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ANTENNA PATTERN SHAPING, SENSING, AND STEERING STUDY

by

D. W. Power

SYLVANIA

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FINAL REPORT

ANTENNA PATTERN SHAPING, SENSING, AND STEERING STUDY

by

D. W. Power

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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

April 15, 1970

NASA Lewis Research Center
Cleveland, Ohio

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FOREWORD

The report covers a nine-month conceptual design study to determine the feasibility of multibeam antenna systems for RF transmission of high power from geostationary satellites. The study was performed under NASA contract NAS3-11524 with Mr. Perry W. Kuhns of the Special Projects Office, Spacecraft Technology Division, NASA Lewis Research Center, as Project Manager. The work was done in the laboratories of Sylvania Electronic Systems located in Needham and Waltham, Massachusetts where the principal technical contributors, in addition to D. W. Power included Dr. S. C. Bigelow, J. D. Killian, P. E. Smith Jr., and G. D. Whiteside.

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TABLE OF CONTENTS

<u>Section</u>		<u>Page</u>
	Foreward	iii
	Abstract	xvii
	Summary	1
I	INTRODUCTION	3
II	COMMUNICATION SYSTEM DESIGN	7
	2.1 System Description	7
	2.2 System Components	7
	2.2.1 Antenna	7
	2.2.2 Principal Rotary Joint	26
	2.2.3 Limited Motion Rotary Joints	35
	2.2.4 Diplexer	35
	2.2.5 RF Switches	37
	2.2.6 Polarizer	41
	2.2.7 Transmission Lines	44
	2.3 Alternate Components	44
	2.3.1 Apertures	44
	2.3.1.1 Dual-Beam Horn-Reflector	44
	2.3.1.2 Focal Point Fed Parabola	53
	2.3.1.3 Planar Array	58
	2.3.2 Principal Rotary Joint - Alternate	60
	2.3.3 Limited Motion Rotary Joints - Alternate	64
	2.3.4 Diplexers - Alternate	64
	2.3.5 RF Switches - Alternate	68
	2.3.6 Polarizer - Alternate	68
	2.4 Power Handling	70
	2.4.1 Multipaction	70
	2.4.1.1 Description	70
	2.4.1.2 Component Power Handling	75
	2.4.2 Ionization Breakdown	81
	2.4.2.1 Description	81
	2.4.2.2 Component Power Handling	83

TABLE OF CONTENTS (Cont.)

<u>Section</u>		<u>Page</u>
	2.4.3 Average Power Limitations	83
	2.5 Gain - Loss Tabulation	83
III	POINTING SYSTEM DESIGN	87
	3.1 System Analysis	87
	3.1.1 Terminology	88
	3.1.2 Gimbal Angle Equations	88
	3.1.3 Offset Equations	92
	3.1.4 Interferometer Equations	92
	3.1.5 Polarization Angle Detector Equations	95
	3.1.6 Control Law and Simplified Servo Equations	95
	3.1.7 Servo Motor Equations	96
	3.1.8 Structure Equations	102
	3.1.9 Satellite Motion Equations - Interferometers	104
	3.1.10 Satellite Motion Equations - Polarization Detector	104
	3.1.11 System Summary	106
	3.2 Equipment Description and Theory of Operation	106
	3.2.1 Sensing Equipment - Prime Mode	106
	3.2.2 Sensing Equipment - Redundant Mode	111
	3.2.3 Steering Equipment	113
	3.3 Equipment Design Parameters	115
	3.3.1 Interferometer Sensors	115
	3.3.2 Polarization Angle Sensors	118
	3.3.3 Servo Control Equipment	119
	3.3.4 Beacon Terminal Equipment	123
	3.3.4.1 Link Computations	123
	3.3.4.2 Beacon Terminal RF Equipment	124
	3.3.4.3 Beacon Terminal Polarization Angle Measurement Equipment	129
	3.3.5 Ground Station Computation	129
	3.4 Alternate Components and Techniques	130
	3.4.1 EW and NS Sensing Techniques	130
	3.4.2 Rotation Sensing Techniques	131

TABLE OF CONTENTS (Cont.)

<u>Section</u>		<u>Page</u>
3.4.3	Steering System Alternates	131
3.5	Error Analysis	133
3.5.1	Atmospheric Propagation Errors	133
3.5.1.1	Interferometer Sensing Errors	133
3.5.1.2	Polarization Sensing Errors	138
3.5.1.3	Redundant Mode Errors	139
3.5.2	Receiver/Detector Errors	139
3.5.2.1	Interferometer Sensing Errors - Prime Mode	139
3.5.2.2	Polarization Angle Sensing - Prime Mode	140
3.5.2.3	Interferometer Sensing Errors - Redundant Mode	140
3.5.2.4	Polarization Angle Sensing - Redundant Mode	141
3.5.3	Servo Errors	141
3.5.4	Thermal Distortion Errors	142
3.5.5	Pointing Error Summary	142
IV	MECHANICAL DESIGN	145
4.1	Hardware Design and Analysis	145
4.1.1	Communication System Components	145
4.1.1.1	Cornucopias	148
4.1.1.2	Limited Motion Rotary Joint	148
4.1.1.3	Diplexer	148
4.1.1.4	RF Switch	149
4.1.1.5	360-Degree Rotary Joint	149
4.1.1.6	Support Structure	152
4.1.2	Pointing System Components	153
4.1.2.1	Interferometer Horns	153
4.1.2.2	Sensing and Steering Electronics	153
4.1.2.3	Slip Rings	154
4.1.2.4	Limited Motion Drive System	154
4.1.2.5	Limited Motion Bearing System	155
4.1.2.6	360-Degree Drive System	155
4.1.2.7	360-Degree Bearing System	155

TABLE OF CONTENTS (Cont.)

<u>Section</u>		<u>Page</u>
4.1.2.8	Support Structure	155
4.1.3	Weight	156
4.2	Structural Analysis	156
4.2.1	Design Criteria	156
4.2.2	Governing Loading Conditions	158
4.2.3	Stress Analysis	159
4.2.4	Vibration Analysis	159
4.2.5	Structural Model for Control Systems Analysis	161
4.3	Thermal Analysis	162
4.3.1	Bulk Temperature Analysis	163
4.3.2	Pointing Error Analysis	169
4.3.2.1	Symmetric Sun Angle Thermal Errors	170
4.3.2.2	Non-Symmetric Sun Angle Thermal Errors	172
4.3.3	Thermal Pointing Error Summary	173
V	DESIGN VERIFICATION RECOMMENDATIONS	175
5.1	Recommended RF Tests	175
5.1.1	Horn-Reflector Evaluation	175
5.1.2	Horn-Reflector Thermal Sensitivity Tests	175
5.1.3	Rotary Joint Development	177
5.1.4	RF Power Handling Tests	178
5.2	Pointing System Verification	178
5.2.1	Sensing System Verification	178
5.2.1.1	Prime Mode Interferometer	180
5.2.1.2	Prime Mode Polarization Detector	180
5.2.1.3	Redundant Mode Interferometer	181
5.2.2	Steering System Analytical Verification	181
5.3	Mechanical Verification	181
5.3.1	Structural Model Tests	181
5.3.2	Thermal Vacuum Test	182

TABLE OF CONTENTS (Cont.)

<u>Section</u>		<u>Page</u>
VI	DESIGN VARIATIONS	185
6.1	Channel Separation vs Channel Isolation	185
6.2	Increased RF Power with Four Channels	185
6.3	Increased RF Power with Two Channels	186
6.4	Improved Satellite Attitude Control	187
6.5	Earth Oriented Satellite with Improved Attitude Control	188
VII	PERFORMANCE SUMMARY AND CONCLUSIONS	191
7.1	Performance Summary	191
7.2	Significant Design Features	196
7.3	Conclusions	197
VIII	REFERENCES	199

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LIST OF ILLUSTRATIONS

<u>Figure</u>		<u>Page</u>
1	RF System Block Diagram	8
2	Conical Horn-Reflector Antenna Defining Relations	9
3	Experimental Multi-Mode Horn-Reflector	11
4	Experimental Multi-Mode Horn-Reflector	12
5	Pattern Plane and Definitions for the Horn-Reflector	13
6	Single Mode Horn-Reflector Radiation Pattern	14
7	Single Mode Horn-Reflector Radiation Pattern	15
8	Single Mode Horn-Reflector Radiation Pattern	16
9	Multi-Mode Horn-Reflector Radiation Pattern	18
10	Multi-Mode Horn-Reflector Radiation Pattern	19
11	Multi-Mode Horn-Reflector Radiation Pattern	20
12	Multi-Mode Horn-Reflector Radiation Pattern	21
13	Multi-Mode Horn-Reflector Circularly Polarized Radiation Pattern	23
14	Multi-Mode Horn-Reflector Circularly Polarized Radiation Pattern	24
15	Multi-Mode Horn-Reflector 360° Radiation Pattern	25
16	Experimental Horn-Reflector Sidelobe Level vs Frequency	27
17	Summary Characteristics of the System Multi-Mode Horn-Reflector	28
18	Four-Channel 360°/Day Rotary Joint	29
19	Four-Channel 360°/Day Rotary Joint and Feed Network	30
20	Four-Channel 360°/Day Rotary Joint Conductor Dimensions	32
21	Rotary Joint Channel "Cutoff" Properties	33
22	Diplexer Configuration	39
23	Rotary Cylinder RF Transfer Switch Configuration	40
24	RF Switch Comparison and Performance	42
25	Circular Polarizer and Input Transition Configuration	43
26	Circular Polarizer Performance	45
27	Pyramidal Horn-Reflector Description and Field Notation	47
28	Horn-Reflector Beam Displacement and Beamwidth vs Flare Angle	49
29	Dual-Beam Horn-Reflector Dimensional Configuration	50
30	Dual-Beam Horn-Reflector Radiation Pattern - Equatorial Plane	51
31	Dual-Beam Horn-Reflector Radiation Pattern - Longitudinal Plane	52

LIST OF ILLUSTRATIONS (Cont.)

<u>Figure</u>		<u>Page</u>
32	Focal Point Fed Parabola Blockage Condition	55
33	Planar Array Antenna Configuration	59
34	RF System Block Diagram - Alternate	61
35	Two-Channel 360°/Day Rotary Joint	62
36	Limited Motion Rotary Joint - Alternate	65
37	Diplexer Alternate	66
38	Diplexer-Isolation vs Filter VSWR	67
39	Circular Polarizer Performance - Alternate	69
40	Existence Regions for Multipactor Effect	71
41	Multipactor Existence Regions	72
42	Parallel Plate Region for Multipactor Breakdown	72
43	Extension of Existence Regions for Multipactor Effect	76
44	Component Power Capacity - Multipactor Limited	77
45	Contours of Constant Voltage Breakdown	82
46	Geometrical Relationships	89
47	Partial System Block Diagram No. 1	93
48	Interferometer and Polarization Sensing Geometries	94
49	Simplified Step Motor Cross-Section	97
50	Step Motor Air Gap Variables	98
51	Partial System Block Diagram No. 2 (Stepper Motor and Structure)	99
52	Partial System Block Diagram No. 3 (Satellite Disturbances)	107
53	Overall System Block Diagram	108
54	Block Diagram of Primary Sensing System	109
55	Block Diagram of Redundant Sensing System	112
56	Detailed Block Diagram - One Axis	114
57	Interferometer Antenna Boresight Squinting	117
58	Response to a Ramp Input	121
59	Beacon Terminal Equipment	128
60	Geometry of Pointing Angle Errors	134
61	Pointing Error Due to Gimbal Angle Errors	135
62	Standard Derivation of Angle of Arrival vs Antenna Separation	137

LIST OF ILLUSTRATIONS (Cont.)

<u>Figure</u>		<u>Page</u>
63	Antenna Assembly	146
64	Antenna Assembly Pointing System Details	147
65	Four-Channel Rotary Joint Assembly	150
66	Four-Channel Rotary Joint Assembly (Power Divider Not Shown)	151
67	Critical Structural Areas	160
68	Bulk Temperature Model	164
69	Bulk Temperature Profile (Components Affecting Pointing Error)	167
70	Bulk Temperature Profile (Thermally Sensitive Components)	168
71	Folded Horn-Reflector Antenna	176
72	Alternate Antenna Assembly (Single Axis Steering)	189

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LIST OF TABLES

<u>Table</u>		<u>Page</u>
1	RF System Design Requirements	4
2	Pointing System Design Requirements	5
3	Mechanical Design Requirements	6
4	Horn-Reflector Performance Characteristics	10
5	Computed Four-Channel Rotary Joint Loss	34
6	Rotary Joint Dissipation	36
7	Diplexer Design Requirements and Computed Performance	38
8	Antenna Efficiency Comparison	57
9	Two-Channel Rotary Joint Loss and Dissipation	63
10	Conceptual Design Gain/Loss Tabulation	84
11	Servo Equipment Parametric Design	122
12	Interferometer Downlink Summary - Redundant Mode	125
13	Interferometer Uplink Summary - Prime Mode	126
14	Beacon Terminal Characteristics	127
15	Alternate Gimbal Arrangements	132
16	Pointing Error Summary	143
17	Pointing Error Summary	144
18	Antenna Weight Estimate	157
19	Resonant Frequency	161
20	Thermal Design Components	163
21	Heat Sources	165
22	Sun Angle Errors	171
23	Thermal Pointing Error Summary	173
24	Recommended RF Component Power Tests	179
25	RF System Design Requirements	192
26	Pointing System Design Requirements	193
27	Mechanical Design Requirements	194

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ABSTRACT

A conceptual design is presented for a steerable satellite antenna system having four 1-kilowatt RF beams of 2.7 degree half-power beamwidth at a frequency of 12.2 GHz. The antenna pattern sidelobes are more than 25 db down and the losses within the system are less than 1 db. The antenna includes a beam pattern direction sensing system and a steering system to point the pattern in the desired direction within 0.1 degree. The design demonstrates the feasibility of steerable multibeam antenna systems for high-power narrow-beam satellite communication.

SUMMARY

This report presents a detailed conceptual design of a mechanically steerable high-power narrow-beam satellite communication antenna. The antenna comprises two subsystems. The communication system switches, combines, and radiates from cornucopia apertures, four 1-kW RF beams of 2.7 degree half-power beamwidths at 12.2 GHz. The pointing system senses the direction of the communication beam pattern and steers it to within 0.1 degree of the desired direction.

Significant technical investigations are reported relating to sidelobe levels and high power breakdown in the communication system. Specified sidelobe levels of -25 db are achieved by a multi-mode horn-reflector configuration. These are verified by scale model tests. Multipaction, ionization breakdown and average power heating are examined. It is concluded that system peak power is limited to 15 kW per channel by multipaction and that system total average power is limited to 6 kW by heating considerations.

Other investigations reported examine sensing techniques and pointing error. Interferometric sensors are selected to measure cross pointing errors. An original polarization angle sensing technique is chosen to measure rotation about the boresight axis. Pointing error analysis treats errors associated with the pointing equipments as well as those attributable to atmospheric anomalies and thermal distortions. Thermal distortions are found to be the largest contributor to pointing error.

The required electrical characteristics are designed into a suitable equipment configuration that is in accord with the design constraints. Appropriate structural analysis sizes critical member sections and establishes overall system weight. Thermal analysis defines limiting bulk temperatures and thermal distortions.

It is concluded that high-power narrow-beam satellite antennas are feasible. The high power RF components and the pointing system components and techniques developed are readily adaptable to other frequencies and configurations.

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SECTION I

INTRODUCTION

The NASA is currently performing a number of studies to aid in evaluation of the use of satellites for high-power narrow-beam radio frequency transmission to earth. In order to optimize the overall system design, detailed subsystem information is required.

The objective of the program reported here is to determine the feasibility of the antenna and to provide detailed information about it. Here the antenna is defined to include that part of the RF transmission system that accepts RF energy from the final power amplifiers and radiates it in a desired pattern, and an antenna pointing system that senses directional error and steers the antenna beam pattern. The program is carried out by means of a detailed conceptual design of a mechanically steerable antenna for transmitting from a geostationary satellite through four 1-kW RF beams of 2.7 degrees half-power beamwidths at 12.2 GHz.

The scope of work embraces conceptual design and supporting analysis of the satellite antenna system and ground equipment detail necessary to system analysis. Also, recommendations are developed for verification of the design. A final inclusion examines design implications of several specified variations in the design parameters, thus providing a broader information base for overall system tradeoffs.

The various design requirements separate into three groups according to whether they impact RF communication system design, pointing (sensing and steering) system design or mechanical design. Table 1 lists those that bear on RF communication system design. As indicated there, alternate configurations are specified for the arrangement of antenna pattern traces at the earth surface. These differ only in the relation of one beam pattern to the other three. In the first configuration, the four patterns are uniformly spaced along an east-west line with centers separated by 0.55 half-power beamwidths. In the second configuration one of the end patterns is relocated directly below the center of the other three, from which it is separated by 0.72 half-power beamwidth. It is clear that both configurations can be realized with five mutually fixed apertures, with alternate apertures for the fourth beam. This is the design approach selected in this program. The requirements for multichannel beams necessitates combining the four channels into two groups of two channels each, and accommodating them in two beams from two of the apertures while the other apertures are without power. The remaining requirements of Table 1 should be self-explanatory. The RF communication system includes as components the transmitting apertures; devices for beam polarizing, switching, and combining; and devices to conduct RF power across interfaces.

Table 2 lists the requirements that relate to the pointing system design. These determine the steering ranges, velocities and accuracy expected of the overall system when subjected to the stated conditions. The pattern direction sensing system is specified to include means for operation in either of two modes. In the prime mode, sensors on the spacecraft are used in conjunction with remote beacons. In the other mode, beacon signals from the satellite are sensed remotely and the error information is telemetered back to the satellite.

TABLE 1. RF SYSTEM DESIGN REQUIREMENTS

<u>PARAMETER</u>	<u>SPECIFIED VALUE</u>
Center Frequency	12.2 GHz
Channel Width	50 MHz
Channel Separation	100 MHz
Number of Channels	4
Power Per Channel	1 kW (CW)
Total Power	2 kW
Operation	Continuous
Polarization	Circular
Beam Width	2.7° HPBW
Beam Separation	0.55 and 0.72 HPBW
Beam Configurations	2
Attenuation	1 db
Phase Delay	1 Degree
Polarization Axial Ratio	0.42 db
Channels Per Beam (Multichannel Mode)	2
Pattern Configuration Change Time	10 Minutes
Near Sidelobes	25 db
Front-To-Back Ratio	40 db
Channel Isolation	40 db
Channel Mixing Products	40 db

TABLE 2. POINTING SYSTEM DESIGN REQUIREMENTS

<u>PARAMETER</u>	<u>SPECIFIED VALUE</u>
Pointing Accuracy	± 0.1 Degree
Spacecraft Angular Error	± 5 Degrees (3 Axis)
Spacecraft Angular Rate	1 Degree/Minute
Spacecraft Station Error	± 0.5 Degree (Lat & Long)
Spacecraft Station Rate	0.1 Degree/Day
Pattern Steering Range (Latitude)	45° N To 10° S
Pattern Steering Range (Longitude)	20° E To 20° W
Pattern Steering Time	15 Minutes Max.
Prime Direction Sensing System	Spacecraft Sensing
Complimentary Sensing System	Earth Sensing
Sensing System Power Level	ERP of 50 dbW (Max.)
Ground Station Location	S/C View > 25° Above Horizon

Table 3 lists the requirements concerning the mechanical design. These define the environment constraining structural and thermal design as well as general physical constraints. The antenna is attached physically to the end of the satellite. It is desired that it be contained in a truncated conical shroud 110 inches in diameter at the base, 37 inches in diameter at the top, and 116 inches high, with a total volume of 220 cubic feet. In orbit the satellite is oriented with its longitudinal axis normal to the equatorial plane. The antenna must therefore be side-looking and capable of continuous rotation about the longitudinal axis to compensate for the solar orientation of the satellite. Steering must be provided about all three mutually orthogonal axes, but only limited motion is required about the other two.

The following sections detail a conceptual design that meets all of the specified requirements. Alternate techniques are also discussed where considered appropriate.

TABLE 3. MECHANICAL DESIGN REQUIREMENTS

<u>PARAMETER</u>	<u>SPECIFIED VALUE</u>
Spacecraft Attitude	Long. Axis Normal To Equatorial Plane Solar Oriented
Antenna Deployment	36-Inch (91 cm) Dia. Mount on Long End
Antenna Orientation	Side-looking
Launch Volume	370 Cubic Feet (10.478 m ³)
Launch Weight	250 Pounds (113.4 Kgs)
Launch Environment	
Shock	30G at 8 MS
Vibration	1 G _{rms} at 10-60 Hz 2 G _{rms} at 60-2000 Hz
Acceleration	6.8 g Max
Acoustic Excitation	136 db/Octave (50-1000 Hz)
Shroud Temperature	120° F
Space Environment	Synchronous Orbit

SECTION II

COMMUNICATION SYSTEM DESIGN

2.1 SYSTEM DESCRIPTION

The recommended RF subsystem developed in the conceptual design study is illustrated schematically in Figure 1. This system which permits independent control of each beam satisfies all design requirements and objectives.

The operation of the system and summary description of the major components is best described by following a typical signal path from the final power amplifier (FPA) to an antenna. Referring to Figure 1 and choosing signal path (1), the operation is as follows: output from the FPA is routed via rigid waveguide to one channel of the four channel 360°/day rotary joint. This joint is a TEM type with isolation of 40 db minimum between channels. Output of the rotary joint connects to two flexible waveguide sections positioned so that they permit the necessary limited motion capability ($\approx \pm 10$ degrees) about the pitch and yaw axes. Rigid waveguide then routes the signal to a mechanical SPDT, RF switch S_1 . In the normal mode this switch connects the signal directly to Antenna A-1. Similar routing to antennas A-2, A-3 and A-4 together with appropriate antenna positioning results in the in-line beam configuration. The alternate beam arrangement is provided through the use of a fifth antenna A-4' in conjunction with antennas A-1, A-2 and A-3. Connection of antenna A-4' is made by proper actuation of switches S_4 and S_5 .

The auxiliary mode, wherein two antennas each radiate two channels, is implemented by using two diplexers and transfer DPDT and SPDT mechanical RF switches as shown in Figure 1. Simply, the switch arrangement ($S_1 - S_4$) routes two non-adjacent channels into a combining diplexer whose output is connected to the proper antenna (A-2 or A-3) by transfer switches (S_2 or S_3). Channels which are non-adjacent in frequency are combined in order to minimize the diplexer requirements.

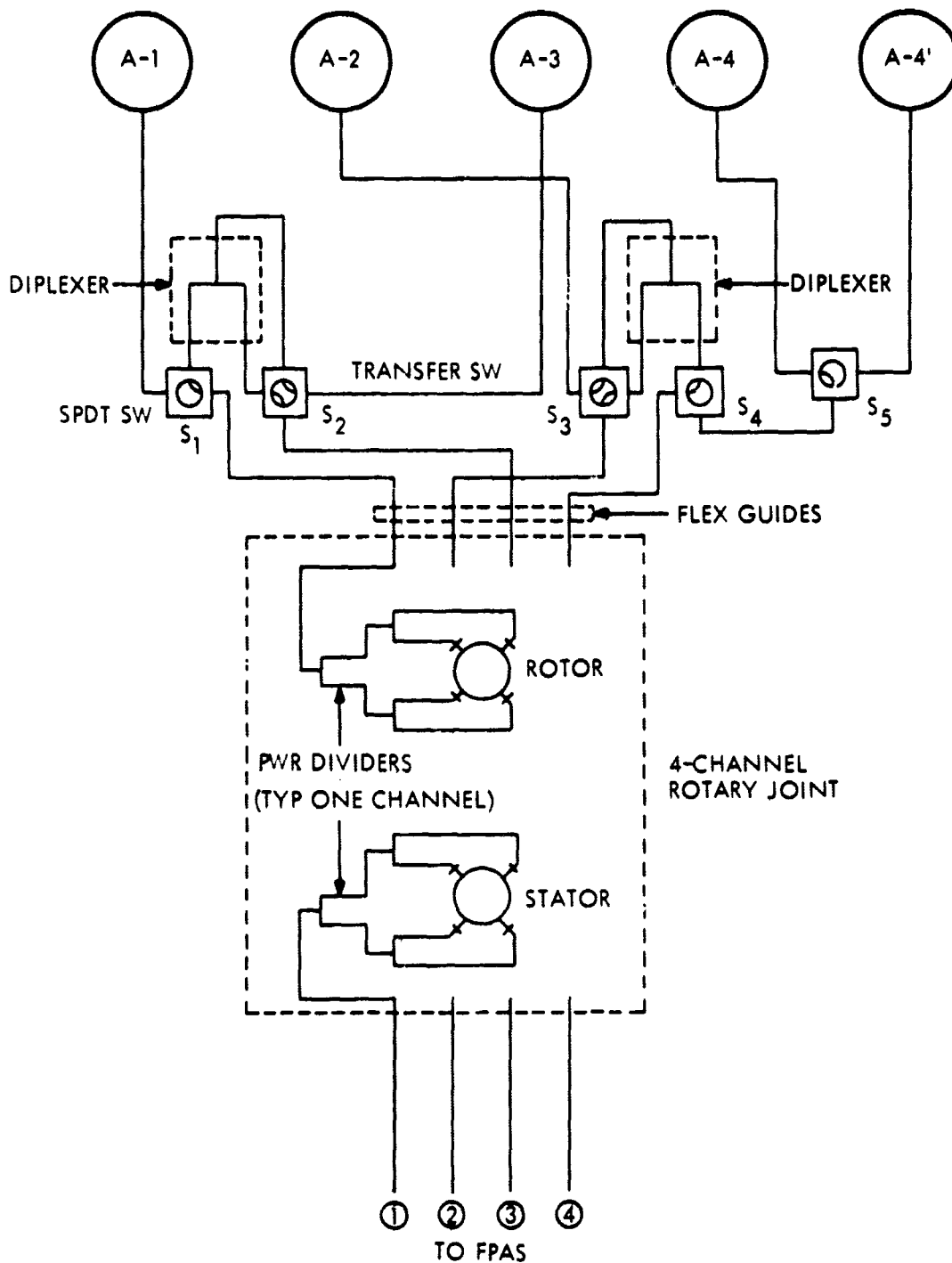
The following section describes in detail the components of the recommended RF subsystem. Subsequent sections describe alternate components considered, power handling considerations, and overall performance of the selected and alternate systems.

2.2 SYSTEM COMPONENTS

2.2.1 Antenna

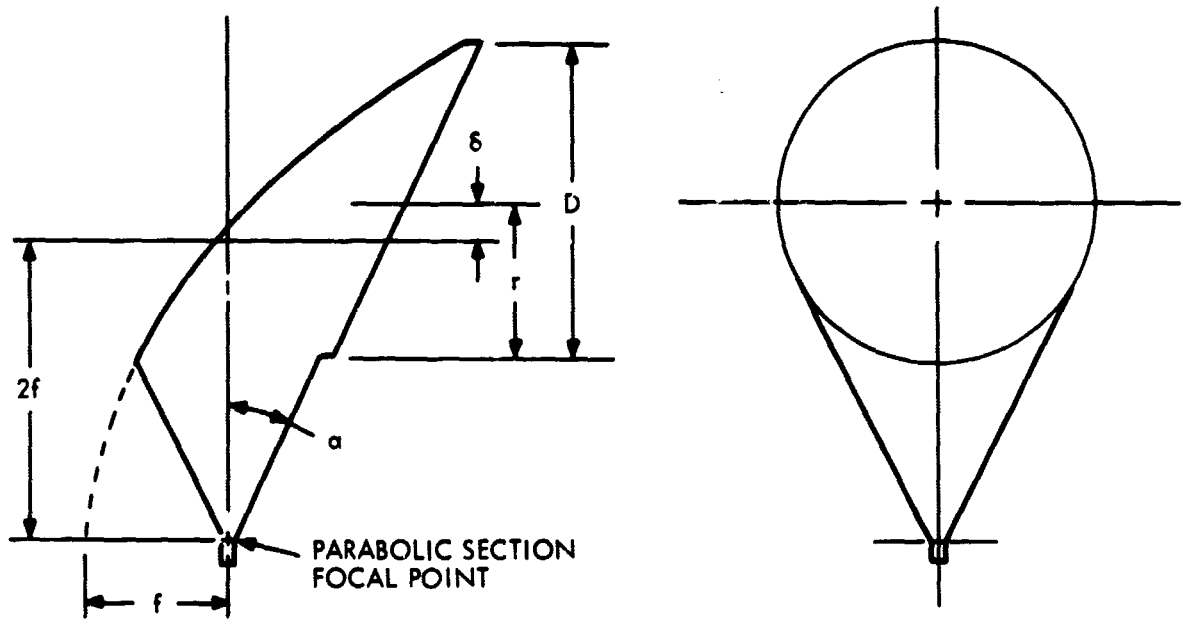
The antenna chosen for the system is a conical horn-reflector driven by a multi-mode feed exciter (TE_{11} and TM_{11} modes in circular waveguide) which affects sidelobe control. The form and relations which define the horn-reflector (less the mode exciter) are given in Figure 2. The decision to use the horn-reflector in this application was based primarily on those advantageous factors listed in Table 4.

Space allotment within the spacecraft necessitated that the horn-reflector providing a 2.7 degrees HPBW (≈ 24 -inch aperture diameter) have a half cone angle α in the vicinity of 25-30 degrees. Because this half angle is considerably larger than that



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Figure 1. RF System Block Diagram



f = PARABOLIC SECTION FOCAL LENGTH

α = HALF FLARE ANGLE

r = APERTURE RADIUS

D = APERTURE DIAMETER

$$r = 2f \tan \alpha$$

$$\frac{\delta}{r} = \frac{\sqrt{\tan^2 \alpha + 1} - 1}{\tan \alpha}$$

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Figure 2. Conical Horn-Reflector Antenna Defining Relations

TABLE 4. HORN-REFLECTOR PERFORMANCE CHARACTERISTICS

<u>Advantages</u>	<u>Significance to RF Subsystem</u>
1. No Aperture Blocking	Reduces Sidelobe Level Maximizes Gain
2. No Spillover Loss	Maximizes Gain
3. High Aperture Efficiency	Minimizes Aperture Size
4. Inherent High Power Capability	-
5. Integral Feed Eliminates Beam Displacement under Various Environments	Stable RF Pointing Direction
6. Can be Multi-mode Excited	Sidelobe Reduction Capability

previously demonstrated in horn-reflectors, and because dual modes were proposed to control sidelobes, detail design of the system antenna was based upon tests made on a small model. This small model had an aperture diameter of eight inches, a half cone angle α of 26.5 degrees, and a focal length of nine inches. This model is illustrated in Figures 3 and 4. The multimode exciter shown in Figure 4 is identical to that used in conical horns¹. It functions by generating a TM_{11} mode which propagates to the horn aperture and adds with the normal TE_{11} mode so that zero electric field occurs at the horn periphery in the E-plane. Proper amplitude of the TE_{11} mode is controlled by the excitation step ratio r_2/r_1 while the proper phasing of the TM_{11} and TE_{11} modes is adjusted through the length ℓ of the oversize circular waveguide section as shown in Figure 4.

