modified to check for physical interference and make the necessary adjustments.

2.0 Modularized Arrays

Several methods have been considered for building the communications array in modular form. They will be discussed in turn.

The first method utilizes a fairly large number of identical modules distributed throughout a square aperture. (See Figure II-4) The element placement on one module will be computed by some form of randomization. All succeeding modules will then have exactly the same number and placement of elements. In general, it is expected that the average element density on these modules will be somewhat higher than the average element density in the center of the completely random array discussed above. There will not be enough of these modules to completely fill the aperture so that some thinning will be achieved on a modular basis. In order to suppress grating lobes as much as possible, the modules will be positioned in the aperture on a random basis. A tapered distribution can also be approximated on a stepwise basis by spreading the modules a little more thinly as the distance from the center of the aperture increases. It is not presently clear whether this can be done by the computer on a random basis or not. It may be necessary to introduce a "human factor" into this determination. The empty areas between the modules will be filled with a continuation of the ground planes used for the modules.

The second method utilizes a circular geometry in which a number of modules on a given radius are all identical except for the relative polarization orientation of the individual radiating elements (See Figure II-5) In this scheme the modules completely fill the aperture which would be circular. At any given radius the density of elements would correspond to the excitation required by the Taylor 30 dB distribution. The elements within the prototype module for each ring would be scattered on a random

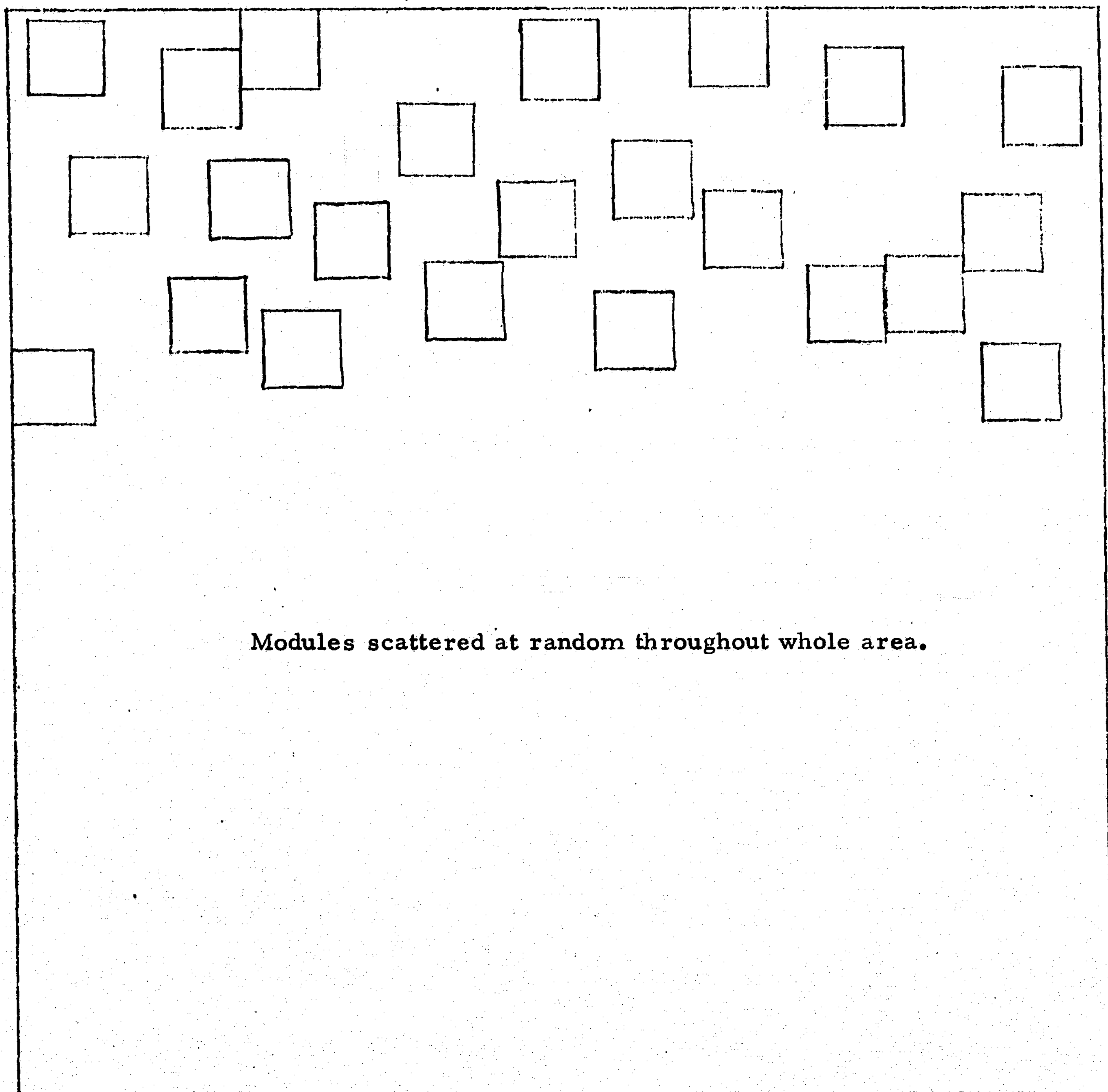


FIGURE II-4

Modularization Technique No. 1

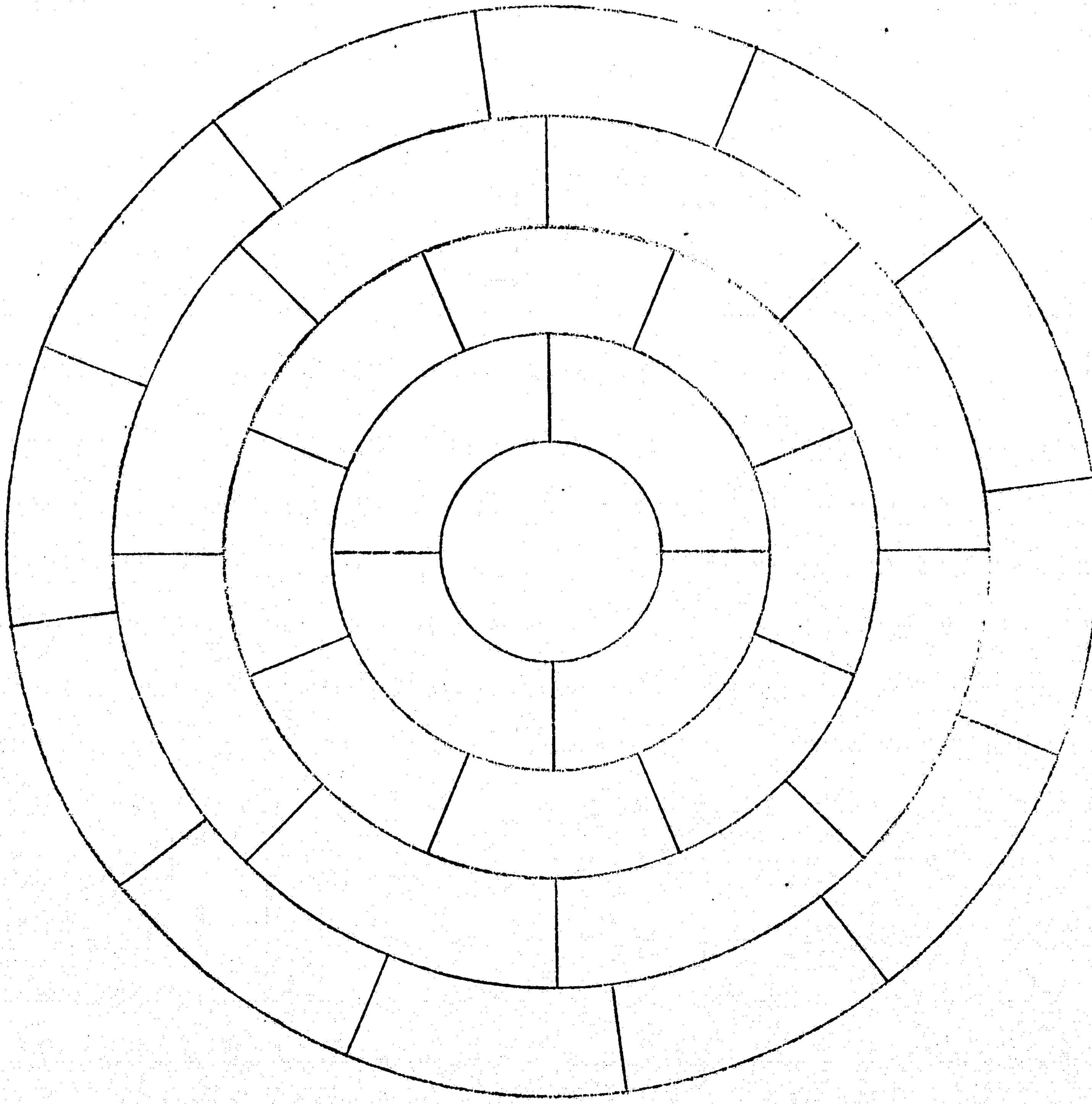


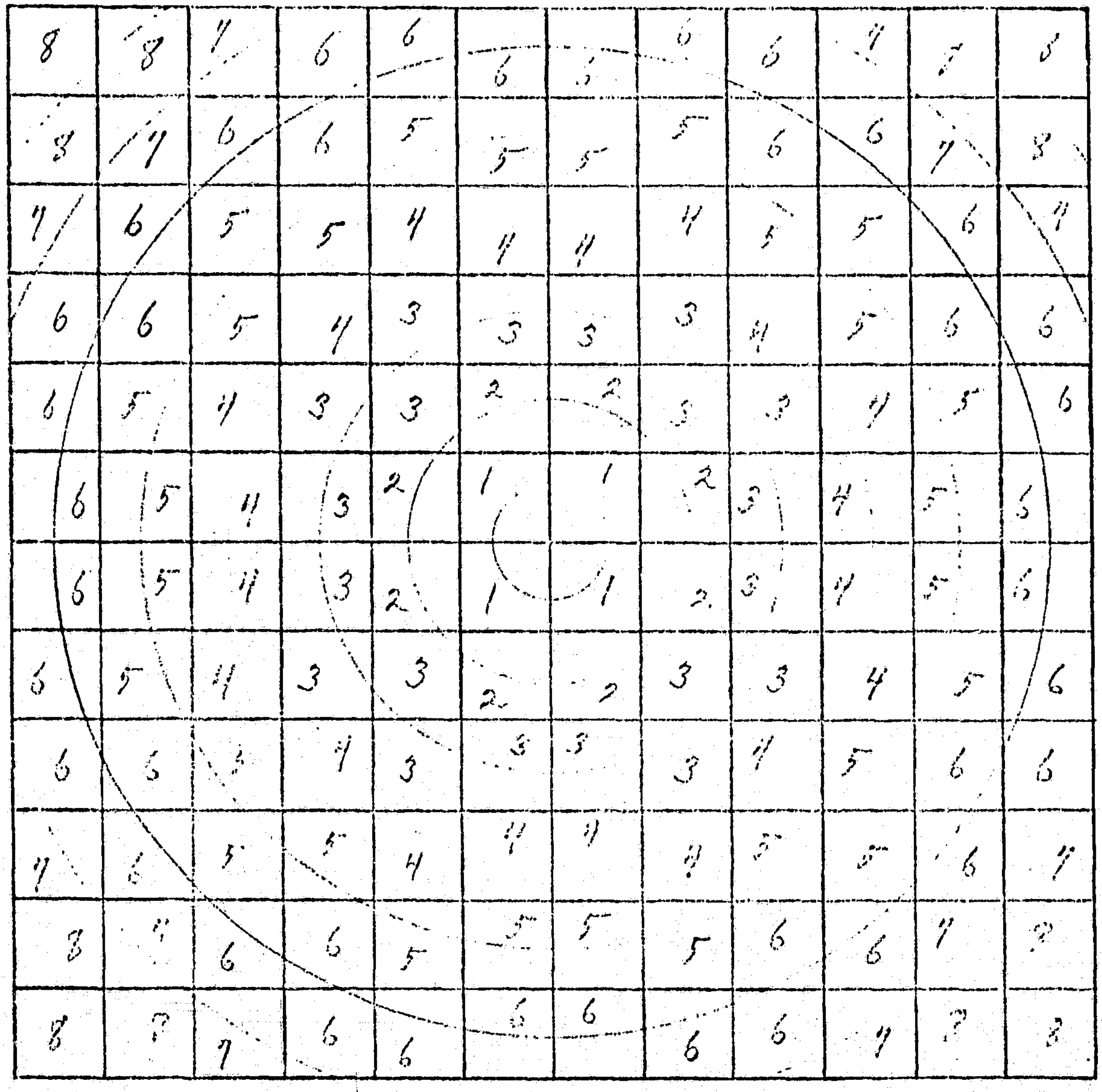
FIGURE II-5

Modularization Technique No. 2

basis subject only to the density requirement. Overall thinning would amount to about 95% so that grating lobes might be a problem. However, an additional variation is introduced by the rotation of each module in a ring by some angle compared to its nearest neighbors. This would tend to break up the periodicity resulting from having elements identically located (in polar coordinates) in all modules in a ring. This geometry has some mechanical disadvantages (i.e., there will be as many different types of modules as there are rings, and polarization* will have to be adjusted on otherwise identical modules) but it has electrical advantages. The electrical performance of this configuration should come close to that of the completely random array discussed above.

The third modularization technique considered consists of a limited number of different types of square modules that fill a square aperture as shown in Figure II-6. The circular Taylor distribution would be superimposed on a square grid and modules approximately the same distance from the center would be labeled as a particular type. 8 to 10 different types of modules should be sufficient to approximate the Taylor 30 dB distribution adequately. The element density would vary from module type to module type. As before, the elements on the prototype module for a type would be randomly spaced, but all succeeding modules of that type would be identical. Mechanically this configuration offers some advantages and it is not anticipated that electrical performance would be degraded severely by the modularization.

* The possibility of using circular polarization is being considered by NASA. If circular polarization is used, then no adjustment would be needed.



Module Type	Quantity
1	4
2	8
3	20
4	20
5	28
6	40
7	12
8	12
Total	144

FIGURE II-6

Modularization Technique No. 3

B. Radar Antenna

1.0 Introduction

The radar is to be used for high-resolution scanning of the earth's surface from a 385 Km, (200 n. m.) orbit. The resolution desired is 100m over a scan range sufficient to map a 100 Km wide swath of the earth's surface. Sidelobe levels and the scan angle have not been defined. However, radar systems usually require low sidelobes. Calculations show that a scan angle of $\pm 15^\circ$ will more than cover the 100 Km swath for the most likely scan geometries; therefore, preliminary designs for the radar antenna will be limited to scan angles no greater than $\pm 15^\circ$.

At the 385 Km mile altitude, assuming the beam is scanned forward at an angle of 45°, a beamwidth of 0.01° is needed to give the desired 100m resolution. This small beam implies a very large antenna, hence methods of reducing its complexity have been considered.

2.0 Conventional Radar

In a conventional radar system, the same antenna is used for both transmit and receive. Thus the return signal as a function of angle has a magnitude that is the square of the one-way antenna pattern. For this reason it is customary to define the resolution limits of such an antenna as being between the $1\frac{1}{2}$dB points rather than between the 3dB points. Thus, the aperture does not need to be quite as large to obtain a particular resolution as would be required on a one-way system, such as for communications or radiometry.

If the antenna is uniformly excited along its length, it will produce the narrowest beam possible without resorting to "super gain" or interferometry techniques. This type of excitation produces -13.2dB 1st sidelobes in the one-way pattern---equivalent to -26.4dB in the two-way pattern. The remaining sidelobes drop off rapidly to values lower than -60 dB in the two-way pattern.

The aperture size required for a specified beamwidth between the $1\frac{1}{2}$dB points is given by

$$d_\lambda = \frac{36.54}{BW_{1.5}}$$

where:

d_λ is aperture size in wavelengths, and

$BW_{1.5}$ is the 1.5dB beamwidth in degrees.

This formula gives an aperture size of 3654 wavelengths for a 0.01° beamwidth between the 1.5dB points. At the operating frequency of 13.9GHz, λ is 2.157 cm, thus making the required aperture 78.82m long. The total area need not be exorbitant, however, since a large aperture is not needed in the in-track plane. An aperture size of 20 wavelengths (0.5m) should be adequate to produce a cosecant squared pattern in that plane. No design effort has been expended to obtain a specific pattern since it does not involve a physically large dimension. The actual aperture size used in this plane may need to be larger than 0.5m to get the necessary signal to noise ratio.

Since the aperture size is so large, no large amount of effort has been expended on the conventional radar concept. Rather, the major effort has been directed to the BISTAR approach which uses a thinned array; it is discussed in the next section.

3.0 BISTAR Radar Concept

The BISTAR (BIStatic Thinned Array Radar) concept has been developed at Hughes to obtain high resolutions from apertures that are significantly thinned compared to a continuously excited aperture.

In the BISTAR approach, the radar transmitter feeds a comparatively small scanning antenna that illuminates the forward area. A separate antenna is used for reception; therefore, the radar is bistatic. The receiving antenna has widely spaced segments over a much larger area and forms a pattern that contains predictably spaced grating lobes much like a multi-element interferometer. In operation, only one grating lobe is illuminated on the ground by the main beam of the transmit antenna. The transmit array is designed such that nulls of its pattern coincide with all of the other grating lobes of the receive pattern, thereby effectively suppressing any radar return from those directions. The transmit and receive antenna beams are scanned in synchronism to provide the sector-scan display.

The system resolution comes primarily from the width of the receive antenna pattern at the 3 dB points. The transmit pattern is much broader than the receive pattern because of its smaller aperture. Thus, the two-way pattern of the system is not much narrower than the receive pattern alone. Hence, on the BISTAR system we cannot use the $1\frac{1}{2}$ dB points on the one-way pattern as the effective beamwidth as is done for conventional radars.

In the aircraft application, for which BISTAR was developed, the receive array has a considerable number of segments and the receive aperture length is many times greater than that of the transmit array. In applying the BISTAR concept to the space shuttle radar experiment, however, it may be advantageous to use a smaller number of segments because it has been found that a small number of segments in the receive array produces a narrower beam for a given edge-to-edge width than does a large number of segments. This results from the fact that the array, when it has only a few segments, takes on some of the characteristics of an interferometer.

A relationship between the 3-dB beamwidth and the total aperture length between the two outside segments can be derived for interferometers and arrays that contain a small number of widely spaced elements.
It is:

$$\theta = \frac{K_n}{l_\lambda}$$

Where θ = Beamwidth between 3 dB points, in degrees
n = number of segments in array
l_λ = aperture length in terms of wavelengths
K_n = a parameter whose value is a function of n

The proper values for K_n have been evaluated for arrays containing from 2 to 6 segments. These values for K_n were derived on the assumption that the spacing between segments was very large in terms of wavelengths and consequently that the beamwidth was very small; small angle approximations were used. The values that have been calculated for K_n are given in Table I along with the aperture size in wavelengths and in meters required to produce a 0.01° and a 0.02° beam at 13.9 GHz.

TABLE II-I

Array Parameters vs. Number of Segments in 13.9-GHz Array

No. of Segments in Array n	K_n degrees	Aperture Length in Wavelengths Required		Aperture Length in Meters Required	
		0.01° Beam	0.02° Beam	0.01° Beam	0.02° Beam
2	28.65	2865	1478	61.79	31.84
3	35.59	3559	1834	76.76	39.55
4	39.14	3914	2017	84.42	43.50
5	41.33	4133	2130	89.14	45.93
6	42.81	4281	2206	92.33	47.58
infinite	50.76	5076	2616	109.48	56.41

The array listed in Table II-I with an infinite number of segments corresponds to a continously illuminated aperture for comparison with the discrete arrays. The aperture lengths given in this table are for the one-way receive pattern only.

The two-element array, or interferometer as it should more properly be called, yields the required beamwidth from the smallest aperture. From the point-of-view of the receive beamwidth, the BISTAR concept as applied to this radar system thus should have only two widely spaced segments in the receive array. The size of transmitter antenna required for BISTAR, however, as noted above, is directly related to the distance between segments in the receive antenna. Thus, if the receive array has only two segments, then the width of the transmit array must be equal to the full distance between them. A better tradeoff may be achieved by using three to five segments in the receive array. The full aperture of the receive array would thereby be increased somewhat, but the width of the transmit array would be reduced by a factor of two to four. Several configurations utilizing small numbers of receive segments will be investigated to determine which one yields the required resolution with the smallest and/or simplest aperture.

4.0 Receive Pattern Investigations for BISTAR Concept

Initial investigations have been confined to a 4-segment configuration (see Figure II-7). These segments would actually be small phased arrays in order to achieve sufficient receiver capture area to obtain a good

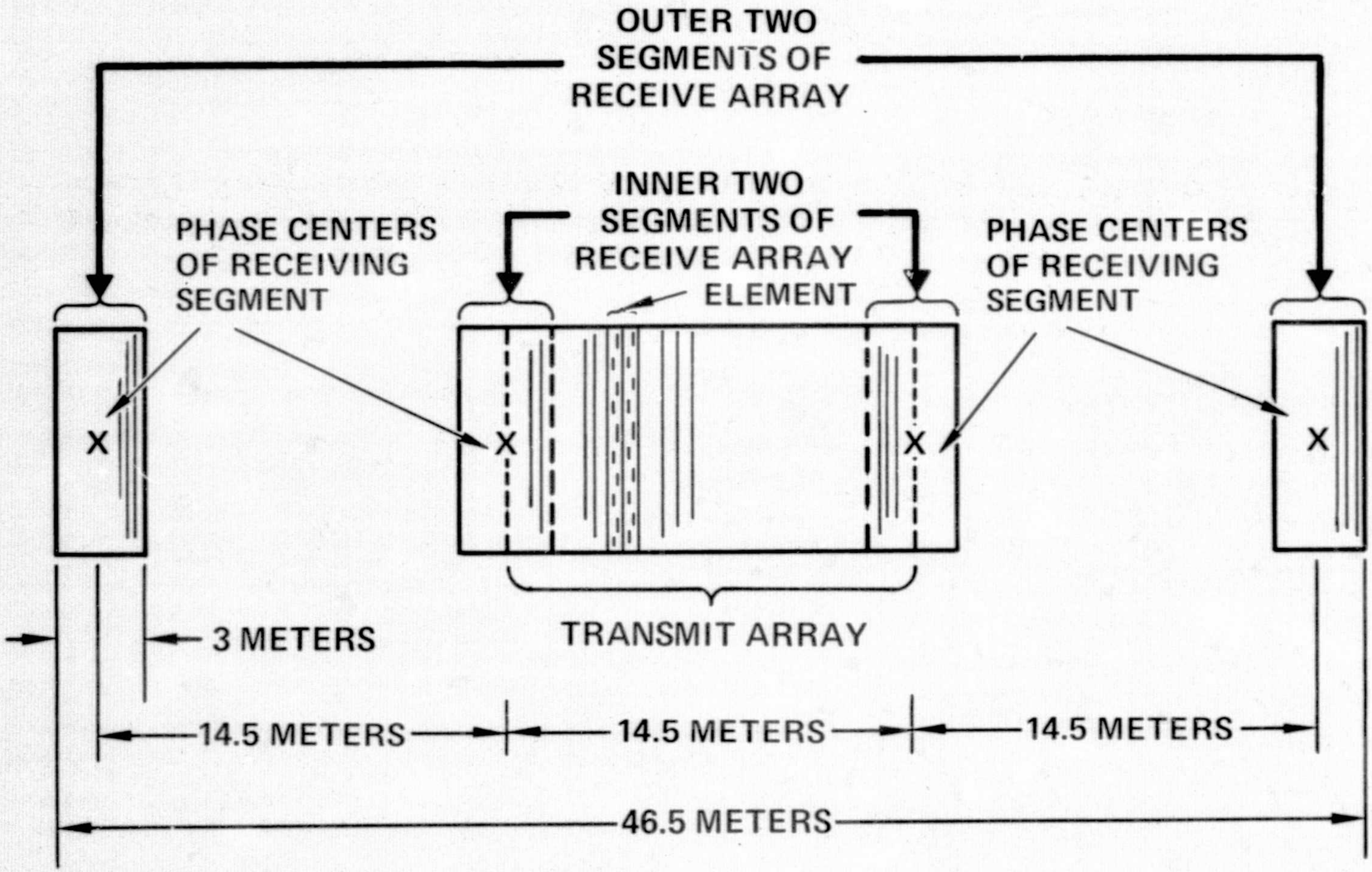


FIGURE II-7

13.9-GHz BISTAR Antenna System

signal-to-noise ratio. These smaller units, called elements, will be used within both the receive segments and the transmit array. Thus, a four-segment array is used to form the receive pattern. Since each segment in this configuration is a 3m wide aperture, it will have a rather narrow beam of its own. The total receive pattern will be a product of the segment pattern and the four-segment interferometer pattern and will consequently be slightly narrower than the interferometer pattern alone.

Since the segment pattern is so narrow, it must scan with the main receive pattern, therefore its elements must be phased. The scan angle, however, is limited to $\pm 15^{\circ}$; hence, the non-scanning element pattern need not have a pattern much wider than that between its 3 dB points for satisfactory performance. For this preliminary design, it has been decided that two waveguide branch lines can be tied together and phased as a unit for the basic element. (See Figure II-8) If the scan angle were limited to an angle smaller than $\pm 15^{\circ}$, then perhaps three or four such branch lines can be phased together to reduce the complexity and weight of the system.