The horn-reflector was evaluated in the 10.5 to 11.5 GHz region with only dominant excitation in order to determine the general linearly polarized pattern characteristics associated with large flare angles (α). Radiation patterns over the majority of this range are constant in form and are illustrated in Figures 6 through 8 at 11.0 GHz for the various planes and polarizations defined in Figure 5. In the planes containing the E-field, peak sidelobes range from -14.8 db to -21.3 db while in orthogonal planes the maximum sidelobe level is -25 db. Cross-polarized response is present in the equatorial plane only and is nominally at 15 db below the peak main response. This performance is similar to that of horn-reflectors with smaller α s (15.75 degrees)² except in two areas. First, the cross-polarized components is significantly higher (15 db versus 20 db) and secondly, the larger α results in an asymmetrical pattern in the longitudinal plane for longitudinal polarization. The higher cross-polarized component is a result of the increased taper differential occurring in the longitudinal plane with increasing flare angle. This characteristic results in transverse beam displacement in the circularly polarized mode which is not deleterious since it can be compensated for mechanically. The asymmetrical pattern in the longitudinal plane is caused by the direct radiation from the conical horn feed section adding to radiation from the parabolic section. Since the TM_{11} excitation provides symmetrical E-plane



Figure 3. Experimental Multi-mode Horn-Reflector

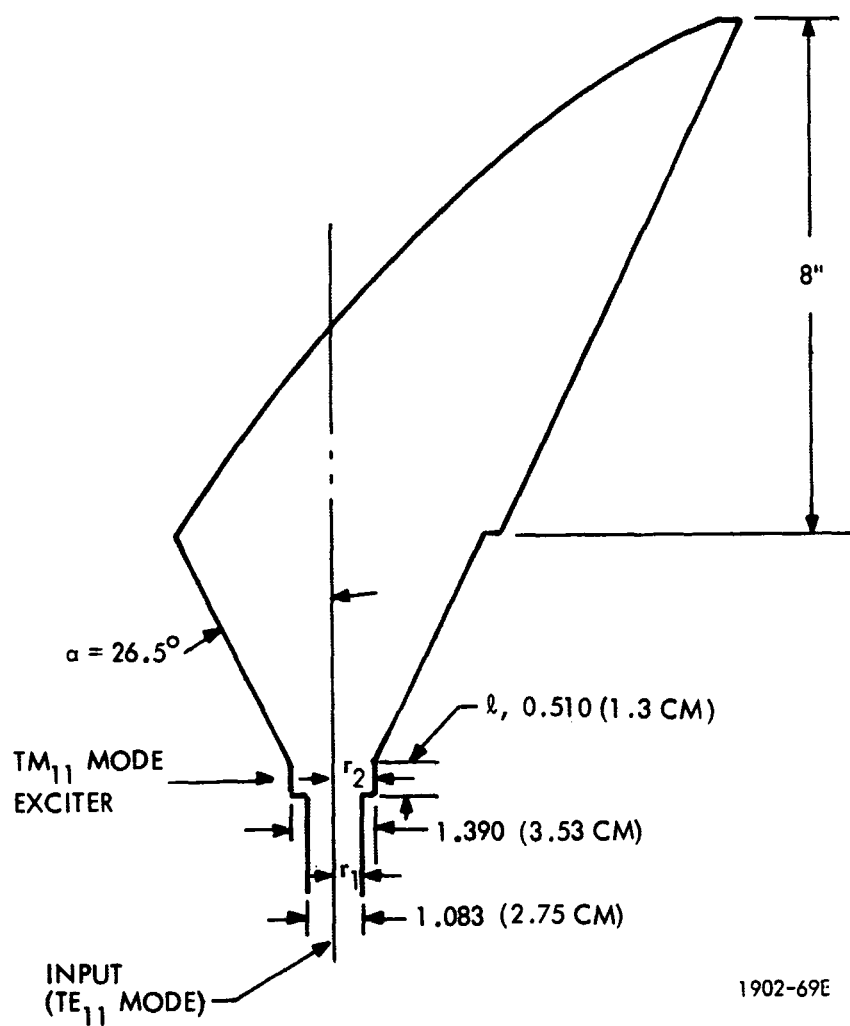
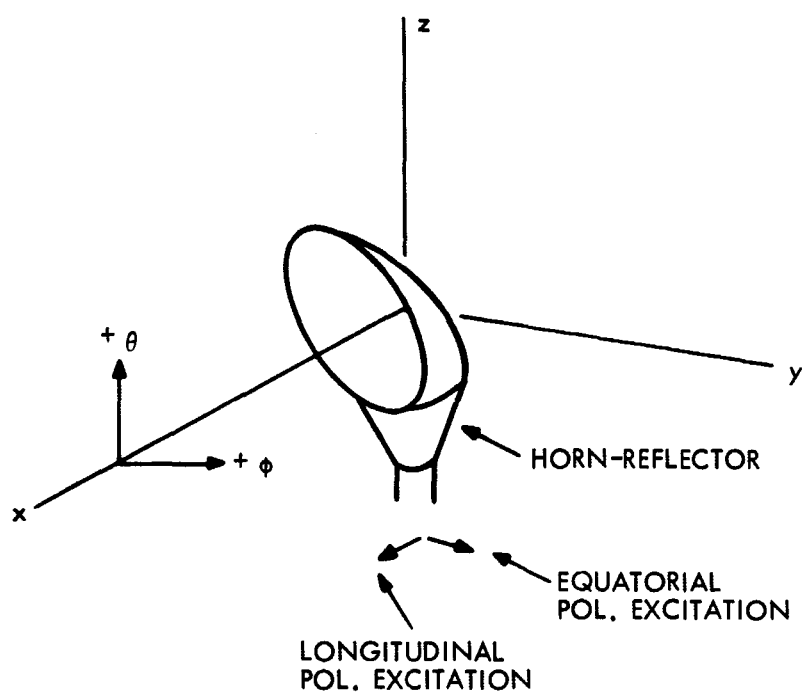


Figure 4. Experimental Multi-Mode Horn-Reflector



xy PLANE - EQUATORIAL PLANE
 xz PLANE - LONGITUDINAL PLANE

POLARIZATION IN xy PLANE - EQUATORIAL POLARIZATION
 POLARIZATION IN xz PLANE - LONGITUDINAL POLARIZATION

639-70E

Figure 5. Pattern Plane and Definitions for the Horn-Reflector

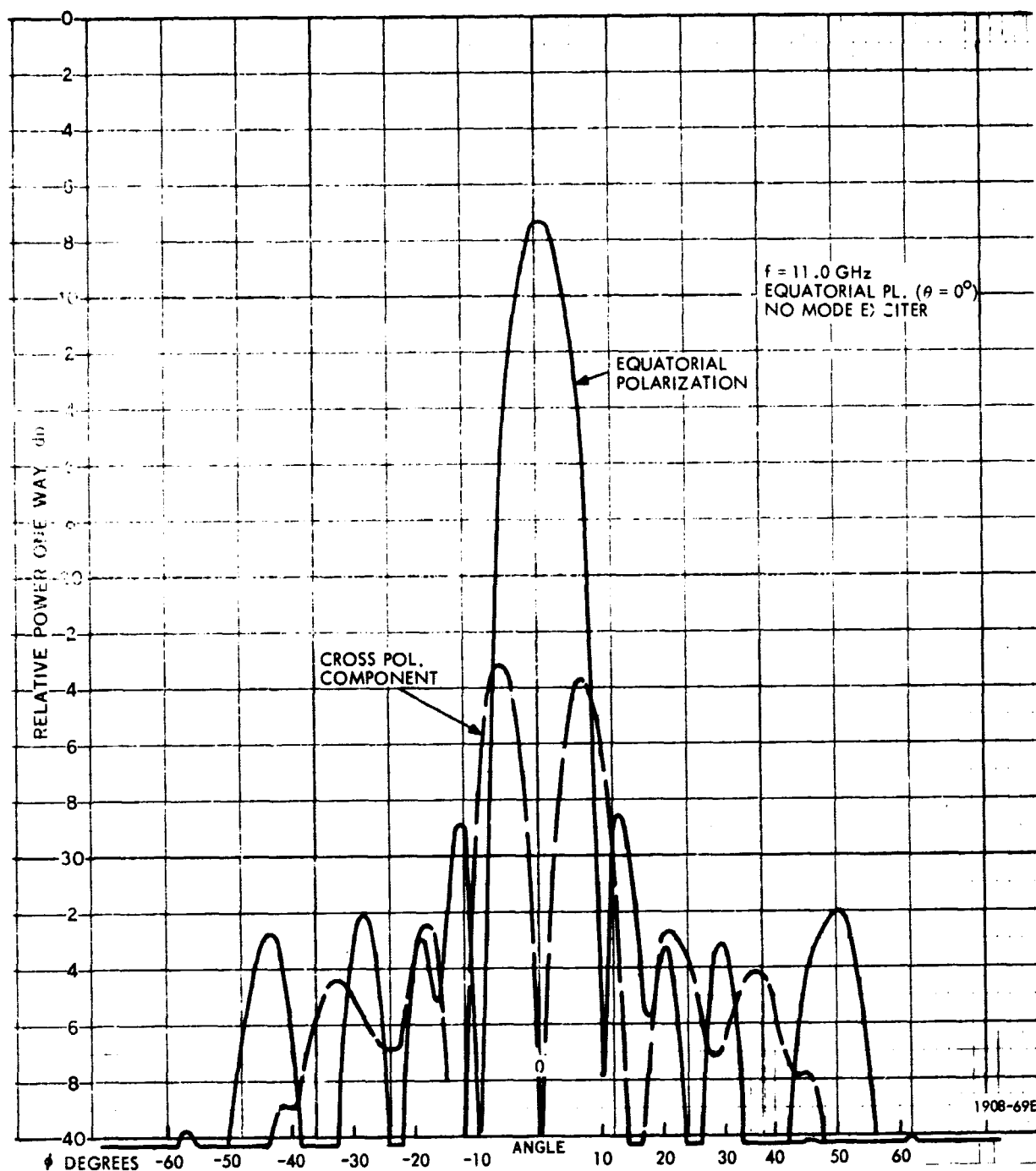


Figure 6. Single Mode Horn-Reflector Radiation Pattern

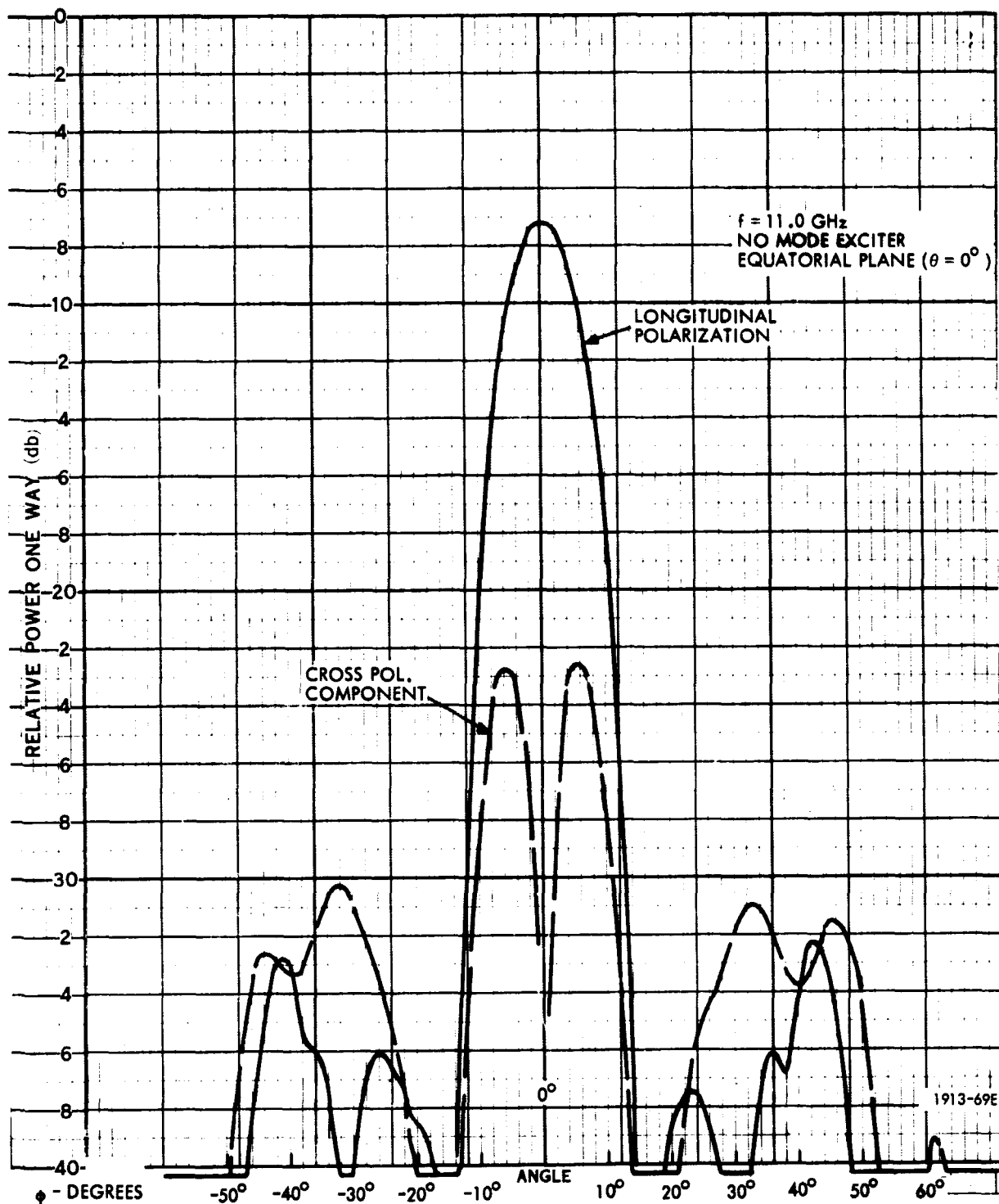


Figure 7. Single Mode Horn-Reflector Radiation Pattern

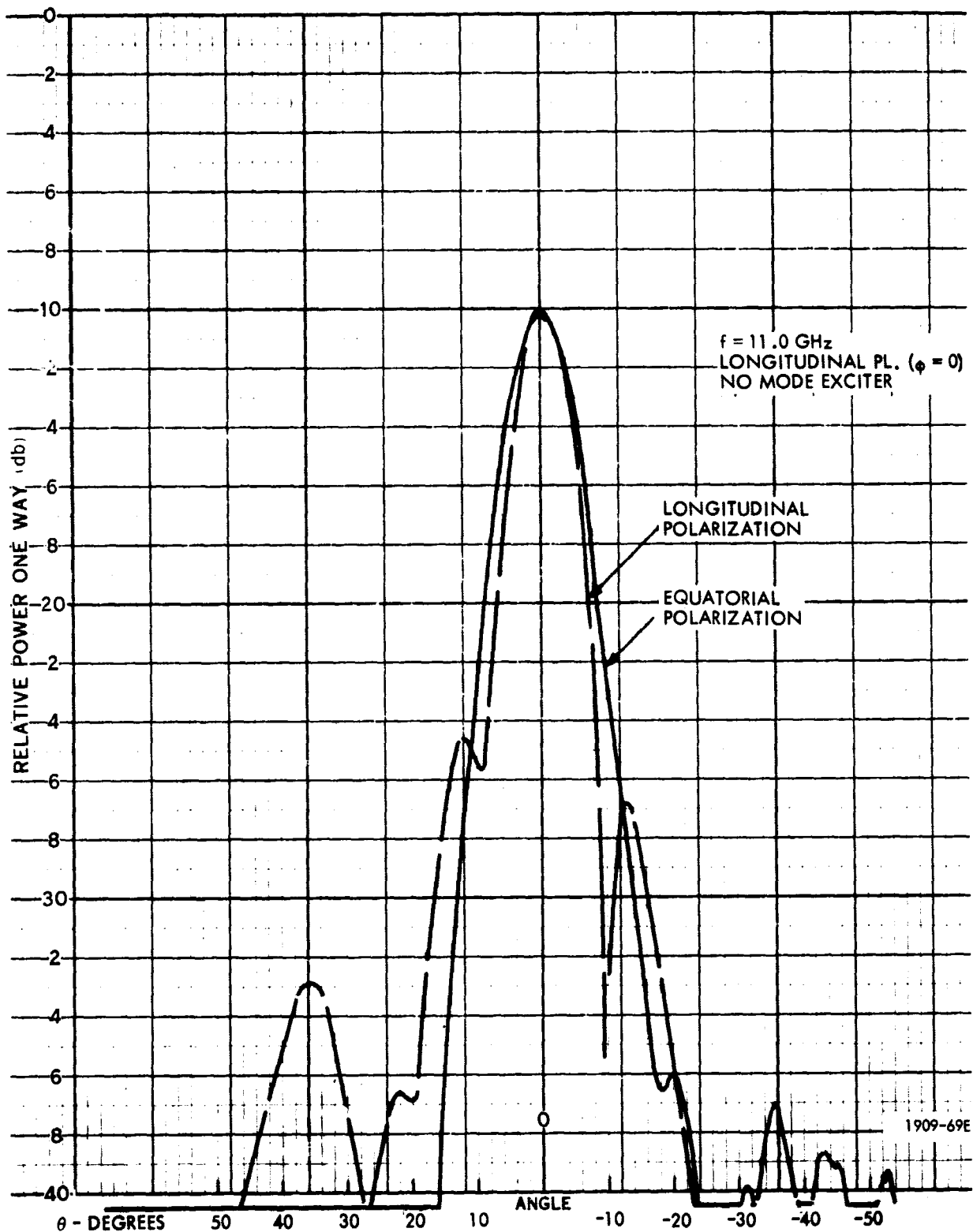


Figure 8. Single Mode Horn-Reflector Radiation Pattern

compensation, its ability to control asymmetrical sidelobe level that occurs in the wide angle horn-reflector was questioned; however, experimental tests indicated that the asymmetry did not destroy the effectiveness of the mode exciter.

Figures 9 through 12 show the 11.0 GHz linearly polarized radiation patterns obtained with the model, using the empirically optimized mode exciter. This exciter reduced sidelobes in the planes containing the E-field to values ranging from -25 db to -32 db and also reduced the cross-polarized component to -17 db.

Circularly polarized performance of the multi-mode horn-reflector was evaluated by computing the CP response from the experimental linearly polarized patterns, assuming a quadrature phase relation between the principal orthogonal components. Using the following notation for the field components:

<u>Component</u>	<u>Excitation Polarization</u>	<u>Measurement Plane</u>	<u>Far-Field Polarization</u>
E_1	Longitudinal	Longitudinal Plane	Longitudinal Polarization
E_2	Equatorial	Longitudinal Plane	Equatorial Polarization
E_3	Longitudinal	Equatorial Plane	Longitudinal Polarization
E_4	Equatorial	Equatorial Plane	Equatorial Polarization
E_5	Equatorial	Equatorial Plane	Longitudinal Polarization
E_6	Longitudinal	Equatorial Plane	Equatorial Polarization

the far field radiation patterns are given by:

$$E_{lr} = K (E_1 + E_2) \quad (1)$$

$$E_{ll} = K (E_1 - E_2) \quad (2)$$

$$E_{er} = K \{(E_3 + E_4) - (E_5 + E_6)\} \quad (3)$$

$$E_{el} = K \{(E_3 - E_4) + (E_5 - E_6)\} \quad (4)$$

where

E_{lr} = desired polarized response in the longitudinal plane

E_{ll} = cross polarized response in the longitudinal plane

E_{er} = desired polarized response in the equatorial plane

E_{el} = cross polarized response in the equatorial plane

K = constant.

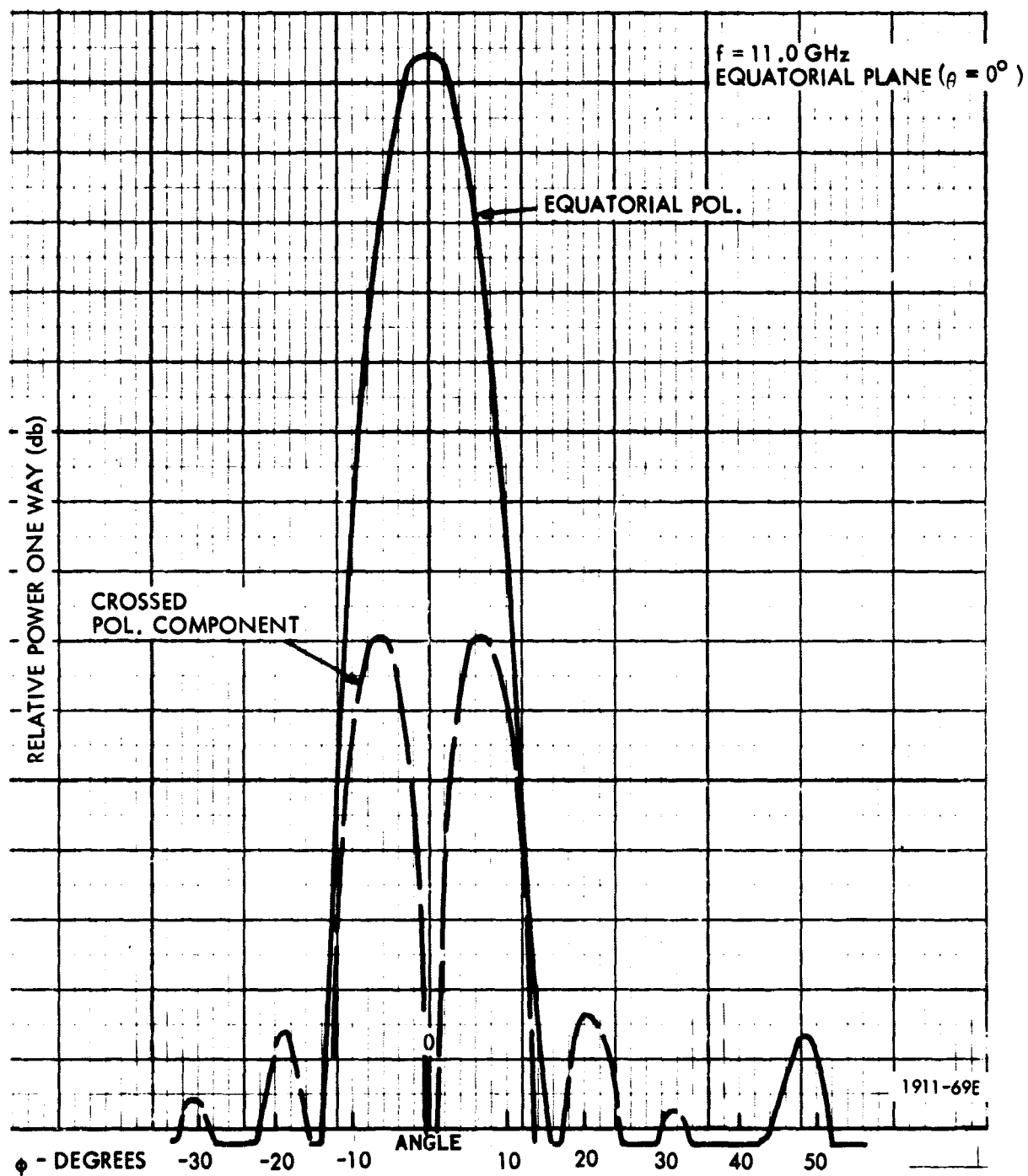


Figure 9. Multi-Mode Horn-Reflector Radiation Pattern

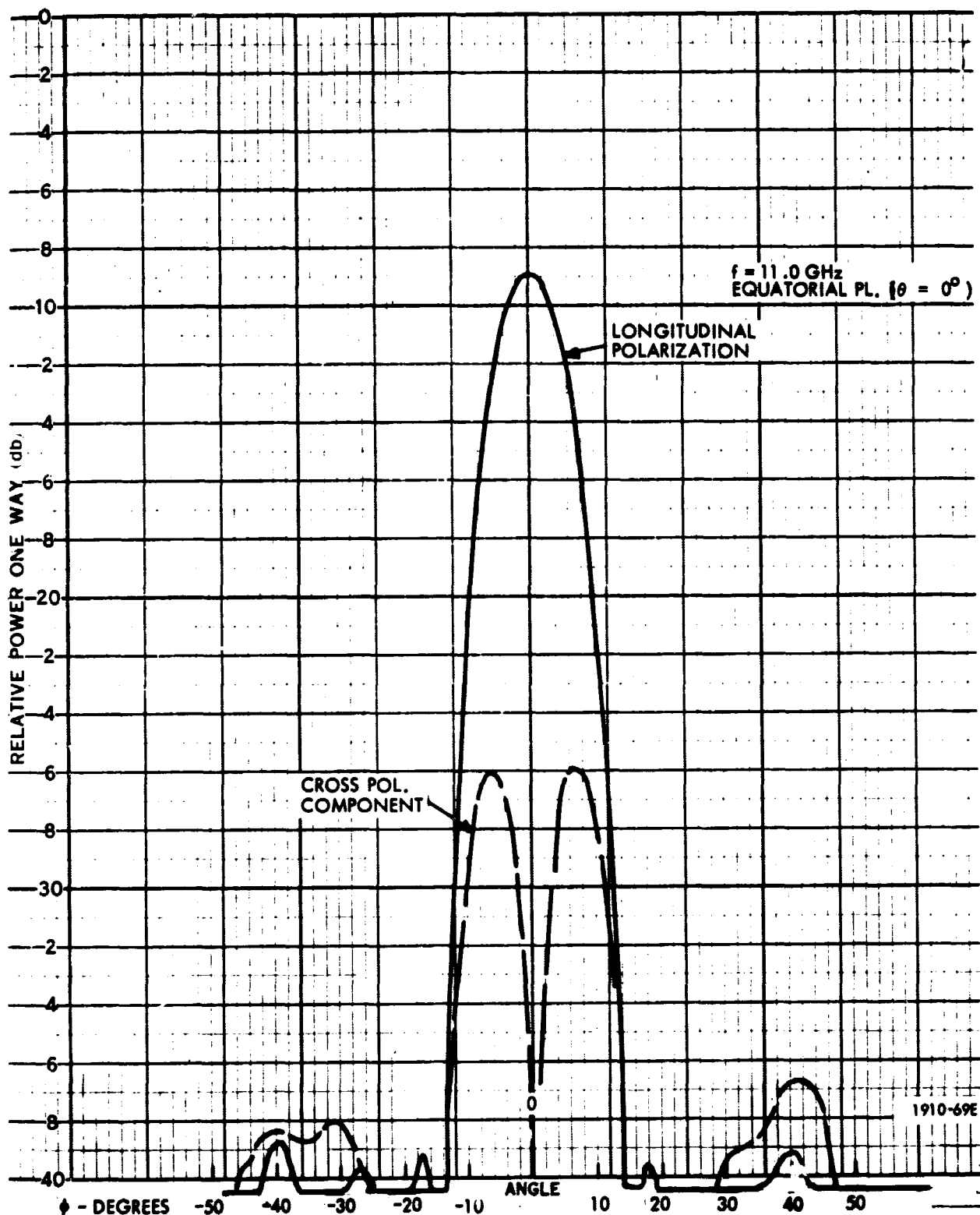


Figure 10. Multi-Mode Horn-Reflector Radiation Pattern

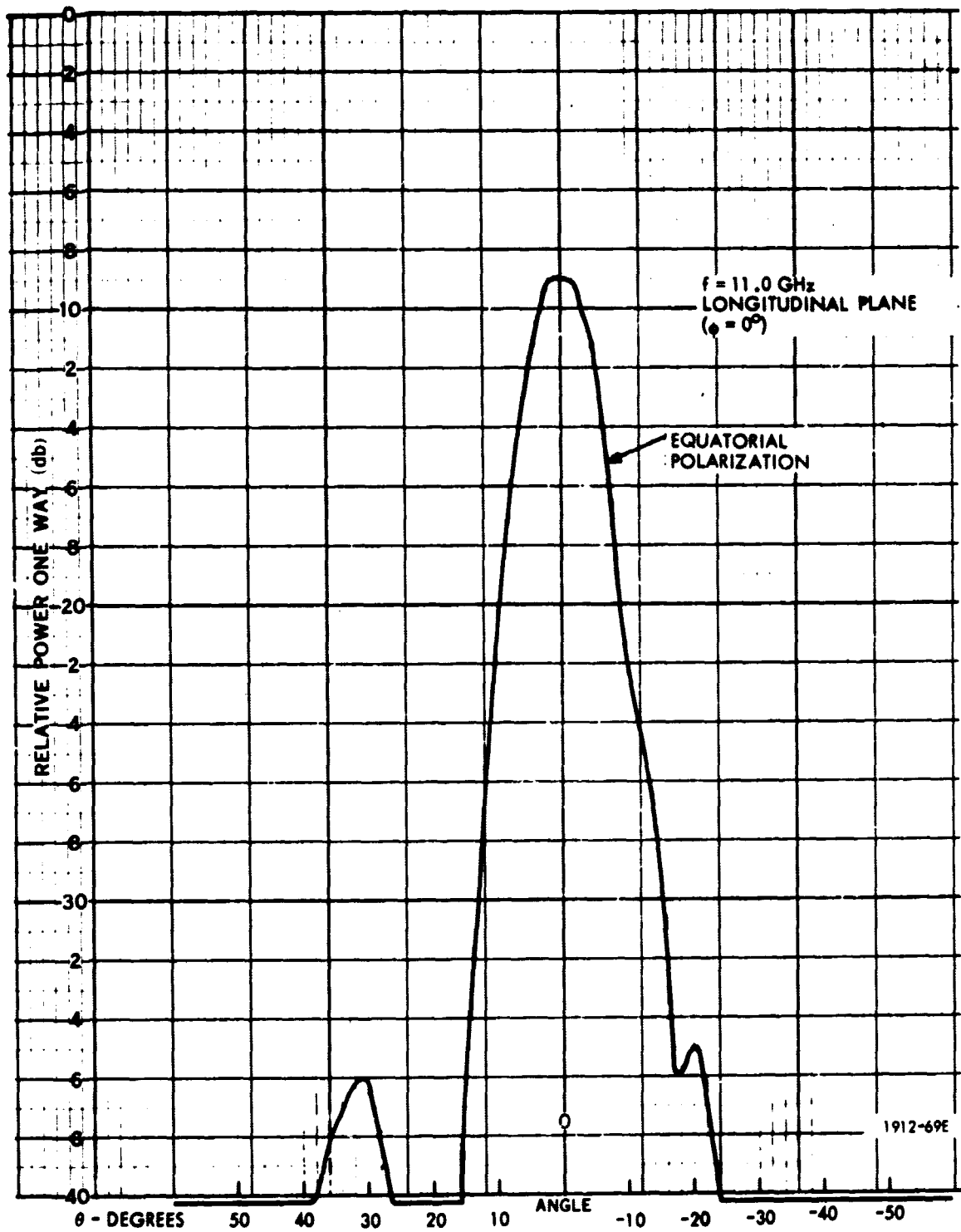


Figure 11. Multi-Mode Horn-Reflector Radiation Pattern

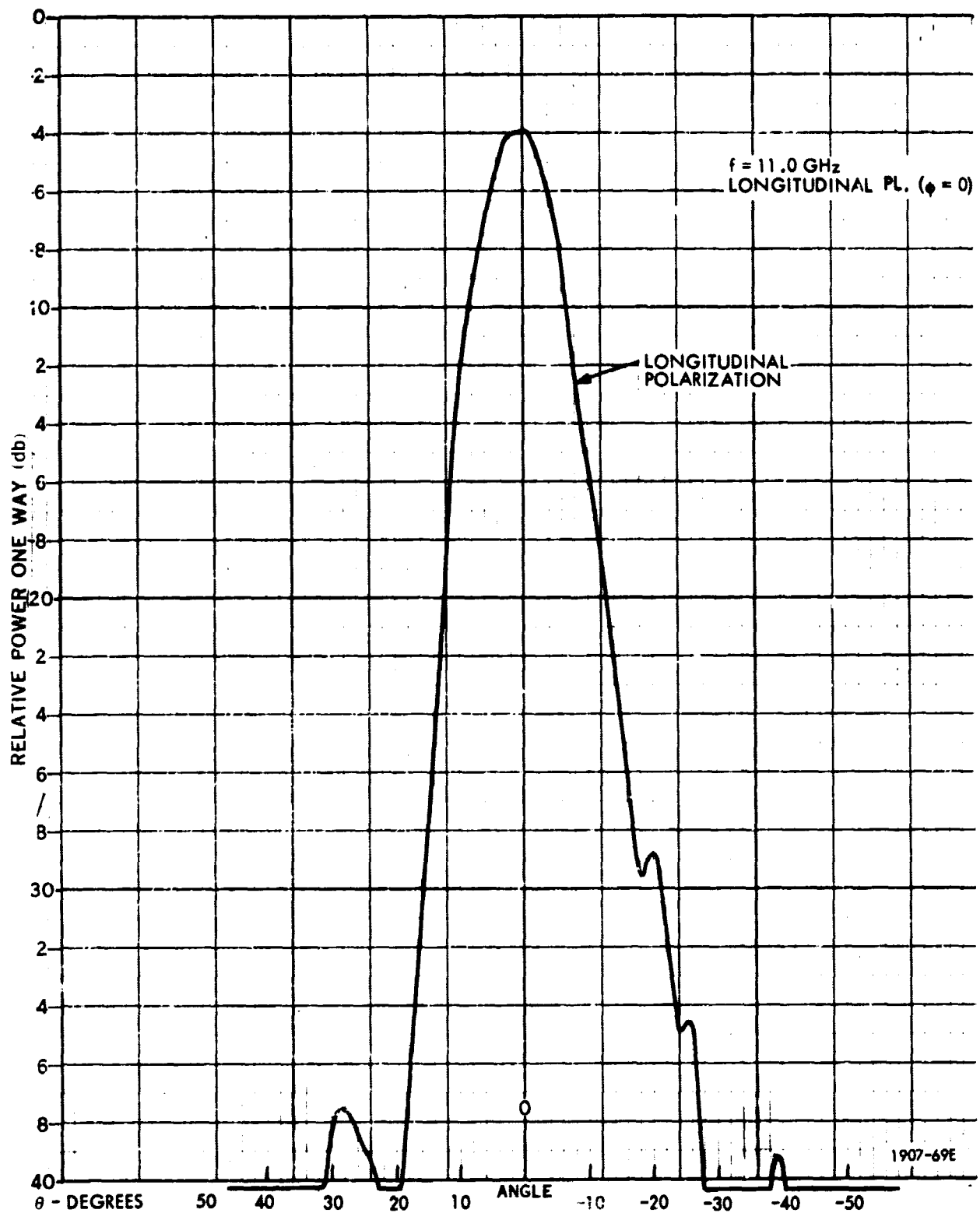


Figure 12. Multi-Mode Horn-Reflector Radiation Pattern

These computed patterns are shown in Figures 13 and 14 for the equatorial and longitudinal planes, respectively. The displacement that occurs in the equatorial plane is inherent with the circular polarization operation of the horn reflector. The non-unity peak of the longitudinal plane pattern is due to the fact that it has been computed in a plane displaced 1.2 degrees from the beam peak. Again, this is due to the inherent beam displacement. Maximum sidelobe levels on the longitudinal and equatorial planes are -25 db and -29 db, respectively. Cross-polarized response is in all cases down at least 25 db from the beam peak.

On the basis of the computed circularly polarized response derived from the model, the beamwidth constant of the multi-mode horn-reflector was determined to be 73. Thus, for the system antenna, the half-power beamwidth (HPBW) and aperture diameter in wavelengths, (D_λ) are related by

$$\text{HPBW} = 73/D_\lambda. \quad (5)$$

For the required 2.7 degrees HPBW this results in an aperture diameter of 27 inches.

The gain of the system horn-reflector was determined by deriving a gain (directivity) constant from the model horn-reflector computed directivity and HPBW when operating in the circularly polarized mode. The resulting gain constant from this computation is 39,600. Thus, the relation between peak gain and HPBW is

$$G = 39,600/B^2 \quad (6)$$

where

G = peak gain relative to circularly polarized isotropic

B = HPBW in degrees.

For the 2.7-degree beam system, a peak gain of 37.3 db is available at the feed input terminal.

Far-out sidelobe and backlobe performance that can be expected of the multi-mode horn-reflector is illustrated in Figure 15. Here, a linearly polarized pattern taken in the longitudinal plane shows that far-out sidelobes are nominally down at least 30 db and back radiation is down more than 40 db.

The ellipticity of the polarization ellipse of the horn-reflector is estimated from the computed circularly polarized response shown in Figures 13 and 14. The axial ratio is given by

$$\text{AR} = 20 \log \frac{V_1 + V_2}{V_1 - V_2} \text{ db} \quad (7)$$

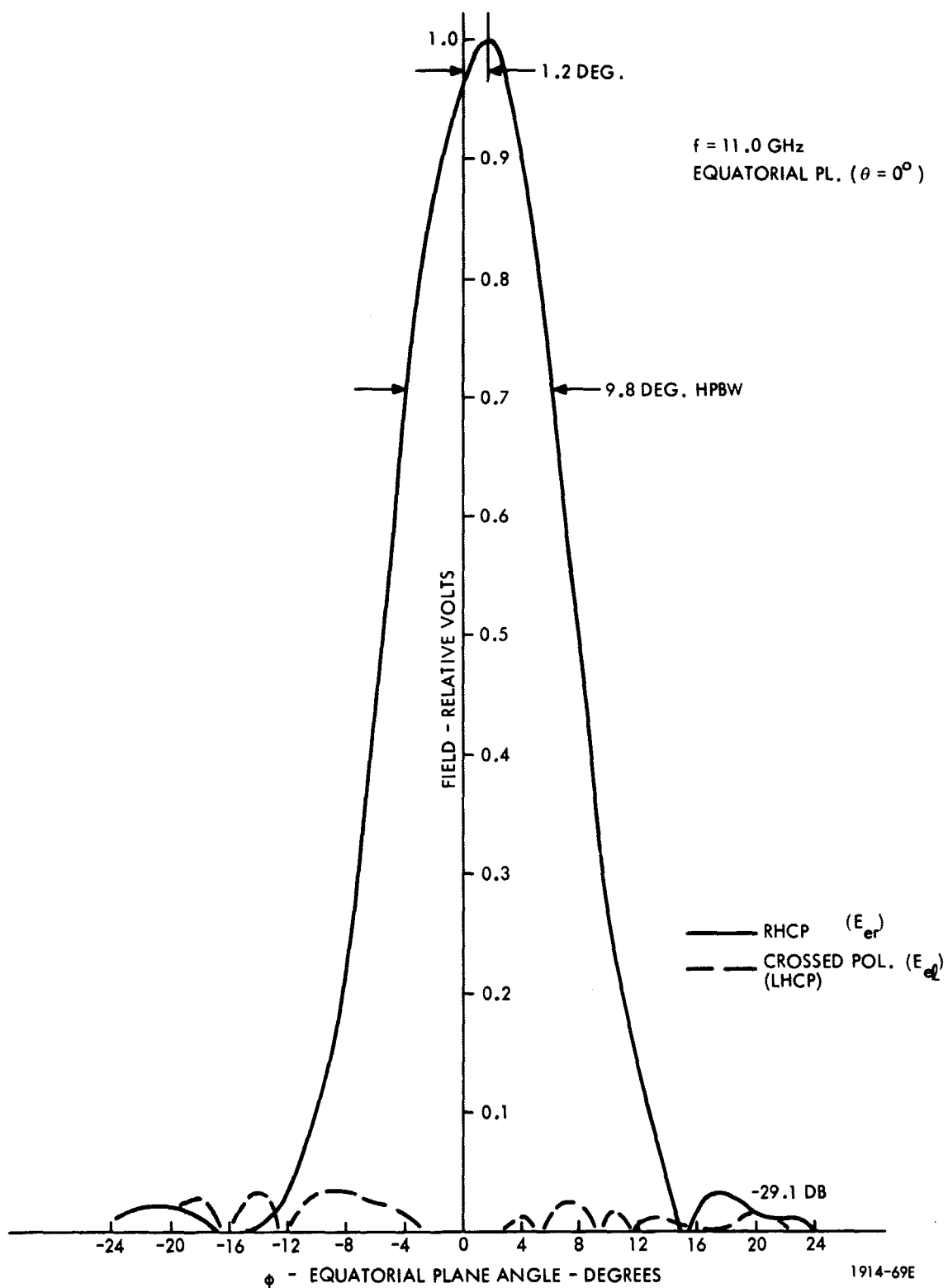


Figure 13. Multi-Mode Horn-Reflector Circularly Polarized Radiation Pattern

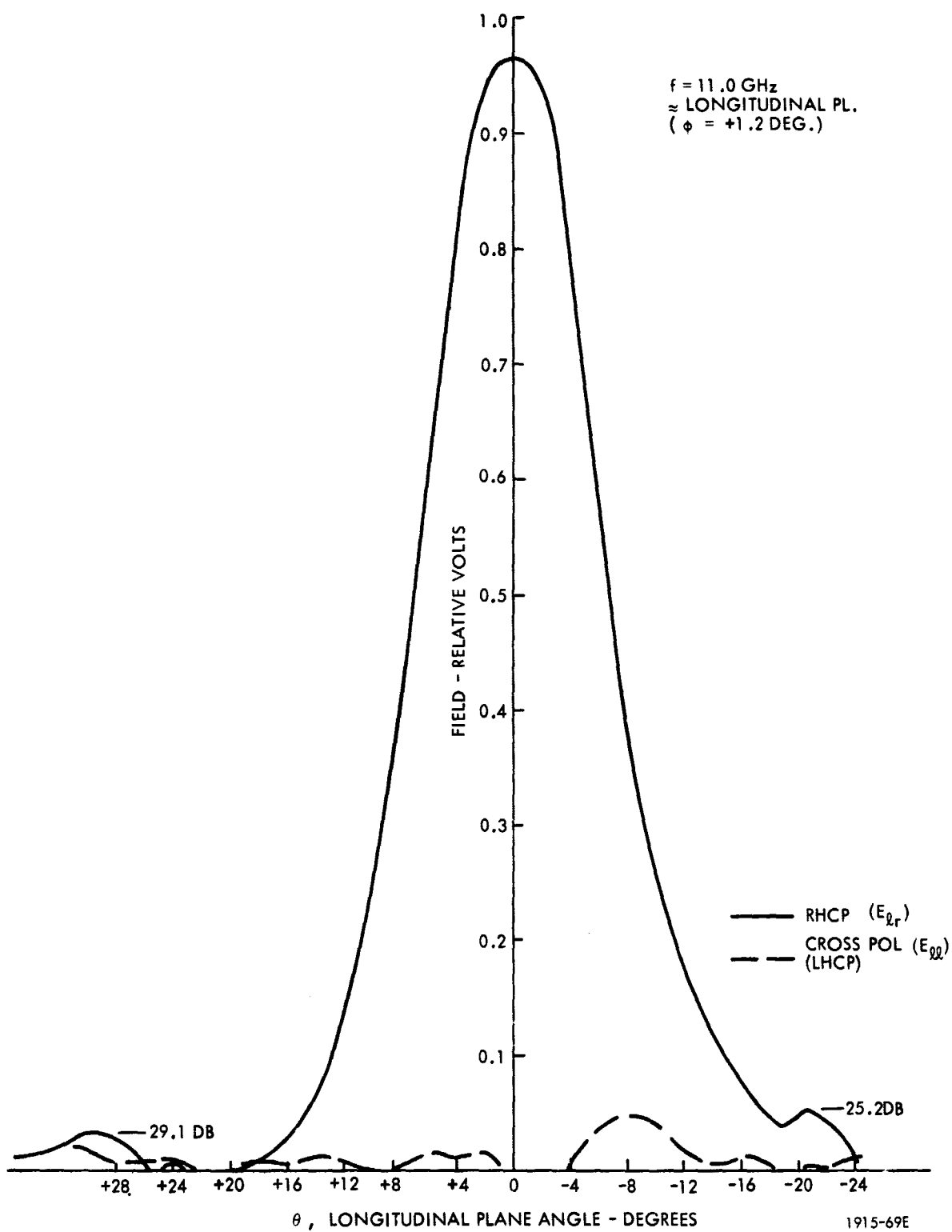


Figure 14. Multi-Mode Horn-Reflector Circularly Polarized Radiation Pattern

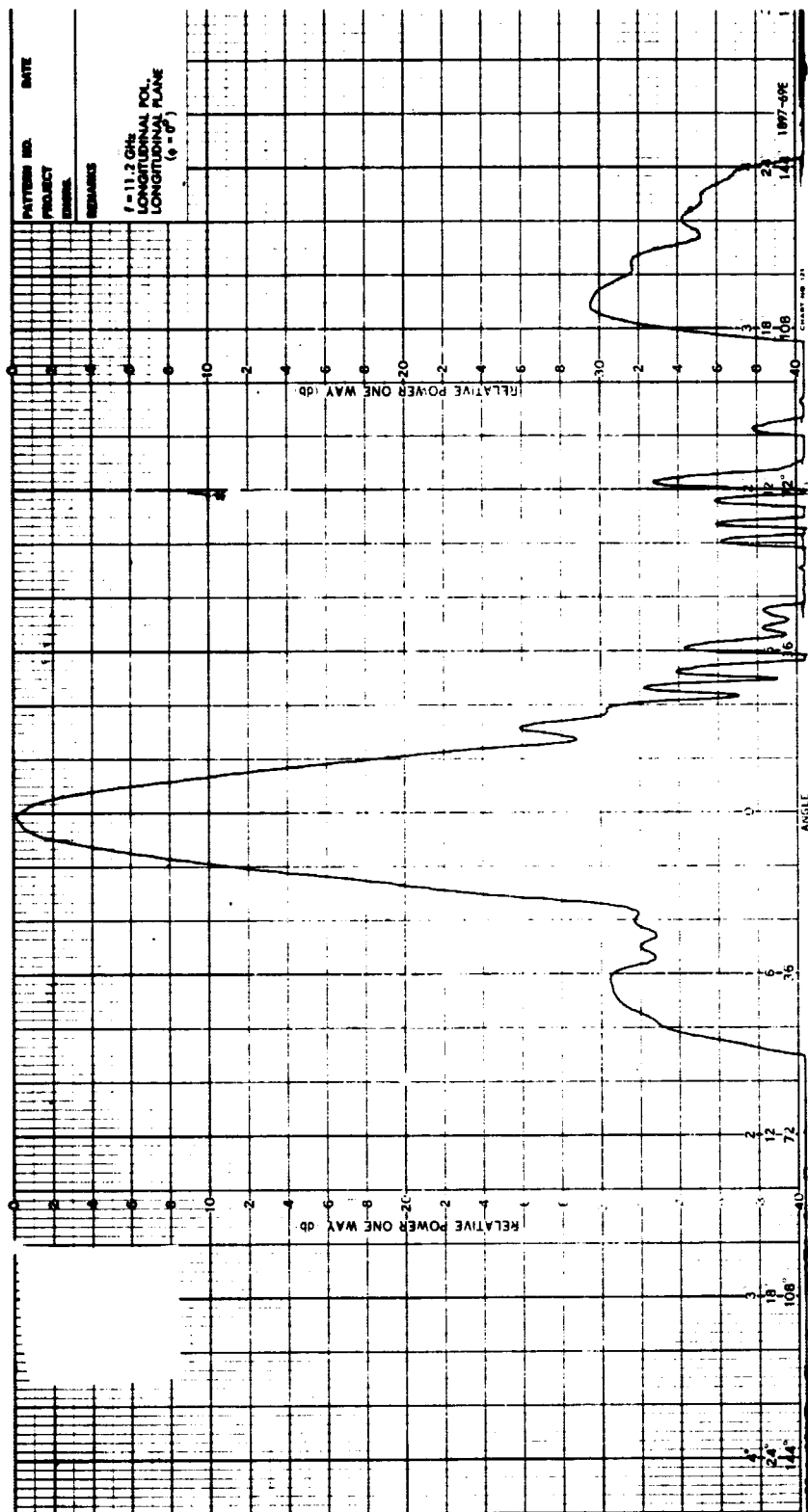


Figure 15. Multi-Mode Horn-Reflector 360° Radiation Pattern

where

V_1 = voltage level of the desired sense

V_2 = voltage level of the cross polarized sense

AR = polarization axial ratio.

Maximum polarization axial ratio within the beam occurs at the half-power points wherein the ratio of the desired to undesired sense of circular polarization (V_1/V_2) is a maximum. For the case in question V_1 and V_2 are nominally 0.707 and 0.02 volts respectively (see Figures 13 and 14) giving an axial ratio of 0.48 db. Axial ratio at other points within the beam is considerably lower since the cross-polarized component approaches zero at the beam center. These estimates of polarization axial ratio assume a perfect circularly polarized wave at the feed input. Performance of the circular polarizer (Section 2.2.6) shows that a 0.2 db axial ratio ratio can be achieved. By phasing the polarizer and aperture characteristics an overall polarization axial ratio of 0.4 db is obtainable over the HPBW and over the entire channel width.

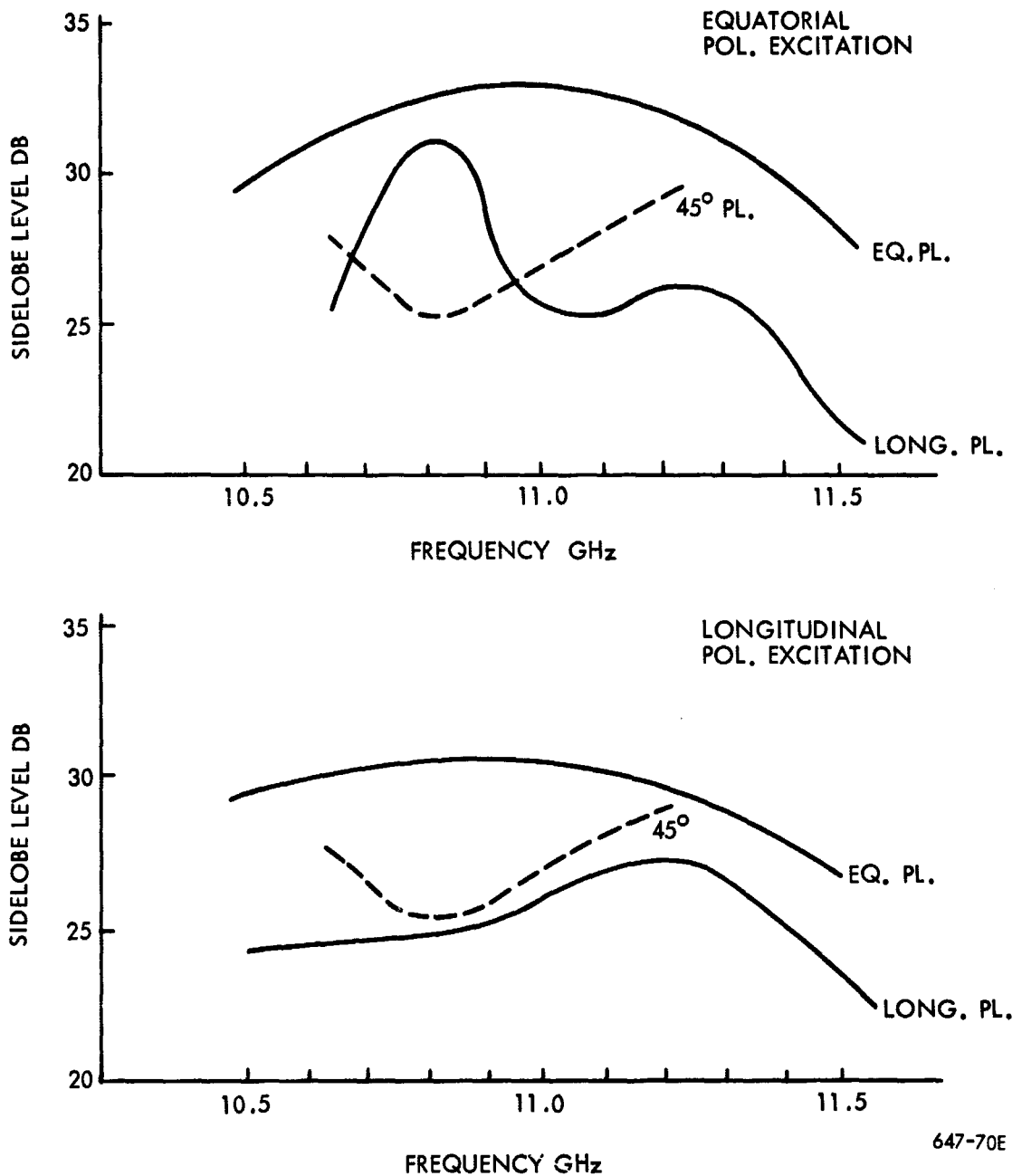
The bandwidth properties of the horn-reflector as determined by the model indicate a usable bandwidth of 400 MHz. This bandwidth is based upon the sidelobe performance for linear polarization being held to a level of at least -25 db as illustrated in Figure 16. Thus the bandwidth capability of the horn-reflector is sufficient to cover both the prime mode (50 MHz) and the auxiliary mode (150 MHz) with considerable margin. Impedance bandwidth of the antenna is also satisfactory in that tests on the model have shown input VSWR of less than 1.1:1 over a bandwidth of 1.05:1.

In summary, a horn-reflector antenna is recommended in this conceptual design. The antenna has an aperture diameter of 27 inches, a half flare angle of 26.5 degrees and electrical dimensions as shown in Figure 17. All sidelobes are suppressed to at least 25 db down and back radiation is in excess of 40 db down. This sidelobe performance is achieved through the use of multi-mode excitation of the horn. Peak gain will be 37.3 db above a circularly polarized isotropic source. Input into the horn will be via a circular waveguide with approximately a 0.7-inch diameter. Independent apertures are used for each beam in order to permit individual optimization and maintain the necessary 40 db isolation between channels.

2.2.2 Principal Rotary Joint

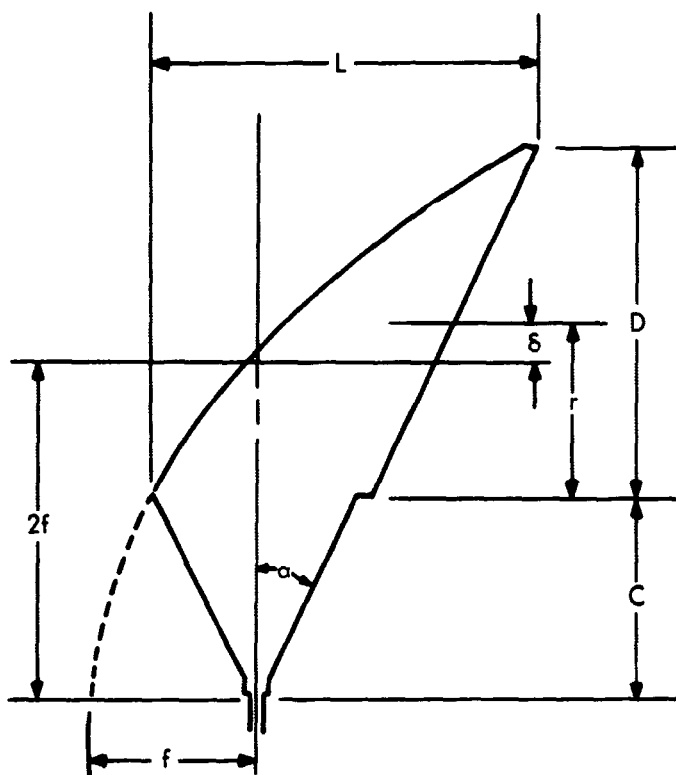
The principal rotary joint permits continuous 360°/day rotation of the antenna system relative to the spacecraft about the pitch axis. Hence the antenna beams can be pointed at earth continuously while the main spacecraft is sun-oriented. The rotary joint is required to accommodate the four separate communication channels from the final power amplifiers.

A four-channel rotary joint with each channel operating in the TEM mode has been selected for the conceptual design. This design, which is illustrated in Figures 18 and 19 was chosen primarily because its low insertion loss, hence power dissipation, was less than in alternative designs (see Section 2.3.2). In addition, the fact that the inter-channel isolation in this design is independent of other components made its use attractive.



647-70E

Figure 16. Experimental Horn-Reflector Sidelobe Level vs Frequency



$\alpha = 26.5^\circ$
 $f = 13.52 \text{ INCHES (34.35 CM)}$
 $D = 27 \text{ INCHES (68.5 CM)}$
 $\delta = 2.72 \text{ INCHES (6.9 CM)}$
 $L = 31.5 \text{ INCHES (80 CM)}$
 $C = 15.75 \text{ INCHES (40 CM)}$

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RADIATION CHARACTERISTICS

HPBW = $73/D_\lambda$ ----- 2.7 DEG

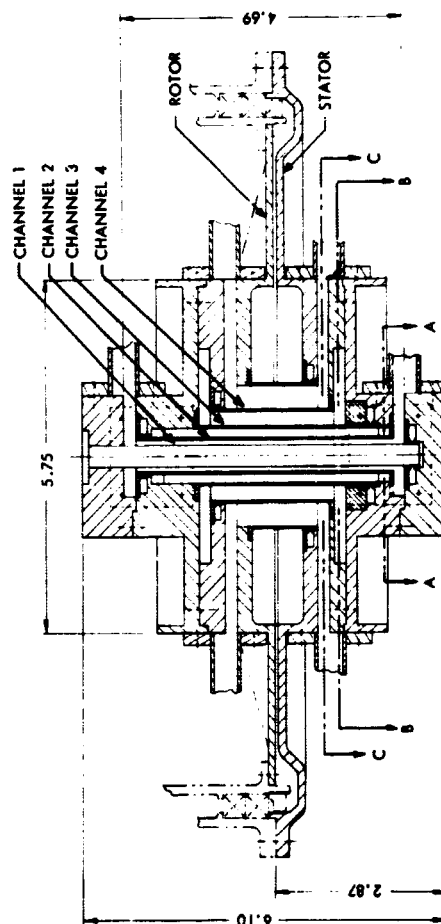
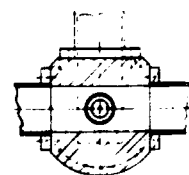
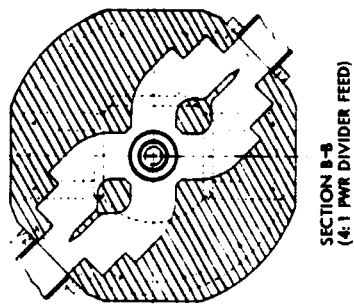
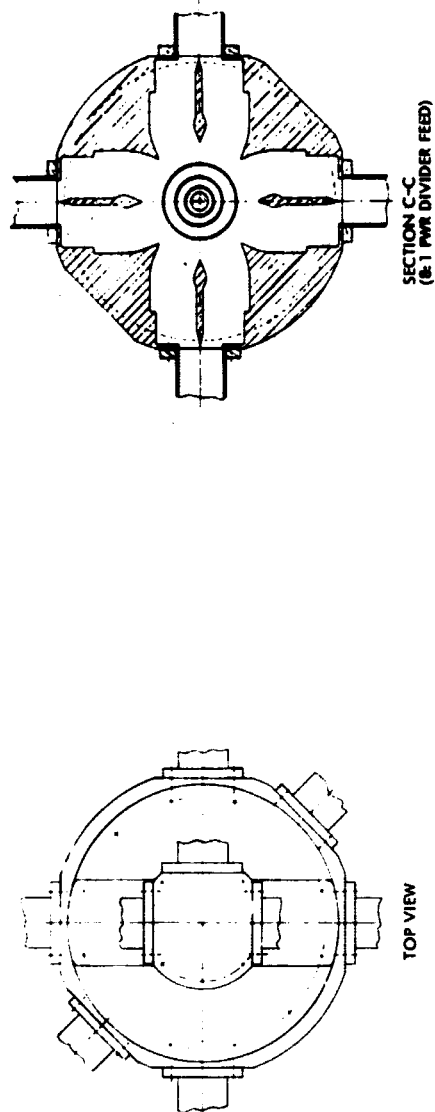
PEAK GAIN $\approx \frac{39600}{(\text{HPBW})^2}$ ----- 37.3 DB

APERTURE EFFICIENCY ----- -75.8 %

SIDELobe LEVEL ----- > 25 DB DOWN

CROSS POLARIZED LEVEL ----- > 25 DB DOWN

Figure 17. Summary Characteristics of the System Multi-Mode Horn-Reflector



FOUR CHANNEL ROTARY JOINT CROSS SECTION

SECTION A-A
(2:1 PWR DIVIDER FEED) 2004-69E

Figure 18. Four-Channel 360°/Day Rotary Joint.

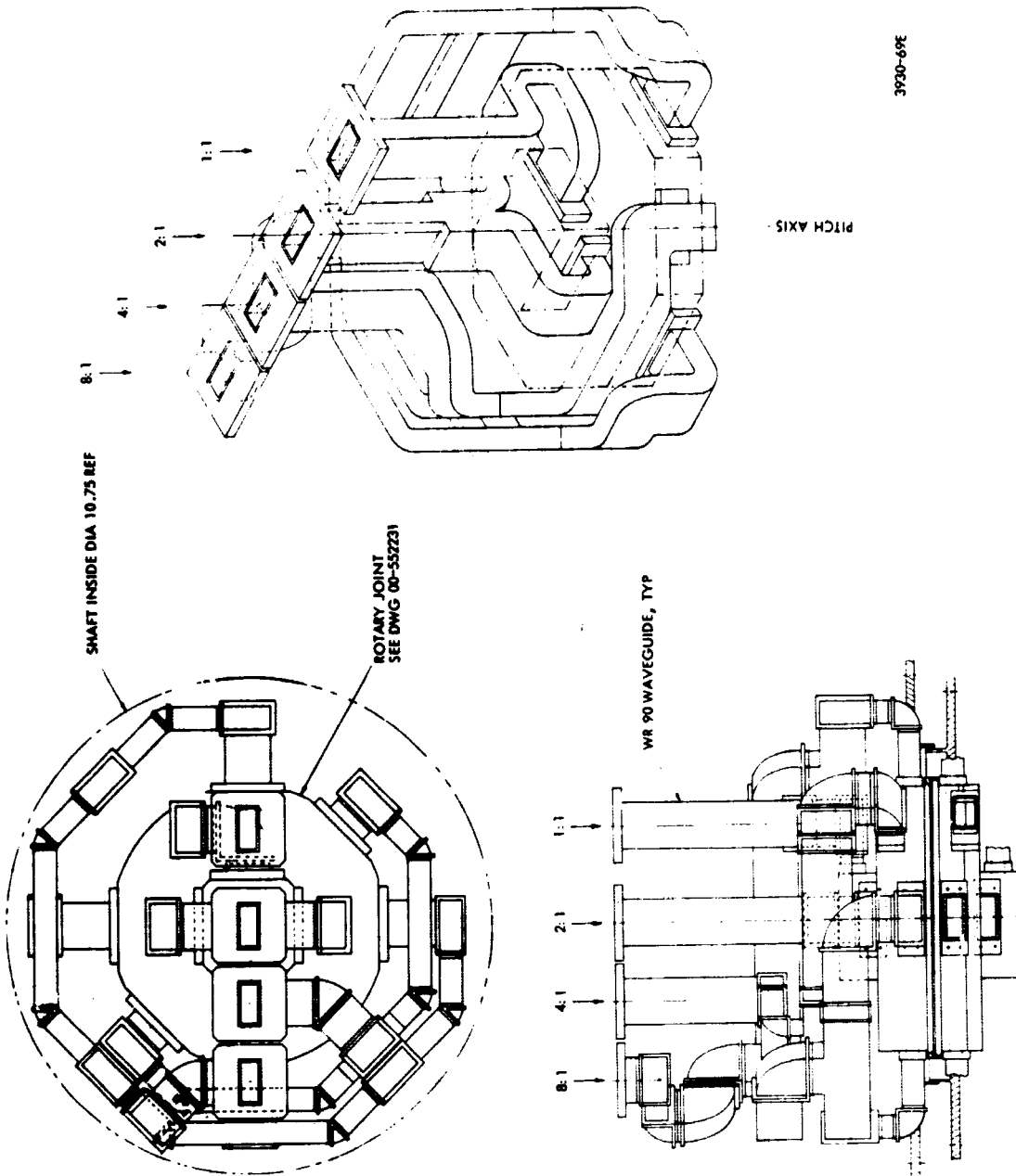


Figure 19. Four-Channel 360°/Day Rotary Joint and Feed Network

The design shown in Figures 18 and 19 uses conductors which have been maximized in diameter and minimized in axial length in order to enhance the average power capability of the joint. The large conductor diameters and short lengths will permit the power dissipated in the joint to be conducted out and radiated to space with acceptable temperature rise within the device (see Section 4.3.1).

Figure 20 shows the major internal dimensions of the rotary joint. All of the TEM lines have nominal 25 ohm characteristic impedances and are fed identically at the stator and rotor ends by half-height waveguides. The 25 ohm impedance was estimated as the lowest value at which satisfactory impedance match with the waveguide could be made. Half-height waveguide was used to minimize the joint length and to aid in impedance matching to the coaxial sections.

Excitation of the TEM sections of the joint is provided by multiple in-phase waveguides about the periphery of each coaxial section. Each waveguide is half-height (0.20 inch) and is 0.900-inch wide. The waveguide feeds for a particular channel are fed from an equal phase and amplitude waveguide reactive power divider. The number of excitation points required for each channel is governed by the higher order mode properties of the TEM section. A greater number of exciters are required for the outermost channels to preclude higher order mode excitation since the diameters of the coaxial line, necessary here, are considerably larger than for the innermost channel, Channel 1. The higher modes of concern that can propagate in the coaxial portions of joint are indicated in Figure 21 together with their cutoff frequencies. The significant modes are of the TE_{n1} type where n is an even integer and equal to half of the total number of phase changes that occur when traversing 360 degrees about the axis of the center conductor.