The pattern of the two-waveguide element is given by

$$E(\theta) = \cos\left(\frac{\pi a}{\lambda} \sin \theta\right) \tag{1}$$

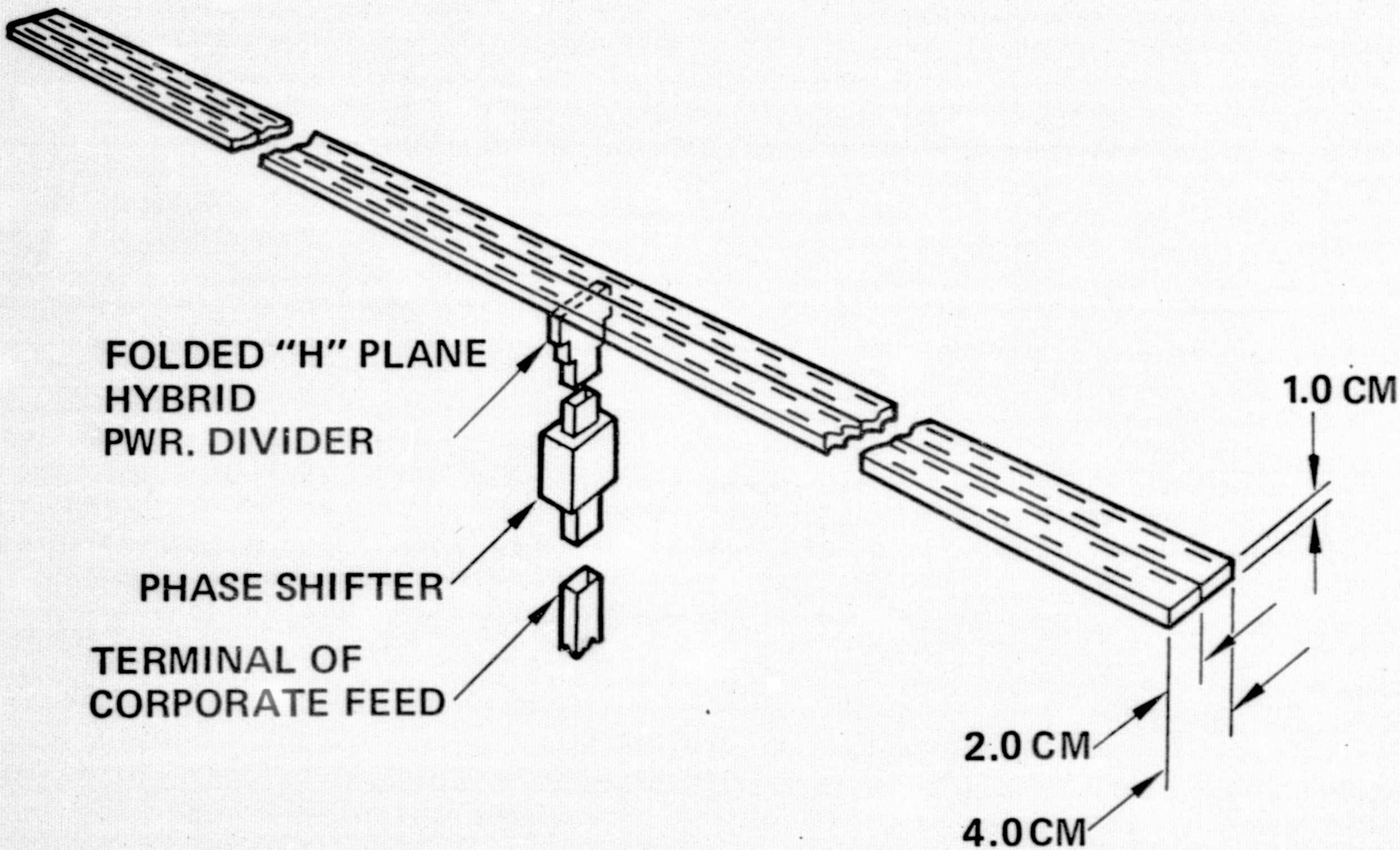
Where:

- a = distance between center-lines of the two waveguides in cm.
- λ = wavelength in cm
- θ = scan angle

The patterns of the individual waveguide slot radiators are not included in this expression since the polarization has not been specified and the slot patterns will be so broad as to have only a negligibly small effect on the final pattern. A convenient value to choose for the distance between centerlines of the two waveguides is 2 cm. At this spacing, and with reasonable wall thicknesses, the waveguides will be operating near the upper end of the band where attenuation losses are minimized. With a 2 cm spacing, the pattern of the two-waveguide pair has a 3 dB beamwidth of approximately 42°, hence should be quite satisfactory for a scan angle of $\pm 15^{\circ}$.

FIGURE II-8

Detail of Two-Waveguide Element



The full width of the two waveguide element is 4 cm; therefore, 75 of them will be required for each receive array segment in this preliminary design which calls for 3m for the segment width. A computer program was used to compute the segment pattern at an assumed scan angle of $\pm 15^{o}$.

The computation is based upon the following generalized array equation which is valid for all uniformly spaced linear arrays.

$$E(\theta) = E_1(\theta) \sum_{n=1}^{N} A_n \cos\left\{ \frac{2\pi a}{\lambda} \left(n - \frac{N+1}{2}\right) \sin \theta + \Psi \right\} \qquad (2)$$

Where:

$E_1(\theta)$ = element voltage pattern for elements used in array
N = Number of elements in array
A_n = Weighting factor of n^{th} element.
Ψ = interelement phase shift of scanned array
a = interelement spacing } Both in same units
λ = wavelength
θ = scan angle off normal to array

For the segment pattern computation $N = 75$, $a = 4$ cm, $E_1(\theta)$ is the element pattern computed in Eq. 1, the A_n were all of equal magnitude, and Ψ was phased so as to scan the beam to an angle of $+15^{o}$. The computations assumed that the phase shifters driving the elements were quantized as would be the case in practice. A quantization level of 3 bits was assumed in line with the current design for the electronics behind the receive array. A plot of the computed pattern is given in Figure II-9. As expected the 3 dB beamwidth is approximately 13 dB down from the peak. Some of the nulls are not very deep because of phase quantization which introduces small phase errors.

The total receive array pattern is the product of the segment pattern just discussed and the four-segment array pattern. Figure II-7 shows that the interelement spacing of the 4 segment array is 14.5m, and as a result the overall length of the array (between centers of the outer two elements) is 43.5M. The pattern of this array was computed using Eq. 2 with $N = 4$, $a = 1450$ cm, $E_1(\theta)$ = the voltage pattern computed for the segment discussed above, the A_n all of uniform amplitude, and Ψ phased to scan the array factor to an angle

No. of Elements = 75

Phase Shifts of Modules Quantized - 3 Bit Phase Shifters Assumed

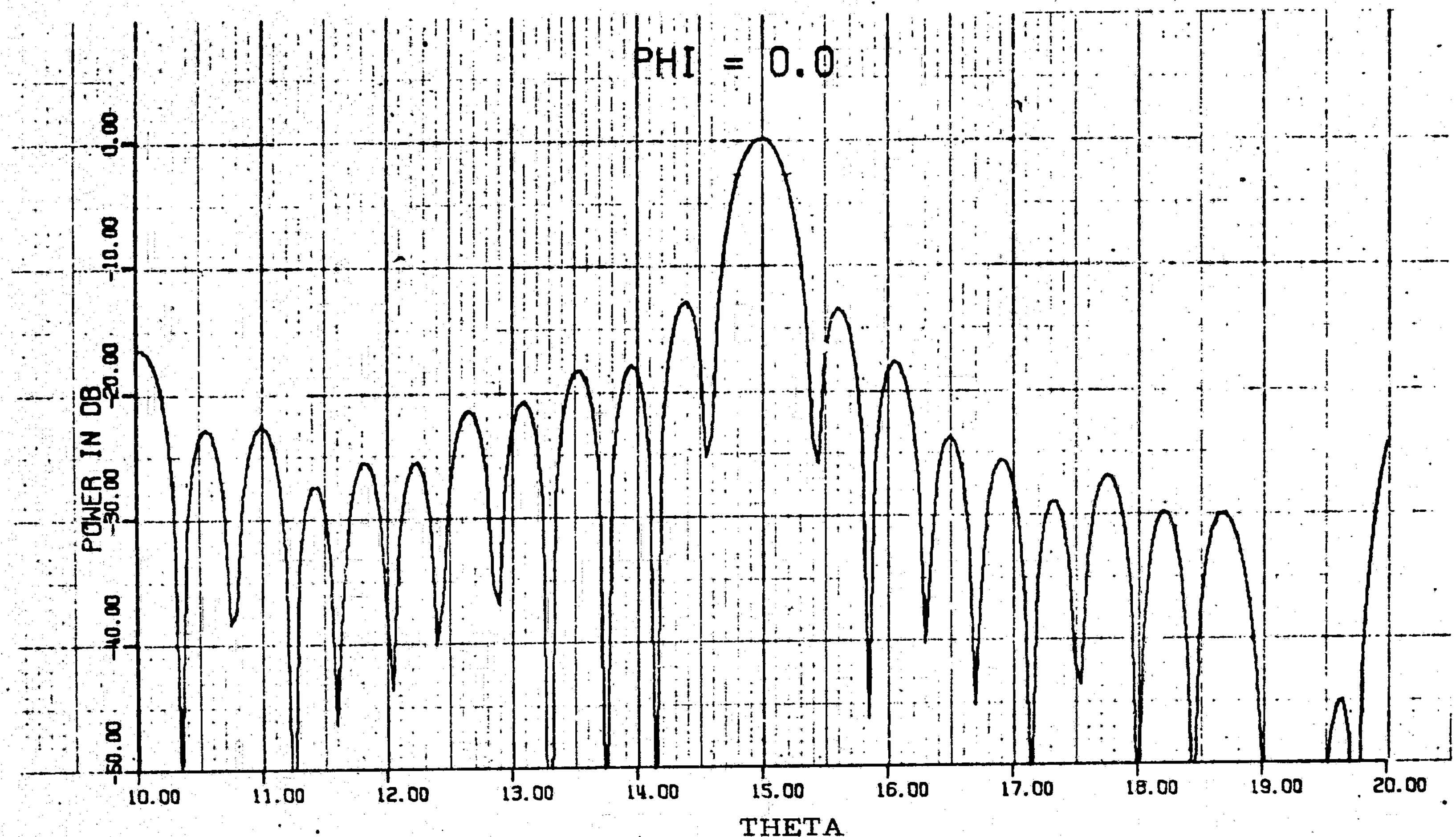


FIGURE II-9

Receive Segment Pattern

of + 15^{o}. Here also 3 bit phase shifters were assumed to simulate actual practice. The pattern was plotted and is given in Figure II-10. The grating lobes mentioned earlier are much in evidence. In fact, there are only two sidelobes between each adjacent pair of maximas. This is to be expected from an array with only four elements. The array factor is multiplied by the segment pattern; therefore, the grating lobes tend to drop off at angles far away from the main beam. The segment pattern is superimposed on the final receive array pattern to show how it acts as an envelope over the array factor which produces the grating lobes. The segment pattern appears to have a much broader beamwidth in this plot than in Figure II-9 because of the much-expanded scale.

5.0 Transmit Pattern for BISTAR

The grating lobes in the receive pattern are about 0.09^{o} apart; therefore, the transmit array must be designed so that its nulls are equally spaced and exactly that angular distance apart. A further requirement is that the angular distance from the peak of its main beam to the first null also be equal to the distance between nulls in order to get the series of nulls started off right. Consideration of various illumination functions reveals that uniform illumination is the only such function that satisfies the second condition set down above. Thus, it is the only function that will suppress all grating lobes but one, as required for the BISTAR system.

The spacing between nulls is controlled by the length of the array, and it can be shown that the length required to make the nulls of the transmit pattern fall precisely on the peaks of the grating lobes of the receive pattern is exactly equal to the interelement spacing of the receive array. Such an array was designed using the same two-waveguide element that was used for the receive segment. An integral number of these elements would not fit in the 1450 cm interelement spacing of the receive array, hence the number was rounded to the next highest whole number.

A pattern was computed using Eq. 2 with the following parameters:

N = 363 elements

$E_1(\theta)$ = the element pattern as given by Eq (1)

a = 4 cm

A_n = All of uniform amplitude

Ψ = phased for a beam scan of $+15^{o}$.

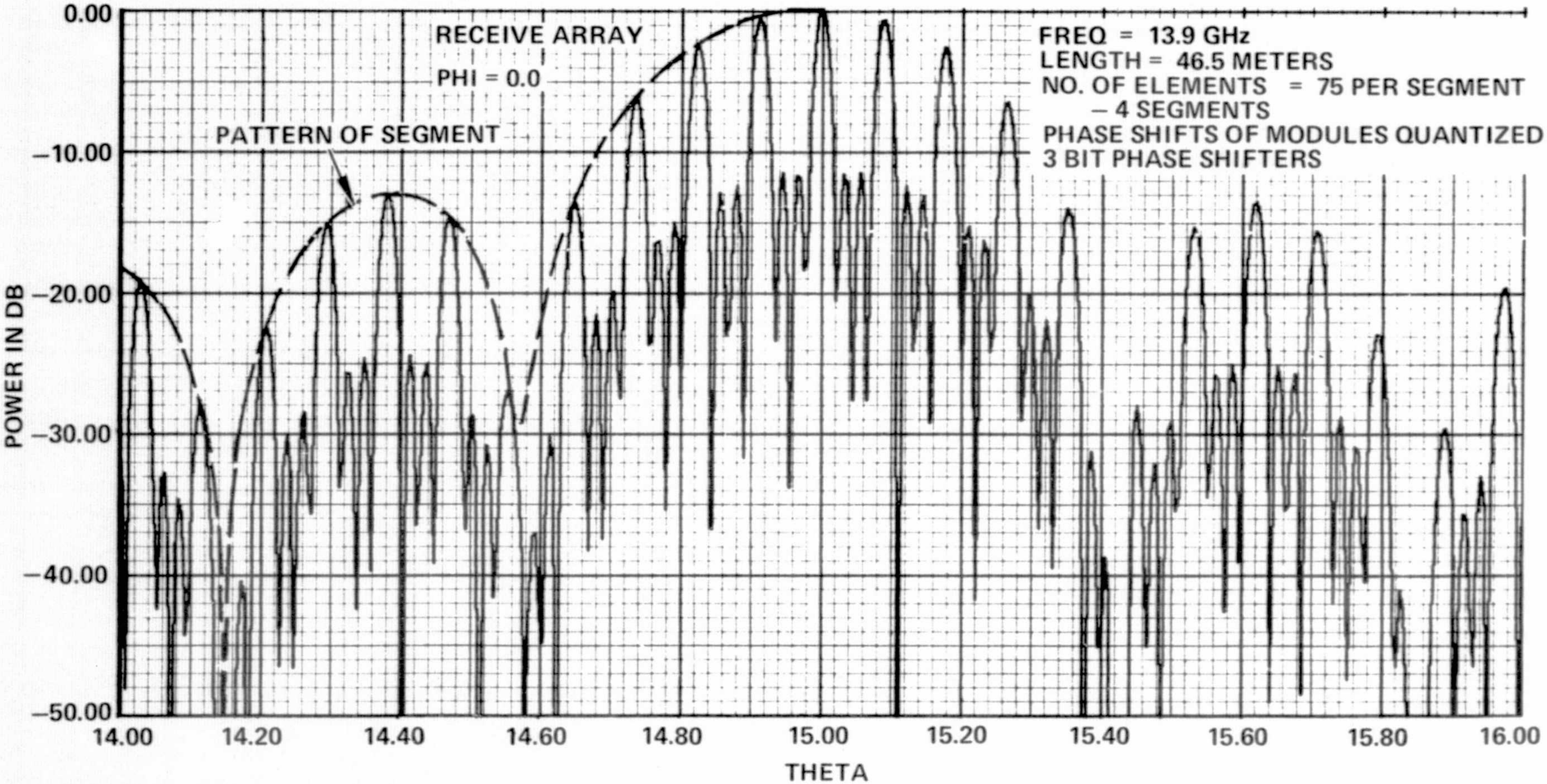


FIGURE II-10

Receive Array

The phase shifters for this array were also assumed to be of the 3-bit quantized type. The resultant pattern is plotted in Figure II-11. It appears to have the desired characteristics except that some of the nulls are not as deep as might be desired because of the quantization phase errors. However, the nulls are very nearly equally spaced, and spaced about 0.09° apart.

6.0 Composite BISTAR Radar Pattern

As noted above, the composite two-way pattern is the product of the transmit and receive patterns. The same computer program that was used to compute those two patterns was modified so that it would store the patterns at the time of computation so that they could be called up and their E-fields multiplied angle-by-angle in a subsequent part in the program. The resulting composite pattern is shown in Figure II-12. As can be seen, the grating lobes are effectively suppressed and all sidelobes are below -50 dB beyond 0.4° on either side of the main beam. The 3 dB beamwidth is about 0.02° as expected for a receive array of 46.5 m extent. The main drawback to this pattern for radar use is the rather high close-in sidelobes. The level of the first few sidelobes is determined primarily by the design of the 4-segment receive array. Thus, in order to lower them, it will be necessary to put a tapered distribution on the receive array. This would tend to broaden the beam and in order to maintain the same beamwidth, a larger aperture would be required for the receive array. Since the transmit array must be equal to the interelement spacing, then it also would have to be lengthened somewhat.

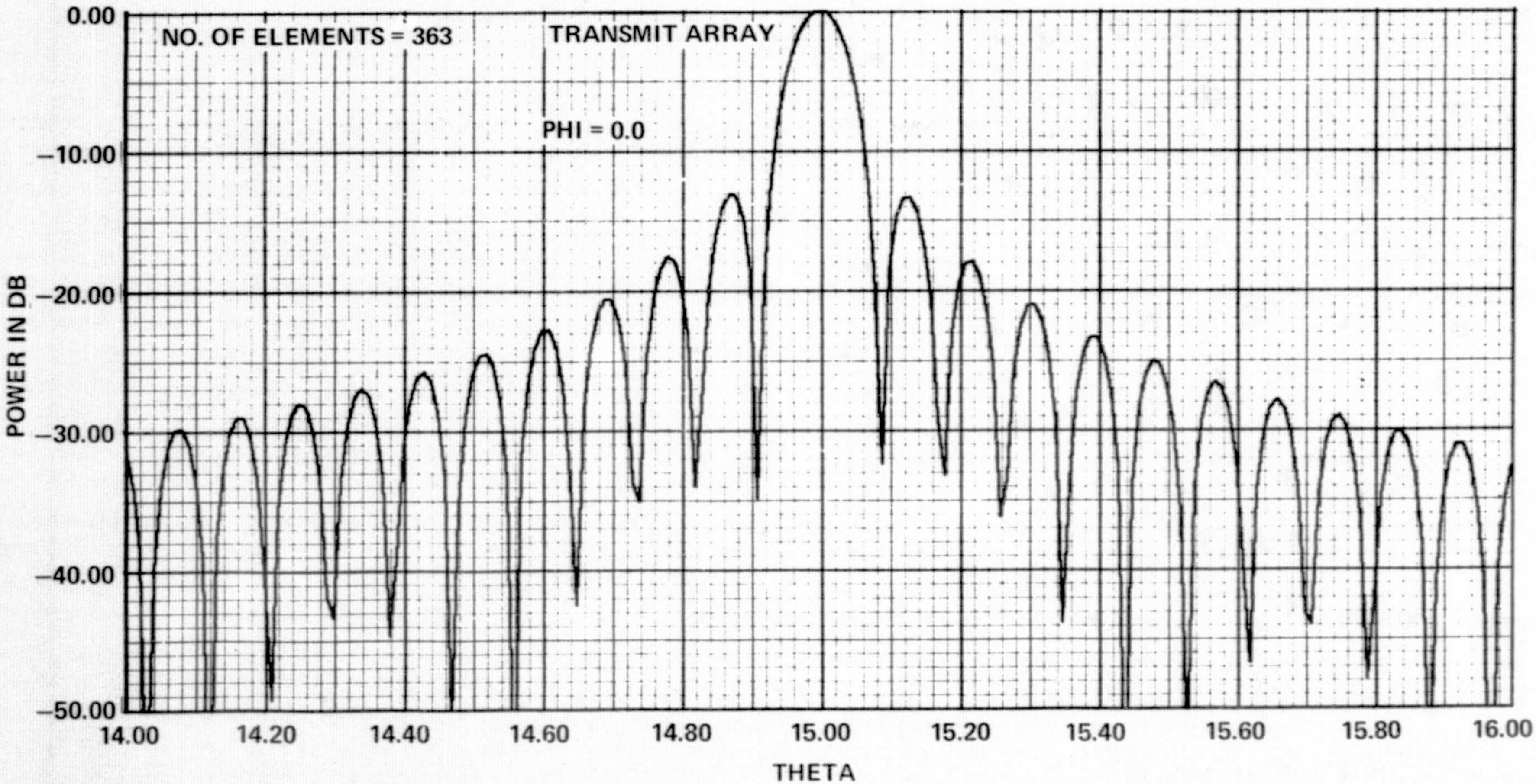


FIGURE II-11

Transmit Array

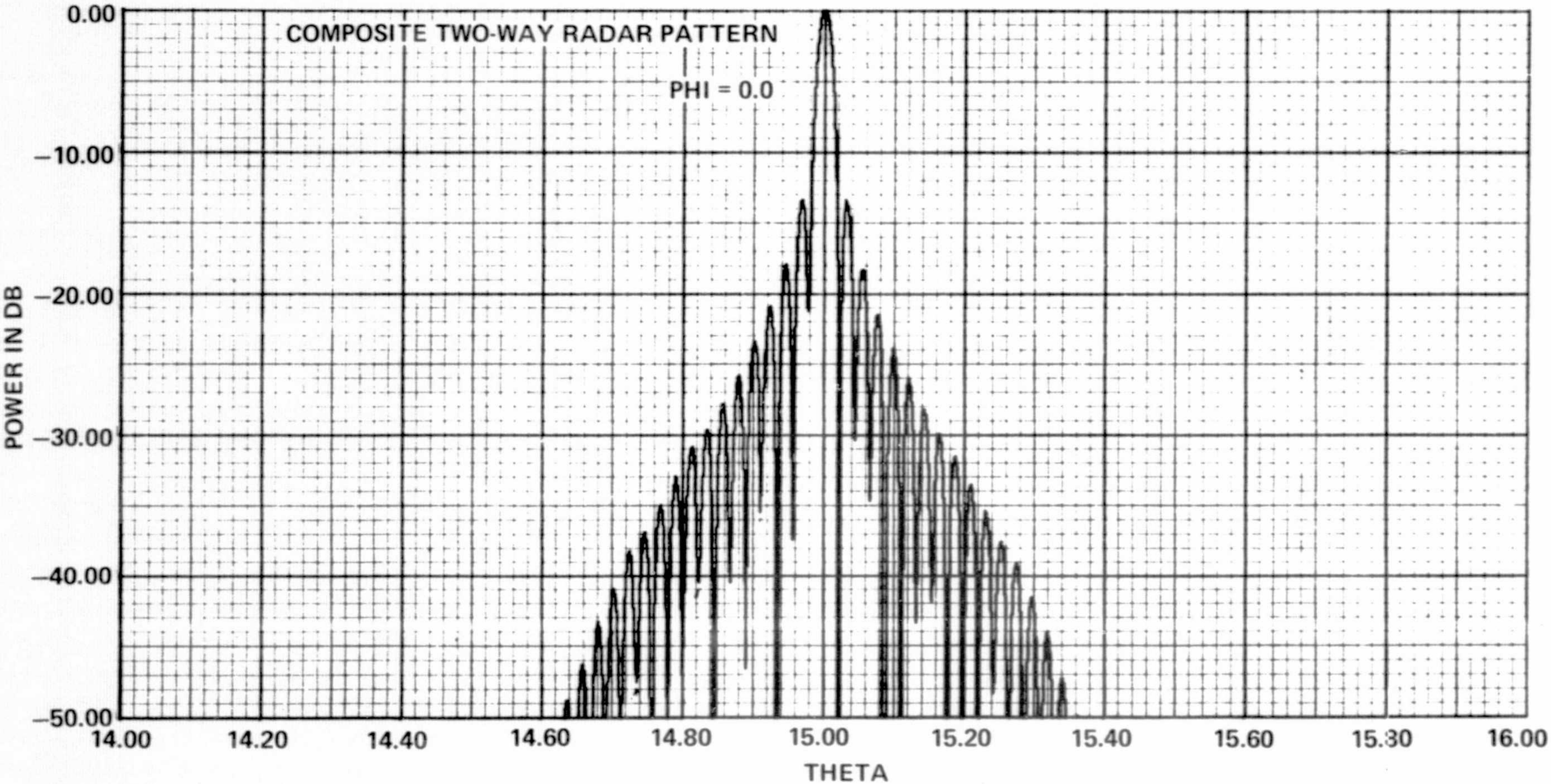


FIGURE II-12

Composite Two-Way Radar Pattern

III. ELECTRONICS CONFIGURATION

Because the performance and physical properties of the required electronic components impact heavily on the various antenna techniques considered during this study, a significant part of the effort has been directed toward the evaluation of possible component configurations. The object of this analysis has been to determine weight, size, power consumption requirements, and anticipated performance for this equipment. This effort included surveys of component state-of-the-art as well as an attempt to project future equipment availability through recommended development work.