To insure that TE_{n1} modes which can propagate in the coaxial section are being excited, $2n$ equi-spaced discrete feed points are required about the periphery. If distributed sources such as open ended waveguides are used it is possible to reduce the number of feed points somewhat. In the recommended design, $2n$ feed probes (where n is the mode number of the lowest frequency propagating TE_{n1} mode) have been provided in all channels except channel 2. The departure in the second channel was due to the 25 ohm requirement and an attempt to maximize line diameters. Note, however, that for Channel 2 (Figure 21), the two waveguides that are used to feed the coax cover the entire rim of the outer conductor. This distributed excitation should minimize the likelihood of TE_{21} excitation in this channel since the two distributed feeds tend to approximate a four point discrete feed. The number of feed guides for each channel of the joint is summarized in Figure 21. No possibility of higher order mode problems exist in the design except for the second channel, and here the possibility is negligible.

The chokes used in the rotary joint are high impedance quarter-wave length short-circuited radial waveguide sections preceded by paralleled quarter-wave low impedance lines which enter adjacent channels of the joint. The choke design is illustrated in Figure 20. It is estimated that this choke design will provide the required 40 db isolation between channels.

Insertion loss of the rotary joint has been computed and is tabulated in Table 5. The loss in the rotary (coaxial) section was computed by conventional transmission line theory. Basic assumptions in the computation were that the joint used aluminum

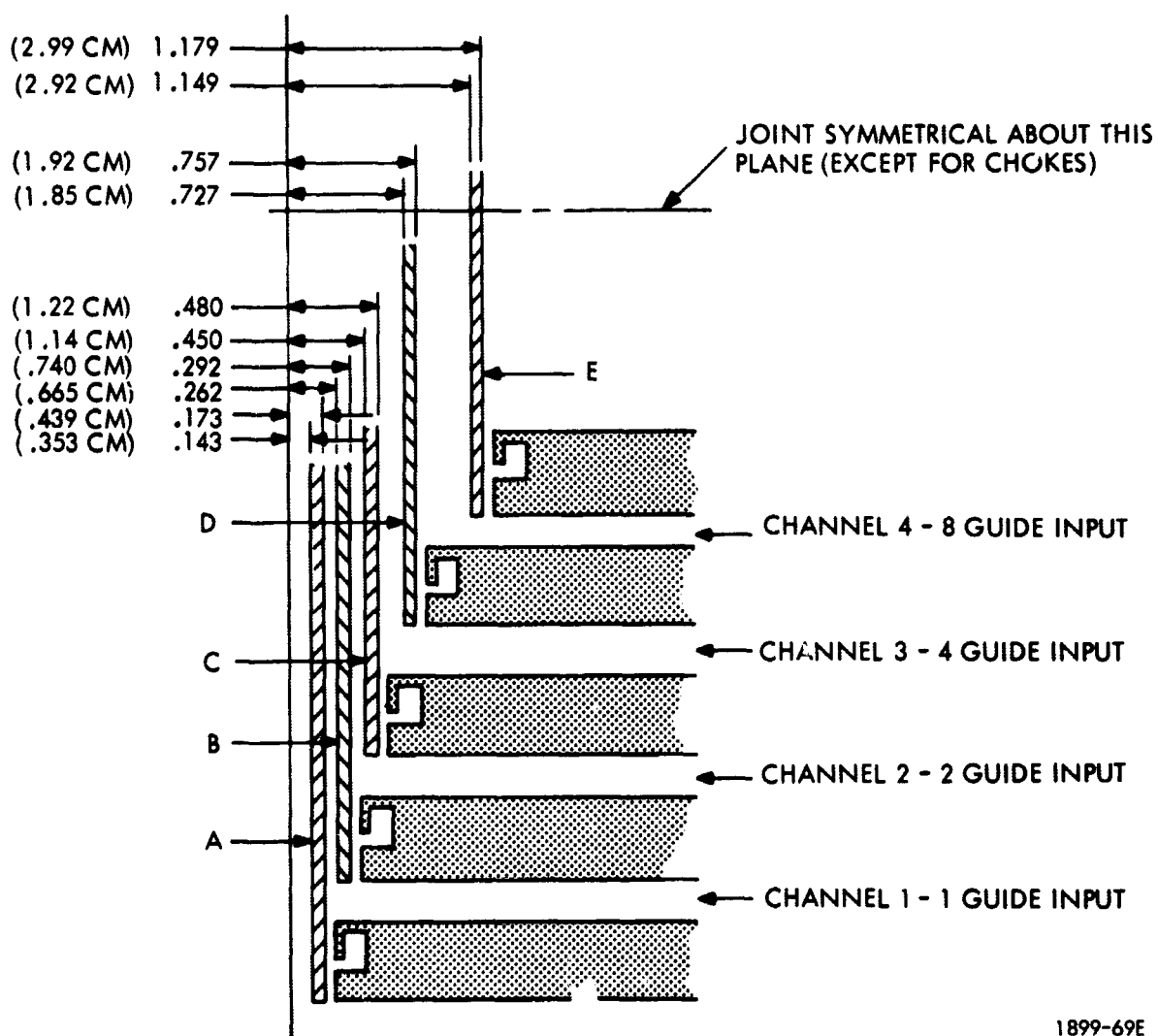


Figure 20. Four-Channel 360°/Day Rotary Joint Conductor Dimensions

JOINT	CHANNEL	RADI		CUTOFF FREQUENCIES			REQUIRED FEED POINTS
		$\frac{a}{\text{INCHES CM}}$	$\frac{b}{\text{INCHES CM}}$	TE_{21}	TE_{41}	TE_{81}	
4 CHANNEL	1	.173 (.439)	.262 (.665)	17.3	-	-	1
	2	.292 (.740)	.450 (1.14)	10.2	20.3	-	2-4
	3	.480 (1.22)	.727 (1.85)	-	12.5	-	4
	4	.757 (1.92)	1.149 (2.92)	-	7.7	15.4	8

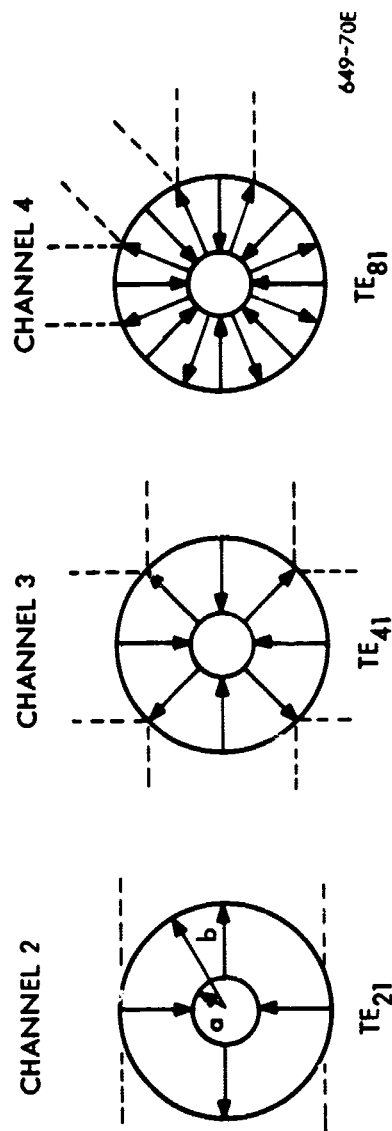


Figure 21. Rotary Joint Channel "Cutoff" Properties

TABLE 5. COMPUTED FOUR-CHANNEL ROTARY JOINT LOSS

Four-Channel				
Channel	Rotary Section Loss (db)	Transition Loss (db)	Feed Network Loss (db)	Total Loss (db)
1	0.052	0.06	-	0.11
2	0.025	0.06	0.06	0.15
3	0.008	0.06	0.11	0.18
4	0.004	0.06	0.22	0.28

* $p = 2.83 \times 10^{-6}$ ohm-cm at 20 °C (aluminum cond.)
t = 250 °F
30 percent increase in attenuation allowed for surface quality

conductors, that the surface quality introduced 30 percent more attenuation than theoretical and that joint average temperature stabilized at a level equal to or less than 250°F. Transition losses at the waveguide to coaxial junction were estimated at 0.03 db and feed power divider losses were computed on the basis of an equivalent length of rigid waveguide. Total loss of the joint ranges from 0.11 db for the innermost channel to 0.28 db for the outermost. Total dissipation within the rotary joint from the single waveguide inputs to the outputs is approximately 151 watts.

Of prime concern is the dissipation of power within the coaxial cylinders of the rotary joint. This is because heat dissipated in this region must be conducted out and radiated to space to maintain safe operating temperatures. While the coaxial diameters have been made as large as possible from an electrical point of view, they are comparatively small in terms of heat transfer conductivity. Dissipation in each cylinder has been computed and summarized in Table 6 for both aluminum and silver conductors when the transition loss is neglected. For aluminum conductors the dissipations range from 0.4 to 8.5 watts with the total being 20.4 watts under the specified 1 KW average power per channel. The total dissipation can be radiated without undue temperature rise within the joint (see Section 4.3.1).

Input VSWR of the rotary joint is estimated to be 1.20:1 maximum for each channel.

2.2.3 Limited Motion Rotary Joints

The motion of the antenna about the roll and yaw axes is governed by the latitude steering requirement (45°NL - 10°SL) and the vehicle stabilization about the roll and yaw axes ($\pm 5^\circ$). The steering requirement translates to an angular motion of approximately ± 4.5 degrees. This motion together with that of vehicle results in total angular motions about the roll and yaw axes of less than ± 10 degrees. Provision for these small movements will be made by flexible waveguide transmission lines about each axis (roll and yaw) for each channel.

The general form of the combined limited motion rotary joint is described in Section 4.1.1. All flexible sections of the four-channel rotary joint are in WR-90 waveguide size with the total length of flexible waveguide for any channel being 12 inches. Interconnection of the yaw and roll flexible sections is made by rigid waveguide components. Note that in this design all angular motion will be about the E-plane of the flexible waveguide.

The attenuation of the flexible waveguide is 0.08 db/foot and therefore the total loss of each two axis limited motion rotary joint, including the rigid waveguide portions will be approximately 0.10 db. VSWR of the limited motion rotary joints will be approximately 1.08:1 maximum.

2.2.4 Diplexer

The two diplexers are used in the auxiliary mode wherein two channels are combined and radiated from a common aperture. The diplexers are inserted in the auxiliary mode by electromechanical switches, as illustrated in the system block diagram Figure 1. To minimize loss in the diplexer, channels 1 and 3 are combined and routed to aperture A-3 and similarly 2 and 4 are combined and radiated from aperture A-2.

TABLE 6. ROTARY JOINT ATTENUATION

FOUR-CHANNEL JOINT

Cylinder ⁴	Dissipation-Watts (1 kW/Channel Input)		Cylinder Length	
	Aluminum ^{1, 3}	Silver ^{2, 3}	(Inches)	(cm)
A	7.1	5.4	3.25	8.25
B	8.5	6.9	2.75	6.45
C	3.1	2.4	1.68	4.26
D	1.3	1.1	1.00	2.54
E	<u>0.4</u>	<u>0.3</u>	0.63	1.60
Total	20.4	16.1		

1) $p = 2.83 \times 10^{-6}$ ohm-cm at 20°C

2) $p = 1.68 \times 10^{-6}$ ohm-cm at 20°C

$t = 250^\circ\text{F}$

3) 30 percent increase in attenuation allowed for surface quality

4) See Figure 2-20 for cylinder location.

This arrangement provides the maximum possible frequency separation between diplexed channels.

The diplexer design requirements are summarized in Table 7. To provide the necessary isolation four cavity sections in each of the two band-pass filters are required. The filters are of the direct coupled type and constructed in silver plated waveguide of WR-90 size. The general physical form of the diplexer is shown in Figure 22. Size of the diplexer is approximately 4 inches by 7 inches by 2 inches.

The diplexer will be tuned to have a 0.01-db Tchebycheff response with a pass band of approximately 78 MHz between the 1.1:1 VSWR points. The insertion loss of the diplexer is dependent upon the ripple bandwidth of the band-pass filters and the unloaded Q of the cavities. The theoretical unloaded Q of silver plated WR-90 is 7345. In practice, approximately 65 percent of this value can be attained. Insertion loss of the band-pass filters is given by:

$$L = \frac{4.343 \sum g_i f_o}{Q BW} \quad (8)^3$$

where

L = insertion loss in db

Q = effective unloaded Q

$\sum g_i f_o$ = summation of low pass element values

BW = ripple value bandwidth in MHz

f_o = center frequency in MHz.

Substituting into equation (8) results in a loss of 0.6 db for each diplexed channel. Hence with isolation requirements of 40 db between channels spaced at four-channel widths, dissipation of 0.6 db will be incurred on a per channel basis.

The drift of the diplexer has been computed to be 28×10^{-6} MHz/MHz- $^{\circ}$ C; this is based upon an aluminum silver plated construction. A temperature rise of 100° F would therefore detune the diplexer by approximately 13 MHz. Since the diplexer has a 1.0 db bandwidth of 78 MHz, this drift can be tolerated without significant increases in loss. Reduction in isolation due to the 13 MHz drift is computed to be 2.6 db. Thus, even with drift of this magnitude, the 40 db isolation requirement can be maintained.

Summary of the overall performance of the diplexer necessary in the auxiliary mode is given in Table 7.

2.2.5 RF Switches

The switches selected to perform the pattern configuration change and auxiliary mode capability are of the electromechanical rotary cylinder type in WR-90 waveguide. A transfer switch of this design is illustrated in Figure 23. The switch rotor consists

**TABLE 7. DIPLEXER DESIGN REQUIREMENTS
AND COMPUTED PERFORMANCE**

Parameter	Requirement	Computed Performance
VSWR at f_1^* , f_3^*	1.20:1 max.	1.10:1 typical
Insertion Loss at f_1	0.5 db max.	0.6 db
Insertion Loss at f_3	0.5 db max.	0.6 db
Bandwidth at 1.0 db points	45 MHz - both channels	78 MHz - both channels
Isolation - over full channel width	40 db min.	43 db
Phase Linearity	$\pm 1.0^\circ$ max.	$\pm 0.45^\circ$

$$^*f_1 = 12.05 \text{ GHz}; \quad f_3 = 12.35 \text{ GHz}$$

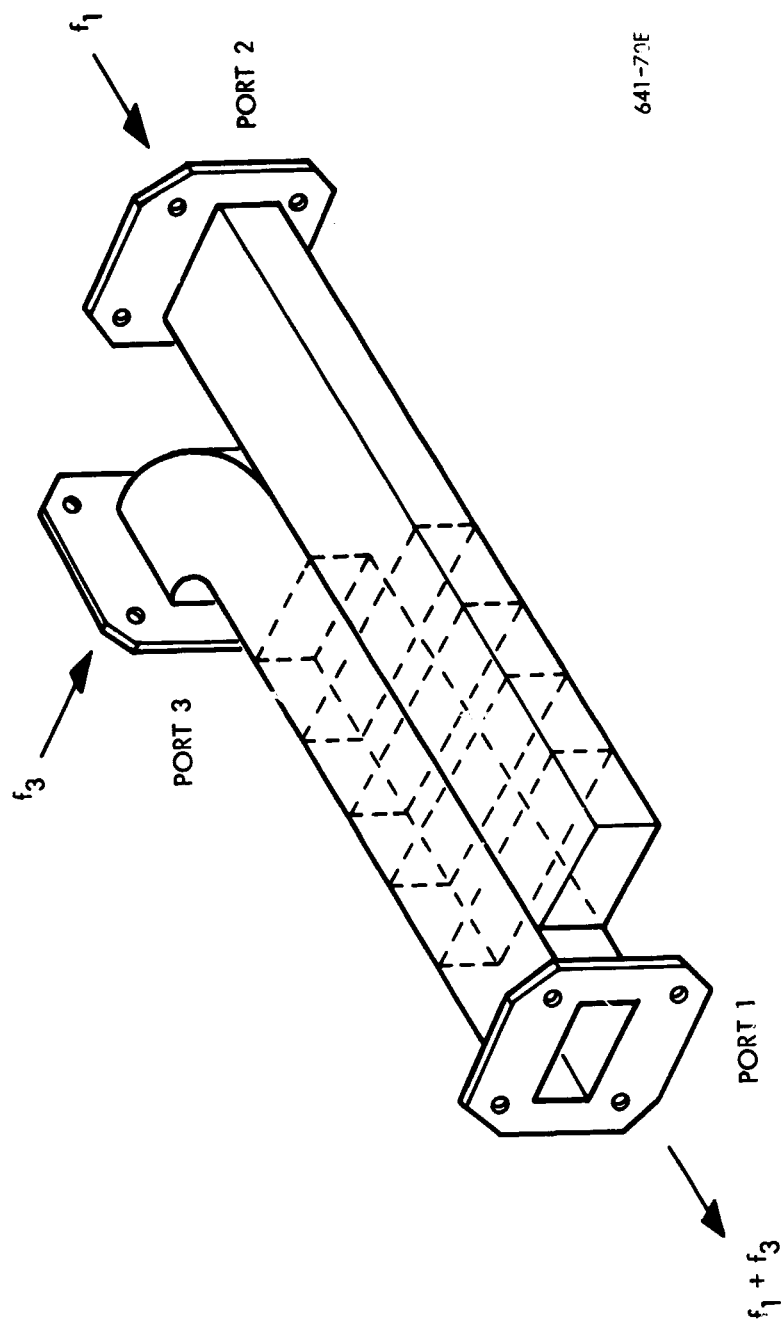


Figure 22. Diplexer Configuration

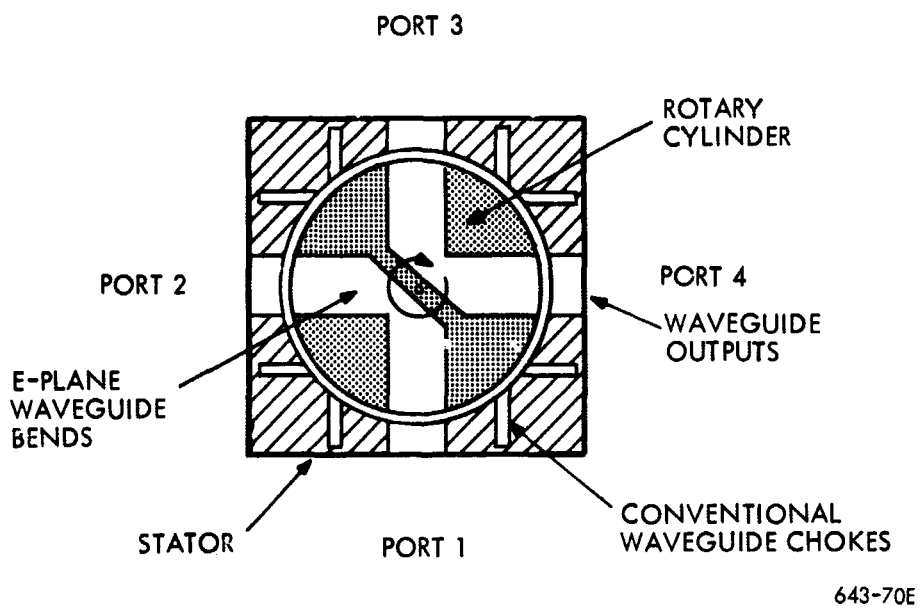


Figure 23. Rotary Cylinder RF Transfer Switch Configuration

of a metal cylinder which has machined into it two E-plane waveguide mitre bends. These bends are arranged so that the terminals are located 90 degrees apart on the surface of the rotor. The rotor is inserted in a cylinder which has four waveguides also spaced at 90-degree intervals about its surface. When the rotor and stator waveguides are aligned, two waveguide paths are available in the switch. Indexing of the rotor by 90 degrees provides the transfer of input and output paths.

The rotor and cylinder are non-contacting and isolation between channels is provided by chokes located on the stator section about each waveguide input (shown in Figure 23). The performance of the switches is summarized in Figure 24.

Indexing of the switch will be accomplished by either a motor or solenoid drive latching system whereby no holding power is required to keep the switch in the desired position. The switch will utilize VacKote lubricated ball bearings with lubricant reservoirs arranged such that lubricant lost via vaporization will not be deposited within the waveguide, but will be directly lost to space.

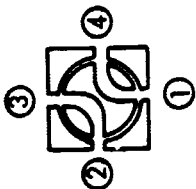
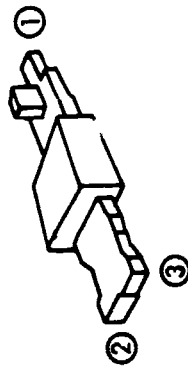
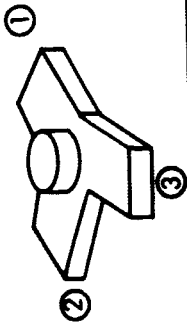
2.2.6 Polarizer

The circular polarizer used in the antenna system is an all metal periodic loaded "pin" type in circular waveguide. This polarizer is located at the throat of the horn-reflector antenna. Input into the polarizer is in WR-90 waveguide. A two-step rectangular to circular waveguide transition is used to couple the rectangular input to the circular guide. Proper field orientation of the electric field relative to the "pins" of the polarizer is achieved by physical orientation of the transition relative to the circular guide section. The polarizer and input transition section are illustrated in Figure 25.

A periodic-type all metal polarizer was chosen for the conceptual design for the following reasons:

1. The metal construction minimizes outgassing problems, particularly in comparison to designs using dielectrics.
2. Average power limitations are minimized due to the fact that lower loss can be achieved than in dielectric designs.
3. Lower polarization axial ratio is obtainable over a greater bandwidth than in a dielectric design.
4. Low input VSWR is easily achieved.

The polarizer shown in Figure 25 will be in a $0.72\lambda_0$ diameter circular waveguide where λ_0 is the band center free-space wavelength. The periodic loading is provided by thin metal pins which are equally spaced along the longitudinal axis of the waveguide. Over the bandwidth of the device (typically 1.14:1), the pins provide near constant differential phase shift between the orthogonally applied fields. Ten pairs of pins spaced $0.266\lambda_0$ apart are used to provide the necessary 90 degrees differential phase shift.

SWITCH TYPE	AVERAGE POWER CAPABILITY (WATTS)	APPROX WEIGHT	LOSS (DB)	COOLING (TERESTIAL APPLICATION AT 1 KW)	NUMBER REQUIRED PER SYSTEM	ISOLATION (DB)
MECHANICAL - ROTARY CYLINDER 	> 1000 (WAVEGUIDE CAPABILITY)	POUNDS Kgm 2.5 (1.13)	.05	NOT REQD.	5	60
SWITCHED DIFFERENTIAL PHASE SHIFT CIRCULATOR ¹ 	≈ 700	3 (1.36)	0.6	REQD.	5	20-30
SWITCHED JUNCTION CIRCULATOR ² 	750	2 (0.91)	0.6	REQD.	7	20

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1) AFTER R.A. STERN & J.P. AGRIOS IEEE TRANS. MTT, DEC 1968

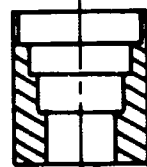
2) EXTRAPOLATED FROM A 15.5 GHz DESIGN

Figure 24. RF Switch Comparison and Performance

σ_1	$.087 \lambda_0$
σ_2	$.108 \lambda_0$
σ_3	$.121 \lambda_0$
σ_4	$.127 \lambda_0$
σ_5	$.130 \lambda_0$
l	$.266 \lambda_0$
d	$.060 \lambda_0$

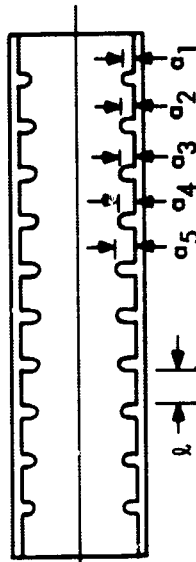
λ_0 = FREE SPACE
WAVELENGTH AT
BAND CENTER

INPUT TRANSITION
ORTHOMODE
TRANSDUCER* (OMT)

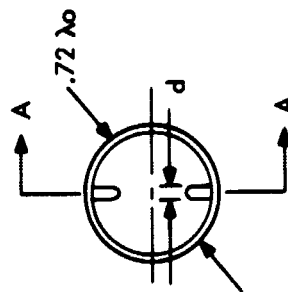


WR-90
INPUT

POLARIZER



SECTION A-A



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* OMT ROTATED 45° ABOUT Z-Z' FROM
ITS NORMAL POSITION

Figure 25. Circular Polarizer and Input Transition Configuration

The main feature of the polarizer is that the heights of the irises are adjusted to provide a cosine reflection coefficient taper. This technique results in a polarizer input reflection coefficient which in form is analogous to the far-field amplitude pattern of an array with a cosine amplitude taper. Spacing between pins is analogous to spacing between the elements in the array. Hence, by appropriate spacing of the pins, the input reflection coefficient of the polarizer can be adjusted to that region analogous to the low sidelobe level region of the array. This approach permits a low VSWR to be attained across the desired bandwidth simultaneously with low polarization axial ratio.

The polarization axial ratio performance of the periodic loaded design is shown in Figure 26. This performance was measured on a scaled unit operating in the 7-8 GHz range. Maximum input VSWR of the device is 1.07:1 over a 1.14:1 bandwidth. VSWR of the companion rectangular to circular waveguide transition is also 1.07:1 over the identical bandwidth. Insertion loss of the combined transition and polarizer is 0.04 db. The overall length of the polarizer and input transition is approximately 3.7 inches (9.4 cm). Construction of the polarizer is anticipated to be of aluminum.

2.2.7 Transmission Lines

The selected transmission line for the main run of the conceptual design is rigid WR-90 aluminum waveguide. This guide exhibits a loss of only 0.04 db/ft. at 12.2 GHz. Since approximately 10 feet of this line is used, it contributes 0.4 db of total loss. On a dissipation basis, this amounts to 90 watts per line per kilowatt. Weight of this line is 0.542 pound per foot or 5.42 total pounds for each run.

The use of oxygen-free copper (OFC) waveguide would reduce the loss to 0.033 db/ft. On a dissipation basis this is equivalent to 75 watts per line per kilowatt. While this is 15 watts less than the aluminum line, it is achieved at the expense of a weight increase. The 10 foot OFC runs would each weigh 17.9 pounds. This small improvement in dissipation at the expense of such a large increase in weight makes the use of OFC waveguide unattractive.

2.3 ALTERNATE COMPONENTS

2.3.1 Apertures

2.3.1.1 Dual-Beam Horn-Reflector

The horn-reflector antenna when operating linearly polarized radiates a beam whose axis is parallel to the focal axis of the parabolic section. When circularly polarized however, the beam is displaced in the ϕ plane (see Figure 13) relative to the focal axis.³ The sense of displacement ($\pm\Delta\phi$) is dependent upon the sense of circular polarization used. Hence, if a horn-reflector is operated in right hand circular polarization (RHCP) and left hand circular polarization (LHCP) simultaneously, two independent beams, angularly displaced from each other, are available. If sufficient displacement between beams could be obtained, i.e., 0.55 HPBW, a common aperture would provide a two channel capability. This technique would simplify the antenna system in that only two apertures would be required to provide four beams.

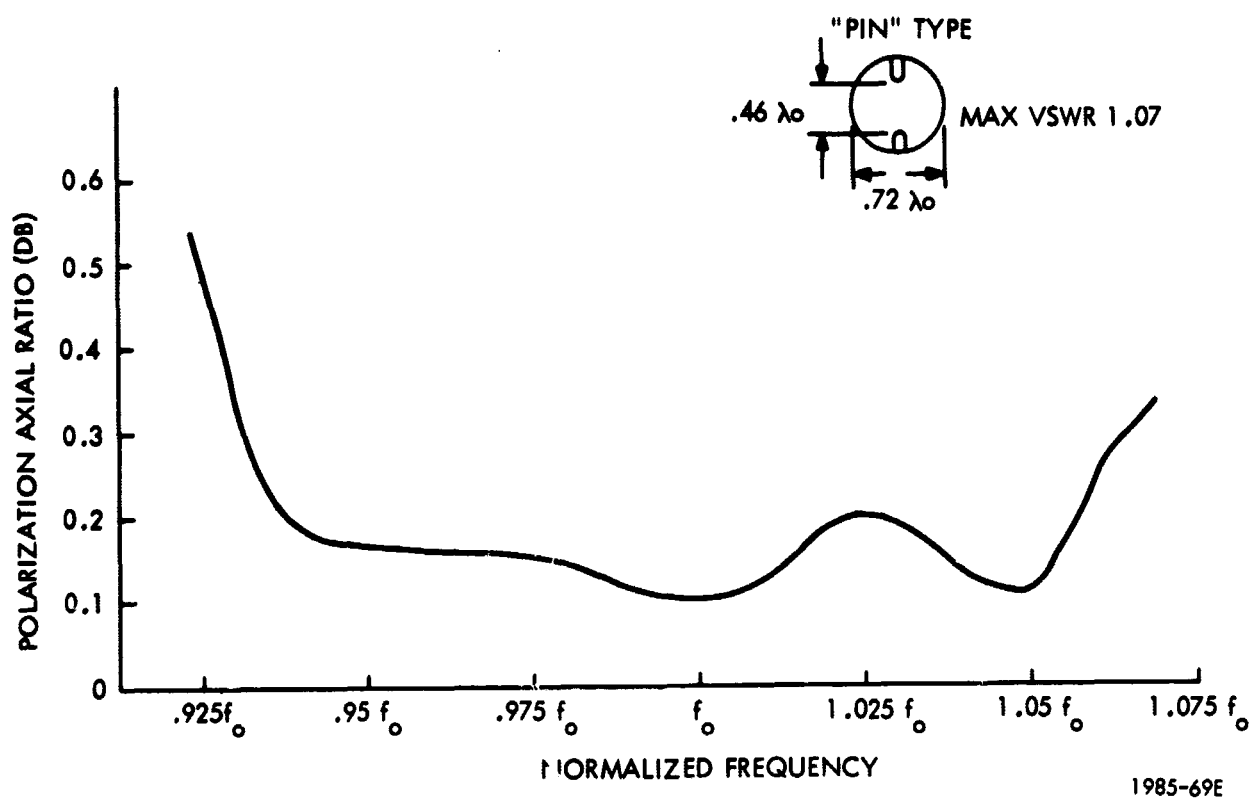


Figure 26. Circular Polarizer Performance

The polarization sensitive beam displacement of the horn-reflector is a function of the ratio of desired to cross-polarized radiation that occurs when operating linearly polarized. The cross-polarized component which occurs only in the ϕ -plane is a result of the inherent space tapering that occurs in the θ -plane on the parabolic surface. The space taper is dependent upon the flare angle (α) of the horn-reflector. Thus, it is necessary to know beam displacement as a function of flare angle to determine if a single aperture can provide simultaneous dual beams spaced 0.55 HPBW apart.

To simplify computations the conical horn-reflector was approximated by a pyramidal type with an aperture sized to provide a nominal 2.7 degrees HPBW. Using the notation in Figure 27 (and the approach formulated by Crawford et. al.³) the radiated field components for linear polarization are:

$$E_1 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \alpha}{2\alpha_0} \cos \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \cos \alpha \sin \theta \ell} d\phi d\alpha$$

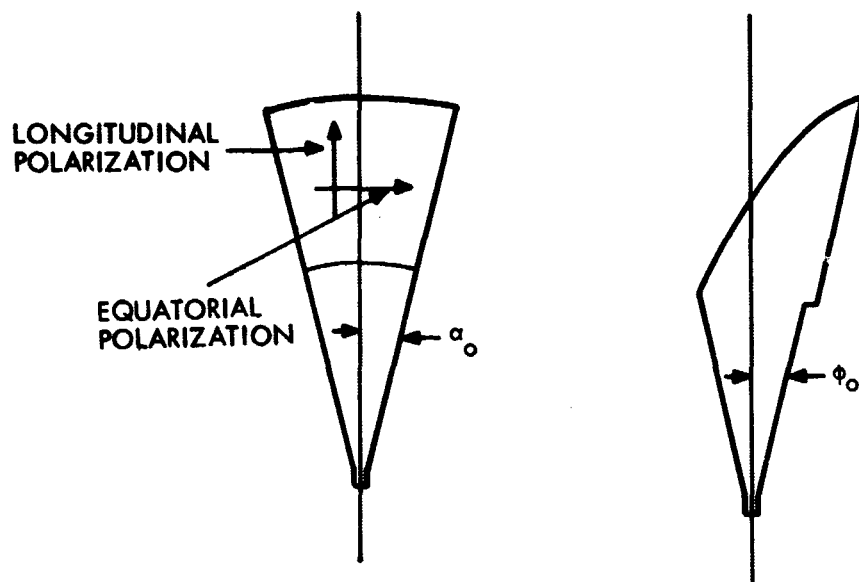
$$E_2 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \phi}{2\phi_0} \cos \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \cos \alpha \sin \theta \ell} d\phi d\alpha$$

$$E_3 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \alpha}{2\alpha_0} \cos \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \sin \alpha \sin \theta \ell} d\phi d\alpha$$

$$E_4 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \phi}{2\phi_0} \cos \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \sin \alpha \sin \theta \ell} d\phi d\alpha$$

$$E_5 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \alpha}{2\alpha_0} \sin \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \sin \alpha \sin \theta \ell} d\phi d\alpha$$

$$E_6 = j \frac{4f^2}{\lambda R} E_0 \int_{-\alpha_0}^{+\alpha_0} \int_{-\phi_0}^{+\phi_0} \cos \frac{\pi \phi}{2\phi_0} \sin \alpha \frac{\cos \phi}{1 - \sin \phi} e^{-j\beta \rho \cos \phi \sin \alpha \sin \theta \ell} d\phi d\alpha$$



POLARIZATION EXCITATION	MEASUREMENT PLANE	FAR FIELD POLARIZATION AND DESIGNATION
LONGITUDINAL	LONGITUDINAL	LONGITUDINAL - E_1
EQUATORIAL	LONGITUDINAL	EQUATORIAL - E_2
LONGITUDINAL	EQUATORIAL	LONGITUDINAL - E_3
EQUATORIAL	EQUATORIAL	EQUATORIAL - E_4
LONGITUDINAL	EQUATORIAL	EQUATORIAL - E_5
EQUATORIAL	EQUATORIAL	LONGITUDINAL - E_6

650-70E

Figure 27. Pyramidal Horn-Reflector Description and Field Notiation

where

f = focal length

λ = wavelength

R = distance to far field point

E_o = relative field amplitude

$\beta = 2\pi/\lambda$

$\rho = \frac{2f}{1-\sin\phi}$

Circularly polarized response in the equatorial plane is given by:

$$E_{er} = K \{ (E_3 + E_4) - (E_5 + E_6) \} \quad (9)$$

$$E_{el} = K \{ (E_3 - E_4) + (E_5 - E_6) \} \quad (10)$$

Using an aperture extent of approximately 24 inches, radiation patterns of the linear components E_1 through E_6 were computed at 12.2 GHz for equal α_o and ϕ_o half flare angles ranging from 14 to 55 degrees. Circularly polarized patterns were computed from the linear components using Equations 9 and 10. The polarization sensitive beam displacement as a function of flare angle ($\alpha_o = \phi_o$) was determined directly from the computed circularly polarized patterns.

The beam displacement characteristic as a function of flare angle is summarized in Figure 28. A beam displacement of 0.275 HPBW for a single beam is attainable when using a flare angle of 55 degrees. Hence a 0.55-HPBW spacing between beams could be achieved from a single aperture. The overall form and size that such a horn-reflector would take is illustrated in Figure 29.

While adequate beam displacement can be achieved using the above techniques it is obtained at the expense of considerable deterioration in the radiation patterns. This is illustrated in Figures 30 and 31 which show the circularly polarized radiation patterns in the equatorial and longitudinal planes, respectively, when the flare angle is the necessary 55 degrees. In the equatorial plane, the desired circular polarization (arbitrarily defined as RHCP) provides low sidelobes (-34 db), however considerable pattern asymmetry exists in the normal sidelobe level regions. This is illustrated by the dotted line pattern which shows a comparison of the left and right hand portions of the pattern. Note that while no sidelobes exist on the $+\phi$ or right hand side of the pattern, the levels are considerably higher than at the corresponding points in the $-\phi$ region. Thus, while no sidelobes exist in the $+\phi$ region, radiation levels are comparable to only 20 db down from the beam peak and are equivalent to sidelobes at this level.

In addition to the deleterious effect on symmetry, the use of a large flare angle results in an asymmetrical cross-polarized component of considerable magnitude.

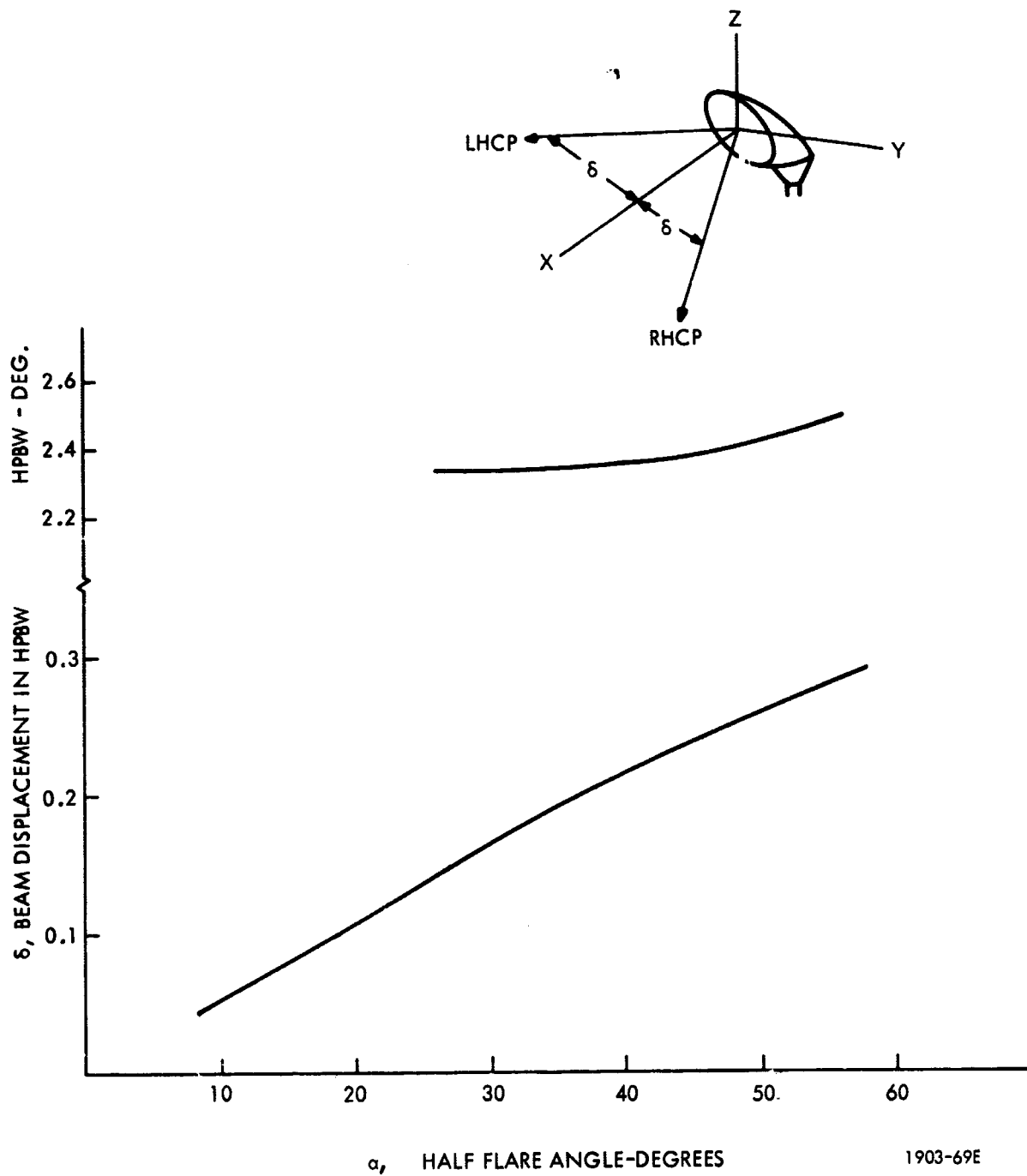


Figure 28. Horn-Reflector Beam Displacement and Beamwidth vs Flare Angle

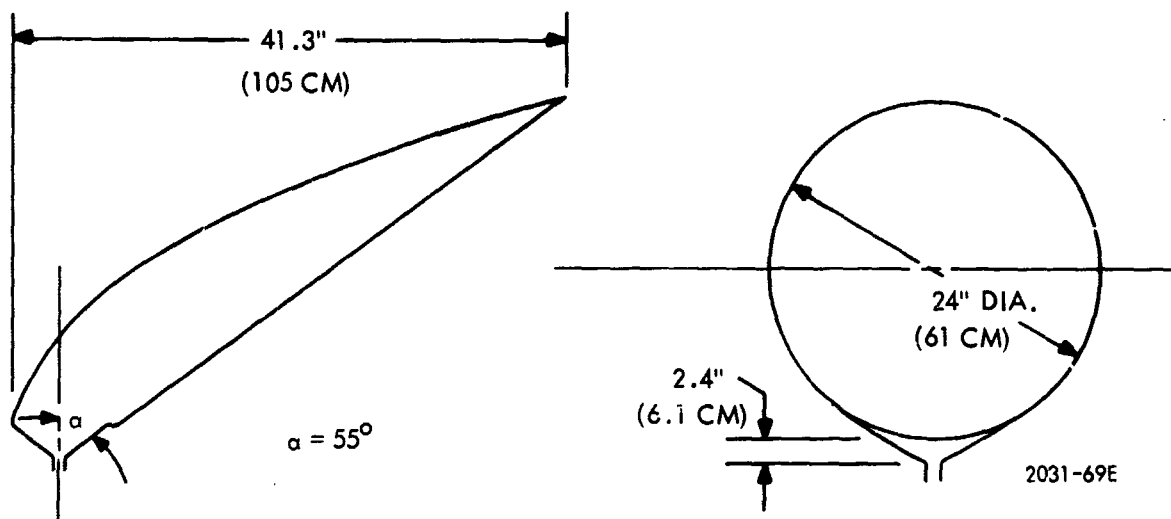


Figure 29. Dual-Beam Horn-Reflector Dimensional Configuration

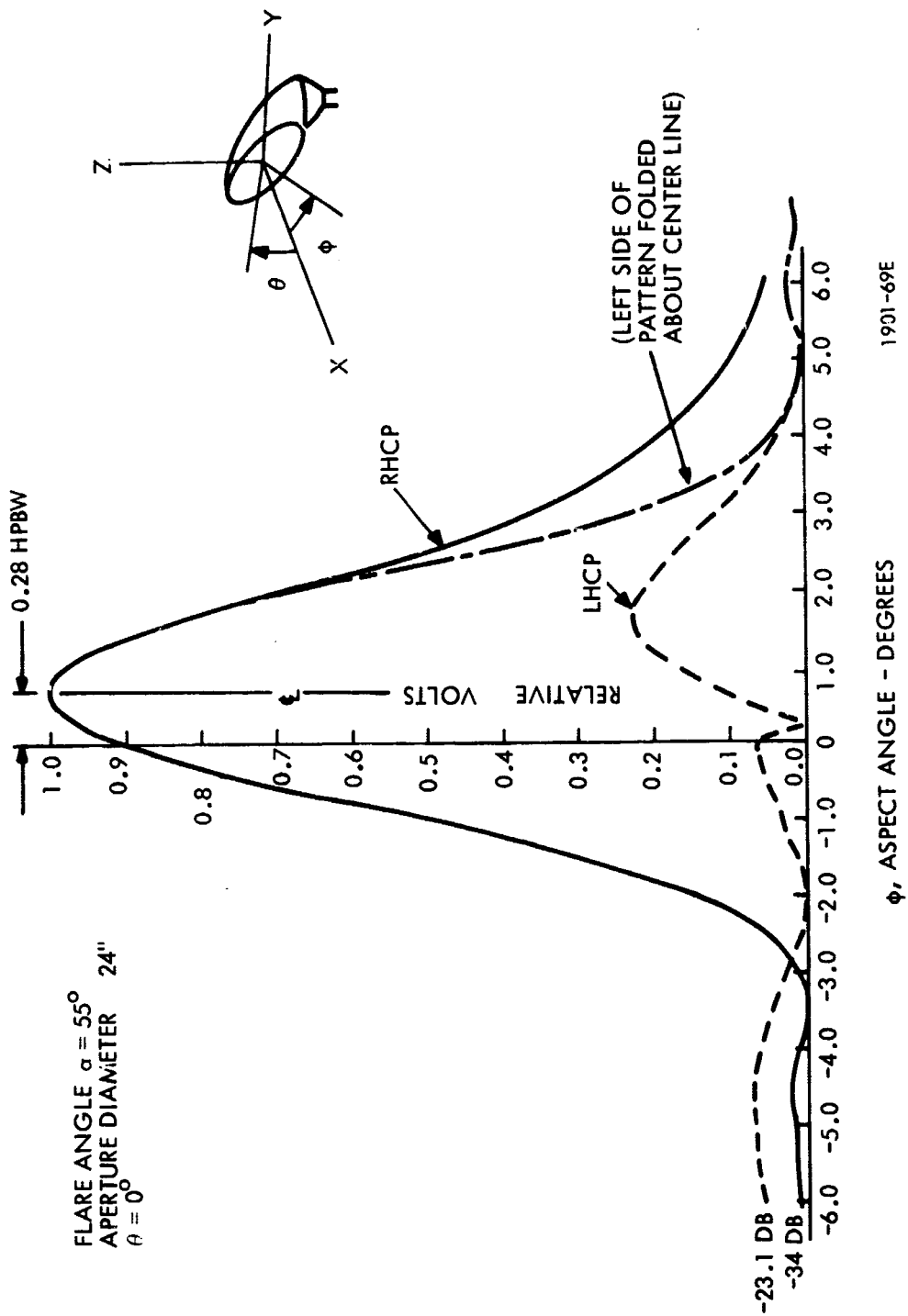


Figure 30. Dual-Beam Horn-Reflector Radiation Pattern - Equatorial Plane

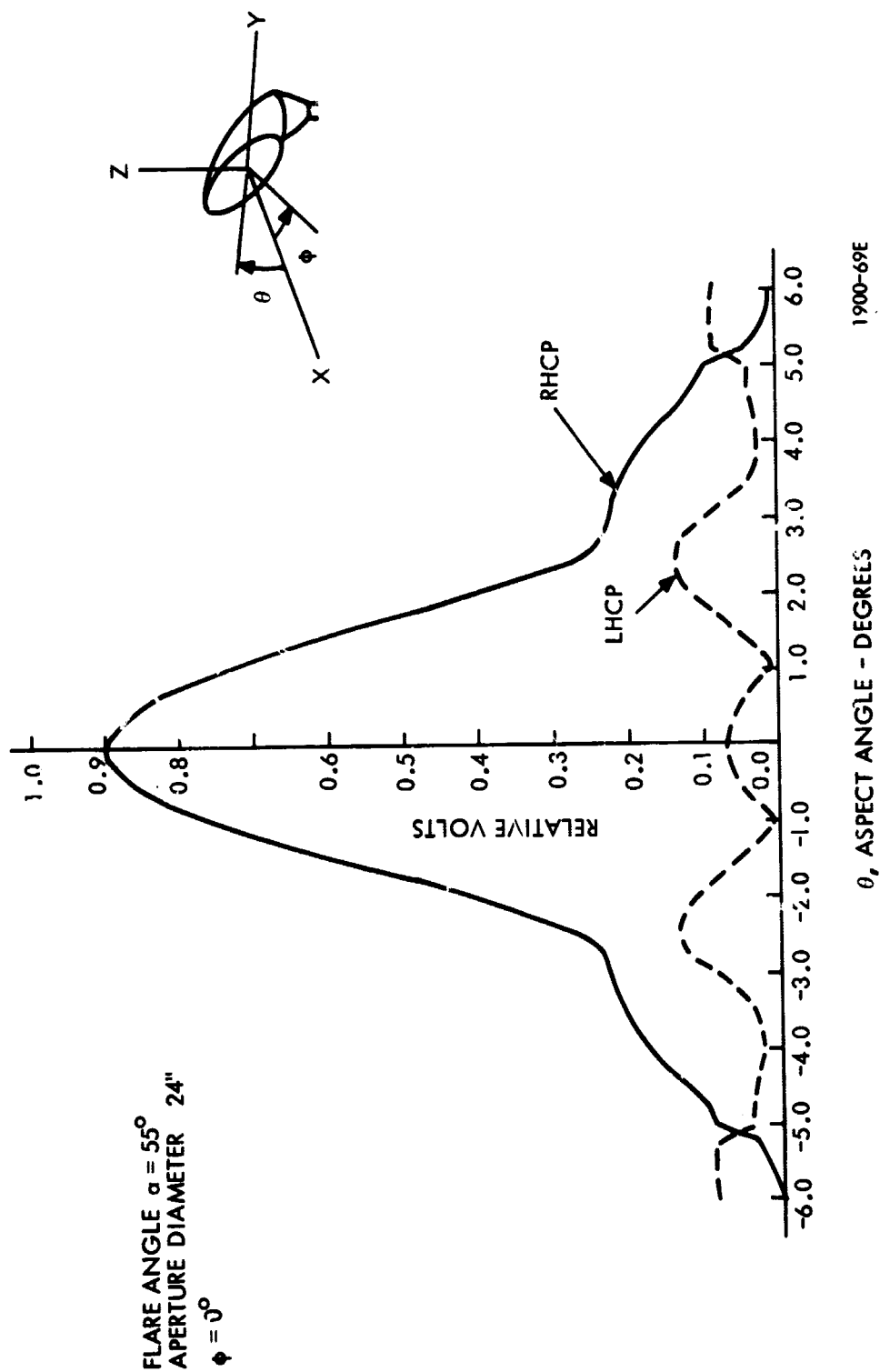


Figure 31. Dual-Beam Horn-Reflector Radiation Pattern - Longitudinal Plane

This is illustrated in Figure 30. The peak cross-polarized level is only 13 db down and falls within the main beam. This value of cross-polarized component is commensurate with a maximum 4.8 db axial ratio within the HPBW. Hence large flare angle designs are not capable of providing polarization axial ratios below the required maximum (0.42 db).

Performance of the dual-beam horn-reflector in the longitudinal plane is also comparatively poor, as illustrated in Figure 31. In the desired polarization (RHCP) considerable shouldering occurs at a 13 db level down from the main beam. This is a consequence of large amplitude taper that is incurred with the large flare angle. The taper introduces large amplitude difference between the longitudinal and equatorial polarized patterns in the longitudinal plane. When these two field components are combined, the broader pattern fills in the null of the narrower giving rise to the shoulder effect. Cross-polarized levels in this plane are also large (-17 db) and at angles which are commensurate with those of the sidelobes for the desired polarizations.

The use of multi-mode excitation of the type discussed previously (Section 2.2.1) in the dual-beam horn-reflector to improve beam symmetry and sidelobe levels was considered. Analysis of the scale model multi-mode horn-reflector performance shows that introduction of TM_{11} mode at a level that provides maximum sidelobe reduction causes a 12.5 percent reduction in beam displacement. This reduction value is for a flare angle of 26.5 degrees and is a result of the fact that the TM_{11} mode reduces the cross-polarized components that cause the beam displacement.

Assuming that the same percentage reduction exists at flare angles in the vicinity of 55 degrees, a maximum displacement of only 0.24 HPBW is available. Use of larger flare angles to compensate for this reduction does not appear feasible in that the beam displacement versus flare angle function approaches zero slope in the vicinity of 55 to 60 degrees (see Figure 28).

Because of the above mentioned difficulties, the use of the dual-beam horn-reflector is not feasible for 0.55 HPBW total displacement between beams unless considerable increase in sidelobes and axial ratios is tolerated. If, however, applications exist where a total displacement of about 0.3 HPBW is required this technique is attractive. For such spacings the required -25 db sidelobe level and 0.42 db axial ratio can be achieved as demonstrated in the experimental work conducted on the horn-reflector (see Section 2.2.1).

2.3.1.2 Focal Point Fed Parabola

Consideration was given to a focal point fed parabola for the system antenna because of its electrical simplicity, light weight, and small occupied volume. Only single-beam apertures were considered in that the small spacing between beams, i.e., either 0.55 or 1.1 HPBWs, preclude the use of dual feeds in a single parabola.

The 0.55-HPBW spacing assumes that immediately adjacent beams are derived from the same aperture while the 1.10 HPBW value assumes that alternate beams in the coplanar configuration are obtained from the same aperture.

The general form of the focal point fed parabola is illustrated in Figure 32. A comparatively large feed is used to illuminate the reflector with a large edge taper to minimize sidelobe level. An edge taper in the range of 14 to 16 db is a practical limit in order that sidelobe energy is not placed upon the reflector. Additionally, the feedline/spars shown in Figure 32 represent what is considered to be a minimum for the support of the feed and routing RF energy to the feed. The RF transmission line, since it is WR-90 waveguide, represents the 0.5-inch width of the structure.

A form of illumination which can be obtained with a conventional feed while providing a 14-db edge taper is from the work of Sciambi⁴.

$$E(r) = A + B [1 - (r/a)^2]^2 \quad (11)$$

where

$E(r)$ = normalized aperture field distribution in volts

A = edge illumination amplitude in volts

$$A + B = 1$$

r = radial distance from the focal axis

a = reflector radius

For the case being discussed, A equals 0.2 giving the 14-db edge taper. The unperturbed sidelobe level that this illumination function provides is 31.7 db. Also for this distribution, the HPBW and aperture diameter are related by

$$\theta = \frac{70.4}{D_\lambda}$$

where

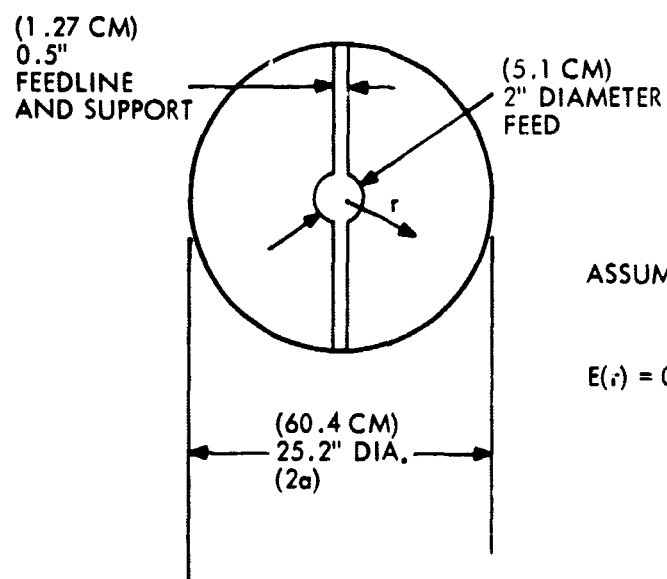
θ = half power beamwidth

D_λ = aperture diameter in wavelengths

For the required beamwidth of 2.7 degrees the aperture diameter is 26λ or 25.2 (60.4 cm) inches at 12.2 GHz. The sidelobe level that this aperture provides can be determined in the manner developed by Gray.⁵ Using this work and Figure 32 we have

$$\frac{E_d}{E_D} = 2 \left(\frac{d}{D} \right)^2 = 2 \left(\frac{2}{25.2} \right)^2 = 0.0126 \text{ volt}$$

$$\frac{E_s}{E_D} \approx 1.44 \frac{W}{D} = 1.44 \left(\frac{0.5}{25.2} \right) = 0.0286 \text{ volt}$$



ASSUMED FIELD DISTRIBUTION

$$E(r) = 0.2 + 0.8 \left[1 - \left(\frac{r}{a} \right)^2 \right]^2$$

644-70E

Figure 32. Focal Point Fed Parabola Blockage Condition

where

- E_d = a voltage proportional to the feed blocking area
- E_D = a voltage proportional to the unperturbed aperture area
- E_s = a voltage proportional to the start blocking area.

The sidelobe level resulting from all blocking is

$$S_l = 20 \log \left\{ \frac{1 - \frac{E_d}{E_D} - \frac{E_s}{E_D}}{\frac{E_d}{E_D} + \frac{E_s}{E_D} + E_\mu} \right\} \quad (12)$$

where

- S_l = sidelobe level in db
- E_μ = the unperturbed sidelobe level (in this case 0.026).

Substitution in Equation 12 gives a sidelobe level of -23.1 db. Thus, the use of a conventional type parabola is not capable of providing the required -25 db sidelobe level. While some improvement could be made through further illumination tapering, it is judged that the sidelobe performance would be marginal. In particular, the desired sidelobe level of -35 db certainly could not be approached.

The use of auxiliary radiators to reduce sidelobe levels was treated briefly and rejected as being unfeasible in this application. This judgement is based primarily on the sensitivity of such systems to thermal distortions.

Movement of the auxiliary radiators relative to the main aperture will upset the necessary main-to-auxiliary aperture phase relations necessary to affect sidelobe cancellation. Similarly, power differential between the transmission lines of the main and auxiliary antennas will cause differential phase shift between them. This phase difference is due to the temperature differences between the lines and will impair blockage compensation in a manner identical to relative main-to-auxiliary aperture movement. In addition, the power lost via the auxiliary radiator reduces the gain available in the communications beam. This loss is typically about 0.2 db. Thus, the loss, thermal sensitivity, and general complexity associated with auxiliary radiators make their use unattractive for blockage compensation.

The efficiency of the focal point fed parabola system is inferior to the single-beam horn-reflector. This is due predominantly to spillover, blocking, and feed phase error loss. Table 8 tabulates the latter efficiencies for a parabolic system using a feed

that provides the $0.2 + 0.8 \left[1 - \left(\frac{r}{a} \right)^2 \right]^2$ illumination and a horn-reflector. Comparison

of these efficiency factors shows that a horn-reflector can provide approximately a 0.7-db higher gain than the focal point fed parabolic system.

TABLE 8. ANTENNA EFFICIENCY COMPARISON

		$\eta_t = \text{Total Antenna Efficiency}$			
		$\eta_t = \eta_a \times \eta_s \times \eta_b \times \eta_{p1} \times \eta_{p2} \times \eta_c \times \eta_d$			
				Parabola 14-db Edge Taper 3% Blockage	Horn-Reflector
Efficiency Factor					
η_a	Aperture	-	-	-	-
η_s	Spillover	0.89			1.0
η_b	Blockage	0.97			1.0
η_{p1}	Reflector Phase Error	-			-
η_{p2}	Feed Phase Error	0.98			1.0
η_c	Cross Polarization	-			-
η_d	Dissipation	-			-

Gain Differential* = Horn Ref Gain - Parabola Gain

$$\text{Gain Differential} = 10 \log \left(\frac{1}{0.89 \times 0.97 \times 0.98} \right) = 0.7 \text{ db}$$

* At equal half-power beamwidths

In summary, focal point fed parabolic systems do not provide the necessary sidelobe levels and are somewhat lower in gain than horn-reflectors. Blockage compensation in the parabola is complex, results in further power loss, and is thermally sensitive. Thus, because of these characteristics the use of the point fed parabola in the antenna system is not recommended.

2.3.1.3 Planar Array

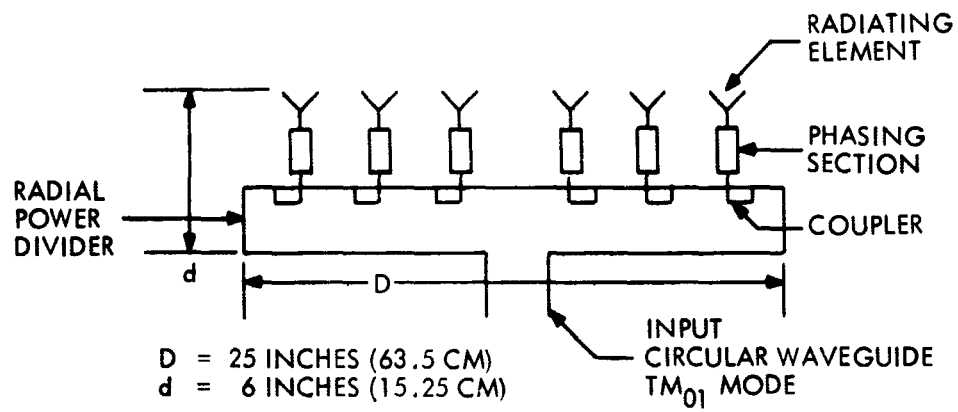
Planar arrays formed from a series of waveguide slot line sources provide a compact (i.e., flat) antenna for this application. Control of the illumination of this type of array is difficult, however, and it is unlikely that sidelobe levels of 25 db down or greater would be attainable.

A new type of planar array, recently developed, circumvents the major problem of illumination control associated with the waveguide slot planar array. This antenna is illustrated in Figure 33. Its main section consists of a radial transmission line driven in the TEM mode. This excitation is achieved by means of a TM_{01} circular waveguide located at the center of the radial transmission line. Individual antenna elements of the array are coupled to the radial transmission lines by directional couplers, hence, the excitation amplitude of each element can be independently and precisely controlled to provide the desired amplitude distribution. In practice, elements are located in concentric rings with the coupling for all elements in a given ring being constant.

Since path length differences exist between the input (TM_{01} circular guide) and the various elements due to the radial line, individual phase shifters are required in each element channel to affect an equi-phase aperture. Either waveguide dielectric vane or strip transmission line phase shifters can be used; however, the latter are more convenient in that they mate easily with the coaxial coupler probes. Use of strip transmission line phase shifters introduces approximately 1 db of loss. Waveguide phase shifters could reduce this value somewhat, however, a total phase shifter loss of less than 0.5 db is not attainable.

The planar array aperture can be designed with an aperture illumination function approximating a Taylor aperture distribution. Using a Taylor distribution with $\bar{n} = 5$, a 25-inch diameter planar aperture would provide a 2.7-degree HPBW, a peak sidelobe level of 37.4 db, and a directivity of 36.8 db. The array would use approximately 900 elements with the element spacing being nominally 0.625λ . Dissipative loss in the radial line, strip transmission line phasing sections, and polarizers is estimated at 1.0 db. Available peak gain from the planar array is, therefore, 35.8 db. Weight of a single planar array would be approximately 35 pounds.

While the planar array exhibits an excellent form factor [25-inch (63.5 cm) diameter and 6-inch (15.25 cm) total depth] and minimum sidelobe level (≈ 35 db down), it does exhibit high dissipative loss, primarily due to the phase shift sections. The 1.0 db loss represents a 206-watt dissipation when operating at 1 kW input. This large dissipated power occurs in a tightly packaged volume and thus represents a difficult thermal design problem. Because of this, the planar array was rejected for use in the conceptual design. Reconsideration of the planar array should be made if system power levels are reduced substantially below 1 kW. Here the excellent sidelobe properties would be available and the thermal difficulties considerably less severe.



2005-69E

Figure 33. Planar Array Antenna Configuration

2.3.2 Principal Rotary Joint - Alternate

The alternative design approach to the four-channel 360-degree per day rotary joint was to use a two-channel design with diplexers at the input and output. The diplexers, which are identical to those described in Section 2.2.4, permit two communications channels to be routed through a single channel of the rotary joint. The system block diagram using this approach is indicated in Figure 34. Operation of this system is basically the same as that described in Section 2.1 with the exception of the routing of signals about the 360-degree per day interface.

The prime reason for consideration of a two-channel joint was to reduce dissipative loss, thereby minimizing the heat transfer problems within the device. Reduction in the number of channels in the joint permits the use of larger diameter conductors and simpler feed networks, resulting in a reduced dissipative loss. Of course, the reduction in loss in the two channel joint is offset by the loss occurring in the diplexing networks necessary to provide four-channel communications capability. Thus, the two-channel joint/diplexer approach permits a trade as to where the dissipation occurs relative to that of a four-channel joint design. Superiority of approaches is based upon the total dissipation that occurs and the ability to radiate the associated heat to space while maintaining safe equipment operating temperatures.

The two-channel rotary joint design is shown in Figure 35. The center channel of the joint operates in the radially symmetric TM_{01} circular waveguide mode while the outer channel is in the TEM mode. Input and output from the TM_{01} section is made with single rectangular waveguides. The TEM section, since it can propagate higher order modes, is driven by four equiphase waveguides at both its input and output. This form of excitation is identical to that described for the four-channel joint. Hybrid power dividers are used to feed each group of four guide feed assemblies. The TM_{01} joint routes signal channels 2 and 4 while the TEM section accommodate channels 1 and 3. Choke designs in the joint are similar to those used in the four-channel design. The TM_{01} section operates at approximately 15 percent above cutoff. The TEM section has a characteristic impedance of 25 ohms. This impedance level is a compromise between susceptibility of the TEM section to higher mode propagation and ease of impedance matching to the waveguide inputs.