Equipment studies were initiated through the configuration of baseline systems which would offer the complete set of performance parameters. In addition, configurations were sought which offered the greatest equipment commonality among the experiments. It was found that a significant penalty is paid in weight and performance to achieve commonality and the desired wideband multiple channel capability.

Because the millimeter wave communication experiment presented the most complex set of requirements, initial efforts were concentrated on its equipment requirements. Evaluation of the 13.9 GHz radar experiment was then introduced with the study of the radiometer systems to follow. A summary of the effort on the electronic equipment for the communications and radar experiments is given here.

In order to maximize equipment usage a structure was sought in which a significant amount of the hardware would be reused for the three experiments. Equipment dedicated for any particular experiment would, therefore, be easily separable, either between flights or through EVA, to configure a future experiment. Although configurations with this versatility are feasible, it is apparent that the imposition of this type of limitation results in heavy weight and performance penalties.

During the initial phases of the program, the equipment was structured solely on the merits of the three experiments considered in this study, i. e., no concern was paid to compatibility with other possible on-board experiments. Although this is an unrealistic situation, the formation of a baseline situation is provided from which trade-offs and compromises can be developed. As experiments are defined and priorities established modifications must be examined to arrive at a useful combine.

A. Communications Array

The baseline electronic configuration for the communication array receiver is shown in Figure III-1. This equipment is capable of independently receiving and steering 3 signal channels distributed over a 1500 MHz bandwidth near 30 GHz. Incoming signals are downconverted to a convenient intermediate frequency and amplified, the channels are then separated and phase shifted, and the beam is formed by the addition of signals in the combiner. Because the system performance, weight, and power consumption are determined by N, the number of array elements, these factors can be discussed on the basis of a single element.

The frequencies indicated on the block diagram of Figure III-1, although not absolute, were chosen on the basis of a number of considerations.

1) A high IF frequency was chosen to achieve the desired wide bandwidth channels. The phase shifter, amplifier, and filter requirements are more easily achieved over the smaller percentage bandwidths.
2) Local oscillator and IF frequencies were chosen to allow maximum component reuse with the radar and radiometry experiments.
3) Intermodulation, interference, and spurious signal problems were examined and minimized with the set of frequencies selected.
4) All frequency conversion and mixing is performed without inversion of information on the signals. In this way, no specific modulation format is favored. Frequency, phase, digital or analog modulation, time and frequency division multiplexing, etc. are all equally acceptable formats for use with the communication system.

The performance of the required components for the baseline system of Figure III-1 will be discussed next, followed by a description of the total system performance and weight and power requirements. This will

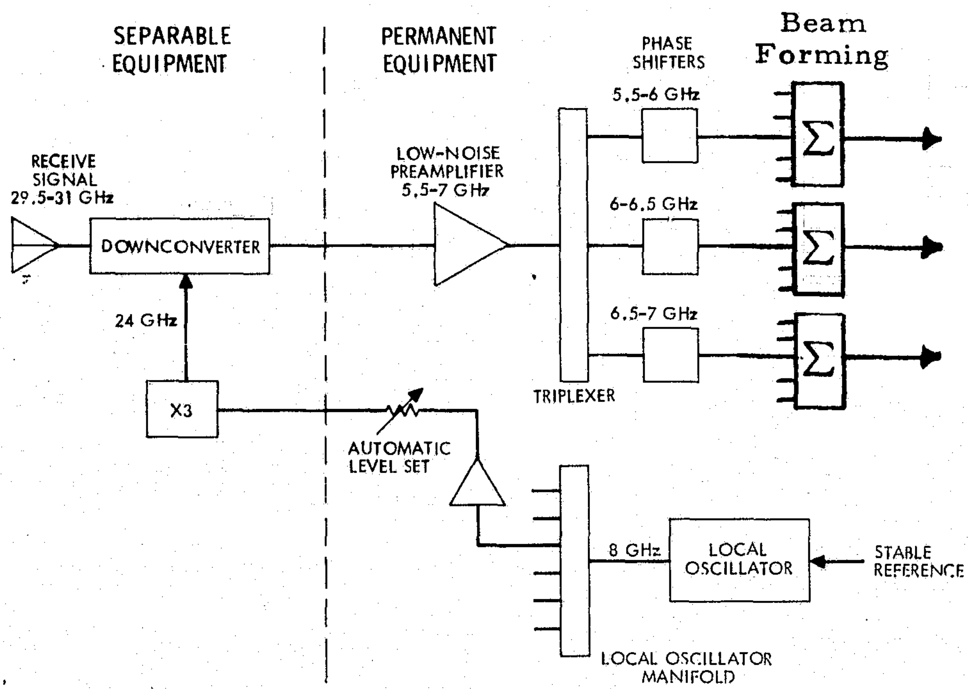


FIGURE III-1

Baseline Communications Receiver Configuration be followed by a discussion of alternative electronic configurations and a comparison of the performance and requirements of these systems with the baseline approach.

1.0 Key Components for Communications Receiver System

1.1 30-GHz Downconverter

The downconverter for the 30-GHz incoming received signal is the most important item in establishing the noise figure or sensitivity of the receiving elements. Because suitable low-noise amplifiers at 30 GHz will not be available* in the numbers required for the array, low-noise RF preamplification is not feasible for this application.

The availability of state-of-the-art 30-GHz downconverters has been examined for use, not only in the baseline configuration, but in the various other configurations that have been considered. Several transmission media, including rectangular waveguide and new forms of construction, (Davis, 1974) are available for fabrication of the converters. Recently developed Schottky barrier diodes, either silicon or gallium arsenide, are available for use as the mixing elements in the various converter designs.

Rectangular waveguide converters have been under development in the millimeter wave region for many years and offer the most conventional approach with the best performance. Conversion losses of these units for single sideband operation typically range from 5 to 6 dB depending on the output IF frequency. Using a high IF near 6 GHz, as required here, will result in a conversion loss of near 6 dB. The receiver noise figure may be calculated using

$$F_r = L_c \quad (N_R + F_{if} - 1) \tag{1}$$

assuming high gain in the IF amplifier. Here

L_c = converter conversion loss
N_R = noise ratio, which will be assumed unity
F_{if} = noise figure of IF amplifier.

Additional contributions due to local oscillator noise and input or filter losses must be added to the calculated result.

* Parametric amplifiers have been developed, but cannot be considered due to cost and complexity.

The main drawback of the rectangular waveguide units is their size and weight. Because of the high IF, single-ended versions can be used here. However, the additional filtering, required to prevent local oscillator radiation, will consume most of any weight reduction achieved. Single-ended or balanced versions of rectangular waveguide downconverters at 30 GHz will weigh 30 to 45 grams and occupy about 20 cm^3. With an IF amplifier noise figure of 4 dB, a SSB receiver noise figure of 10 dB can be achieved.

Of the new transmission media presently being developed in the industry, the dielectric image line waveguide (Chrepta and Jacobs, 1974) appears extremely promising for use with millimeter wave integrated circuits. The advantage of this medium is its small size and weight accompanied by low insertion loss at millimeter waves. Also, of the converters studied, the image line is best suited for batch fabrication, an important consideration for the numbers envisioned here.

Currently, integrated silicon dielectric single-ended and balanced mixers are under development at Hughes and other laboratories; however, most of these efforts are directed toward 60-GHz components. The single-ended mixers, as shown in Figure III-2, utilizing Schottky barrier diodes exhibit less than 7-dB conversion loss at frequencies from 30 to 60 GHz. The configuration of a 30-GHz balanced downverter as presently envisioned is shown on the sketch of Figure III-3. It is estimated that this converter will require a volume of 2.5 cm^3 and weigh about 6 grams including the aluminum ground plane and supporting structure. A slight additional weight of .15 gram may be required for cover and hermetic sealing.

Converters with good performance have also been constructed using microstrip techniques and hybrid (microstrip/waveguide combination) approaches. Laboratory models of these components have demonstrated conversion losses as low as 5.5 dB. (Glance and Snell, 1974) Microstrip techniques can be used at 30 GHz on fused quartz substrates with acceptable insertion loss, without encountering moding problems. Expected weight for these structures is about 15 grams.

Typical local oscillator power requirements for the converters considered here are about 10 milliwatts. Because of the large numbers required for the array, it is desirable to operate with the minimum

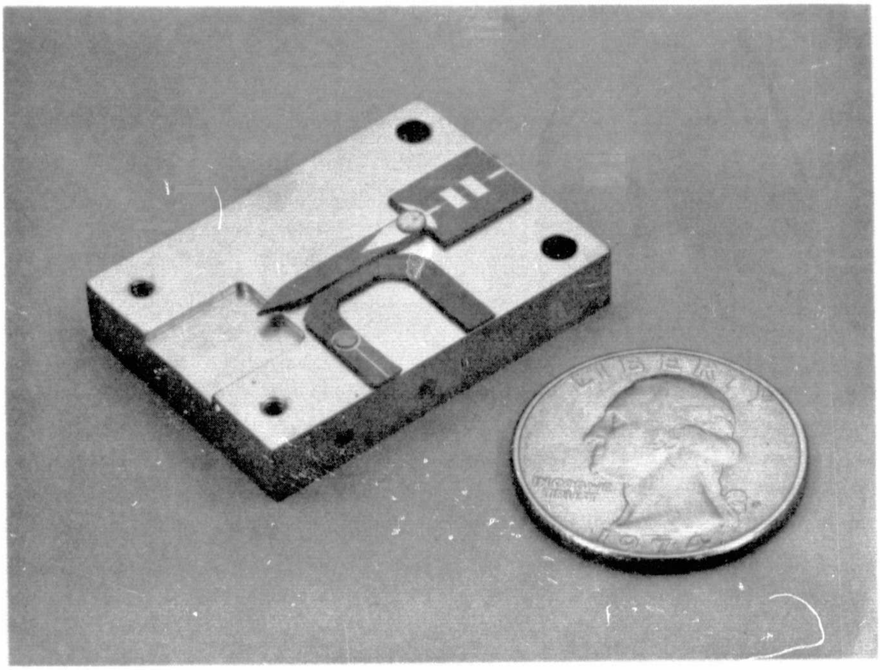


FIGURE III-2

60-GHz Single-Ended Mixer

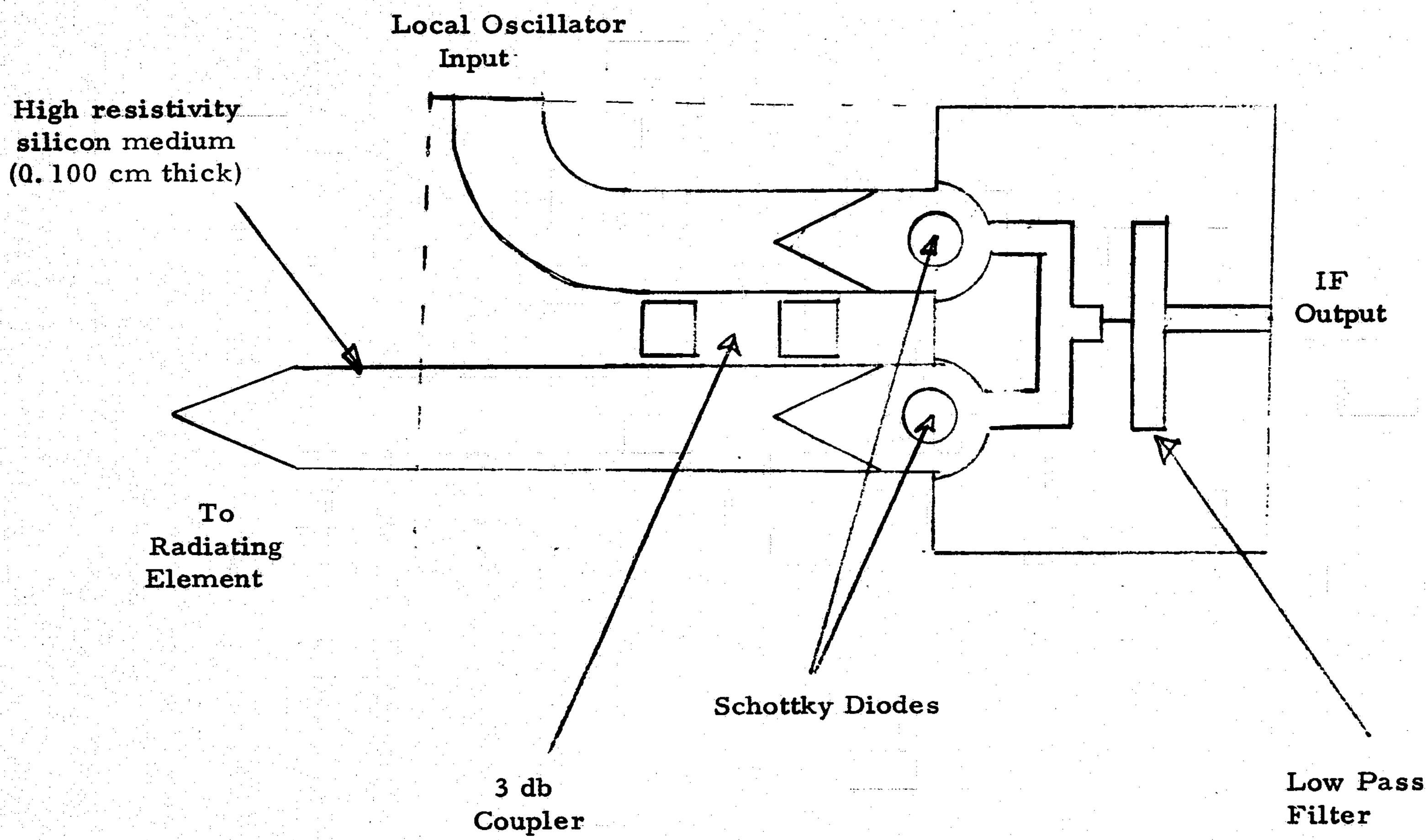


FIGURE III-3

Image Line 30-GHz Downconverters

L. O. drive possible consistent with low conversion loss and sufficient dynamic range. Using Schottky barrier diodes, the L. O. power requirement can be reduced to about 1 milliwatt per diode with the use of a DC bias. DC biasing of the diodes also provides a means to fine tune the IF output impedance of the converter. Typically, 0.7 volts at less than 1 milliampere per diode will provide a suitable bias. The additional requirement for the distribution of this bias among the array elements is felt to be justified in the light of the considerably reduced L. O. power requirement.

Using these components as balanced converters (using matched diode pairs) as shown in Figure III-3 offer several advantages over the single-ended versions. Local oscillator isolation from the RF port and L. O. noise rejection are excellent while a wideband match can be obtained for both the local oscillator and signal ports. Because of the wide frequency separation between L. O. and signal planned here, however, these features may be obtained using a single-ended converter as depicted in Figure III-4. Filters are required at the local oscillator ports to prevent the signal from flowing into the L. O. arm and to match to the diode. A bandpass filter is also required in the signal arm to prevent radiation of L. O. power into the array element. The use of a slot radiator for the array element, which is cut off at the local oscillator frequency, may be used to accomplish this purpose.

The amount of local oscillator leakage into the signal ports must be considered for all of the candidate converter configurations. For example, if the local oscillator (24 GHz) power is 1 milliwatt and the L. O. isolation is 30 dB, -30 dBm will be radiated at each array element. Isolation between elements will be sufficient to prevent sufficient L. O. coupling to produce significant phase errors in the received beams. But, the total radiation from several thousands of elements, each having an element gain of up to 10 dB at the L. O. frequency, could develop an ERP of +10 to +20 dBm at the L. O. frequency.* If these levels of radiation into space are excessive (or unlawful), reductions through additional filtering must be provided.

A system can be considered which utilizes a harmonic mixer, that is, the converter is driven directly with the 8 GHz L. O. and no multiplier is required. This approach typically results in a conversion loss

* This could be used as a beacon.

MODULE ELECTRONICS

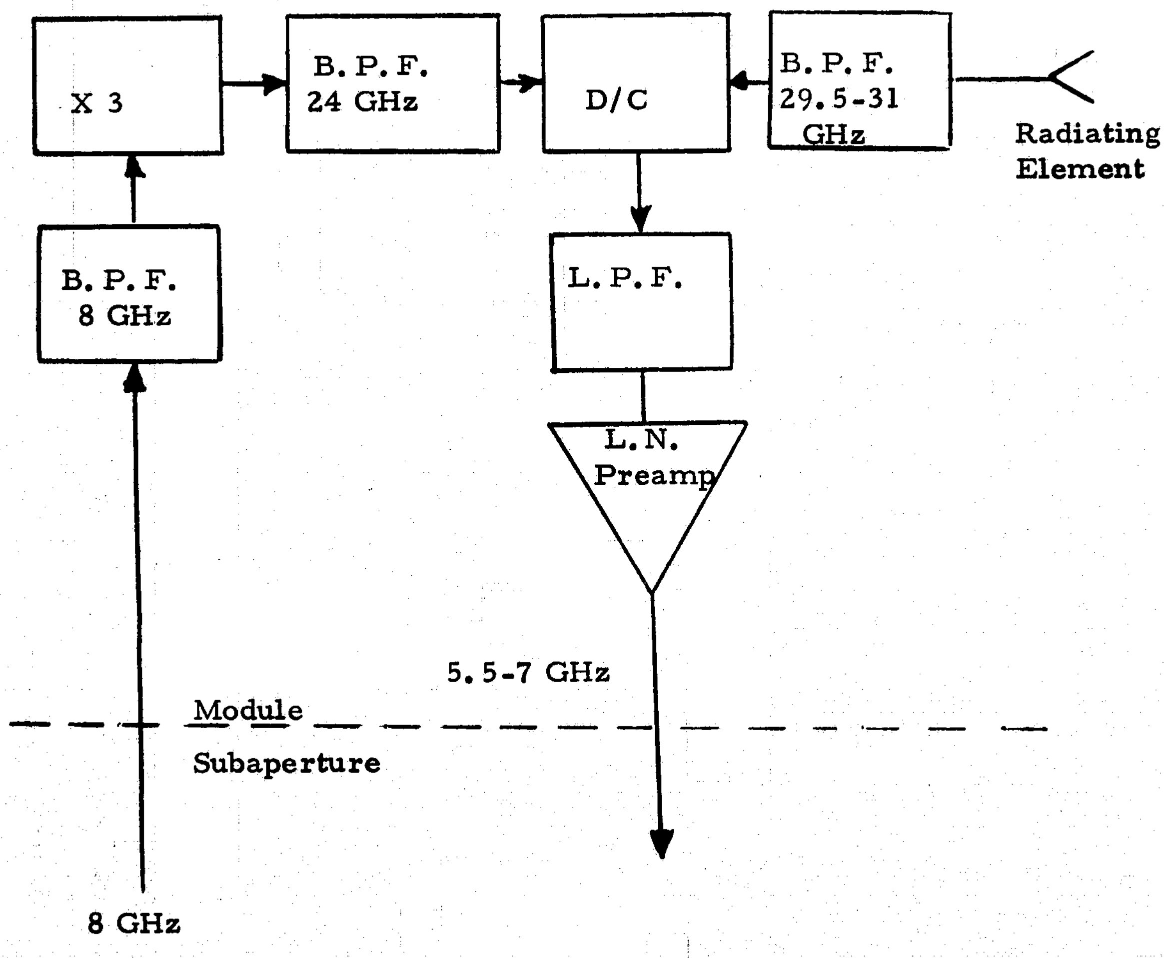


FIGURE III-4

Downconverter Electronics Block Diagram

degradation of at least 3 dB as compared to fundamental mixing. The local oscillator power requirement would not be reduced since additional power necessary for harmonic mixing is about the same as that required to drive the X3 multiplier. So, the use of the harmonic technique chiefly eliminates one component, the multiplier, while increasing the expected conversion loss from 7 to over 10 dB.

1.2 Local Oscillator Source

The circuit layout for a single-ended downconverter integrated with a X3 local oscillator multiplier on a fused quartz microstrip substrate, is shown in Figure III-5. The entire circuit is contained on an area less than 4 cm^2. It is probe-coupled to a rectangular waveguide transmission line or radiating element. Assembled with a sealed cover and base plate this assembly weighs less than 30 grams. Ideally, one stage of IF amplification is integrated with the converter assembly and a total weight of less than 50 grams is expected.

In order to avoid manifolding high-frequency local oscillator signals to each of the many receiver elements, a multiplier approach is used in the baseline system. Using the 8-GHz L.O. driver frequency shown, a X3 multiplier is required at each converter in the communications array. The frequencies involved in the multiplication process and the mixing products are least likely to interfere, or to be interfered with, by the remainder of the system. Also, the stable 8-GHz reference signal generated for this purpose is suitable for use in some of the other two experiments. High-power, high-efficiency components, such as traveling-wave tubes, are available for generation of the required L.O. power and low-loss power dividers have been developed to manifold 8-GHz power to the large number of elements.

X3 multipliers in the frequency range of interest have been developed and demonstrated at the frequencies of interest with up to 30% efficiency. (Schneider and Snell, 1971). Integrated and hybrid construction techniques discussed for use with the downconverter are also suitable for the multiplier. As indicated previously, the X3 multiplier and downconverter are fabricated as an integral unit. In this way, the filtering for the multiplier is combined with the signal-separation filter of the converter.

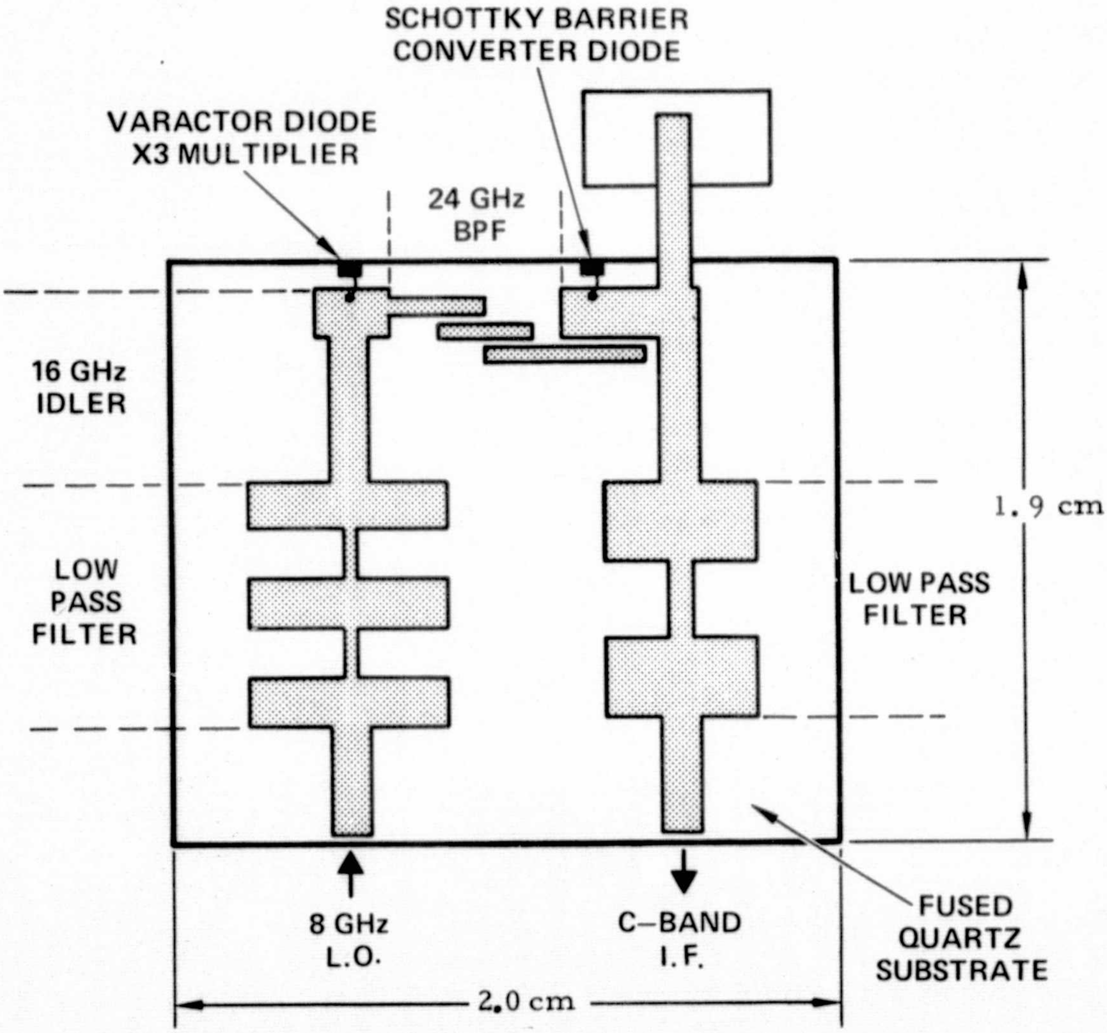


FIGURE III-5

Integrated Downconverter

A functional block diagram of the complete front-end converter assembly for the receive modules, including filters and low-noise preamplifier, was shown on Figure III-4. The 8-GHz input drive signal is introduced through a microstrip filter to the multiplier diode. The bandpass 24-GHz output filter is then constructed of silicon image line, of microstrip, or possibly rectangular waveguide depending on the converter design. Using image-line construction, the expected weight of the complete multiplier/converter assembly would be approximately 15 grams including filters and structure. The use of waveguide construction allows higher performance filters but increases the weight of the total unit to about 45 grams. The chief disadvantage of the waveguide construction is, as indicated previously, higher production cost.