Loss performance and power dissipation within the joint are summarized in Table 9. Maximum total loss in the TM_{01} and TEM sections are 0.07 and 0.17 db, respectively. Total dissipation within the rotary section of an aluminum unit under 4 kW total input (1 kW per channel) is 4.5 watts, which is approximately one-fourth of that in the four-channel rotary joint. This lower dissipation, together with the fact that large conductors are used, minimize the thermal design problems. VSWR of the joint is expected to be 1.2:1 maximum.

While this joint provides lower dissipation, its use requires two diplexers per each joint channel. For the communications channel isolation and spacing requirements (40 db and 200 MHz, respectively) the diplexers each introduce a 0.6-db loss. This represents a total dissipation of 130 watts per kilowatt input for each diplexer. Under normal conditions this gives a total dissipation of 1040 watts via the diplexers when the four communications channels are operating at their 1 kW levels. Analysis of the thermal problem associated with the diplexers (Section 4.3.1) has shown that it is more

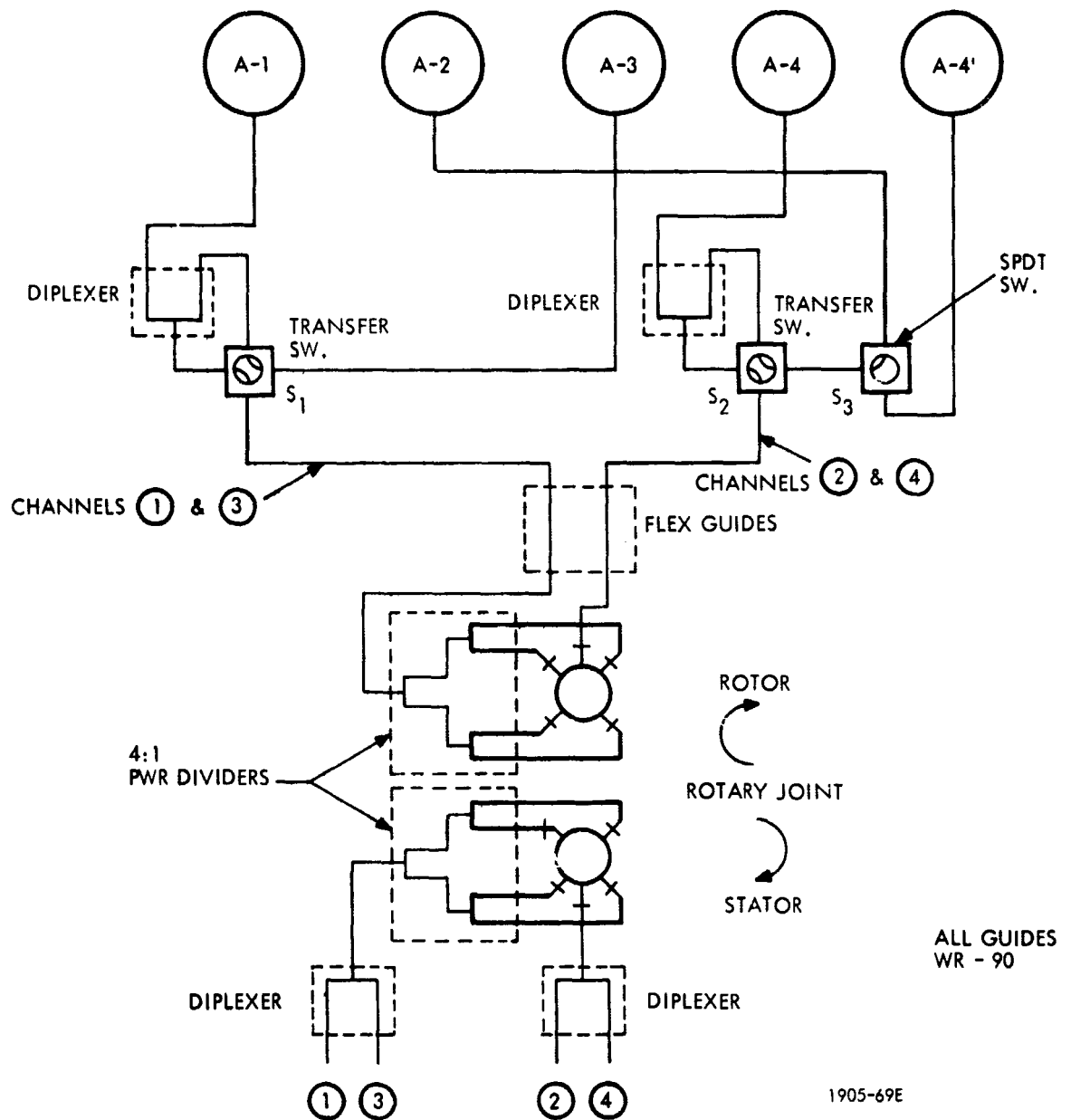


Figure 34. RF System Block Diagram - Alternate

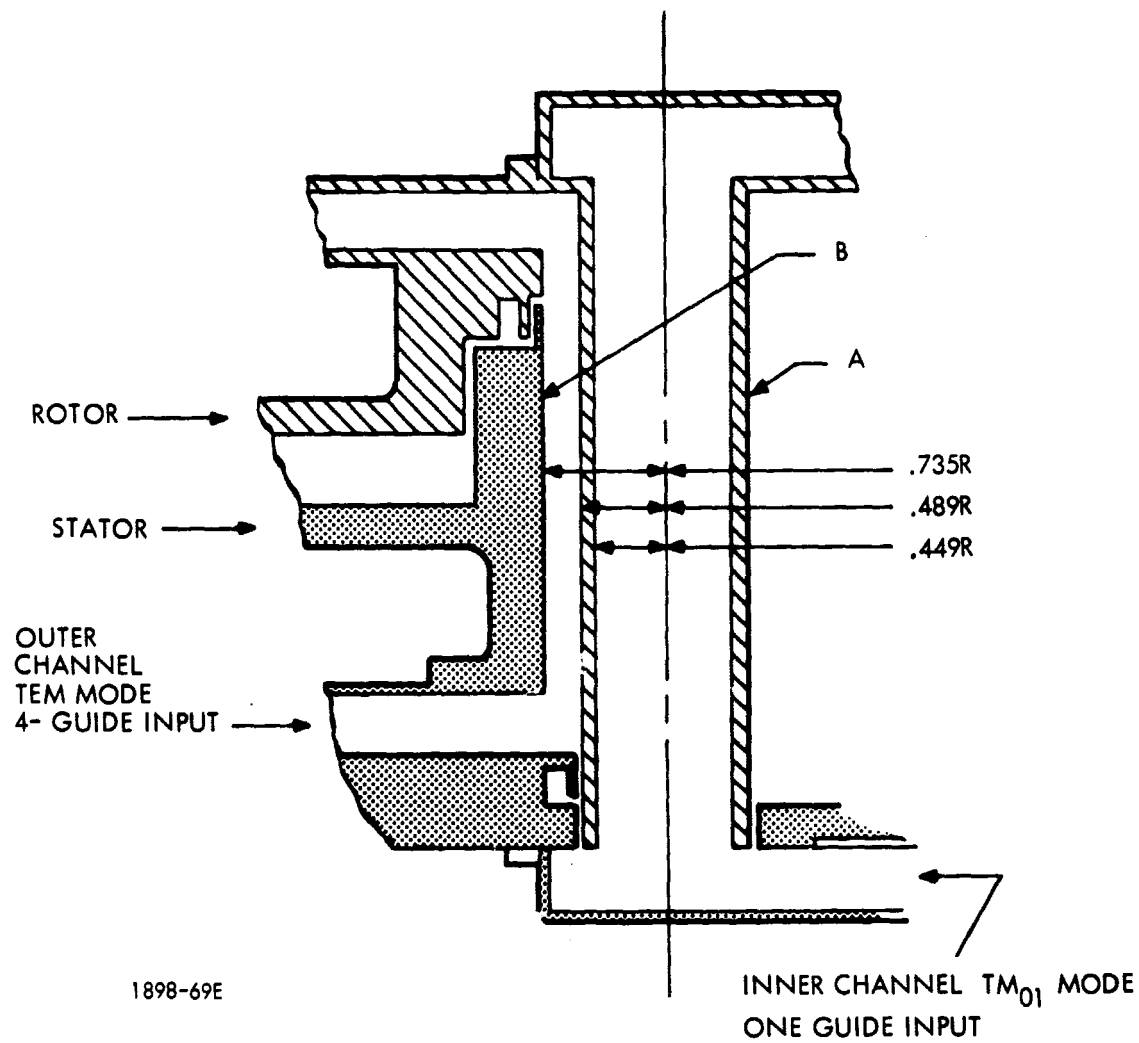


Figure 35. Two-Channel 360°/Day Rotary Joint

TABLE 9. TWO-CHANNEL ROTARY JOINT LOSS AND DISSIPATION

Rotary Section Loss - db	Channels 1-3	Channels 2-4	Dissipation-Watts (2 kW Total Inputs)		Cylinder Length Inches
			Aluminum ¹	Silver ²	
Transition Loss - db	0.0035	0.006	-	-	-
Feed Network Loss - db	0.06	0.06	-	-	-
Total Loss - db	0.07	0.18	-	-	-
Cylinder A	-	-	3.7	2.9	1.0
Cylinder B	-	-	0.8	0.6	0.62
Total Dissipation	-	-	4.5	3.5	-

Notes 1) $p = 2.83 \times 10^{-6}$ ohm-cm at 20°C

2) $p = 1.68 \times 10^{-6}$ ohm-cm at 20°C

3) Assumed Temperature of Joint, 250°F

4) 30% increase in attenuation allowed for surface quality

severe than that in the four-channel joint. Therefore, the four-channel joint system, which eliminates diplexers in the prime mode, was selected for the conceptual system.

2.3.3 Limited Motion Rotary Joint, - Alternate

Motion about the roll and yaw axes is limited, as discussed previously in Section 2.2.3. Accommodation of this motion in the RF transmission lines can be achieved using single-channel joints stacked along the rotational axis with transmission input lines so arranged to prevent physical interference over the required angular interval. This arrangement is illustrated in Figure 36. Each of the rotary joints is of the TM_{01} design identical to the center channel of the two-channel joint described in Section 2.4.2. Insertion loss of each joint is 0.07 db.

Since for a given channel two joints are required, the loss due to joints exclusive of interconnecting transmission line loss is 0.14 db. This compares unfavorably with the 0.08-db total loss of the flexible transmission line limited motion joint. Additionally, the larger amount of rigid transmission line necessary to provide interference-free interconnection between the roll and yaw axis joint would further increase the loss of the system using TM_{01} joints relative to the flexible guide techniques.

2.3.4 Diplexers - Alternate

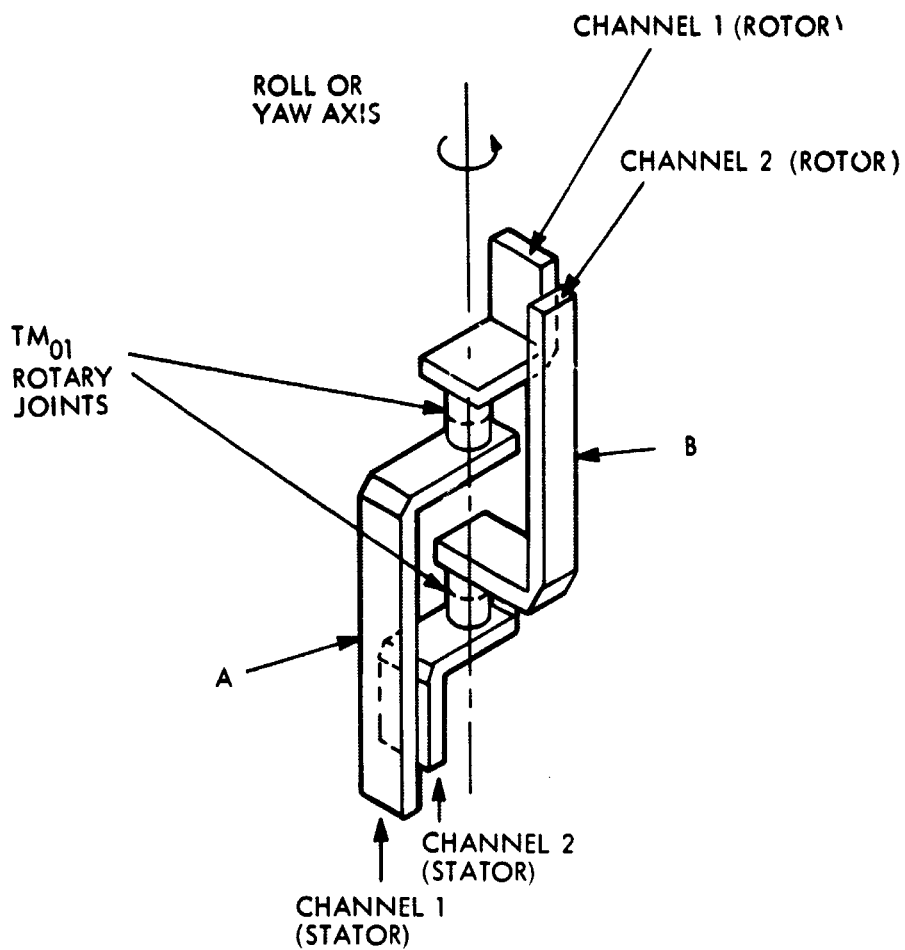
The alternative diplexer design that has been considered is illustrated in Figure 37. Here, the filter networks, either band-pass or band-reject are located between two quadrature hybrids. The two filter networks in each leg between the hybrids are identically matched in their rejection and pass characteristics. Signal paths for the diplexer when using band-reject filters are illustrated in Figure 37.

The insertion loss that can be achieved in this design is comparable to that in the H-plane direct-coupled design. Power capability is higher in that a 2:1 power division occurs at the hybrid and each filter section is required to carry only one-half of the channel power. Isolation between the two channels in this approach is inferior to that of the H-plane direct-coupled design.

Isolation in the hybrid diplexer is a function of the inherent isolation in the hybrid and the reflected energy from the band-pass or band-reject filters. Generally, available quadrature hybrids provide isolation of approximately 35 db. With precision tuning of these devices it is possible that isolation levels of about 45 db could be obtained.

If band-reject filters are used, the isolation between ports 1 and 2 of the diplexer in Figure 37, at frequencies where the filters are in a pass condition, is a function of the filter VSWR. This is a result of the fact that reflected energy at the hybrid output ports terminates in the normally isolated arm of the hybrid. VSWR of the filter sections in their pass condition can be held to approximately 1.05:1 maximum. This results in an isolation of only 32.5 db. Figure 38 shows the isolation as function of filter VSWR. To obtain a 40 db isolation, a near perfectly matched filter ($VSWR \approx 1.01$) is required.

In summary, the diplexer using hybrids and matched filters is advantageous from a power handling viewpoint, while its isolation performance is poor. Since the diplexer is used in a reduced capability auxiliary mode, a trade-off between power handling and



- NOTES:
- 1) ROTATION LIMITED BY INTERFERENCE OF GUIDE B WITH GUIDE A
 - 2) TWO CHANNELS ABOUT ONE AXIS IS SHOWN

642-70E

Figure 36. Limited Motion Rotary Joint - Alternate

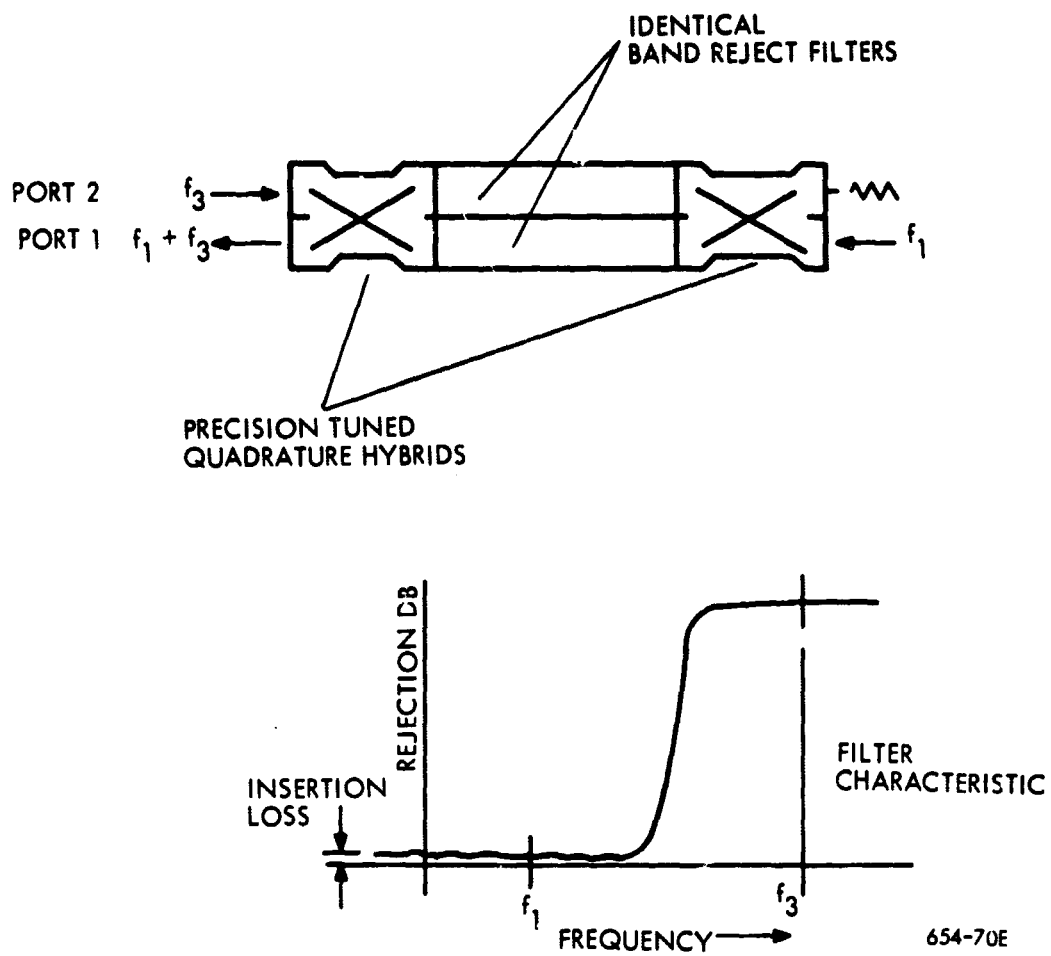


Figure 37. Diplexer Alternate

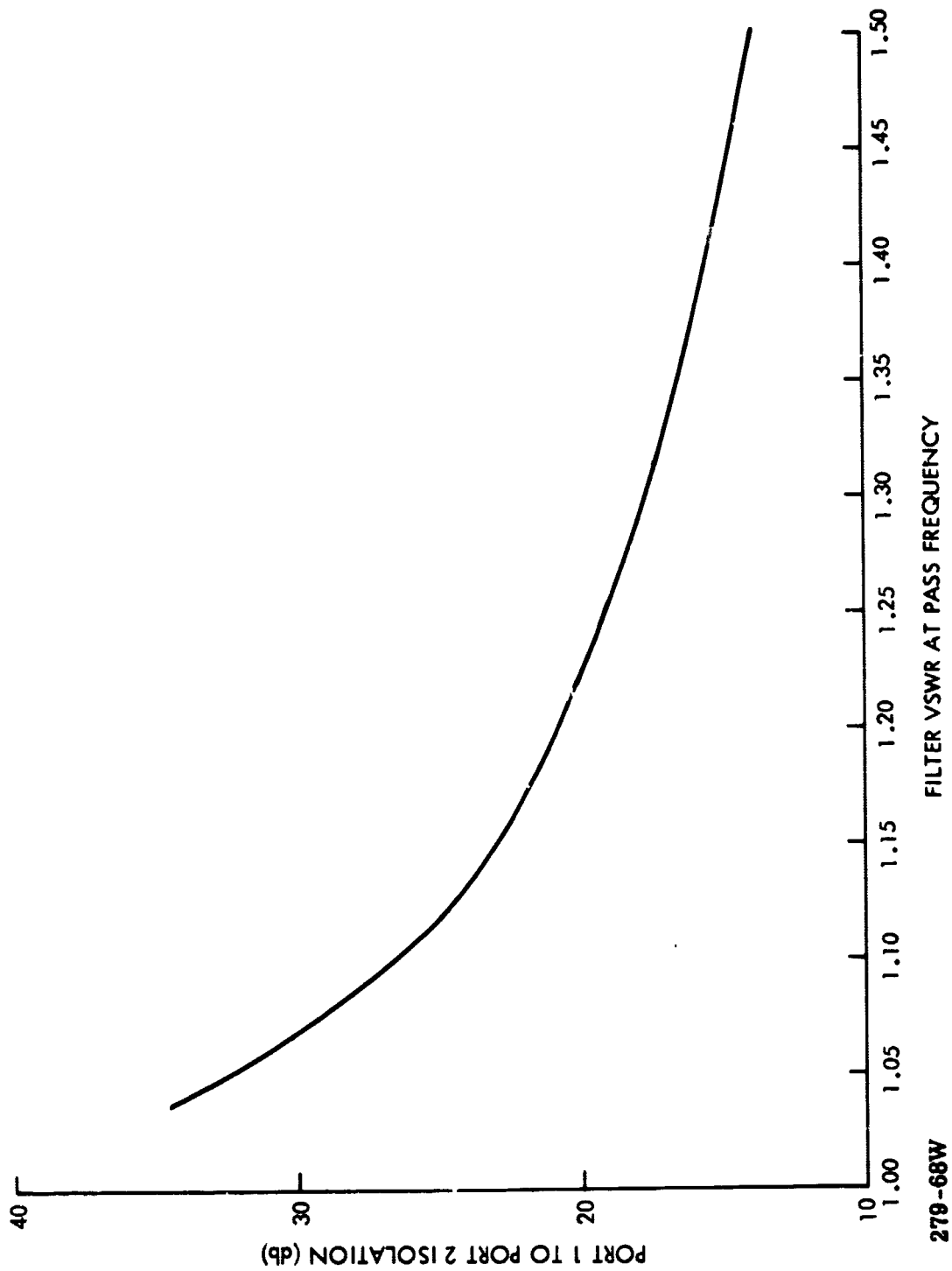


Figure 38. Diplexer Isolation vs Filter VSWR

isolation shortcomings must be made. Power handling difficulties are multipactor oriented (see Section 2.4.1). Although the H-plane direct-coupled diplexer theoretically cannot handle in excess of 0.56 kW without multipaction, it has not been demonstrated experimentally. Further work is required to make the necessary trade-off decision. In the chosen conceptual design, interchannel isolation was judged to be a more significant factor in the design.

2.3.5 RF Switches - Alternate

The use of electronic switches to control the composite pattern form and to implement the auxiliary mode was investigated. The use of electronic switches was considered in an attempt to eliminate or reduce the number of moving parts requiring special lubrication.

Semiconductor switching was discounted because of the high average power, the high operating frequency and the special cooling requirements of the switching diodes.

Ferrite switches exhibited higher power capability, however, they have high loss, poor isolation characteristics, and inherent thermal sensitivity and heat transfer problems. Two forms of the ferrite switch are available and are illustrated in Figure 24.

The differential phase shift circulator switch consists of a quadrature hybrid and a "magic-T" hybrid connected together by two ferrite phase shifters. A 90-degree differential phase shift exists between the two-phase shifters and the lag or lead condition between the two can be changed by controlling the magnetic field about the ferrite. DPDT action can be obtained from this ferrite switch and thus only five would be required. Designs of this type have been operated in X-band at 700-watt CW levels. The high insertion loss (0.6 db) necessitates either forced air or liquid cooling in terrestrial applications.

The junction circulator also shown in Figure 24 consists of an H-plane Y junction of rectangular waveguides with a ferrite cylinder located at the junction. The direction of circulation, i.e., from the input 1 to terminal 2 or terminal 3, is controlled by the magnetization direction in the ferrite. This magnetization is controlled by a solenoid which is pulsed (switched) to the desired polarity. The switched junction circulator is capable of only SPST action and, therefore, seven of these switches would be required to provide the necessary switching requirements. Loss, power capability, and isolation of the two ferrite switches are given in Figure 24 and compared to that of the mechanical type switch.

The ferrite type switches, while having no moving parts, are inadequate in loss, isolation, and power handling in terms of the system requirements.

2.3.6 Polarizer - Alternate

The alternate considered for the polarizer is also a periodic loaded design. The design uses "vaness" as opposed to "pins" to obtain the 90 degree differential phase shift between applied orthogonal field components. The configuration and performance is given in Figure 39.

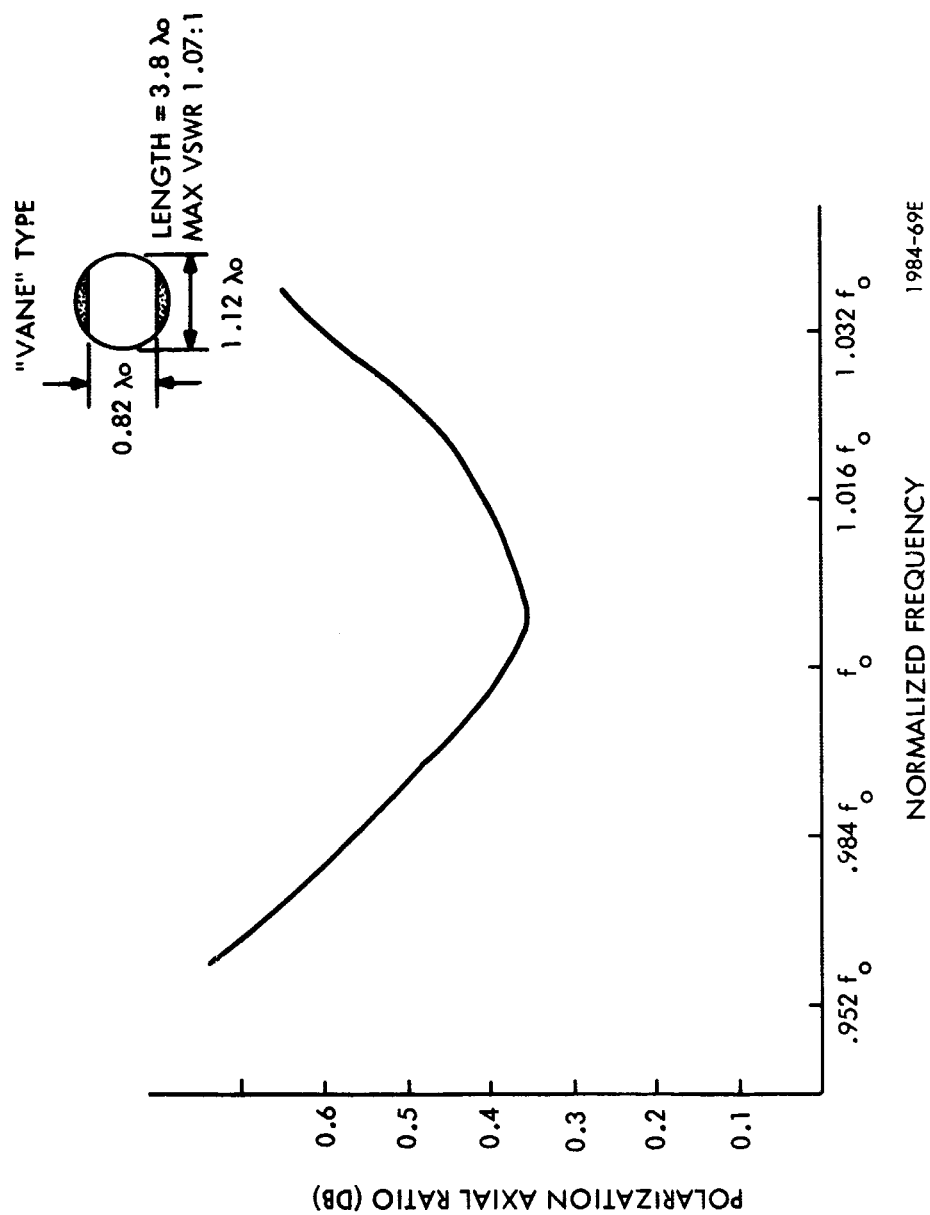


Figure 39. Circular Polarizer Performance - Alternate

The polarization axial ratio performance of the "vane" polarizer is poorer than that of the "pin" type. The former type, however, permits the use of a larger diameter circular waveguide. This factor may be of some significance in terms of both average power and ionization breakdown capability. However, at 1 kW levels this is not a consideration since the "pin" design is operating well below its breakdown limit. Thus, the vane design, while not used in the conceptual system because of its poorer axial ratio, may find application in higher power systems.

2.4 POWER HANDLING

2.4.1 Multipaction

2.4.1.1 Description

Under high vacuum conditions the mean free path of electrons in a containing vessel becomes large. Thus, when an alternating potential is applied across the vessel, electrons originating at one surface will be accelerated toward the other. Under certain conditions of amplitude and phase of the applied field, the electrons will traverse the gap in an odd number of RF cycles, and will have gained sufficient velocity to dislodge electrons from the second surface upon impact. These secondary electrons will be accelerated by the field toward the first wall, where upon collision, more secondary electrons are emitted. This process continues and culminates in an RF breakdown condition. This type of RF breakdown mechanism is commonly called multipactor breakdown.

As implied above, for multipaction breakdown to occur, certain relationships between frequency, applied potential, and conductor spacing must exist. These relations are described by theoretical multipactor effect existence regions derived by Kossel & Krebs⁶ and are shown in Figure 40. In computing the curves in Figure 40, an electron emission velocity of 5 eV (electron volts) has been assumed. The curves given by Kossel and Krebs extend only to transit times of $7/2$ of an RF cycle.

The data given in Figure 40 is insufficient for the APS⁴ RF design in that all possible breakdown voltages for typical X-band frequency-electrode separation products (9000 to 13000 MHz-cm) are not given. In view of this shortcoming, extension of curves in Figure 40 for higher modes (i. e., $n/2$ cycle transit times where $n = 9, 11, 13$, etc.) was computed as follows.

Figure 41 shows the lower portions of the $3/2$ and $5/2$ multipactor existence curves. At points 1 and 2 in this figure, the electron velocities at impact, v_{xd} , are equal. Similarly, for the higher modes ($7/2, 9/2, 11/2$, etc.) the electron impact velocities necessary for multipactor are clearly the same. Thus, once this velocity is determined, the corresponding frequency-electrode separation and applied potential for a mode can be computed.