With 1 milliwatt of 24 GHz L. O. power required per diode, 5 milliwatts of 8-GHz signal will be required at the input to the multiplier. The efficiency is reduced from obtainable values of 30% by losses in the filters and by the fact that the unit is operating at a relatively low level.

The local oscillator distribution system will contribute a significant amount of weight to the system. It has been estimated that the power dividers and traveling-wave tubes will require approximately 15 grams per element, in a system with greater than 1000 elements. This estimate was based on a requirement for 10 milliwatts of 8-GHz power at each of the receiver assemblies and does not include the power lost in the interconnecting transmission lines.

The power-division system is most suitably distributed throughout the entire array. Because the array elements are grouped into modules, a divider per module represents a convenient configuration. Each of these module dividers would then be driven by the true-time delay phase element.

The detailed design of the power-divider system for the local oscillator must await further analysis of the remainder of the system and experiments. There is concern for maintaining accurate phase tracking among the elements over a wide range of environmental conditions. Also, it is desirable to use this manifold system economically for the remainder of the experiments. Because the other experiments will use fewer elements, a system which does not waste RF power, or can even be partially removed, is needed. In addition, the baseline transmitter system will use the same 8 GHz reference signal which can, therefore, be included with the L.O. manifold.

1.3 Low-Noise Preamplifiers

The downconverters must be followed by low-noise preamplifiers in order to achieve the optimum system noise figure. Devices have become available within the past few years which offer excellent noise performance at the high IF considered here. The GaAs Field Effect Transistor (FET) offers the best noise performance at frequencies from 5.5 to 7.0 GHz, although some bipolar transistors may be used with good results up to 6 GHz. Because the GaAs FET is a relatively new component, some experimental investigation of its utilization in this role is recommended. A noise figure of 4 dB at a gain of 10 dB will be available from the required number of uniform devices for use in this type of program.

A number of firms are presently marketing GaAs FET's and FET amplifiers for use at frequencies from 4 to 12 GHz. These include Fairchild, Avantek, Watkins-Johnson, Plessey and NEC. Several other firms including Hughes, Hewlett-Packard, Varian and Raytheon have active research programs in this area and can be considerd as possible future sources. Device noise figures of less than 4 dB have been demonstrated using available commercial devices, while laboratory devices have exhibited less than 3 dB.

It is expected that more than one amplifier stage will be required to achieve an optimum noise figure. Also, in order to achieve a proper match between the converter and preamplifier over a large bandwidth, it is necessary to mount the preamplifier as close as possible to the converter, preferably as an integral part of the converter assembly. Unfortunately, this introduces a power dissipating component into the array module separable equipment. The alternative, however, appears to be a severe degradation in receiver performance, not only in bandwidth, but in problems arising fromplacing a critical interconnect between the converter and preamplifier.

The FET is a particularly critical item for matching to the converter because optimum noise figure is achieved with a relatively high input VSWR (greater than 2:1 typically). Brute force methods of obtaining low input VSWR to the preamplifier consist of using an isolator or a balanced pair of devices. These approaches are both being employed presently with success, but are undesirable for this system because of increased weight, power consumption, and component count. Preamplifier development closely tied to that of the converter is, therefore, recommended. A single preamplifier stage could then be part of the downconverter assembly and any remaining stages are then installed in the permanent phase shifter structure.

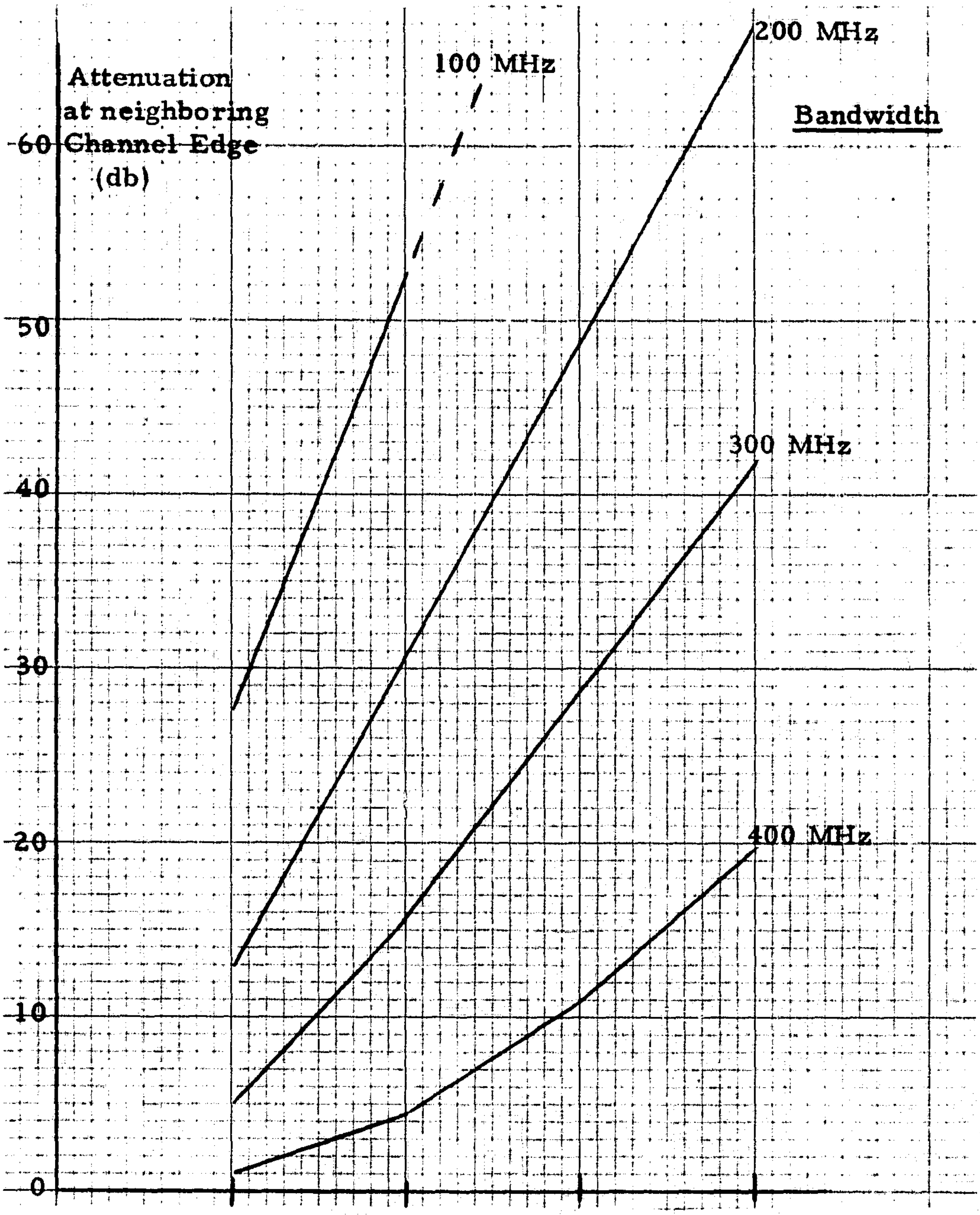
Assuming a 4-dB preamplifier noise figure and a 7-dB converter loss, a SSB noise figure of about 11 dB can be reasonably achieved. Additional filtering requirements ahead of the converter will degrade this figure by 1 to 2 dB. The low-noise preamplifier, using a GaAs FET will require less than 40 milliwatts of prime power per stage.

It is also possible that bipolar transistors may be available for use in the low-noise preamplifier during the time frame of this program. These are expected, however, to offer a noise figure 1 to 2 dB above that available using the GaAs FET. Another device which may be considered is the tunnel diode amplifier (TDA). Although performance equivalent to the GaAs FET can be achieved using the TDA, the required weight and cost are unattractive.

1.4 Three-Channel Multiplexers

The 5.5 - 7 GHz IF signal from the converter/preamplifier will be directed from the array modules to the subaperture electronics. It is here that the three channels will be separated and individually phase shifted to eventually form the three beams. A variety of filter triplexing approaches have been studied to provide the information necessary to evaluate system configurations. These have included construction using rectangular waveguide, airstrip, microstrip on alumina substrates, and coaxial and combline structures using both air and teflon dielectrics. The type of information resulting from this analysis is indicated in Figures III-6, III-7, III-8, and III-9.

The multiplexer calculations were performed for channel bandwidths of 100, 200, 300, and 400 MHz. The maximum channel bandwidth of 400 MHz allows a guard channel of 200 MHz between channels. For a 3-channel system distributed over 1500 MHz, a 400-MHz channel bandwidth is about the limit without putting excessively stringent skirt requirements on the filters. One and two channel systems can be used with 500 MHz channel

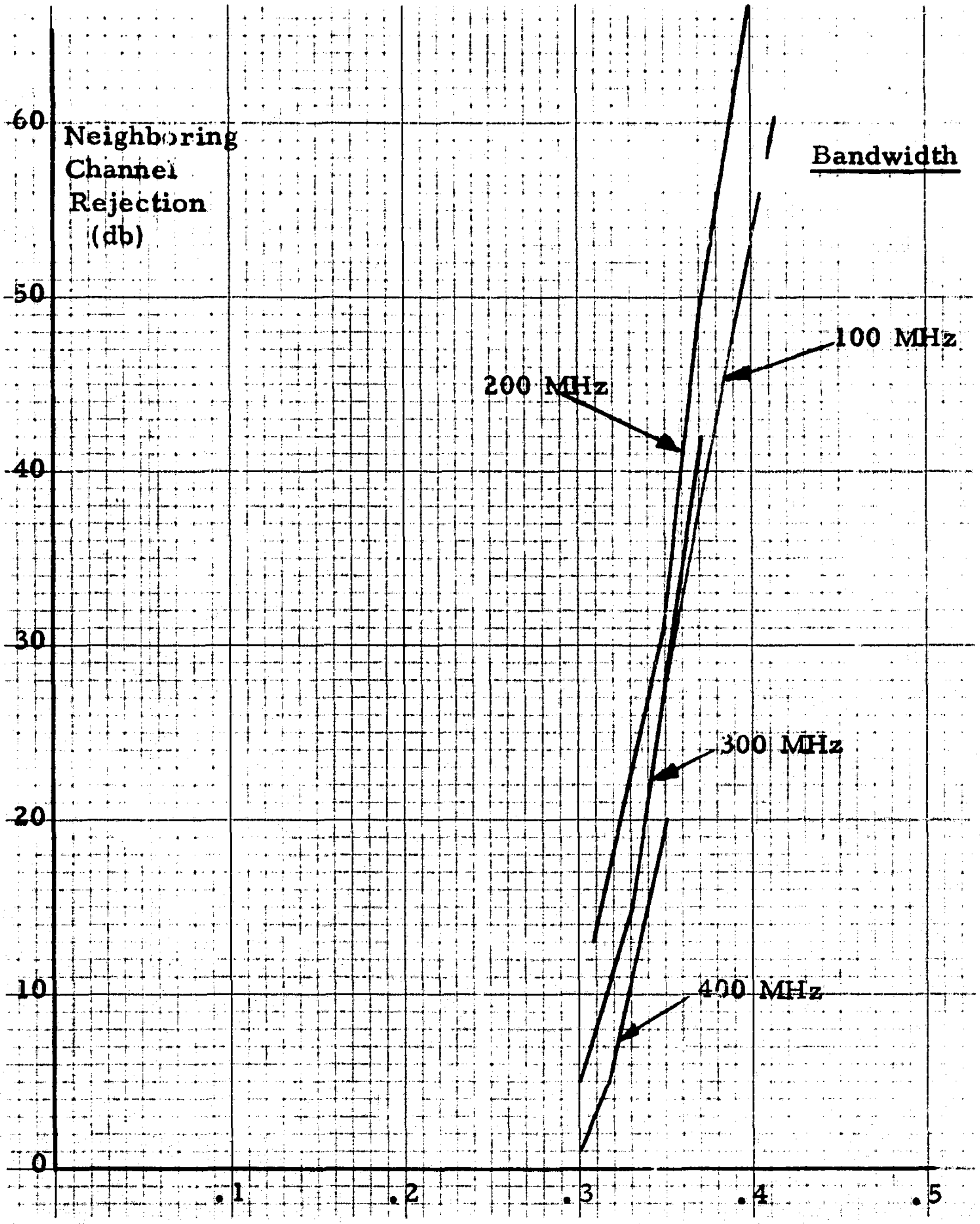


Coaxial or Combline	Wt.	34	40	44	48	grams
	Vol.	16	20	24	28	cm^3
Airstrip	Wt.	71	91	110	130	grams
	Vol.	28	36	44	53	cm^3
Microstrip	Wt.	18	25	32	38	grams
	Vol.	7	9.5	12.5	16	cm^3
Waveguide	Wt.	125	175	230	275	grams
	Vol.	85	145	210	275	cm^3

Assumes 0.1 dB ripple design
Center Frequency = 6.25 GHz

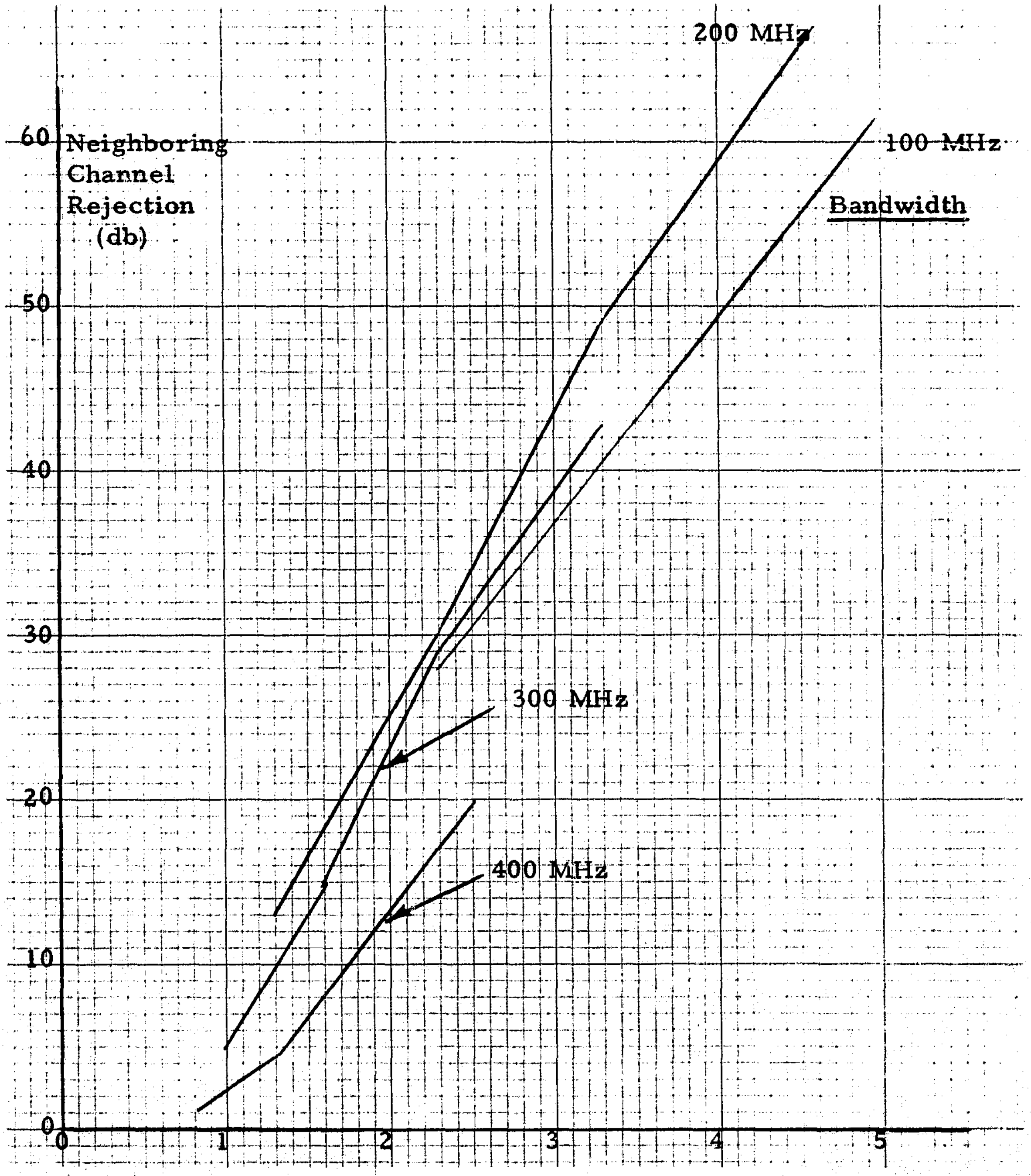
FIGURE III-6

Multiplexer Designs



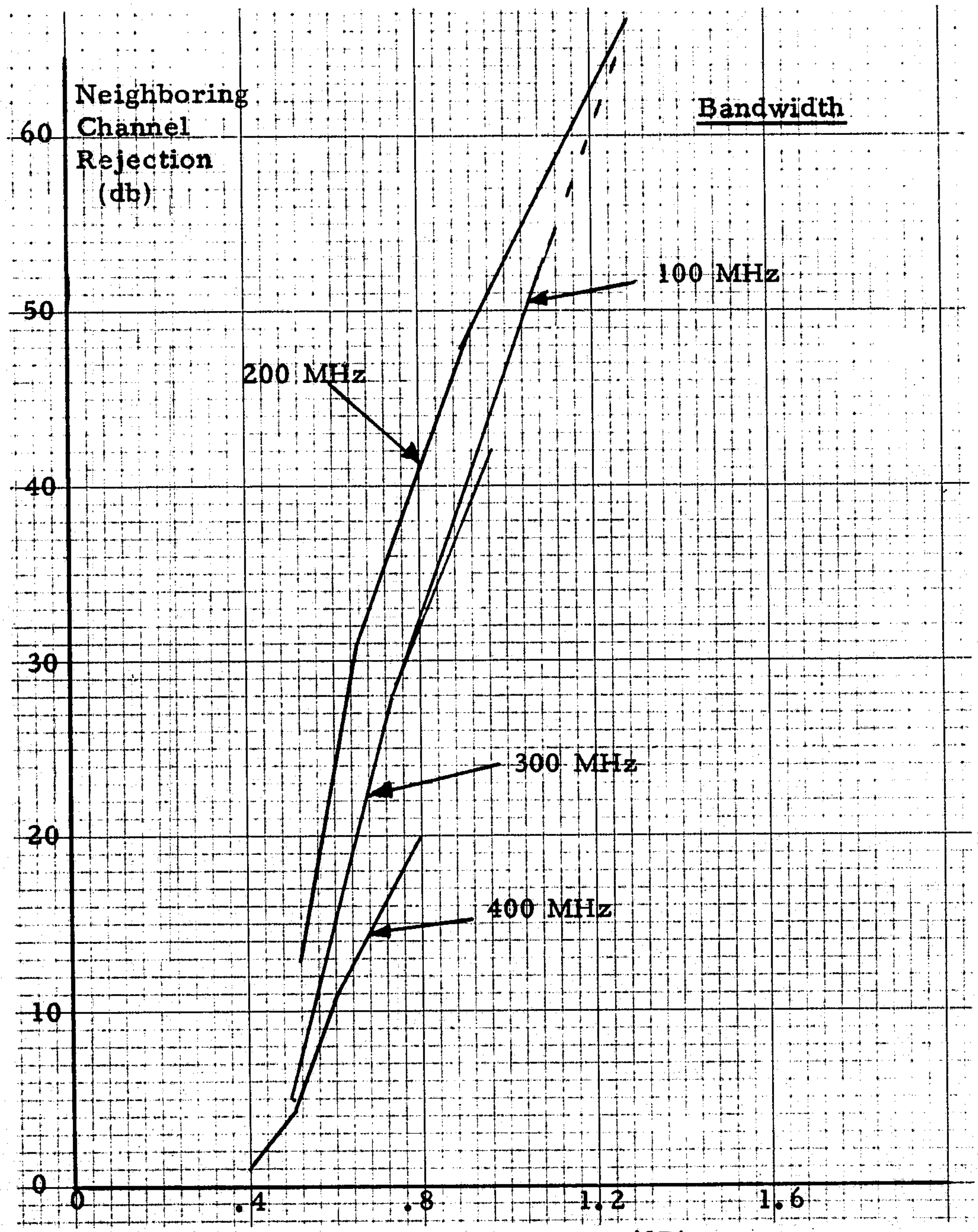
Midband Insertion Loss (dB)
Center Frequency = 6.25 GHz
(Rectangular Waveguide)

FIGURE III-7
Three-Channel Multiplexer



Midband Insertion Loss (dB)
Center Frequency = 6.25 GHz
.025 in. Alumina Substrate

FIGURE III-8
Microstrip Three-Channel Multiplexer



Midband Insertion Loss, (dB)
Center Frequency = 6.25 GHz

FIGURE III-9

Three-Channel Multiplexer - Airstrip

(Also suitable for Coaxial and Combline Structures) bandwidths because sufficient guard channels can be inserted to relieve the filter skirt requirements. A maximum ripple of 0.1 dB was allowed across the band and the bandwidth was defined as the ripple bandwidth.

The skirt selectivity, or attenuation of a neighboring channel signal at a particular channel edge, is plotted as a function of weight and size for different approaches on Figure III-6. It is seen that the alumina microstrip multiplexers are the most attractive in terms of size and weight. The weights include the common point divider and three filters for all of the units. The normally required circulators or isolators are not included in these estimates, but are included in the final weights calculated in Section V. The curves were generated as a function of the required number of filter elements and are, therefore, not meant to indicate continuity between the weights listed.

Insertion losses as a function of interfering channel rejection are shown on Figures III-7, III-8, and III-9 for three different construction types. The performance shown for the airstrip units can also be used for the coaxial and combline structures. The rectangular waveguide yields the best insertion loss performance and the microstrip units the poorest.

All of the performance and physical parameters indicated in the figures are a result of theoretical calculations based on presently available filter design techniques. These parameters were verified by comparison to existing multiplexers fabricated by several different manufacturers.

Other than utilizing the 0.1 dB ripple designs, no further provisions were included to linearize phase. Typical devices of this type will exhibit phase deviations from linear of about $\pm 5^{\circ}$ across the band. Group delay equalizers can be added to the multiplexers, but these are normally heavy and expensive. It may be most suitable to provide pre-distortion at the transmitting ground station if stringent equalization is indeed required.

Some conclusions may be drawn from the multiplexer analysis which have led to our present view on the system implementation. The microstrip multiplexer appears to be the only version offering suitable size and weight for the system considered. However, the insertion loss exhibited by the microstrip filters is excessive. For this reason, it appears worthwhile to examine operation of a simple 3-way divider operating directly into the IF combiner networks. In this approach, 5 dB of loss will be introduced into each channel, but it is felt that an

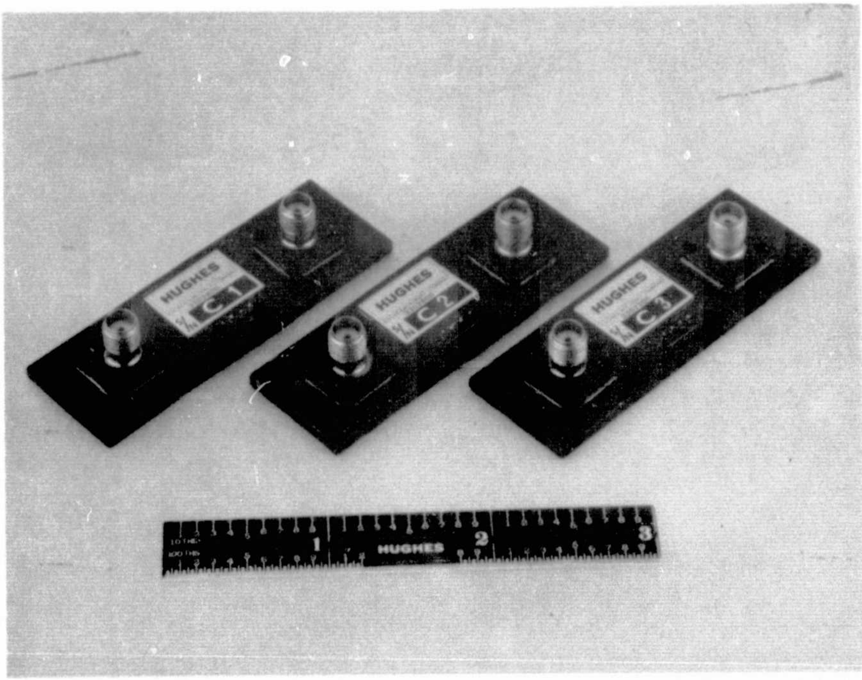


FIGURE III-10

X-Band Three-Bit Diode Phase Shifter
(Coaxial Connectors)

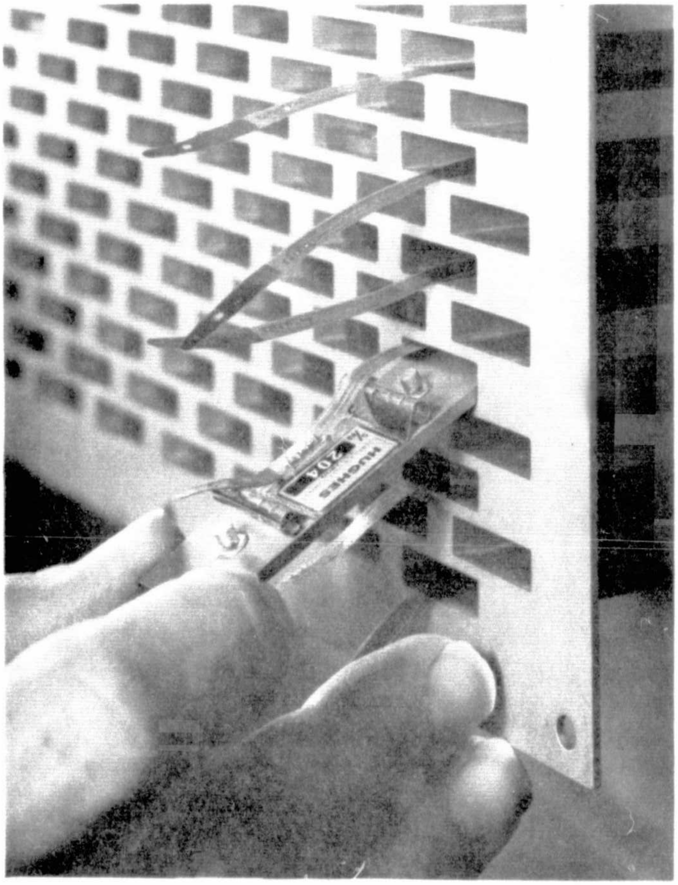


FIGURE III-11

X-Band Phase Shifter in Array Configuration

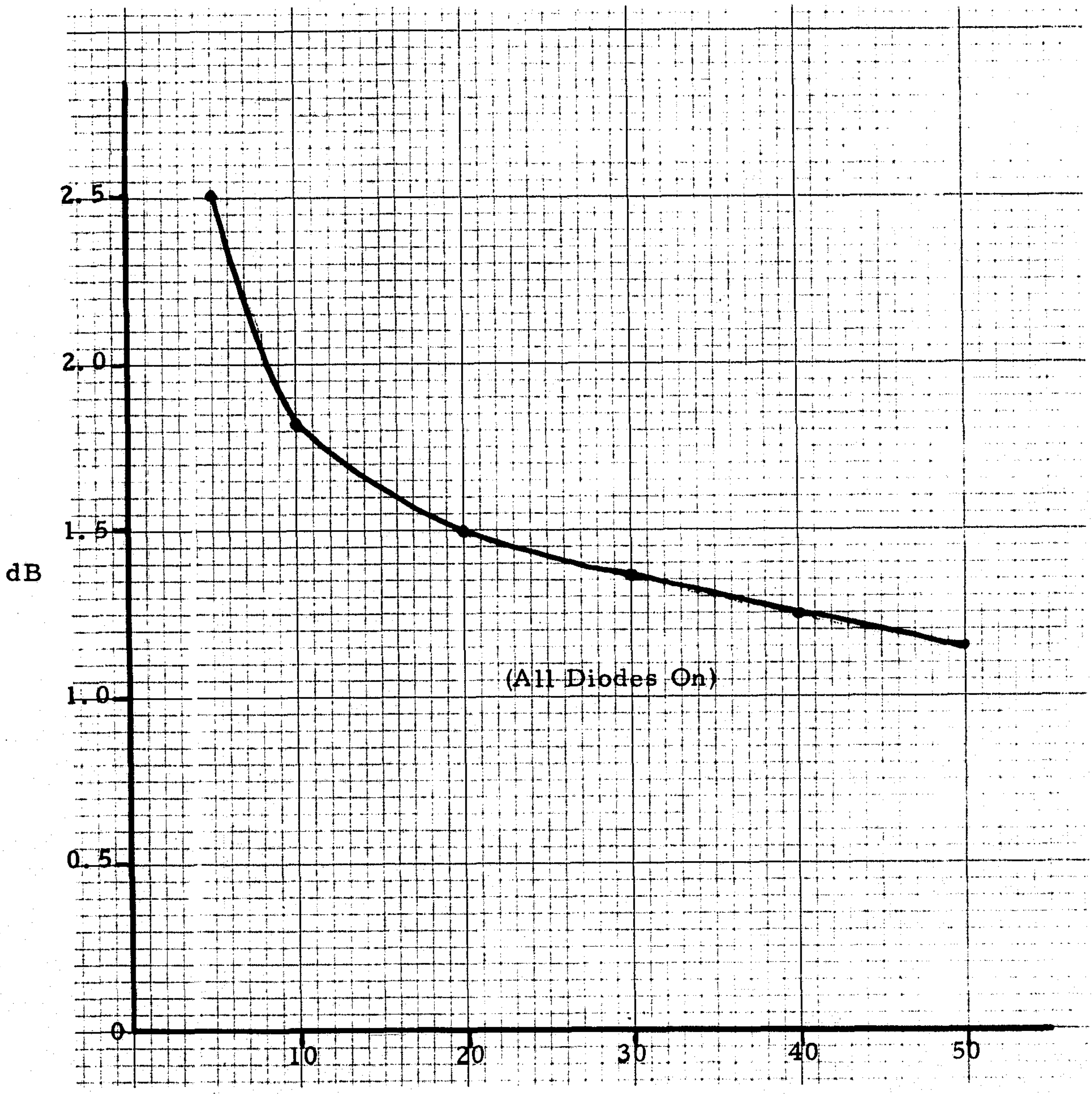


FIGURE III-12

3-Bit Phase Shifter Insertion Loss
as a Function of Diode Current

The driver power consumption for the phase shifters in the system can be estimated on the assumption that on the average one-half of the bits are biased on at any one time. Because each 3-bit phase shifter required 6 diodes, operated at 20 milliamperes per diode, for example, 60 milliamperes are required for each phase shifter.

RF power handling capability is not an important consideration for the receive phase shifters, but, phase bit accuracy, linearity, and tracking from unit to unit must be controlled. In addition, some concern must be directed toward the possible generation of intermodulation signals when multiple frequencies are present. This latter problem is of no concern when the input signal levels at the phase shifters are below -50 dBm.

2.0 Spurious Signal Generation

The level of spurious signals, generated chiefly through intermodulation in the various nonlinear devices, have been examined. The results of this examination is summarized here.

The only products which were found to merit serious consideration were the third-order products produced by the three channels and fourth-order products introduced through transmitter leakage (19.7 to 21.2 GHz). These signals were traced through the system, over a range in power levels, and the output signal to spurious signal ratios were estimated. Presently-available empirical data was used to estimate the products for the converters and preamplifiers. An example of the behavior of the products in a 30-GHz converter is given in Figure III-13. This behavior is typically a strong function of the local oscillator level.

If the three signal frequencies are designated f_1, f_2, f_3 and the transmitter and L. O. frequencies are designated f_T and f_L, respectively, the products of concern are:

$$(2f_1 - f_2) - f_L$$

$$(2f_3 - f_2) - f_L, \text{ etc.,}$$

and $(2f_T - 2f_L)$, the fourth-order product.

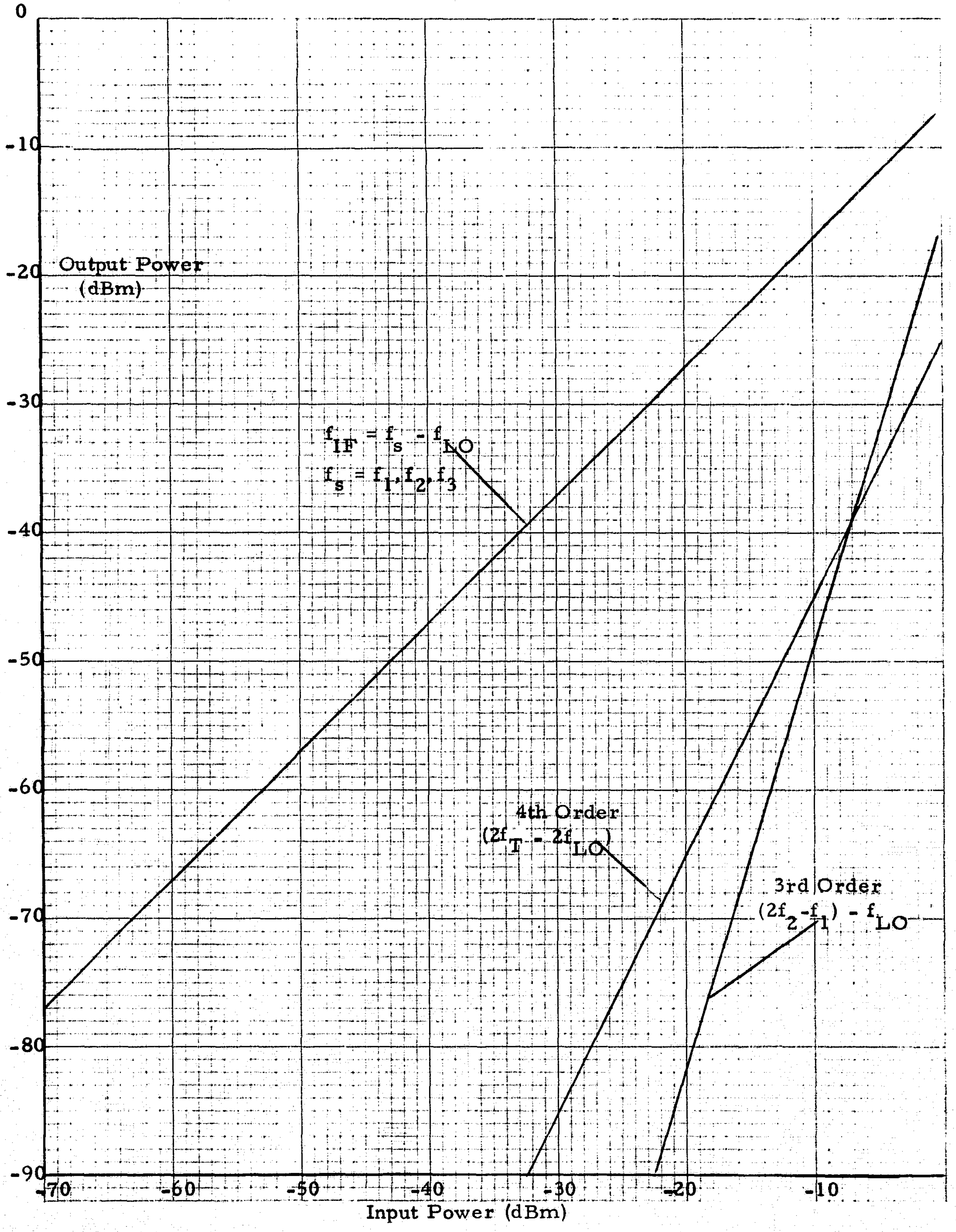


FIGURE III-13

Intermodulation Products - 30-GHz Downconverter

The low-noise preamplifiers, using GaAs FET's also can generate third-order products which lie within the IF passband

$$(2f_{i1} - f_{i2})$$

$$(2f_{i3} - f_{i2})$$

$$(2f_{i2} - f_{i3}), \text{ etc.},$$

where $f_{i1}, f_{i2}, f_{i3} = (f_1, f_2, f_3 - f_L)$, respectively.

A plot of the intermodulation performance of a single-stage GaAs FET amplifier, operating at 7 dB gain, is shown on Figure III-14. The third-order products are shown for two equal level carriers at the power indicated by the fundamental signal curve. Increasing the gain by adding more stages would essentially decrease the input power level for a given intermodulation level, by the added gain.

It is estimated that for incoming 30-GHz signals below -50 dBm at the input to the downconverter, all third-order products will be at least 50 dB below the signals at the IF combiner. The fourth-order products depend strongly on whether balanced or single-ended converters are used. However, it is estimated that with transmitter leakage levels below -50 dBm into the converter no problems will occur for the common modulation formats. In the analysis, the fact that spurious signals are not properly phase shifted, in general, to add in the beam-forming network was considered. Detailed limitations on signal levels can be determined from an analysis of the frequencies involved, modulation format used, and beam steering requirements.

3.0 Transmitter Electronics

Electronic configurations for the communication array transmitter are limited in number because of the performance required of key components. Converters and amplifiers operating at the required power levels cannot be used with multiple signals without producing intolerable intermodulation products. Consequently, these components must be duplicated in each channel. A transmitter configuration capable of radiating and independently steering three beams in the 20-GHz frequency range is shown

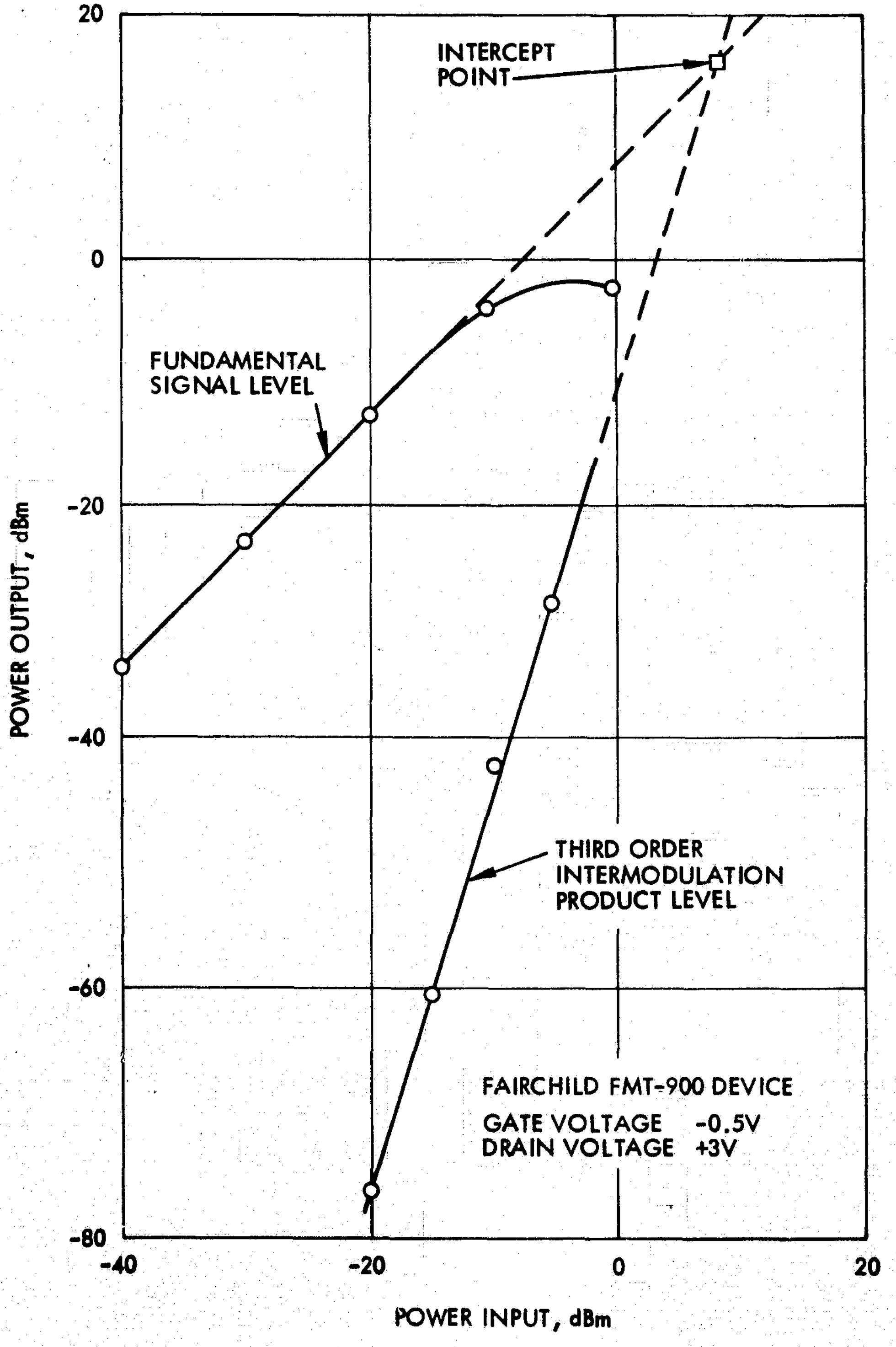


FIGURE III-14

GaAs FET Intermodulation Performance

in Figure III-15. The signal for each channel is divided before upconversion and phase shifting as shown. Multiplexing occurs following all amplification and frequency conversion to eliminate the intermodulation problems. As shown, the system consists of three power dividers, one for each channel or array port.

The upconverters can be fabricated using the same techniques discussed for the receiver downconverters. An 8-GHz stable reference signal is doubled to provide the required 16-GHz local oscillator for the converter. Resistive-type upconverters can be expected to deliver up to 10 milliwatts near 20 GHz with the proper drive levels. Higher power levels can then be obtained using solid-state power amplifiers in the position indicated on the figure. The signals for the 3 channels are combined in the filter triplexer and delivered to the radiating element.

In order to develop 5 to 10 milliwatts from the wideband upconverters, up to 100 milliwatts of power at 8 GHz will be required in the X2. In addition, up to 50 milliwatts of power will be required for each channel at the 3.7 to 5.2 GHz signals. These power levels are most suitably generated ahead of the power dividers using traveling-wave tubes. Some of this power may be distributed in the distribution system using solid-state amplifiers. In the 3.7 to 5.2-GHz frequency range, bipolar transistors can be used to generate up to several watts. At 8 GHz, Gunn diode amplifiers can deliver over 0.5 watts, avalanche diodes can produce up to 5 watts, or perhaps a field-effect power transistor could be used to produce several hundred milliwatts. An exact distribution of power levels and amplifiers must await a more firm system design. For purposes of weight and power estimates at the present, traveling-wave tubes were used.

Pin-diode 3-bit digital phase shifters, similar to those discussed for the receiver are used here. These can easily be used at power levels up to several watts before encountering any distortion problems.

Solid-state power amplifiers for operating near 20 GHz are most suitably constructed using Gunn or avalanche diodes. Gunn diodes are capable of up to 200 milliwatts at this frequency, while silicon avalanche diodes (IMPATT's) can deliver well over 500 milliwatts, and GaAs

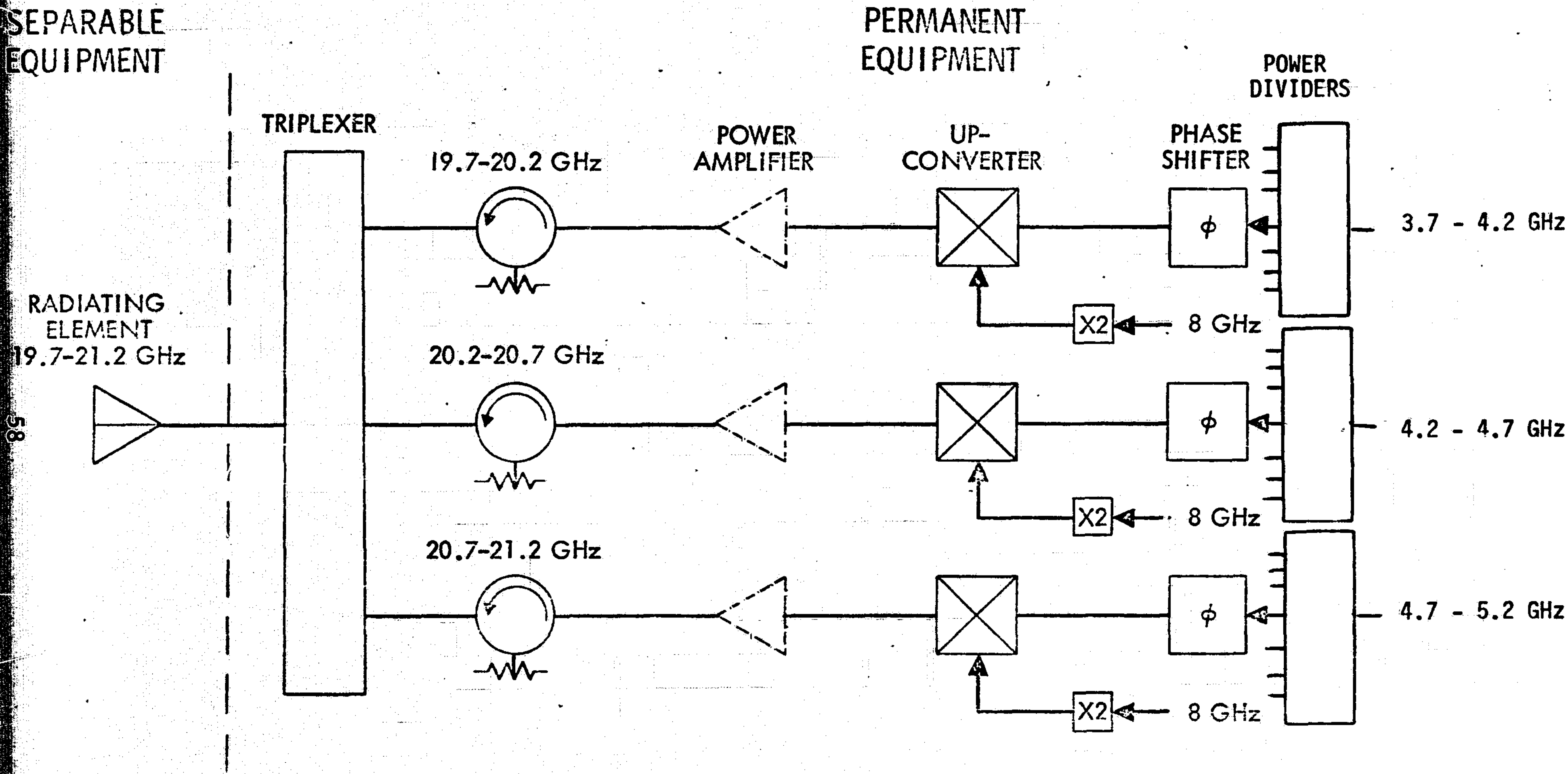


FIGURE III-15

Three-Channel Communication Transmitter

IMPATT's may deliver over 1 watt. It is possible that a power GaAs FET may be available during the time frame of this program with over 100 milliwatt capability. The Gunn and IMPATT amplifiers weigh 30 to 50 grams per amplifier stage, depending on heat-dissipation requirements.

The filter triplexers for use at the output of the transmitters must use waveguide techniques in order to achieve an acceptable insertion loss. Using filters of an elliptical design with acceptable bandpass characteristics, the triplexer is expected to weigh approximately 150 grams. Addition of the three input isolators will increase the weight of this output assembly to about 240 grams. The insertion loss from the input of the isolator to the radiating element will be 1.5 to 2 dB.

Estimates for the ERP of the communication array transmitter are made on the assumption of a 12-dB element gain. Therefore, a 4000-element array radiating 5 milliwatts per element, for example, would have an ERP = 25 dBw as follows:

Radiated power per element	-23 dBw
Element gain (at edge of scan)	12 dB
Array gain (4000 elements)	36 dB
ERP	25 dBw

This magnitude of ERP is adequate for communication with permanent ground terminals as estimated from the following link calculations:

Transmitter ERP	25 dBw
Space Loss (400 Km)	-172 dB
	-147 dBw
Noise power	-228.6 dBw/°K-Hz
Bandwidth (300 MHz)	85 dB
Signal-to-Noise	20 dB
	-123.6 dBw
Required System G/T	23.4 dB

Ground system noise temperature is given by

$$T_s = T_a + (L - 1)\, T_L + LT_R$$

where T_R = receiver input temperature
T_a = sky temperature
L = insertion loss between antenna and receiver
T_L = loss temperature

Assuming $T_a = 50^oK$, $L = 0.2$ dB, $T_L = 290^oK$, and $T_R = 1500^oK$ (7.8 dB noise figure),

$$\begin{aligned} T_s &= 50 + (1.047 - 1)290 + (1.047)(1500) \\ &= 50 + 13.6 + 1570 = 1633.6^oK \\ &= 32.1 \text{ dB} \end{aligned}$$

Therefore, the required G/T can be achieved with an antenna gain of G_r = 32.1 dB + 23.4 dB = 55.5 dB. At 20 GHz this gain will result in the use of a 4-meter diameter parabola.

4.0 Component Summary

A survey of available devices indicates converters, preamplifiers, and phase shifters are available with the properties discussed. To allow for the large-scale production of the items required for the baseline system, in the necessary quantities, dedicated development programs will be necessary. Small, lightweight, economical devices with optimum performance can be obtained in most cases with an effort of 6 to 12 months.

A weight of approximately 50 grams is realizable for the down-converter assembly, which includes necessary filters, X3 multiplier, and 1 preamplifier stage. A receiver noise figure of 11 dB can be achieved over 1500 MHz bandwidth with a local oscillator power requirement of 10 milliwatts at 8 GHz. In addition, a maximum of 2 milliwatts of DC bias will be required for the converter and 40 milliwatts for the preamplifier.

The output of the converter assembly will be followed by one or two stages of low-noise amplification requiring 60 milliwatts and 15 grams per stage. A filter multiplexer will follow, if required. Using micro-strip, a 3-way multiplexer will weigh 100 to 150 grams, including an isolator, and contribute 2 to 4 dB of insertion loss per channel. A coaxial or combline triplexer will weigh 150 to 200 grams and exhibit 0.6 to 1.2 dB of insertion loss. Considerable weight can be saved, therefore, by using a simple 3-way divider assembly and accept an insertion loss of 5 to 6 dB.

Pin-diode 3-bit digital phase shifters are most suitable for the IF chosen. These will weigh 30 grams and require an average of 30 to 60 milliwatts drive power to achieve 1.5-dB insertion loss. The IF combiners used to form the beams require 15 grams per element per channel.