Referring to Figure 42, consider an electron leaving surface A with an energy U of 5 eV and directed along the X axis at $t = 0$. Velocity and electron energy are related by:

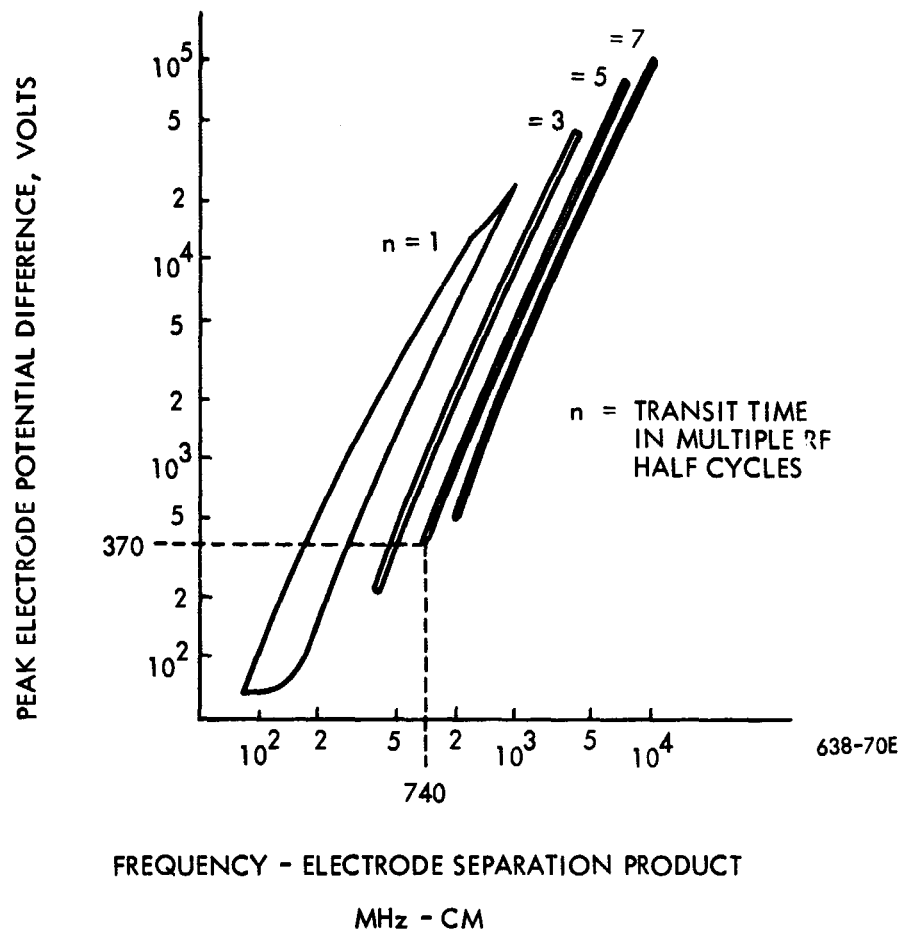


Figure 40. Existence Regions for Multipactor Effect

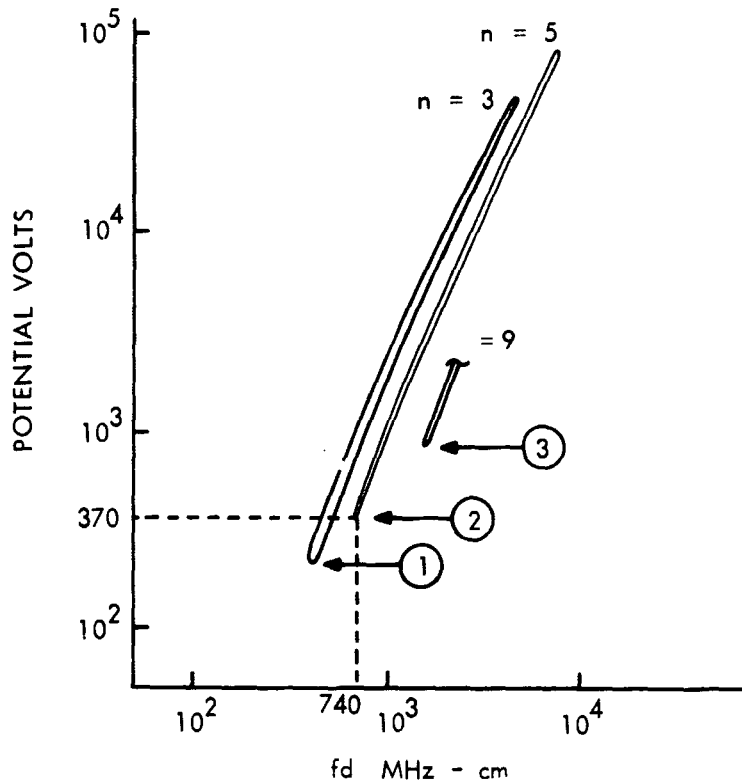


Figure 41. Multipactor Existence Regions

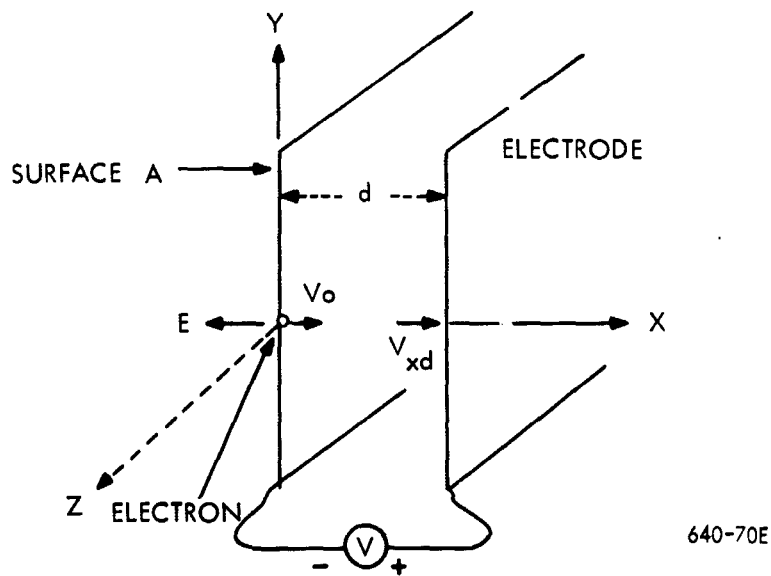


Figure 42. Parallel Plate Region for Multipactor Breakdown

$$U = (1/2) mv^2 \quad (13)$$

where

m = electron mass, 9.107×10^{-31} Kg

v = electron velocity, meters/sec.

U = electron energy in joules

For U equal to 5 eV, $v = v_0 = 1.34 \times 10^6$ meters/sec. To compute the velocity at impact, v_{xd} , we proceed as follows:

$$E = \frac{V}{d} \sin wt \quad (14)$$

where

$V \sin wt$ is the applied voltage - volts

E = electric field - volts/meter

d = the electrode separation - meters

$w = 2\pi f$

t = time - seconds

f = frequency in CPS

$$\frac{dv_x}{dt} = \frac{eE}{m} \quad (15)$$

where

e = electron charge in coulombs

Substituting Equation (14) into Equation (15) and integrating, we have

$$\frac{dv_x}{dt} = \frac{eV}{md} \sin wt$$

$$v_x = \frac{eV}{md} \int \sin wt \, dt + c$$

$$v_x = - \frac{eV}{wmd} \cos wt + c$$

when

$$t = 0; v = v_0; \therefore$$

$$v_x = -\frac{eV}{w d m} \cos wt + v_0 + \frac{eV}{m d w}$$

$$v_x = -\frac{eV}{2\pi m f d} \cos wt + v_0 + \frac{eV}{2\pi m f d} \quad (16)$$

From Figure 40, choosing $f d = 740 \text{ MHz-cm}$, $V = 370 \text{ volts}$, and $wt = 5\pi$ and then substituting in Equation (16) we have

$$v_{xd} = 4.15 \times 10^6 \text{ meters/sec.}$$

where v_{xd} is v_x the electron velocity at impact when $x = d$.

Integrating Equation 16 with respect to t we have

$$\int v_x dt = -\frac{eV}{2\pi w m f d} \int \cos wt dt + (v_0 + \frac{eV}{2\pi m f d}) \int t dt$$

$$x = -\frac{eV}{2\pi w m f d} \sin wt + (v_0 + \frac{eV}{2\pi m f d}) t + c$$

when

$$t = 0, x = 0, \therefore c = 0 \text{ and}$$

$$x = -\frac{eV}{2\pi w m f d} \sin wt + (v_0 + \frac{eV}{2\pi m f d}) t \quad (17)$$

Let t equal the transit time. For breakdown to occur, t must equal an odd multiple of half period, i. e.,

$$t = \frac{n}{2f} \text{ where } n = 1, 3, 5, 7, 9, \text{ etc.}$$

Consider the case when $n = 9$. Substituting $t = \frac{9}{2f}$ and $x = d$ in Equations (16) and (17) we have

$$v_{xd} - v_0 = \frac{2eV}{m d w} \quad (18)$$

$$2fd - 9v_0 = \frac{9eV}{m d w} \quad (19)$$

Solving Equations (18) and (19) simultaneously for fd and V we have for a $9/2$ cycle transit time:

$$fd = 1237 \text{ MHz-cm}$$

$$V = 620$$

These values set the lower most limit on the existence curve for $9/2$ cycle transit time, that is, point 3 in Figure 41.

This procedure has been repeated for transit times/commensurate with fd products somewhat beyond 10^4 MHz-cm. A similar technique was used to generate the remaining portions of the curves. The resultant data is given in Figure 43. Note that in Figure 43 only the lower envelope of the existence curves have been plotted (except for $n = 1$), since these determine the practical power capability of a microwave device.

2.4.1.2 Component Power Handling

The multipactor-limited peak power handling capability of four representative microwave components has been considered. These structures are illustrated in Figure 44.

For the waveguide, the peak voltage is related to the average power by:

$$P = \frac{a V_p^2 \sqrt{1 - \left(\frac{\lambda}{\lambda_{co}}\right)^2}}{1508 d} \quad (20)^7$$

where

P = average power in watts

a = guide width - inches

d = guide height - inches

V_p = peak voltage in waveguide - volts

λ = free space wavelength

λ_{co} = waveguide cutoff wavelength

Using the WR-90 waveguide we have from Figures 43 and 44:

$$d = 0.4 \text{ inches (1 cm)}$$

$$fd = 12400 \text{ MHz-cm}; V_{\text{peak}} = 6 \times 10^3 \text{ volts}$$

$$a = 0.9 \text{ inches (2.29 cm)}$$

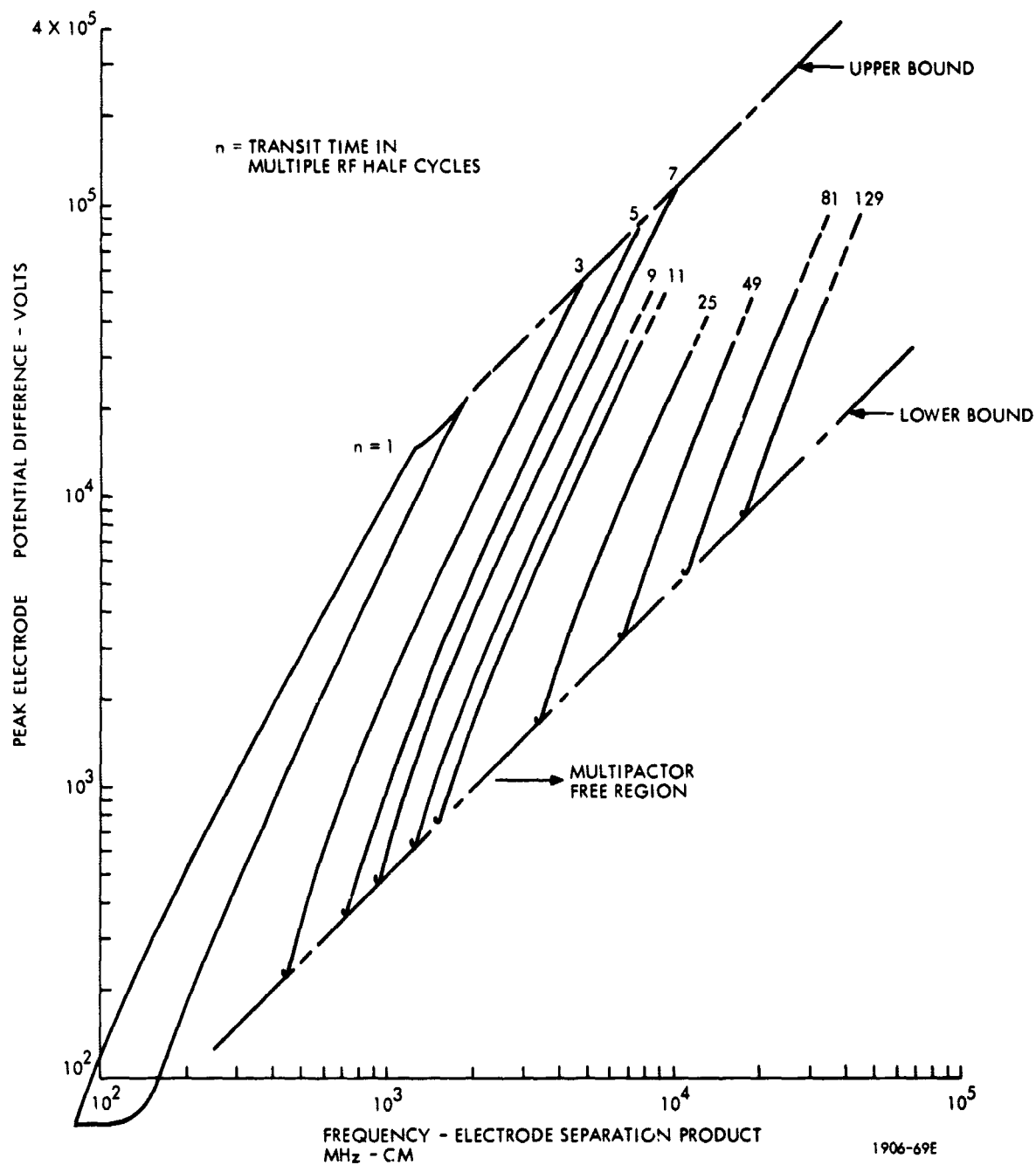
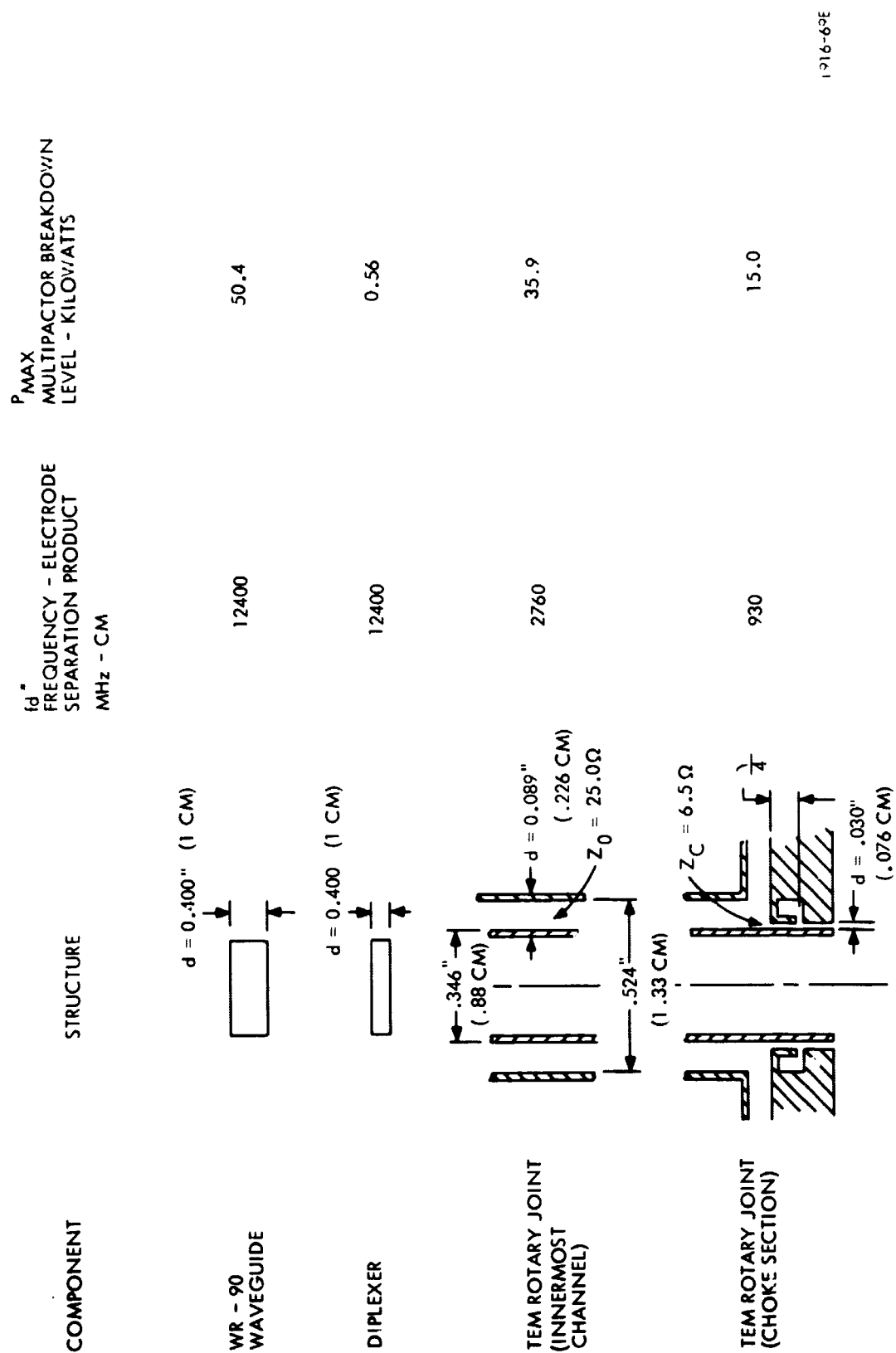


Figure 43. Extension of Existence Regions for Multipactor Effect



1916-60E

Figure 44. Component Power Capacity - Multipactor Limited

$$\lambda = 0.968 \text{ inches (2.46 cm)}$$

$$\lambda_{CO} = 1.8 \text{ inches (4.57 cm)}$$

Substituting in Equation (20), the maximum multipactor limited power handling capacity of WR-90 waveguide is

$$P_{\max} = 50.4 \text{ kW}$$

The voltage buildup that occurs within the diplexer because of the resonant cavities can be computed from the low-pass element values. The peak voltage within the WR-90 waveguide diplexer is given by:

$$V_d = V_p \frac{\lambda_o}{\lambda_g} \left[\frac{2g_1 w_1}{\pi w} \right]^{1/2} \quad (21)$$

where

V_d = peak voltage in the diplexer

V_p = peak voltage in the unperturbed WR-90 waveguide

λ_o = free space wavelength in centimeters

λ_g = guide wavelength in centimeters

g_1 = low-pass element value of the section of the diplexer being evaluated

w_1 = 1.0 for equal ripple designs

w = equal ripple bandwidth divided by the center frequency.

For the diplexer used in the conceptual design, the third cavity exhibits the highest low-pass element value (1.3212) and, therefore, the highest voltage across it for a given input power.

Substitution in Equation (20) for a 1 kW power in WR-90 waveguide gives a peak voltage V_p in the guide of 890 volts. The other pertinent characteristics of the diplexer handling channels 1 and 3 are for channel 1:

$$f_o = 12050 \text{ MHz (center frequency)}$$

$$\lambda_o = 2.49 \text{ cm}$$

$$\lambda_g = 2.97 \text{ cm}$$

$$g_1 = 1.3212$$

$$w_1 = 1.0$$

$$w = 68 \text{ MHz}/12050 \text{ MHz}$$

Substituting these values in Equation (21) gives

$$V_d = 890 \left[\frac{2.49}{2.97} \right] \left[\frac{12050(2)(1.3212)}{\pi 68} \right]^2$$

$$V_d = 9100 \text{ volts}$$

This value of peak voltage occurs where the fd product is 12400 MHz-cm. Referring to Figure 43 shows that this level is sufficient to cause multipaction in the higher modes.

The maximum capability of the diplexer in terms of multipaction breakdown is determined in a similar manner. Figure 43 indicates that a maximum voltage of 6000 volts peak can be sustained in the WR-90 guide of the diplexer. Substitution of this value for V_d gives the corresponding maximum guide voltage in an unperturbed waveguide. Then, using this value of V_d in Equation (20) gives the maximum power that can be sustained without multipaction; this level is 0.56 kW.

The theoretical limit of 0.56 kW for the diplexer design limits the system power level in the auxiliary mode wherein two channels are radiated from the same antenna. Alternate designs using hybrids such as described in Section 2.3.4 do have higher power capability by virtue of their inherent power division in the filter sections. While such designs can theoretically handle the 1 kW level, they cannot meet the isolation requirements. Thus, in actual system usage, a trade-off between the isolation and power handling characteristics will be necessary in choosing the system diplexer.

The rotary joint channel with the lowest multipaction breakdown level is the innermost channel. This is because this channel has the smallest frequency electrode separation product (fd). The fd product for this channel is 2760 MHz-cm and the characteristic impedance is 25 ohms. By comparison, channels 2, 3, and 4 have fd products of 4900, 7650, and 12105 MHz-cm, respectively. Characteristic impedance of each of these latter channels is also 25 ohms.

For the innermost channel where fd is 2760 MHz-cm, the maximum permissible peak voltage V_j between the center and outer conductor prior to multipaction is shown in Figure 43 to be 1340 volts. The power in the coaxial section of the joint is related to V_j by:

$$P = \frac{1}{2} \frac{V_j^2}{Z_0} \quad (22)$$

where

P = power flowing through the joint in watts

V_j = peak voltage between the center and outer conductors in volts

Z_0 = characteristic impedance of the rotary joint in ohms.

Substituting the 1340 volt value for V_j in Equation (22)

$$P = \left(\frac{1}{2}\right) \frac{(1340)^2}{25}$$

$$P = 35.9 \text{ kW}$$

Thus, considerable margin exists in the rotary joint in terms of multipaction breakdown when operating at 1 Kw levels.

The rotary joint choke that is most susceptible to multipaction is that used in the innermost channel. This is a result of fixed minimum choke spacings for all joints (nominally 0.030 inches) with the innermost channel providing the highest choke impedance. Since the maximum voltage in the choke is directly proportional to the choke impedance, this inner channel is the most critical.

The choke structure used in the rotary joint is shown in Figure 44. For a given main line RMS current, I , the peak voltage in the choke across the minimum gap is given by

$$V_c = 2\sqrt{2} Z_{oc} I \quad (23)$$

where

V_c = peak voltage across the minimum gap of the choke in volts

Z_{oc} = characteristic impedance of the choke

I = RMS current in the mainline of the coaxial rotary joint

The 0.030-inch choke section is commensurate with an fd of 930 MHz-cm. From Figure 44 this fd product sets the multipaction voltage level at 455 volts. Substitution of this latter value for V_c in Equation 23 together with 6.5 ohms for Z_{oc} gives a mainline current I of 24.8 AMPS RMS.

Since power in the mainline is related to current by $I^2 Z_o$, this current is equivalent to a power capability of

$$P = I^2 Z_o$$

$$P = (24.8)^2 (25)$$

$$P = 15.4 \text{ kW}$$

Again, this power capability in the choke gives substantial margin for protection against multipaction.

In summary, with the exception of the diplexer used in the auxiliary mode, none of the antenna system RF components are multipaction limited at the 1 kW level. The capability of the various components is tabulated in Figure 44.

2.4.2 Ionization Breakdown

2.4.2.1 Description

Ionization breakdown or diffusion-controlled breakdown, as it is sometimes called, occurs when electrons are produced by ionizing collisions between electrons and gas molecules. The electrons are lost by diffusion to the walls of the component in which the breakdown occurs. This type of breakdown is dependent upon the spacing (d) of the surfaces between which the potential is applied, the wavelength, type of gas present, gas pressure, and the type of surface material.

Normally, ionization breakdown is of concern in terrestrial or airborne applications where substantial amounts of gaseous material is present. In space, however, atmospheric pressure is extremely low $\leq 10^{-12}$ Torr and, therefore, ionization breakdown due to atmospheric gas is unlikely. Artificial atmospheres that occur in space due to outgassing of RF components can cause considerably higher pressures than the 10^{-12} Torr and are therefore, of concern.

To determine the outgassing/ionization breakdown problem associated with all metal RF components (dielectric outgassing has been excluded) the worst case pressures within the aluminum waveguide and rotary joint due to outgassing of absorbed gases (predominantly oxygen) was computed. This "worst case" analysis assumed the following: the component under consideration underwent a pressure change from atmospheric to $< 10^{-12}$ Torr instantaneously; the outgassing was considered to occur within a closed container, i. e., both the waveguide and rotary joint were sealed immediately upon the start of the outgassing process. Under these conditions, steady state pressures of 1.37×10^{-2} Torr and 4.28×10^{-2} Torr were computed for the waveguide and rotary joint, respectively. Detail of this computation is given in Section 4.1.1 of this report.

Data on ionization breakdown between parallel plates for pressure spacing products pd of less than one Torr-cm occurring simultaneously with fd products in the vicinity of 12000 is not available in the literature. These pd and fd products are those that occur in the RF components of the recommended design. To provide an approximation of the ionization breakdown properties of RF components at 10^{-2} Torr pressure, the work of Meuhe⁹ which gives contours of constant voltage breakdown for hydrogen gas between parallel plates was extrapolated to cover the regions of pressure interest ($\approx 10^{-2}$ Torr) and component frequency - electrode separation products ($\approx 3 \times 10^3$ to 12×10^3 MHz-cm). This extrapolation is shown in Figure 45 and was achieved by first extending the demarcation line between the multipaction and diffusion (ionization) breakdown regions to smaller pd levels. The demarcation line was defined by connecting points of the 500 and 150 volt contours where the slope changed from zero to a positive value. Higher voltage contours were obtained from the work of MacDonald et. al¹⁰ on ionization CW breakdown in X band, in air. The latter work provided three points (indicated by + in Figure 45) in the diffusion controlled region. Constant voltage contours were then drawn through these points parallel to the 500 volt contour. This procedure is judged to be accurate to ± 10 percent in that breakdown properties for two similar behaving but different gases (air and hydrogen) have been combined.

Use of the extended ionization data (Figure 45) in evaluating the RF component ionization breakdown properties is covered in the following subsection.

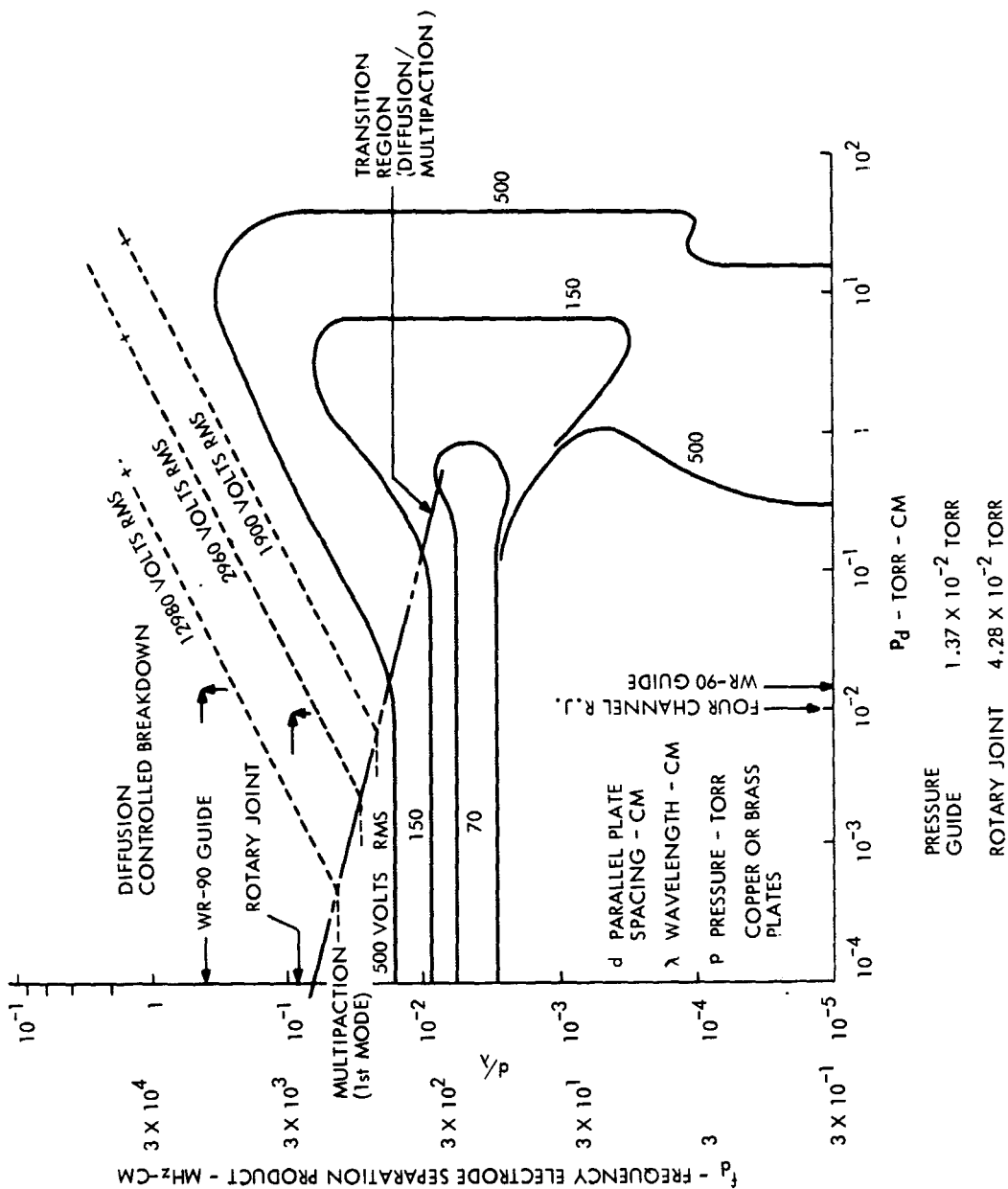


Figure 45. Contours of Constant Voltage Breakdown

2.4.2.2 Component Power Handling

As indicated previously, the worst case pressure in the waveguide and rotary joint components is 1.37×10^{-2} Torr and 4.28×10^{-2} Torr, respectively. Frequency-electrode separation products (fd) for the waveguide and rotary joint (innermost channel of the four-channel joint) are 12.2×10^3 and 2.79×10^3 MHz-cm, respectively. These operating points have been entered in Figure 45 and show that a voltage in excess of 12980 volts RMS is necessary to cause ionization breakdown in the WR-90 waveguide while approximately 7000 volts RMS is necessary to cause breakdown in the rotary joint. When operating at 1 kW, the WR-90 waveguide, the diplexer, and the rotary joint incur maximum peak voltages of 900, 9100, and 224 volts, respectively. Since these voltages are considerably less than those necessary for breakdown, ionization breakdown associated with outgassing of absorbed gases is not a problem in the RF components of the antenna system.

Refined pressure computations on the above components was also made using molecular flow equations. In this technique the components are not "sealed" and thus, vent to space. Results of this analysis give pressures in the waveguide and rotary joint of 4.97×10^{-3} Torr. This pressure is approximately an order of magnitude less than that of the "worst case" computation and is more representative of the real case. In view of this, the ionization breakdown is extremely remote.

In the analysis, the effect of lubricant outgassing has not been considered. Lack of ionization breakdown information on space lubricant vapors prevents such analysis. Experimental work is required in this area before power breakdown estimates can be made.

2.4.3 Average Power Limitations

The average power capability of the RF components is determined by the magnitude of their dissipative loss and their ability to radiate heat such that unacceptable temperature rise is avoided. Temperature rise must be limited to keep materials within their capability and to keep dimensional change in critical areas such as bearing shafts to a minimum.

The design of all microwave components has been based upon minimizing loss and incorporation of features which permit good conductive heat paths from the dissipative areas to space. For example, in the rotary joint, coaxial lines have been minimized in length, thus permitting heat to be easily conducted out via the center and outer conductors to a radiative structure and then radiated into space.

Detail listing of the power dissipated in the various components and the computed temperature rise is given in Section 4.3.1. In all cases acceptable temperature rises are obtained when operating at the required 1 kW per channel level.

2.5 GAIN-LOSS TABULATION

The gain-loss summaries for the prime and auxiliary mode are given in Table 10. In the prime mode the available peak gain is 36.7 db for Channel 1. This gain is typical for all channels operating in the prime mode (i. e., single channel per antenna)

TABLE 10. CONCEPTUAL DESIGN GAIN/LOSS TABULATION

Aperture A-1 Channel 1	Aperture A-3 Channel 1
Antenna Peak Gain	Antenna Peak Gain
37.3 db	37.3 db
Losses	Losses
Rigid Transmission Line ($\ell = 62''$ at 0.04 db/ft)	Rigid X Mission Line ($\ell = 55''$)
0.21	0.18
Flexible Transmission Line ($\ell = 12''$ at 0.08 db/ft)	Flexible X Mission Line ($\ell = 12''$)
0.1	0.1
Waveguide Bends	Waveguide Bends
0.2	0.2
Rotary Joint (Rotary Section)	Rotary Joint (Rotary Section)
0.06	0.06
Rotary Joint (Feed Network) ($\ell = 60''$ at 0.04 db/ft)	Rotary Joint (Feed Network)
0.20	0.20
Switch (1)	Diplexer
0.05	0.6
	Switch (2)
	0.1
Total Loss	Total Loss
0.82	1.44
Available Peak Gain	Available Peak Gain
36.5 db	35.9 db

irrespective of the pattern configuration. Similarly, the 36.1 db gain for the condition where two channels are radiated from the same antenna is typical for all antennas when operating in this manner. The 0.6 db increased loss in the auxiliary mode is a result of the diplexer necessary to implement this function.