5.0 Weight and Power Comparison

It is suitable to examine the weight and power requirements of the baseline transmitter and receiver configurations before presenting other variations on the system. Quantities are listed for the full three-channel receive and transmit system, as well as for operation with a reduced number of channels. The weight and power consumption for the receiver elements are listed on Table III-I, and those for the transmitter on Table III-II. The receiver elements are essentially capable of achieving a noise figure of 11 dB.

On examination of Table III-II, it is seen that the weight and power requirements of the transmitting elements increase rapidly with increasing output power. The inclusion of power amplifiers in the elements does not improve the situation. The principal reason for this is none of these components (Gunns, IMPATT's) operate efficiently at these power levels.

Receiver weights on Table III-I are shown for the cases where micro-strip filter triplexers or 3-way dividers are used. For these estimates, a 300-MHz channel bandwidth, with 30 dB of out-of-band rejection, was used. Use of combline, airstrip, or even waveguide filters would simply increase the weight by the amounts indicated in Figure III-6. An additional stage of amplification is included in the cases using the 3-way dividers to compensate for the 5 dB of additional insertion loss. In this way, the noise figure at the input to the converter can be maintained near 11 dB.

Table III-I

Baseline System Communication Array Receiver Electronics

	Weight/Element			Power Consumption per Element	Noise Figure +
	Separable Equipment	Permanent Equipment	Total Weight		
<u>3-Channel Receiver</u>					
Filter Triplexer (microstrip)	50 grams	310 grams	360 grams	.40 W	11 dB
Divider Triplexer	50	245	295	.52*	11.5
<u>2-Channel Receiver</u>					
Filter Triplexer	50	235	285	.31	11
Divider Triplexer	50	185	235	.39*	11
<u>Single Channel Receiver</u>	50	95	145	.23	11

*Power requirement may be reduced with 1 dB increase in noise figure.

+Includes losses between array element and converter.

Table III-II

Baseline System Communication Array Transmitter Electronics

	Weight per Element	Power Consumption	Radiated Power
3-Channel Transmitter			
Low power upconverter	555 grams	.74 Watts	1 milliwatt
High power upconverter	735	4.50	10
Gunn/IMPATT Amplifier			
Low power	735	3.0	10
High power	1090	8.1	100
Single Channel			
Low power U/C	150	.25	1 milliwatt
High power U/C	210	1.50	10
Gunn/IMPATT Amplifier			
Low power	210	1.0	10
High power	270	2.70	100

6.0 Configuration Trade Offs

The baseline system configuration achieves the desired multichannel, multiple beam performance with a receiver noise figure of around 11 dB. However, the weight and power consumption for the transmit and receive elements are undesirably large and methods of reducing these requirements must be sought. A large portion of the weight is contributed by the requirement for signal separation in the 3-channel system and by the many power dividers and combiners. Approaches which reduce the requirement for these items are, therefore, of interest.

A large number of component configurations have been evaluated and compared. Some of these were basically variations of the baseline system, while others represented a significantly different approach. Two of these approaches are summarized here.

6.1 Millimeter Wave Multiplexed System

In this configuration, shown on Figure III-16, the incoming receive signals are separated at 30 GHz; phase shifting and power combining take place before downconversion. Ferrite phase shifters are used because significantly lower insertion losses are available than with the diode devices at 30 GHz. Three-bit ferrite devices can be developed which weigh about 30 grams and exhibit about 1.5 dB of loss. A multiplexer system, including filters and an isolator, for the 3-channel system will weigh about 100 grams and have 2 dB of loss. The multiplexer weight and loss will increase significantly, if stringent requirements are placed on out-of-band rejection and phase linearity.

The 30-GHz combiner will be a critical item for this configuration. In order to eliminate long waveguide runs, it is most suitable to combine a suitable small number of elements, say 16, and then perform the downconversion. The remainder of the power combining system then operates at IF where losses are lower and phase tracking through lengthy lines are more easily achieved.

Assuming a noise figure of 11 dB at the output of the combiner, i.e., input to the downconverter, a receiver noise figure of 16 dB can be expected for the 3-channel system. The chief advantage of this appraoch is its relative simplicity and low power consumption requirements.

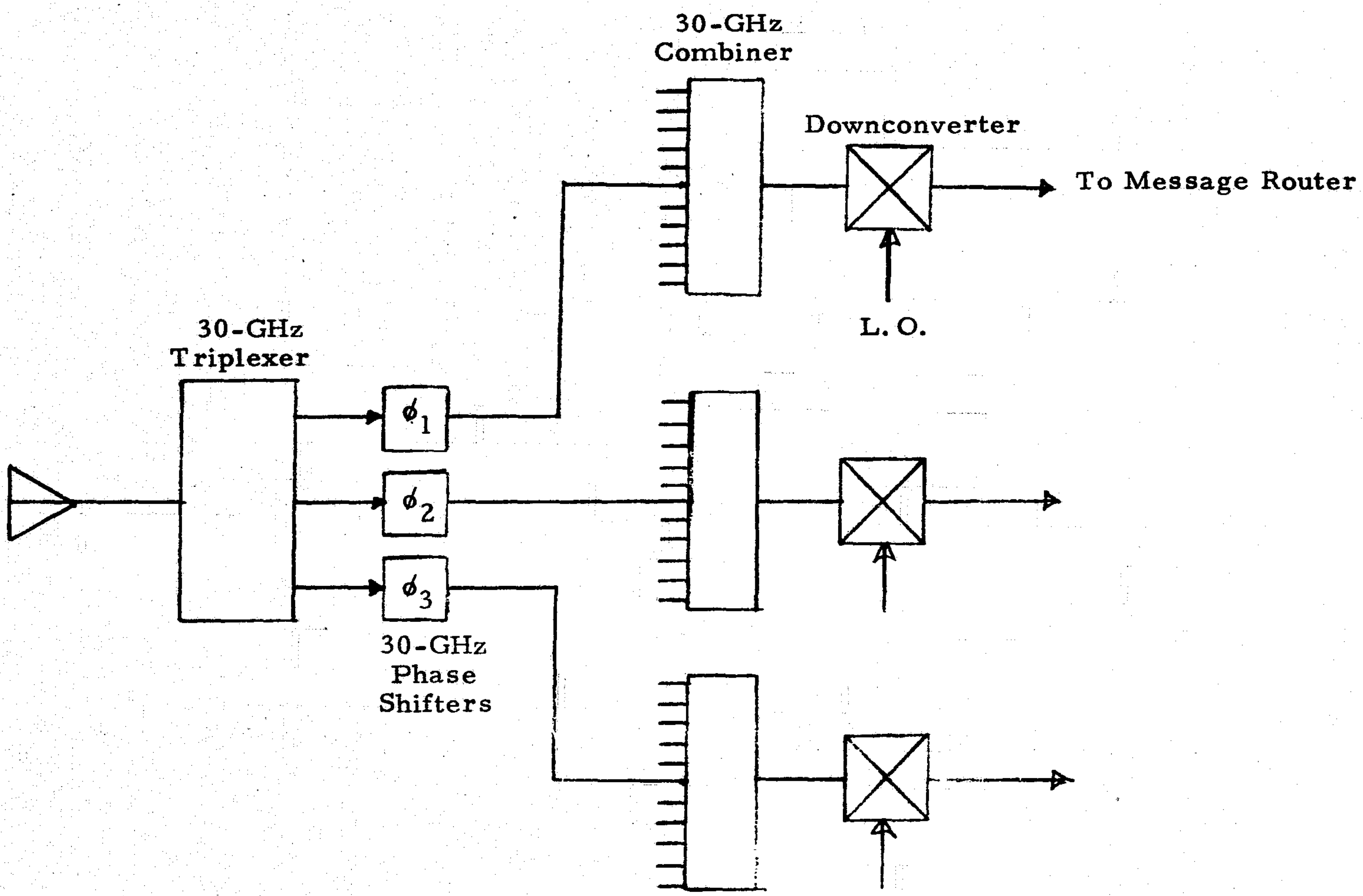


FIGURE III-16

Millimeter Wave Multiplexed System

Because the ferrite circulators are latched devices, the actual power requirement is determined by the rate at which the phase shifts must be updated. It should be also noted that very little commonality of equipment exists with the radar and radiometric experiments when this configuration is used.

6.2 Single Combiner Receive System

The system shown on Figure III-17 is an example of an approach which utilizes a single IF combiner for a 3-channel system. Here the receive signals are downconverted to IF and fed to a phase shifter assembly. This assembly consists of a 3-way divider, the three phase shifters, and a 3-way combiner. All three signal channels are present in each of the phase shifters, but each phase shifter inserts the correct phase shift to form a beam to only one of the channels. The three channels are separated into the proper beams by the filter triplexer after combining. Because only a single triplexer is required, some expense may be incurred to utilize a high-performance unit here.

Used with simple 3-way dividers and combiners, the phase shifter assembly will introduce over 11 dB of loss to the signals. The preamplifiers must, therefore, have near to 30 dB of gain to achieve a noise figure of less than 12 dB for the system. This will require at least four stages and, therefore, some additional weight and power consumption will be introduced. Filter multiplexers may be used in place of the 3-way dividers and combiners to reduce the insertion loss of the assembly to about 5 dB, again at the expense of some weight increase. This would, however, reduce the gain requirements of the preamplifier which results primarily in a slight power savings.

The estimated weight and power requirements for the Millimeter Wave Multiplexed approach and the configuration using a single IF combiner are compared on Table III-III. It is seen that performing the phase shift and multiplexing functions at 30 GHz results in a system with a relatively high noise figure, low power requirements, and a modest weight. The single combiner system achieves essentially the same noise figure as the baseline system, but at a slight reduction in weight. This approach requires the development of the phase shifer assembly, and some further examination of its intermodulation performance is required.

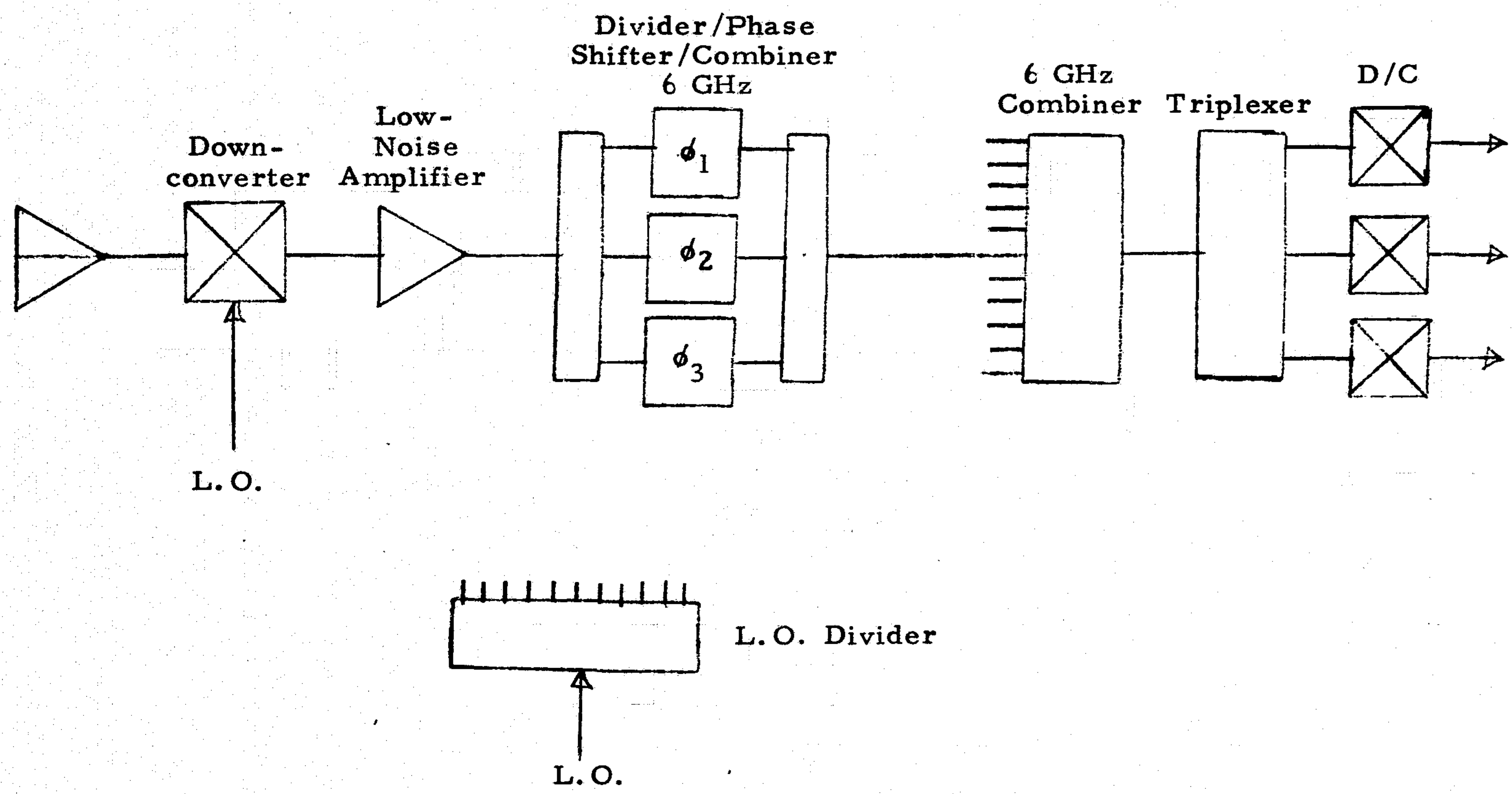


FIGURE III-17

Single Combiner System

TABLE III-III

Communications Array Receiver Systems

	Weight				
	Separable Equipment	Permanent Equipment	Total	Power Consumption	Noise Figure
Millimeter Wave Multiplexed System					
3 channels	0	250 grams	250 grams	.10 watts minimum*	16 dB
2 channels	0	175	175	.06 " "	16 dB
1 channel	0	60	60	.03 " "	14 dB
Single I.F. Combiner System					
(Filter Multiplexer)					
3 channels	50 grams	230	280	.40	11 dB
1 channel	50 grams	95	145	.23	11 dB
Divider Assembly)					
3 channels	50 grams	175	225	.44	12 dB
1 channel	50	95	145	.23	11 dB

*Actual power consumption is determined by the beam switching requirements.

In addition to the variations discussed here, others were evaluated which eliminated or combined the function of components. These chiefly resulted in a performance degradation not justified by the achieved weight reduction. A configuration was examined, for example, in which the L.O. power divider was eliminated by feeding the L.O. signal through the IF combiner. However, the weight of the circulator and filter required to separate the L.O. from the IF signal in the element weighs more than the weight saved by eliminating the L.O. divider. Also, techniques were examined in which the local oscillator is phase shifted rather than the received signal or IF. These approaches generally did not appear to offer any advantage in performance or weight over those discussed here.

6.3 RF Multiplexed Transmitter System

The possible configurations for the transmitter elements are somewhat limited because of the component requirements. One approach which must be considered, however, is the RF-multiplexed/RF-phase shifted system shown on Figure III-18. In this approach the 20-GHz communication signals are amplified and then distributed to the elements through the 20-GHz power dividers. Traveling-wave tubes are most suitable for the power amplification. Available tubes for this function include the Hughes 268H, which is a 2-watt device weighing about 1 kilogram including power supplies. A tube develoment program could make available a device capable of several hundred watts with no advancement of the state of technology, if required. These tubes operate at about 18% efficiency. The tubes would be most economically distributed appropriately throughout the power divider system. Elements would then be grouped with a power divider driven by a single TWT.

Ferrite phase shifters capable of several watts of power, without distortion, are used for beam steering. These introduce about 1 dB loss and weigh approximately 40 grams each. A diode phase shifter could be considered for this application, with an insertion loss of 2.5 dB and weight of about 20 grams. Such a device is not presently available, but is within the present state of technology.

The 20-GHz triplexer includes the channel combining filters and isolators. This item is expected to exhibit 2 dB of loss and weigh 120 grams. It is possible to eliminate the triplexer by using a separate

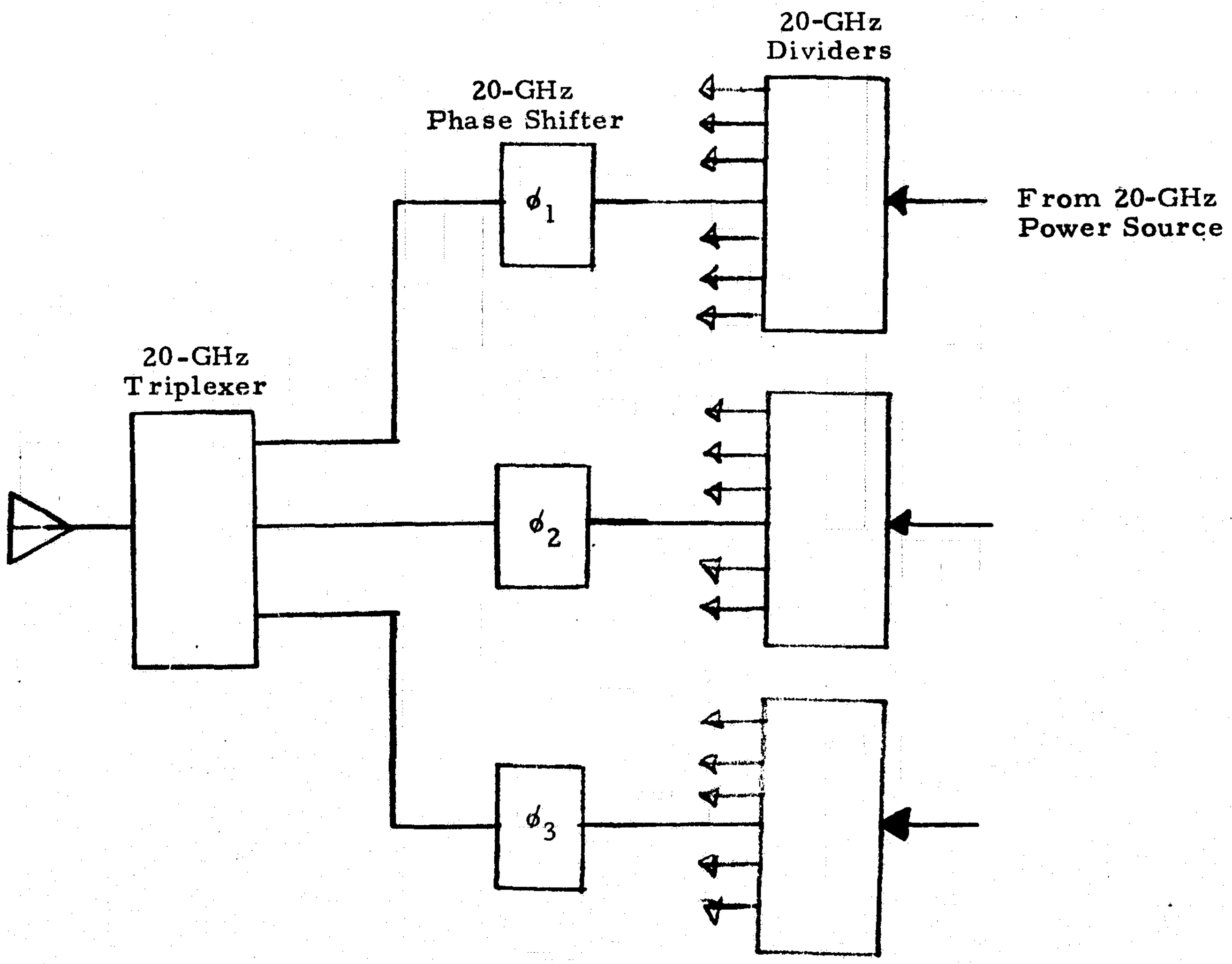


FIGURE III-18

RF Multiplexed Transmitter

radiating element for each channel. Because the array is thinned, space is available for the installation of the extra radiating elements. As is the case with all of the configurations in which the signals are multiplexed and phase shifted at RF, very little commonality exists with the other experiments.

Weight and power consumption, as well as radiated power per frequency channel are listed on Table III-IV for the transmitter configuration of Figure III-18. Because ferrite phase shifters are used, the actual power consumption is a function of the beam steering rate. It is seen that weight and power consumption for this configuration are attractive compared to the baseline approach. The weights shown do not include the radiating elements or transmission lines.

The quantities shown on the table assumed a network of 16-way dividers weighing 240 grams and with 1.5-dB insertion loss as the basic building block. Two-watt traveling-wave tubes, preceded by an upconverter driver (essentially an element of the baseline system), were used to develop the radiated power levels shown.

The development of the lightweight (40 grams) ferrite phase shifters is the key item for this configuration. Although cost has not yet been included in the analysis of the system, the ferrite devices are expected to be more expensive than the diode units.

B. 13.9-GHz Radar System Implementation

The electronics configurations developed for the communications array were chosen for a number of reasons which do not apply to the radar system. Several of these reasons are reviewed here:

1) The millimeter wave transmit and receive frequencies make it undesirable to power combine or divide directly at RF. Insertion loss and weight for these components are unattractively large at 20 and 30 GHz.
2) The wide bandwidth desired for the communication system requires the use of high intermediate frequencies. Consequently, IF component performance is somewhat degraded.

TABLE III-IV

Communications Array Transmitter Systems
(Figure III-18)

	Weight	Power Consumption	Radiated * Power
RF Multiplexed/ Phase Shifted System			
3 Channels			
High Power	750 grams	2.75 Watts	40 milliwatts
Low Power	450	0.20	1
1 Channel			
High Power	210	0.80	63
Low Power	110	0.13	1.6
1 Radiating Element per Channel (Triplexer not required)			
3 Channels			
High Power	630	2.75	63
Low Power	285	0.20	1.6
1 Channel			
High Power	210	0.80	63
Low Power	110	0.13	1.6

* Radiated Power per element per channel.

3) The number of array elements must be extremely large to obtain the desired beam performance from the communication array. Great attention must, therefore, be directed toward the reduction of weight and power consumption on a "per element" basis.

The electronics configurations for the array radar are most efficiently developed under a different set of ground rules. However, it is the intention of this study to also evaluate a system which offers considerable component commonality with the communication and radiometer experiments. Development efforts on key components for this approach could then be optimized and experiment installation and transformation would involve minimum hardware change. The degree of attainment of these features must be considered in the light of achieved experiment performance.

In order to utilize the local oscillators, power combiners, phase shifters and amplifiers used for the communications array, the electronic configuration of Figure III-19 would be used for the radar. The radiating elements consist of linear arrays of slots as depicted in the figure. Approximately one-fourth of the elements will be used for both transmit and receive and will, therefore, require the circulator. Of the remaining elements, about half will be receive only and the others will be transmit only. Several variations of the approach shown in Figure III-19 can be formulated, but the one shown is representative.

The 13.9-GHz mixers or downconverters are available in a variety of form factors depending on requirements. Without stringent bandwidth requirements, a device weight of 15 grams and conversion loss of 6 dB are available. The requirement for the high IF output frequency does degrade the available conversion loss somewhat, from the 5 to 5.5 dB values which are available for low IF operation. The mixers are followed by low-noise, 5.9-GHz IF amplifiers with 4-dB noise figures resulting in a receiver-element noise figure of about 10 dB.

On the transmit side of the element, the 5.9-GHz reference signal is phase shifted and upconverted to the 13.9-GHz transmitter frequency. With sufficient 8-GHz and 5.9-GHz power levels available, (about 50

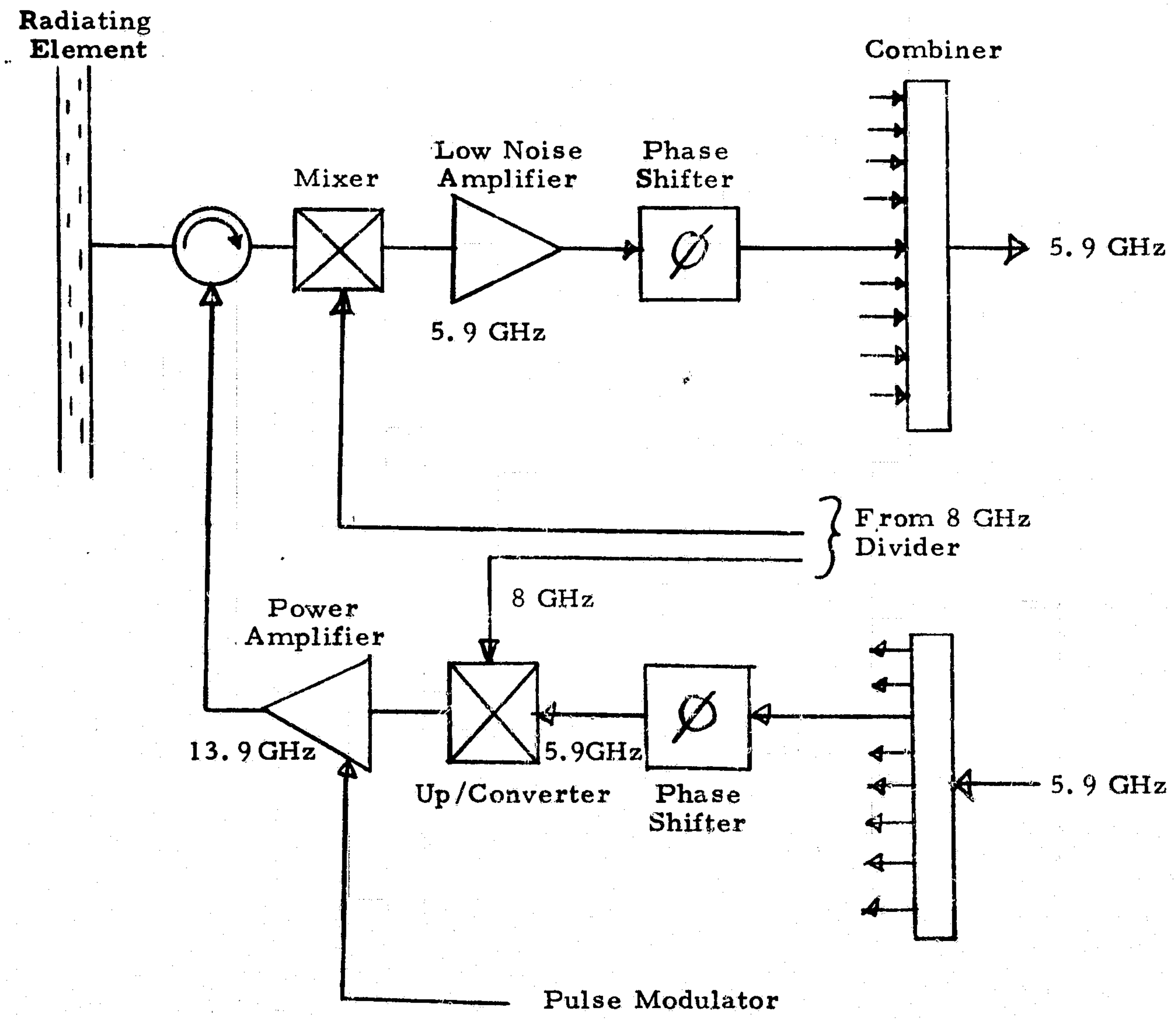


FIGURE III-19

13. 9 GHz Radar Electronics
Configuration I. 5. 9 GHz IF System

milliwatts) a miniature upconverter can deliver about 10 milliwatts at 13.9 GHz. Higher power levels are achieved by installing a power amplifier at the output of the upconverter.

Several solid-state devices can be considered for utilization in the pulsed power amplifier. The capability and features of Gunn (transferred electron or TEA) and IMPATT (IMpact Avalanche Transit Time) amplifiers for 13.9 GHz are summarized on Table III-V. The capabilities indicated have all been reliably demonstrated and are commercially available from a number of manufacturers. Output levels under pulsed operation can be adjusted by optimizing device design for a particular set of conditions, i.e., pulse width and pulse repetition frequency. Gains for all these devices range from 6 to 10 dB per amplifier stage. Therefore, multiple stage units will be required to go from the 10-milliwatt upconverter output to an output level of several watts. All of these devices are two terminal reflection types requiring circulators with a resulting weight per stage of 30 to 50 grams. Microwave-integrated circuit techniques can be used to reduce this weight to about 20 grams per stage with adequate development. Provision for adequate heat removal under high power and duty factor conditions will contribute additional weight.

The Configuration I element can be most easily improved in receiver performance by operating at a lower IF. Reducing the IF to the neighborhood of 100 MHz will result in a noise figure of near 7 dB at the input of the mixer, which is a considerable improvement. Use of the low IF requires manifolding of local oscillator power in the 13-GHz frequency range, which is somewhat more lossy. Also, phase shifters at the VHF and UHF frequencies, which would be required, are larger and heavier.

If the transmitter electronics is implemented using the approach of Figure III-19, but with an offset frequency below 1 GHz in the upconverter, the filtering problems become important. Weight and size of the upconverter will be increased with the necessary additional filtering. With the use of an intermediate frequency in the VHF range, phase shifting of the local oscillator signal may be also considered and the IF phase shifters eliminated.

TABLE III-V

Solid State Amplifier Capability, 13.9 GHz

DEVICE	CW POWER	PEAK POWER	MAX PULSE WIDTH @ MAX PEAK POWER	MAX DUTY	TYPICAL EFFICIENCY	BIAS VOLTAGE
Gunn	.5 Watt	10 Watt	1 μsec	1%	5%	9 V CW
		2 Watt	4 μsec	5%		30 V pulse
IMPATT						
Silicon (Double Drift)	2 Watt	15 Watt	5 μsec	10%	10 - 15%	130 V
		10 Watt	10 μsec	10%		
GaAs	5 Watt	8 Watt	100 μsec	50%	20 - 30%	45 V

The simplest approach to the design of the radar would involve only phase shifting at the radiating elements. All of the mixing, frequency conversion, amplification and duplexing is done after the divider/combiner as shown on Figure III-19. The development of low-loss 13.9-GHz phase shifters and the combiner/divider are required to insure adequate system performance. Three-bit diode phase shifters will offer less than 2 dB of loss, while the loss of the combiner is expected to be around 3 dB at this frequency. Ferrite phase shifters may be used to achieve 0.8-dB insertion loss, but at the expense of increased weight.

Because the high-powered transmitter is duplexed into the power divider, a receiver-protection device is required. The RPD can be a diode limiter or switching device which, typically, will add 0.5 dB of insertion loss. Assuming 0.4-dB loss for the duplexer circulator, a total insertion loss of 5.9 dB is encountered in front of the receiver low-noise amplifier. This figure does not include the insertion loss of the transmission lines leading to the radiating elements.

Some expense may be profitably incurred in the low-noise amplifier since only one such device will be required per system. Even a parametric amplifier, although quite expensive, offering a noise figure near 2 dB, may be considered for this application. Devices such as tunnel diode amplifiers (TDA) or possibly a GaAs field-effect transistor (FET) could be used to obtain a 5-dB receiver noise figure. Therefore, the net front-end noise figure would be about 8 dB using a parametric amplifier, 11 dB with a tunnel diode or FET amplifier, and above 13 dB with a mixer alone.

The single transmitter power source would most suitably be a traveling-wave tube. Space-qualified tubes at this frequency, according to available information, are limited to about 20 watts (Hughes 851 H, e.g.). Present tube technology is capable of generating many kilowatts of peak power. Approximately 6 dB of insertion loss must be included between the transmitter and radiating element to account for the circulator, power divider, and phase shifters.

A number of variations can be configured which will offer improvements in performance over the basic system of Figure III-20. These variations can be discussed by examination of Figure III-21. The poor

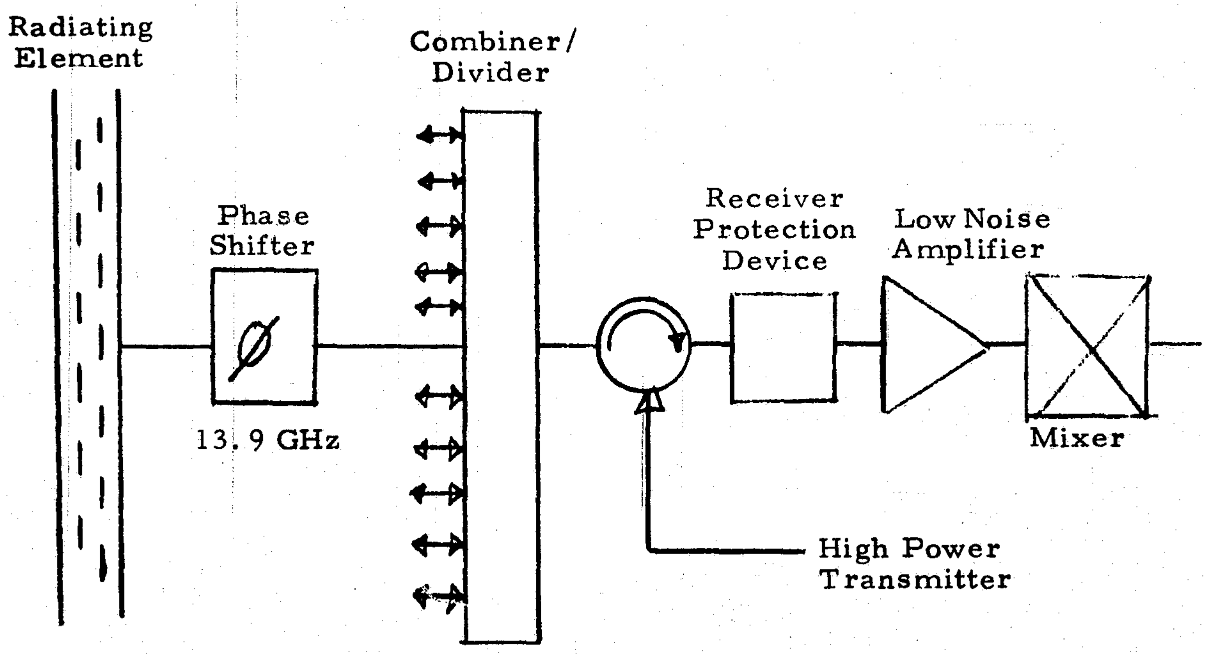


FIGURE III-20

13.9 GHz Radar Electronics

Configuration II. RF Phase Shifted/RF Combined

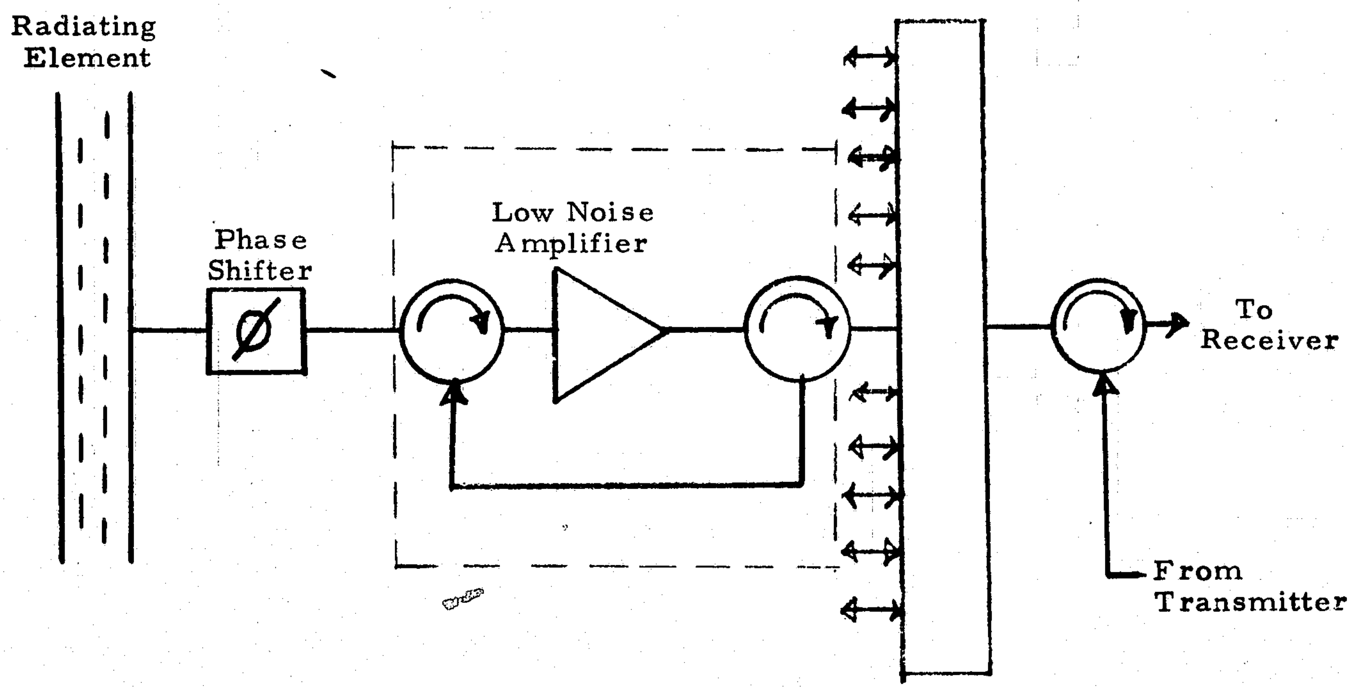


FIGURE III-21

13.9 GHz Radar Electronics

Configuration III. Distributed Receiver Amplifier System receiver performance is the result of placing the phase shifter and combiner losses in front of the receiver. Performance can, therefore, be significantly improved by placing a low-noise receiver, a TDA or FET, in each element as shown in Figure III-21. Duplexer circulators are now required in those elements which must transmit as well as receive. Assuming 2-dB loss in the phase shifter, 0.4-dB loss in the input circulator, and sufficient gain in the LNA to yield a 5-dB input noise figure, an element noise figure near 7.5 dB can be achieved.

Transmitter efficiency can be improved by distributing solid-state power amplifiers in the elements as well. This approach is indicated in Figure III-21(a), where the portion of the element within the dashed box on Figure III-21 is replaced as shown. The advantage of transmitter amplifier distribution is that the power-divider losses are incurred at a lower power level with the gain inserted in the elements. Also, a configuration may be structured to eliminate one of the circulators in the elements by using a separate transmitter divider replacing a portion of the element as shown in Figure III-21(b). This approach amounts primarily to a trade-off between the weights of circulators and power dividers.

The performance and weight of the various configurations can be compared on the basis of present component technology. An array configuration is assumed in which 300 receiver elements are distributed in four sections of 75, as shown in Figure III-22. 363 transmit elements are located at the center of the array, as indicated by the dashed region on the figure. It is seen, therefore, that about 75 of the elements are common to both arrays and the electronics contained behind these elements must be capable of both transmit and receive. The area containing the transmit/receive elements is cross-hatched on the figure.

With the use of linear-slotted arrays for each of the radiating elements, an element gain of 12 dB will be assumed. Transmitter ERP can be compared for the candidate configurations by examination of the sum:

Number of elements (363)	25.6 dB
Element gain	12.0 dB
Power/Element	<u>P dBw</u>
ERP (peak)	(37.6 + P) dBw

Therefore, used with the output of the upconverter alone (10 milliwatts), the ERP = 17.6 dBw. Inserting a 5-watt peak IMPATT amplifier in each

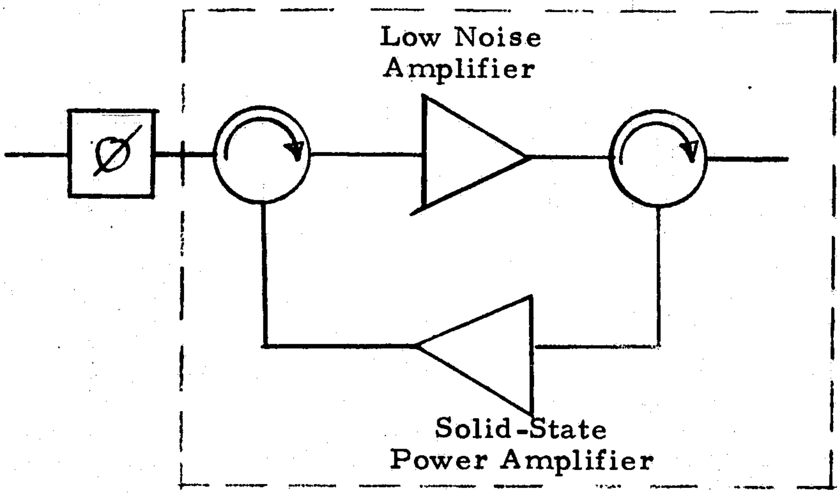


FIGURE III-21(a)

Details of Distributed Receiver Amplifier System

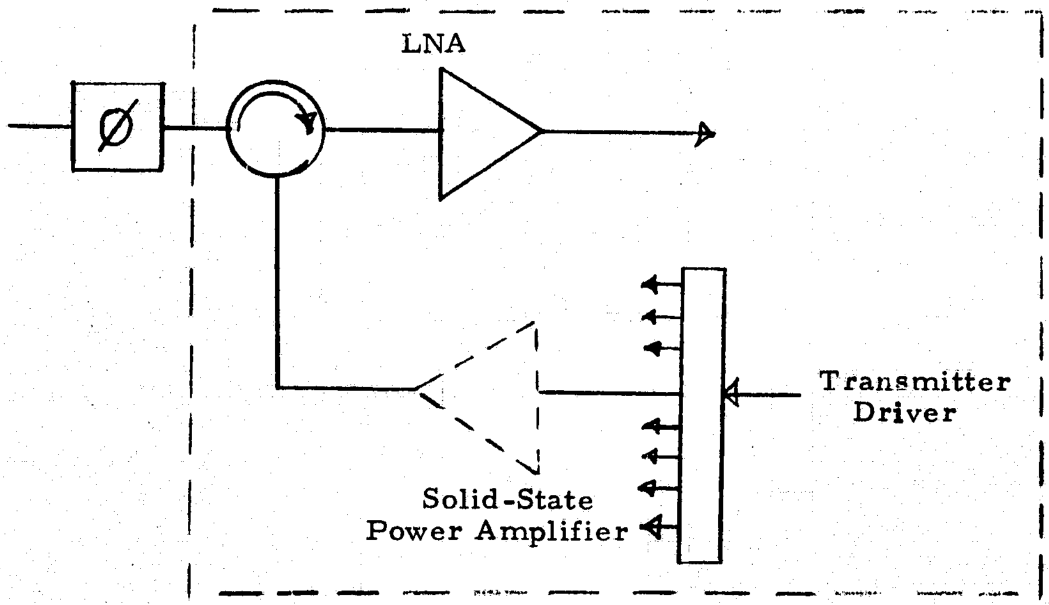


FIGURE III-21(b)

Details of Distributed Receiver Amplifier System

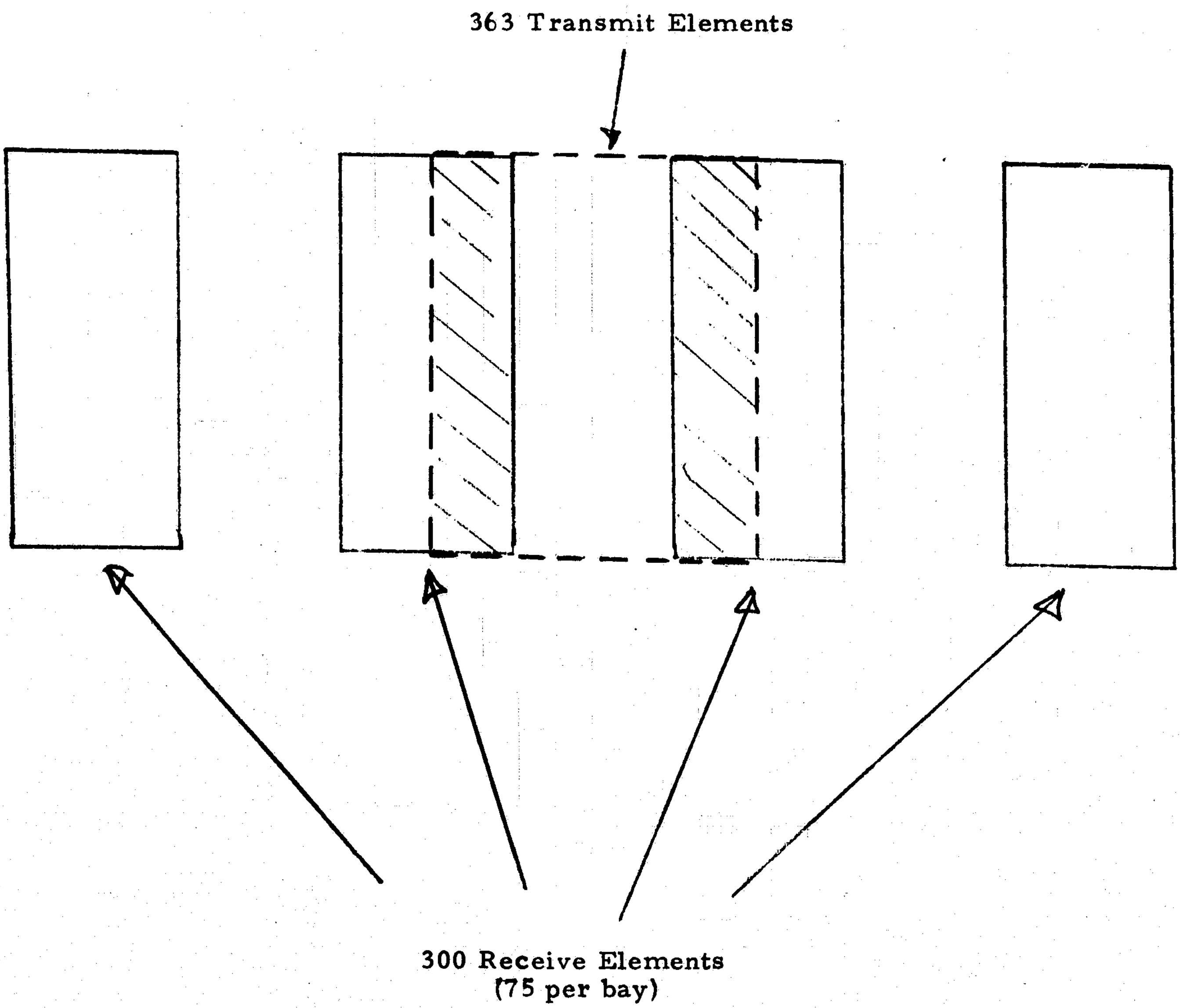


FIGURE III-22

Distribution of Transmit and Receive Elements element, an ERP = 44 dBw peak is achieved. These values must, of course, be multiplied by the duty factor in order to determine the actual ERP for radar system calculations.

Receiver performance is most easily compared at this time by examination of the single-channel noise figure at the input to the element. Other important factors which will eventually determine receiver sensitivity, e.g., processing bandwidth and antenna temperature, can be considered fixed.

IV. MECHANICAL DESIGN
COMMUNICATION ANTENNA

A. Introduction

The first mechanical design was concerned with the deployment of a ten-meter aperture. The philosophy included the use of astronauts in the deployment of the antenna structure.

The present designs include a six-meter antenna that occupies part of two pallets for stowage and the deployment is mechanically initiated. After deployment the pallets are vacant. Since only part of each pallet is used during stowage, there is room for another experiment to be stowed since the MLAE antenna hugs the periphery of the pallets.

B. Method of Design

The mechanical design of the primary array structure is the result of a relaxation process whereby a given beam arrangement is chosen, then the electronic equipment platform cross-section is chosen so that its stress is at the design maximum. This platform stress is a function of the edge-support beam cross-section properties so these calculations are made for several support beam sizes. The combined weights of the electronic support platform and the elastic support beams, and the beam bending stress are then calculated. The optimum design for this beam arrangement is the combination which results in minimum structure weight with the structural elements working near their design limits.

The number and placement of the structural beams may be altered and the above process repeated to optimize the total structural design.

C. Basis for Design

The basis for the designs have been taken from the following documents:

1) Spacelab Payload Accomodations Handbook, Intermediate Issue, Revision A, April 1974.
2) Space Shuttle System Payload Accomodations, Level II Program Definition and Requirements, Volume XIV, Revision C, July 1974.
3) Spacelab Payload Accomodation Handbook (Preliminary), October 1974.

D. Results of Design

One of the several designs is shown in Figure IV-1. The antenna when deployed lies along the rear vertical stabilizer (not shown in drawing). In the figure, the antenna is shown both in the stowed and operating positions. The corresponding weights for this design are given in Table IV-I. The detailed calculations are given in Appendix A.

TABLE IV-I

Weight Estimate Summary
6-Meter Diameter Communications Antenna

No.	Item	$W_{lb.}$	W_{kg}
1	Electronics Plate	207	93.9
2	Subaperture Beams	173	78.5
3	Edge Rim	80	36.4
4	Radiating Aperture	150	68.0
5	Electronics	3,310	1,500.0
6	Pallet Tie-downs (12 ea.)	80	36.3
7	Deployment Trunion Brackets	57	25.8
8	Deployment Motors	24	10.9
9	Rotary Joints, Cable Harness	108	49.0
10	Thermal Protection	36	16.3
11	Misc. Hdwe (10% of above, less electronics weight)	92	42.0
		4,317	1,957.0

It will be noted that the total weight allowed for the electronics is 1,500 Kg. The weight can be allocated to the particular system described in Section III, and the number of elements can then be computed.

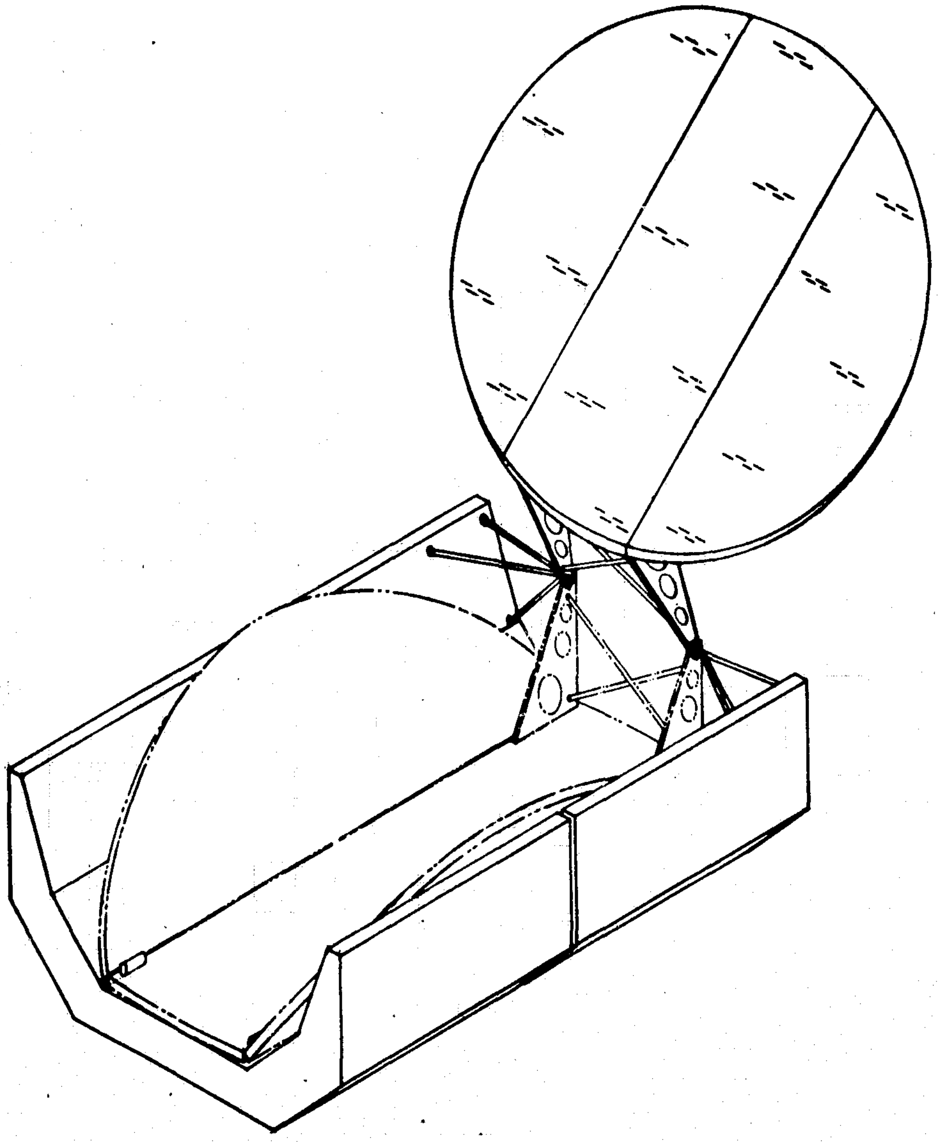


FIGURE IV-1

Six-Meter Antenna
(Stowed and Deployed)

V. NEW TECHNOLOGY

During the course of the program no new technology was generated.

VI. PLANS FOR NEXT REPORTING PERIOD

The mechanical designs for the radar antenna will be started. The radar equation will be used to obtain the appropriate transmitter power required for adequate S/N in the receiving aperture.

Various designs of radiometric antennas will be traded off to obtain suitable configurations.

C-2

VII. CONCLUSIONS AND RECOMMENDATIONS

Viable designs have been generated for the communications and radar antennas. Parametric studies have been concluded for several electronic systems for the communication and radar systems. Weight and power calculations indicate that these systems can be flown on the shuttle. The recommended antennas for the communications and radar systems are shown in Tables VII-I and VII-II, respectively. The weights for corresponding electronics are shown in Tables VII-III and VII-IV.

The designs that have been generated need be analyzed for costs. In order to establish costs some of the long lead items such as mixers in the millimeter wave region need be breadboarded and configured for low-cost and low-weight fabrication.

An initial estimate of cost and schedule for the components for the communications array is shown in Table VII-V.

The interface with the Spacelab itself and other experiments must be investigated so as to maximize the performance of every equipment for multiple functions.

Aperture Distribution: 30 dB Taylor
Method of Tapering: Thinned (46% left and Space Tapered)
Diameter: 6 Meters
No. of Elements: 3894
Phase Shifter Quantization: 3 bit
Sidelobe level = 35 dB RMS (estimate)
Frequencies: 20 and 30 GHz
Radiators: Interleaved for 20 and 30 GHz
Polarizations: To be specified
Element Gain: 12 dB at edge of scan ($\pm 15^{\circ}$)

TABLE VII-I

Communications Antenna Array

Aperture Distribution: Uniform
Length: 46.5 meters
Width: 5 meters
No. of Elements: Transmit 363, Receive 300
Phase Shifter Quantization: 3 bit
Frequency: 13.9 GHz
Polarization: To be selected

TABLE VII-II

Radar Antenna Array

	Weight (Electronics Only)	Power Consumption	G/T*	ERP (Per Channel)	Number of Elements +
Three Channel Transmit/Receive	1500 Kg	2060 Watts	8.8 dB/°K	14.4 dBw	1750
Three Channel Receive/ One Channel Transmit	1500 Kg	2510 Watts	12.0	17.6	3600
One Channel Transmit/Receive	1500 Kg	2260 Watts	13.1	18.7	4700

* Assumes antenna temperature of 290°K, Receiver Temperature of 2700°K (F = 11 dB) including losses. Total system weight, including electronics, is 2000 kilograms.

+ Each electronic element includes one complete transmit/receive unit with number of channels required.

TABLE VII-III

Summary of Communication Systems

30-GHz Receivers, 20-GHz Transmitters

	WEIGHT (ELECTRONICS ONLY)	POWER CONSUMPTION
ERP, 40dbW PEAK (20 dbW AVERAGE)		
11 db NOISE FIGURE *	100 Kg	1882 WATTS
6 db NOISE FIGURE *	110	1906
ERP, 30 dbW PEAK (10 dbW AVERAGE)		
11 db NOISE FIGURE *	79 Kg	248 WATTS
6 db NOISE FIGURE *	89	272

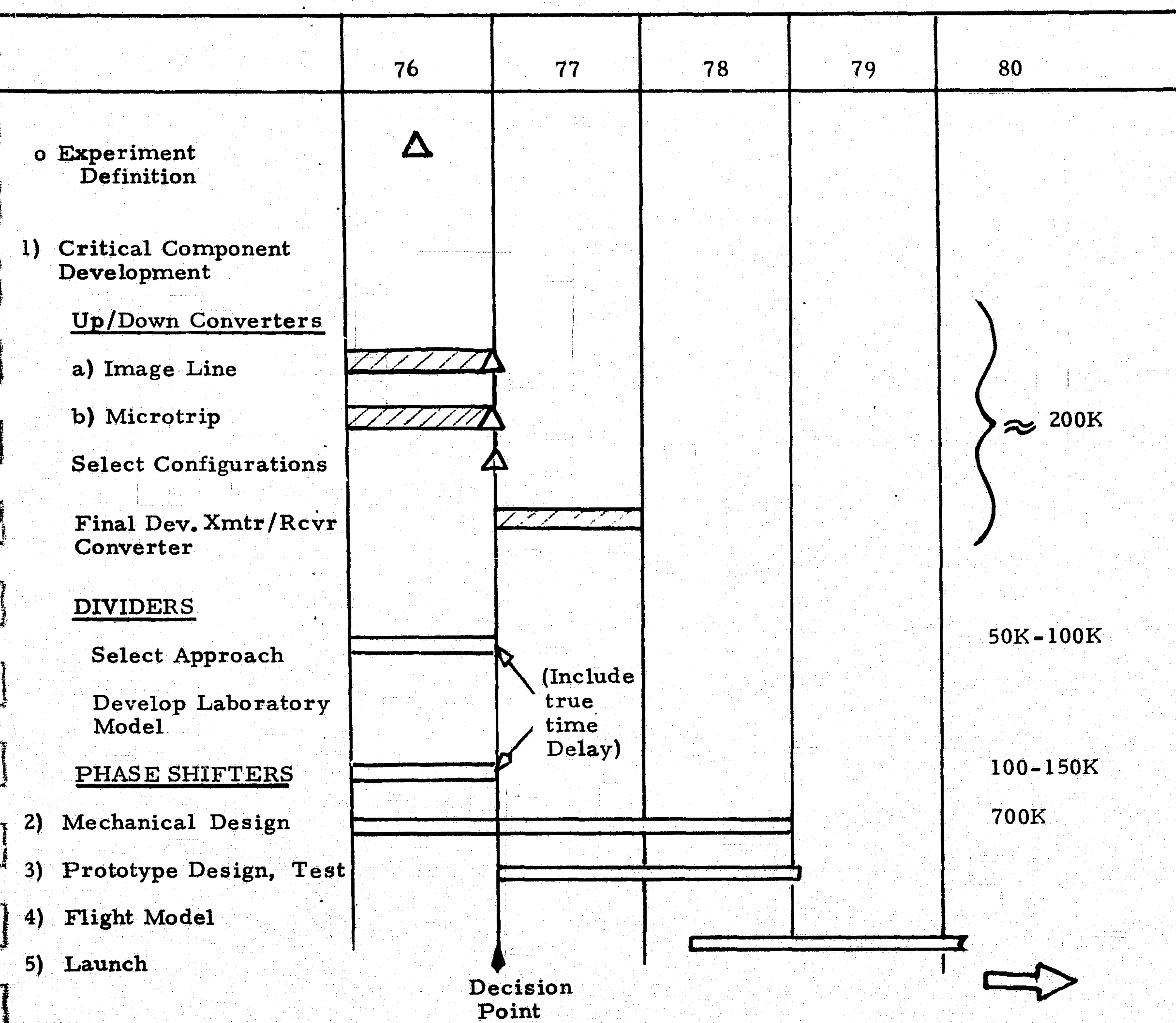
* RECEIVE ANTENNA GAIN = 37 db
CROSS TRACK SCAN ± 15°

TABLE VII-IV

Summary of 13.9-GHz Radar Systems

TABLE III-V

Communication Array
Development Schedule



REFERENCES

Chrepta, M.M. and H. Jacobs, "Millimeter Wave Integrated Circuits," 1974 IEEE Microwave Symposium Digest, p. 198.

Davis, R., "New Transmission Media Emerging for Millimeter Wavelengths," Microwaves, Sept. 1974.

Glance, B. and W.W. Snell, "Low-Noise, Integrated, Millimeter-Wave Receiver," BSTJ 53, Sept. 1974, p. 1321.

Hansen, R.C., "Tables of Taylor Distributions for Circular Aperture Antennas," IRE Transactions on Antennas & Propagation, Vol. AP-8, Jan. 1960, pp. 23-26.

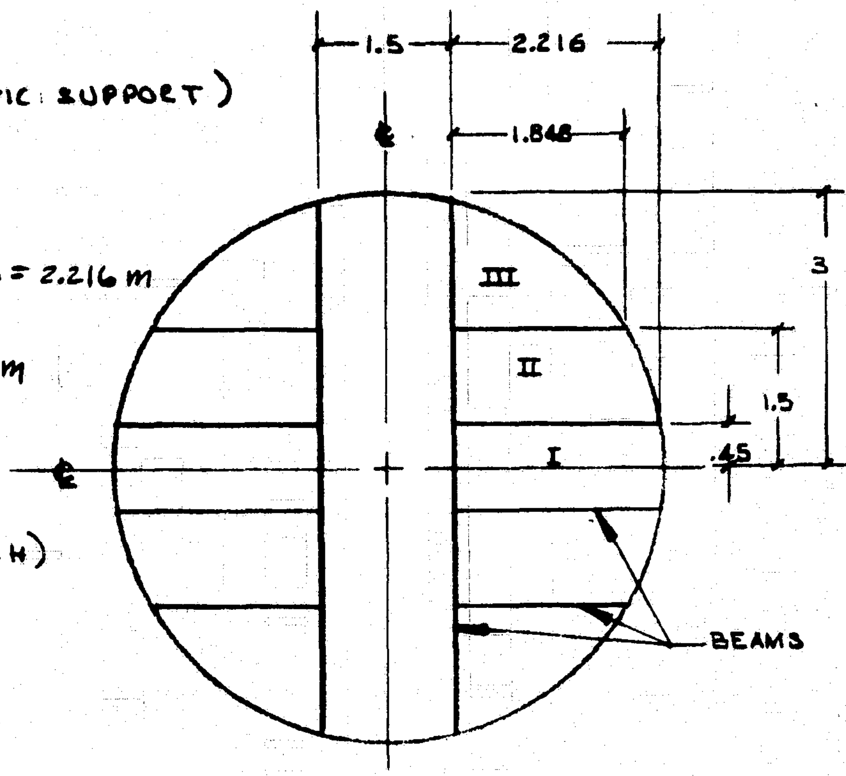
Schneider, M.V. and W.W. Snell, "A Scaled Hybrid Integrated Multiplier from 10 GHz to 30 GHz," BSTJ 50, July-August 1971, p. 1933.

APPENDIX

SIX-METER ANTENNA

Mechanical Design

(PLATE WITH ELASTIC SUPPORT)



PANEL I:

$a = \left[9-(.45)^2\right]^{\frac{1}{2}} - .75 = 2.216\ m$

$b = 2(.45) = 0.9\ m$

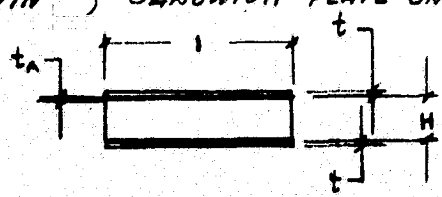
$\frac{b}{a} = 0.406$

PLATE: (SANDWICH)

$\sigma_p = M_p \frac{c}{I} Q_p$

$M_p = \beta_P\, q_D\, a^2$

$q_{PD} = \rho_p 2t + \rho_c H + 2\rho_A t_A$ $\#/IN^2$, SANDWICH PLATE ONLY.



$\rho_P = 0.1\ \#/IN^3$

$\rho_c = \frac{3.1}{1728}\ \#/IN^3$

$\rho_A = 0.06\ \#/FT^2$

LET $t = 0.03H$,

$q_{PD} = \left(.2(.03) + \frac{3.1}{1728}\right)H + \frac{.12}{144} = 0.007794(H+0.107)\ \#/IN^2$

β_P FROM REF 1, TABLE 3.

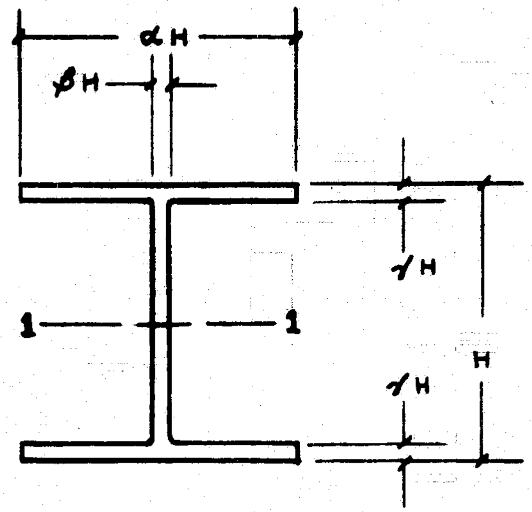
$c \approx \frac{H}{2}$, $I \approx \frac{tH^2}{2}$, $\frac{c}{I} = \frac{1}{tH}$ IN^{-3}/IN

SUBAPERTURE SUPPORT BEAM

$$A = [\beta + 2(\alpha - \beta)\gamma]\, H^2 \ \text{IN}^2$$

$$I_1 = [\beta + 2(\alpha-\beta)(3 - 6\gamma + 4\gamma^2)\gamma]\frac{H^4}{12} \ \text{IN}^4$$

$$\frac{c}{I_1} = \frac{6}{[\beta + 2(\alpha-\beta)(3 - 6\gamma + 4\gamma^2)\gamma]\, H^3} \ \text{IN}^{-3}$$



BEAM CROSS-SECTION

FOR UNIFORMLY LOADED BEAM OF LENGTH ℓ, SIMPLY SUPPORTED ENDS:

$$M_B = \frac{W\ell Q}{8} \ \text{IN \#}$$

$\sigma_B = M_B \frac{c}{I}$ APPLIED STRESS

$\sigma_A = \frac{F_{TY}}{1 + MS_y}$ ALLOWABLE STRESS

$$\sigma_B = \sigma_A: \quad H_{B\,MIN} = \left\{ \frac{3W\ell Q(1 + MS_y)}{4F_{TY}[\beta + 2(\alpha-\beta)(3 - 6\gamma + 4\gamma^2)\gamma]} \right\}^{\frac{1}{3}} \ \text{IN}$$

$$\left(\sigma_B = \frac{3W\ell Q}{4[\beta + 2(\alpha-\beta)(3 - 6\gamma + 4\gamma^2)\gamma]H^3} \ \text{psi} \right)$$

SUPPORT BEAMS:

$$\sigma_B = M_B \frac{c}{I} Q$$

$$M_B = \beta_B \, g_B \, a^3$$

β_B FROM REF 1 TABLE 6

$$g_B \approx \frac{W_{TOTAL}}{A_{TOTAL}} = \frac{2000(2.2046)}{\frac{\pi}{4}\left(\frac{6}{.0254}\right)^2} = 0.10 \ \#/IN^2 \equiv g_P$$

$$\frac{c}{I} = \frac{1}{.06 H_B^3} \quad IN^{-3}$$

$$Q = 3(12.2) = 36.6 \ g's$$

$$\lambda \equiv \frac{EI_B}{a D_P}, \quad I_B = 0.03 H_B^4, \quad D_P = \frac{E_P t H_P^2}{2(1-\nu^2)} = \frac{.03 H_P^3}{2(.91)} E_P$$

$$@ \ E_P = E_B, \quad I_B = a\lambda \frac{D_P}{E_B} = a\lambda(.0165) H_P^3$$

$$@ \ \sigma = \frac{F_{TY}}{1+MS_y}; \quad \sigma_P = \frac{35000}{1.15} \ psi, \quad \sigma_B = \frac{42{,}000}{1.15} \ psi$$

$$\sigma_P = \frac{\beta_P g_P a^2 Q}{.03 H_P^2}; \quad H_P = \left[\frac{1.15 \beta_P (.10) 36.6}{35000(.03)}\right]^{\frac{1}{2}} a = 2.4926 \, a_m \beta_P^{\frac{1}{2}} \ IN$$

$$H_B = \left(\frac{a\lambda D_P}{E_B \, .03}\right)^{\frac{1}{4}} = \left[\frac{a_m \lambda (.03) H_P^3}{.0254(.03) 1.82}\right]^{\frac{1}{4}} = 2.1566 (a_m \lambda H_P^3)^{\frac{1}{4}} \ IN$$

$$W_P = \frac{\pi}{4}\left(\frac{6}{.0254}\right)^2 g_{PO}$$

$$W_B = \rho A \Sigma \ell = .1(.1728 H_B^2) \frac{4}{.0254}\left[\Sigma a_m\right] = 2.7213\left[a_1 + a_2 + \frac{2(6)}{4}\right]$$

$$W_B = 19.223 H_B^2, \qquad W_S = W_P + W_B \ LBS$$

$$\sigma_B = 4.0508 \times 10^7 \frac{\beta_B}{H_B^3} \ psi$$

$H_P = 2.4926\, a_m \phi_P^{\frac{1}{3}}$ IN; $H_B = 2.1566\,(a_m \lambda H_P^3)^{\frac{1}{4}}$ IN; $\sigma_B = 4.0508 \times 10^7 \frac{\phi_B}{H_B^3}$ psi.
$W_P = 341.57\,(H_P + 0.107)$ LBS; $W_B = 19.223\, H_B^2$ LBS,

TABLE 1. SUBAPERTURE STRUCTURE, BASED ON PANEL I DIMENSIONS

λ	ϕ_P	ϕ_B	H_P, IN	H_B, IN	W_P, LB	W_B, LB	W_S, LB	σ_B, PSI
0.5	0.03733	0.01821	1.0672	2.3232	401	104	505	58,830
1	0.02448	0.02078	0.8642	2.3584	332	107	439	64,170
2	0.01652	0.02245	0.7099	2.4201	279	113	392	64,156
4	0.01203	0.02339	0.6058	2.5553	243	126	369	56,786
6	0.01046	0.02372	0.5649	2.6834	230	138	368	49,726
→ 10	0.00916	0.02399	0.5287	2.9009	217	162	379	39,806
30	0.00783	0.02427	0.4888	3.5997	203	249	452	21,076
100	0.00736	0.02437	0.4739	4.7523	198	434	632	9,197
13.5	0.00892	0.02403	0.500	3.000	207	173	380	36,052

TABLE 2. SUBAPERTURE STRUCTURE, BASED ON PANEL II DIMENSIONS

λ	ϕ_P	ϕ_B	H_P, IN	H_B, IN	W_P, LB	W_B, LB	W_S, LB	σ_B, PSI
0.5	(SAME AS TABLE 2)	(SAME AS TABLE 2)	0.9786	2.1304	371	87	458	58,823
1			0.7925	2.1627	307	90	397	64,160
2			0.6510	2.2192	259	95	354	64,153
4			0.5555	2.3432	226	106	332	56,783
6			0.51802	2.4607	213	116	329	49,723
10			0.4848	2.6601	202	136	338	39,804
30			0.4482	3.3009	190	209	399	21,076
100			0.4345	4.3578	185	365	550	9,197

WEIGHT ESTIMATE

1. ACTIVE ELECTRONICS, W/HDWE

$W_1 = 4412\,(.7516) = 3310\text{ LBS}\,(1500\text{ kg})$

2. RADIATING APERTURE (20 gHz W/G I.D. = .17 × .42)

LET: $H_p = .3\,\text{IN},\ t_p = .01\text{ IN}$

$$g' = 2(.01)2 + \frac{3.1}{1728}(.3) + \frac{.12}{144} = 0.0033715\ \#/\text{IN}^2$$

$$W_2 = A g' = \frac{\pi}{4}\left(\frac{6}{.0254}\right)^2 .0033715 = 148\text{ LBS}\,(67.0\text{ kg})$$

3. ELECTRONICS SUPPORT PLATE

W_3, FROM PP 3, = 207 LB (93.9 kg)

4. SUPPORT BEAMS

W_4, FROM PP 3, = 173 LB (78.5 kg)

5. EDGE RING ($H = H_B$, $\frac{H_B}{2}$ WIDE CHANNEL, $H_B = 3$ IN)

$$W_5 = \pi\left(\frac{6}{.0254}\right).1\left[2(3)(.06)3\right] = 80\text{ LB}\,(36.4\text{ kg})$$

6. PALLET LATCH/TIE DOWNS (12 @ BASE, 4 @ TOP EDGE)

MOTOR: 2" DIA × 3" LG, (1 LB) 40 IN OZ

GEAR DRIVE, SAME (1 LB)

LATCH: $A_{MIN} = \frac{P}{\sigma} \Rightarrow \frac{WQ}{N\sigma} = \frac{4400(36.6)1.15}{16(42100)} = 0.2756\text{ IN}^2$

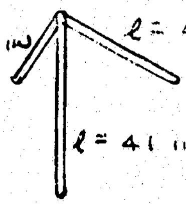
SAY 1 IN^2 ea; $L \Rightarrow 10$ IN ea, $w \to (1\#\text{ ea})$, M

MOTOR MTG, 1 #, Σ = 4 #, SAY 5 # ea

$W_6 = 16(5) \simeq 80\text{ LBS}\,(36.3\text{ kg})$

7. DEPLOYMENT TRUNION

PALLET BRKTS: (AL TUBES)



$P' = \frac{\pi^2 E I}{l^2} \geq A \sigma_A Q$

LET $P = W$ (1 g max accel)
$= 4400$ LBS

$t = .1D, \quad A = \pi D t = \pi .1 D^2$

$I = \frac{\pi D^3 t}{8} = \frac{\pi}{80} D^4$

$\sigma_A = \frac{42000}{1.15}$; $W = \rho A l = .1(.1)\pi D^2 l$ LB

$4400 = \frac{\pi^2 \times 10^7 \pi D^4}{80 l^2}$; $D' = (0.001135 l^2)^{\frac{1}{4}}$ IN

l	D'	D	W
41	1.18	0.62	1.8
46	1.25	0.62	2.3

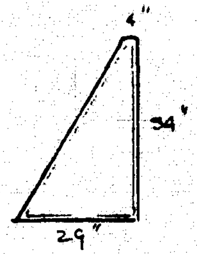
$W \approx 2[2(1.8) + 2.3]\ 150\% = 18$ LBS (8.0 kg)

ARRAY BRACKETS

$A_W \approx 29(27) = 783$ IN^2

$A_F = \overset{1''}{B}(54 + 29 + 61 + 4) = 148$ IN^2

$W = t \rho A = .1(.1)(783 + 148) = 9.3$ # ea



BRACES $(\ell = 80 \text{ IN})$

$P' = 4400$ LBS $\quad D' = 1.65$ IN

$W = \pi\, 1.65\,(.165)\,.1\,(80)(150\%) = 10.3$ # EA

$W_P = 2(9.3 + 10.3) = 39.2$ LB (17.8 kg)

$W_7 = 18 + 39 = 57$ LBS (25.8 kg)

8. DEPLOYMENT MOTORS (2 @ TRUNIONS, 2 UNFURL) (EST)

$W_8 \approx [2(5^{\#}) + 2(1^{\#})]\,200\% = 24$ LBS (10.9 kg)

9. ROTARY JOINTS, CABLES (6 M × N = 4400 WIRES)

$W_9 \approx 5^{\#} + (4400)\frac{6}{.0254}(1 \times 10^{-4}\ \#/\text{IN}) = 108$ LB (49.0 kg)

10. THERMAL PROTECTION (25 LAYER SUPERINSULATION BACK SURFACE, PAINTED APERTURE)

$W_{10} \approx \frac{\pi}{4}\left(\frac{6}{.0254}\right)^2 \{.05(.004) + .05[25(.00025)]2\} = 36$ LBS