SECTION III

POINTING SYSTEM DESIGN

The pointing system causes the communication antenna pattern coverage to be properly located and oriented on the surface of the earth. It includes sensing elements, which determine if the actual pattern location and orientation is as desired. It also includes steering elements, which cause the pattern location and orientation to go to and remain at the desired location in the presence of disturbances (due to the motion of the satellite platform on which the antenna subsystem is mounted).

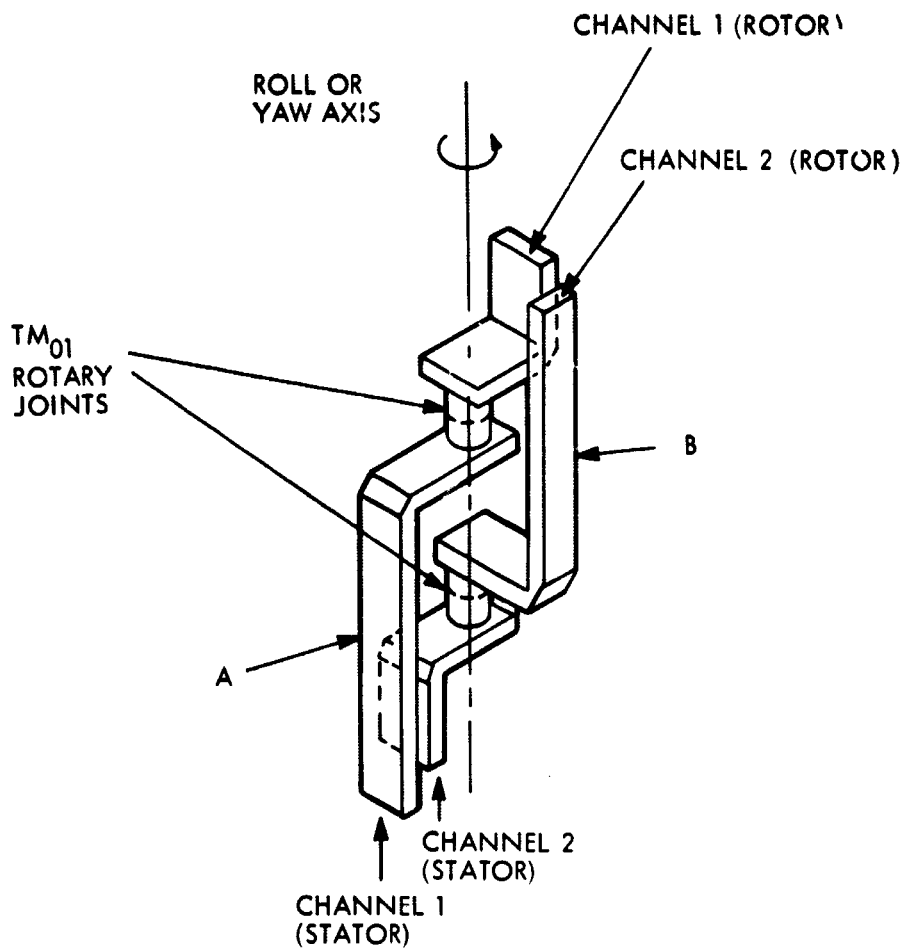
Except for perturbations or disturbances, the satellite body does not rotate about its body axes during the course of a day. Since the satellite orbits the earth once a day, the antenna must rotate once per day relative to the satellite body for the pattern to remain stationary on the surface of the earth. In addition, the boresight axis must be moved in the East-West and North-South directions to change pattern coverage. Rotation about the boresight axis is needed to align the pattern centerline with the latitude circle at the selected pattern location. These three angular motions are accomplished by a three-axis gimbal system between the satellite body and the antenna. Rotation about each of the gimbal axes is accomplished by a motor whose output is controlled by servo-amplifiers such that the actual pattern location and orientation is rotated to the desired location and orientation.

The sensing system is redundant and thus provides directional information in either of two operating modes. In the prime mode, reference beacon signals are transmitted from the ground and sensed on the satellite; in the secondary mode, reference signals are transmitted from the satellite, sensed and processed on the ground, with directional update information then transmitted to the satellite via the uplink.

The sensing techniques in the prime and redundant modes are similar. Pattern motion in a linear direction on earth is sensed by a crossed arm interferometer that provides error information in two orthogonal directions. Steering from the interferometer null position is provided by bias offset information. Pattern rotation on earth is sensed by a polarization angle detector. Offset rotation is accommodated by mechanically varying the polarization angle of the transmitted signal or the receiving sensor.

3.1 SYSTEM ANALYSIS

This section describes the mathematical relationships of the elements which comprise the pointing system. The description includes the interactions between the satellite motions, gimbal angles, and the location of the beacon in terms of the antenna reference frame. The resulting functions constitute a mathematical model of the pointing system which can be used to investigate the dynamic accuracy of the system when excited by the various pointing commands or satellite body perturbations. The system is described in terms of the prime sensing mode with later sections detailing operation in the secondary or redundant sensing mode.



- NOTES:
- 1) ROTATION LIMITED BY INTERFERENCE OF GUIDE B WITH GUIDE A
 - 2) TWO CHANNELS ABOUT ONE AXIS IS SHOWN

642-70E

Figure 36. Limited Motion Rotary Joint - Alternate

3.1.1 Terminology

One of the specified pattern configurations consists of four individual beams in a line. The line between the antenna and the midpoint of the beams is termed the antenna boresight axis. The line connecting the four centers of the circular patterns is termed the pattern centerline. Pattern location and orientation are achieved by causing the boresight axis to pass through a specified point on the surface of the earth and by causing the pattern centerline to be tangent to the latitude circle which passes through the specified point. The interferometer output signals are measures of the beacon coordinates in terms of the antenna reference frame. The polarization detector output is a measure of the angular orientation of the antenna reference frame about the boresight axis. Each pair of interferometers defines one of the two axes of the antenna reference frame which are mutually orthogonal to the boresight axis.

The satellite body orientation and location is defined in terms of an "inertial" reference frame. The center of the inertial reference frame is coincident with the nominal or desired satellite location in space. The z axis passes through the center of the earth, the y axis is parallel to the earth's axis of rotation, and the x axis is orthogonal to the y and z axes. (Figure 46)

Points on the earth are described in terms of latitude and longitude angles. Latitude corresponds to the usual geographic latitude, positive angles being northward from the equator. Positive longitude angles are measured from the meridian of the subpoint to the East.

Gimbal angles are measured from the following zero references. The gimbal closest to the satellite (inner gimbal) corresponds to antenna rotation about the satellite's axis. When this angle is zero the boresight axis is aligned with the satellite z axis. Rotation about the middle axis causes the pattern to move to the north or south on the earth. Zero angle corresponds to the boresight axis lying in the equatorial plane. Rotation about the outer axis causes pattern centerline rotation; zero angle is the value which causes the line connecting the centers of one pair of interferometer apertures to be parallel to the earth's axis. Thus, with all gimbal angles at zero and with the satellite aligned with the inertial reference frame the boresight will pass through the subpoint and the centerline will be tangent to the earth's equator.

3.1.2 Gimbal Angle Equations

Expressions for the gimbal angles required to cause the boresight to pass through a desired point, D, with the correct centerline tangency, are derived for the case where the satellite is assumed to be located coincident with the inertial reference frame. The effect of satellite motion relative to the inertial reference frame will be described later. The geometrical relationships are shown in Figure 46. Point D, is an arbitrary antenna boresight point on the earth's surface. Line ℓ_1 is a straight line tangent to the latitude circle on which D is located. The different reference frames of interest are shown with arrowheads denoting positive directions.

The coordinates of D in reference frame 1 are denoted by x_{d1} , y_{d1} , z_{d1} . In matrix notation the vector x_{d1} , y_{d1} , z_{d1} will be denoted D_1 . From the figure it is seen that

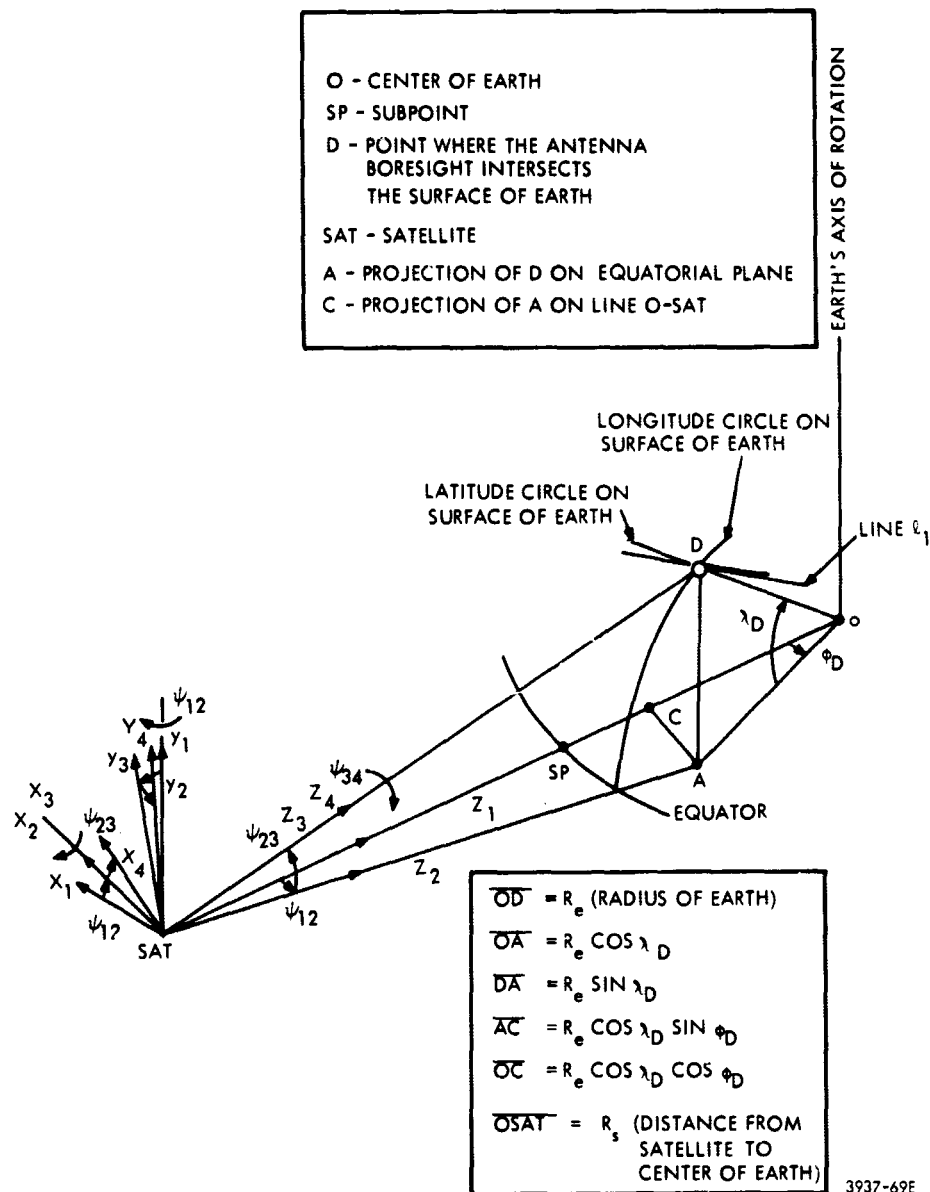


Figure 46. Geometrical Relationships

$$D_1 = \begin{bmatrix} -R_e \cos \lambda_D \sin \phi_D \\ R_e \sin \lambda_D \\ R_s - R_e \cos \lambda_D \cos \phi_D \end{bmatrix} \quad (23)$$

The coordinates of D in reference frames 2, 3, and 4 are:

$$[D_2] = [R_{12}] \times [D_1] \quad (24)$$

$$[D_3] = [R_{23}] \times [R_{12}] \times [D_1] \quad (25)$$

$$[D_4] = [R_{34}] \times [R_{23}] \times [R_{12}] \times [D_1] \quad (26)$$

where

$$[R_{12}] = \begin{bmatrix} \cos \psi_{12} & 0 & \sin \psi_{12} \\ 0 & 1 & 0 \\ -\sin \psi_{12} & 0 & \cos \psi_{12} \end{bmatrix} \quad (27)$$

$$[R_{23}] = \begin{bmatrix} 1 & 0 & 0 \\ 0 & \cos \psi_{23} & -\sin \psi_{23} \\ 0 & \sin \psi_{23} & \cos \psi_{23} \end{bmatrix} \quad (28)$$

$$[R_{34}] = \begin{bmatrix} \cos \psi_{34} & \sin \psi_{34} & 0 \\ -\sin \psi_{34} & \cos \psi_{34} & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad (29)$$

The value of the inner gimbal angle (ψ_{12}) that will cause the boresight to pass thru the projection of D on the equatorial plane can be expressed in terms of the latitude and longitude of point D (λ_D, ϕ_D) as

where

$$\psi_{12}^d = \tan^{-1} \left[\frac{\cos \lambda_D \sin \phi_D}{R_1 - \cos \lambda_D \cos \phi_D} \right] \quad (30)$$

where

$$R_1 = R_s / R_e \quad (31)$$

From the figure it can be seen that Z_{2A} , the distance from SAT to A, is given by

$$Z_{2A} = \frac{R_s - R_e \cos \lambda_D \cos \phi_D}{\cos \psi_{12}} \quad (32)$$

and that the second gimbal angle can be defined by

$$\psi_{23}^d = \tan^{-1} \left(\frac{R_e \sin \lambda_D}{Z_{2A}} \right) \quad (33)$$

when the boresight passes through point D.

Combine (32) and (33) so that

$$\psi_{23}^d = \tan^{-1} \left(\frac{\sin \lambda_D \cos \psi_{12}}{R_1 - \cos \lambda_D \cos \phi_D} \right) \quad (34)$$

The desired value for ψ_{34} , the third gimbal angle, is that which causes the pattern centerline to be tangent to the line ℓ_1 which is tangent to the latitude circle at D. This result is achieved by rotating the reference frame about z_3 until the line ℓ_1 lies in the x_4 - z_4 plane. An equivalent requirement is that the y coordinate of line ℓ_1 in the fourth reference frame be zero for all values of x_4 . This condition is used to evaluate ψ_{34} in the following manner.

The equation of line ℓ_1 is expressed in the second reference by

$$\left\{ \begin{array}{l} y_2 = R_e \sin \lambda_D \\ z_2 = Z_{2A} - x_4 \tan (\phi_D + \psi_{12}) \end{array} \right\} \quad (35)$$

where it is a function of y_2 and z_2 only. Using the transpose of the R_{23} rotation matrix, ℓ_1 can be written in terms of the third reference frame and solved for y_3 and z_3 . Then using R_{34} and previous relationships, an expression for y_4 can be obtained with x_3 as the only variable. If the coefficient of x_3 is equated to zero, y_4 is zero for all values of x_3 ; that is, the line ℓ_1 lies in the x_4 , z_4 plane. This results in

$$\psi_{34}^d = \tan^{-1} \left[\sin \psi_{23} \tan (\psi_{12} + \phi_D) \right] \quad (36)$$

Equations (30), (34) and (36) define values for gimbal angles that will cause the pattern to be properly oriented on the surface of the earth when the satellite body is aligned with the satellite inertial reference frame. The gimbal angles are superscripted d, (ψ_{ij}^d) to denote the desired reference frame which locates the pattern boresight in the desired location and orientation on the earth. A superscript a denotes the actual gimbal angles.

3.1.3 Offset Equations

The beacon is located at point B, having coordinates $[B_4]$ in the antenna reference frame given by

$$[B_4] = [R_{34}] \times [R_{23}] \times [R_{12}] \times \begin{bmatrix} R_e \cos \lambda_B \sin \phi_B \\ R_e \sin \lambda_B \\ R_s - R_e \cos \lambda_B \cos \phi_B \end{bmatrix} \quad (37)$$

The subscript d is used to identify the value of B_4 corresponding to the desired gimbal angles ψ_{12}^d , ψ_{23}^d , ψ_{34}^d . The coordinates of B_4^d are termed the "Beacon offsets."

The coordinates of the beacon in the actual antenna reference frame when the actual gimbal angles are ψ_{12}^a , ψ_{23}^a and ψ_{34}^a and the satellite is aligned with the inertial reference frame are referenced by the superscript a, eg, B_4^a .

The relationship between the variables ψ_{12}^d , ψ_{23}^d , ψ_{34}^d and earth coordinates of D and B is shown by the first block of the signal diagram in Figure 47.

3.1.4 Interferometer Equations

The interferometer detector measures the phase difference (PD) between the signals received by two separate apertures due to the difference in the path lengths to the transmitter. The path length difference (Δ) shown in Figure 48 is related to the phase difference (in radians) by

$$\begin{aligned} \text{PD} &= 2\pi \Delta / \lambda \\ &= 2\pi (D/\lambda) \sin \theta \\ &= 2\pi (D/\lambda) (y_4 / R_{BS}) \end{aligned} \quad (38)$$

where λ is the wavelength of the beacon signal. The "offset" phase angles corresponding to the desired beacon location are then given by

$$\text{PD}_x^d = 2\pi (D/\lambda) \left(x_{B4}^d / R_{BS} \right) \quad (39)$$

$$\text{PD}_y^d = 2\pi (D/\lambda) \left(y_{B4}^d / R_{BS} \right) \quad (40)$$

The actual value of the phase difference is related by similar equations to the actual location of the beacon in reference frame 4 coordinates. Thus

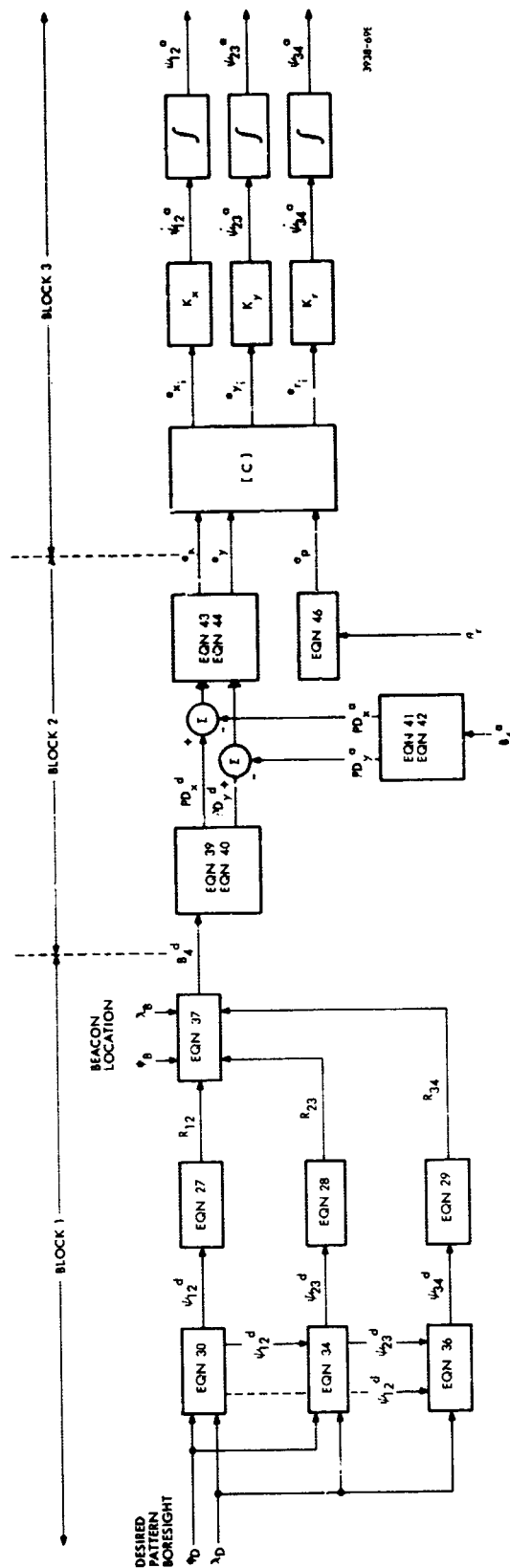


Figure 47. Partial System Block Diagram No. 1

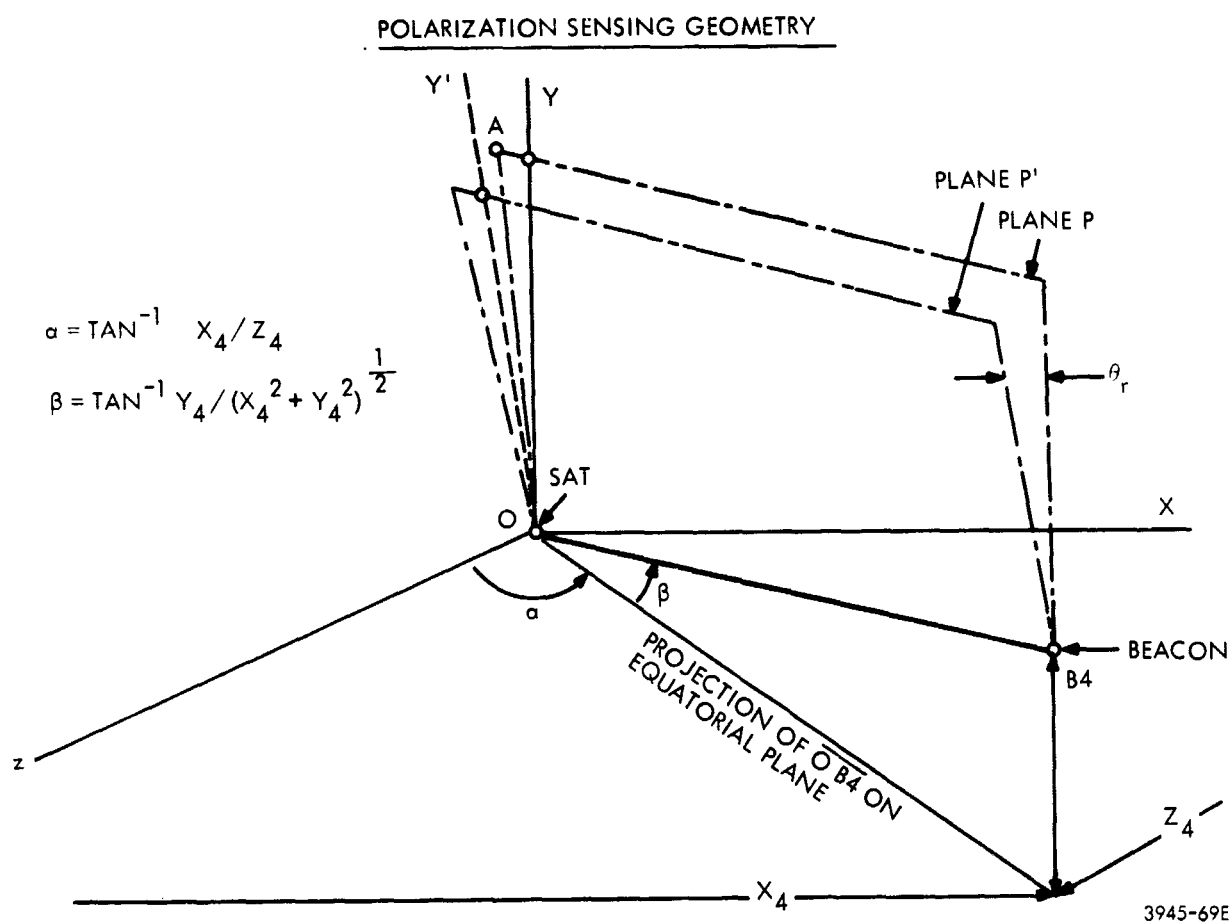
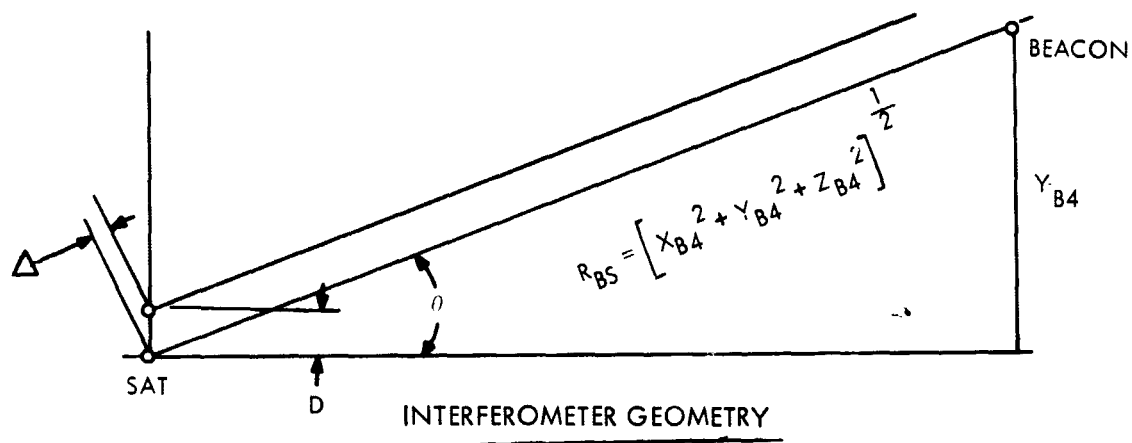


Figure 48. Interferometer and Polarization Sensing Geometry

$$PD_x^a = 2\pi (D/\lambda) \left(x_{B4}^a / R_{BSa} \right) \quad (41)$$

$$PD_y^a = 2\pi (D/\lambda) \left(y_{B4}^a / R_{BSa} \right) \quad (42)$$

The distance $R_{BSa} = R_{BS}$ for rotations about the inertial axes in the absence of translation. When satellite translation occurs,

$$R_{BSa}^2 = x_{B4}^a{}^2 + y_{B4}^a{}^2 + z_{B4}^a{}^2 \quad (43)$$

The difference, $PD^d - PD^a$, is measured in the interferometer subsystem. The outputs of the two channels are voltages

$$e_x = K_i(D/\lambda) (y_{B4}^d/R_{BS} - y_{B4}^a/R_{BSa}) \quad (44)$$

$$e_y = K_i(D/\lambda) (y_{B4}^d/R_{BS} - y_{B4}^a/R_{BSa}) \quad (45)$$

3.1.5 Polarization Angle Detector Equations

The desired polarization angle offset (θ) will always be zero since the polarization angle of the transmitted beacon signal is computed and adjusted on the ground such that the spacecraft polarization angle sensor functions as a null seeking device.

The output of the polarization angle detector when the actual rotation differs from the desired rotation is given by

$$e_p = K_p \sin \theta_r \quad (46)$$

The block diagram of the interferometer detectors and polarization angle detector is shown in the second block of Figure 47.

3.1.6 Control Law and Simplified Servo Equations

The three voltages e_x , e_y , e_p found previously are error signals used as inputs to the servo amplifiers which cause rotation of the three gimbal motors. The actual functional relationships will be described in detail in later sections, however, the average gimbal angular rate is approximately proportional to the servo amplifier input voltage and the servo amplifier inputs are related to the error signals. One possible relationship is

$$\begin{bmatrix} e_{xi} \\ e_{yi} \\ e_{ri} \end{bmatrix} = C \times \begin{bmatrix} e_x \\ e_y \\ e_p \end{bmatrix} \quad (47)$$

where the terms of the control law matrix $[C]$ are functions of the desired gimbal angles ψ_{12}^d , ψ_{23}^d , ψ_{34}^d and are chosen to insure an optimum system response to the disturbances due to satellite body motions relative to the inertial reference frame. The functional relation between gimbal angles, satellite motion, and detector outputs will be described in succeeding sections. Incremental variations of gimbal angles, in the absence of satellite motion, are defined by the matrix equation

$$\begin{bmatrix} dx_{B4}^a \\ dy_{B4}^a \\ d\theta_r^a \end{bmatrix} = [F] \times \begin{bmatrix} \psi_{12}^a \\ \psi_{23}^a \\ \psi_{34}^a \end{bmatrix} \quad (48)$$

where the elements of F are partial derivations of x , y , or θ with respect to ψ_{12} , ψ_{23} , and ψ_{34} . If $[C]$ is chosen to be equal to the inverse of $[F]$ then the system will behave, for incremental motions, as though it were three uncoupled control systems. The block diagram corresponding to this control law matrix and simplified servo response is shown in the third block of Figure 47.

3.1.7 Servo Motor Equations

The approximate motor behavior was described in Section 3.1.6 while this section presents a more exact set of describing equations for the variable reluctance type of stepping motor selected for this application.

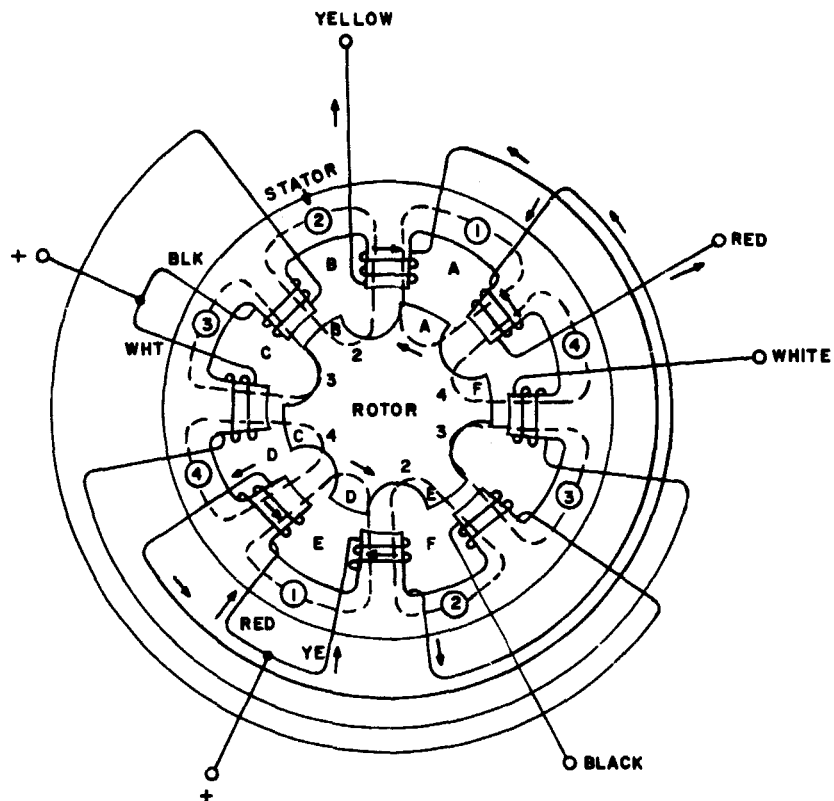
A simplified cross-sectional view of a reluctance servo motor (two-phase) is shown in Figure 49. The flux and current distributions resulting from connecting the terminals labeled red and yellow to power negative are shown by identifying arrows. The stable positions of the rotor corresponding to flux paths (1) through (4) are with rotor pole A opposite interpolar space (A), then (B) opposite (B), etc. Note that the rotor turns 15 degrees clockwise while the flux distribution moves 45 degrees counter-clockwise for each change of the excitation state. Following an energy method of analysis², the energy stored in the airgap is found by evaluating the H_r field due to stator currents; it is computed as though the airgap were uniform. The B_r field is then found by multiplying the H_r field by a permeance which is a function of the rotor position. In order to yield tractable results, a second order sinusoidal function is used for the permeance,

$$\mu = \mu_0 - \mu_2 \sin 6 (\theta - \phi) \quad (49)$$

θ is measured relative to a reference point on the stator and ϕ is the motion of a reference point on the rotor relative to the reference point on the stator; μ_0 and μ_2 are constants.

The assumed current density and H_r field distributions are shown in Figure 50, together with the permeance function.

The expression for H_{ry} plotted in Figure 51 is written as



NODES SHOWN + ALWAYS CONNECTED. NODES LABELED RED, WHITE, ETC ARE CONNECTED TO POWER RETURN IN THE FOLLOWING SEQUENCE TO CAUSE CCW ROTATION OF ROTOR. THE PATH CARRYING THE PRINCIPAL FLUX FOR EACH MOTOR CONNECTION IS AS FOLLOWS.

ENERGIZED

YELLOW & RED →
 BLACK & YELLOW
 WHITE & BLACK
 RED & WHITE

FLUX PATH NUMBER (DASHED LINE)

1 →
 2
 3
 4

3944-69E

Figure 49. Simplified Step Motor Cross-Section

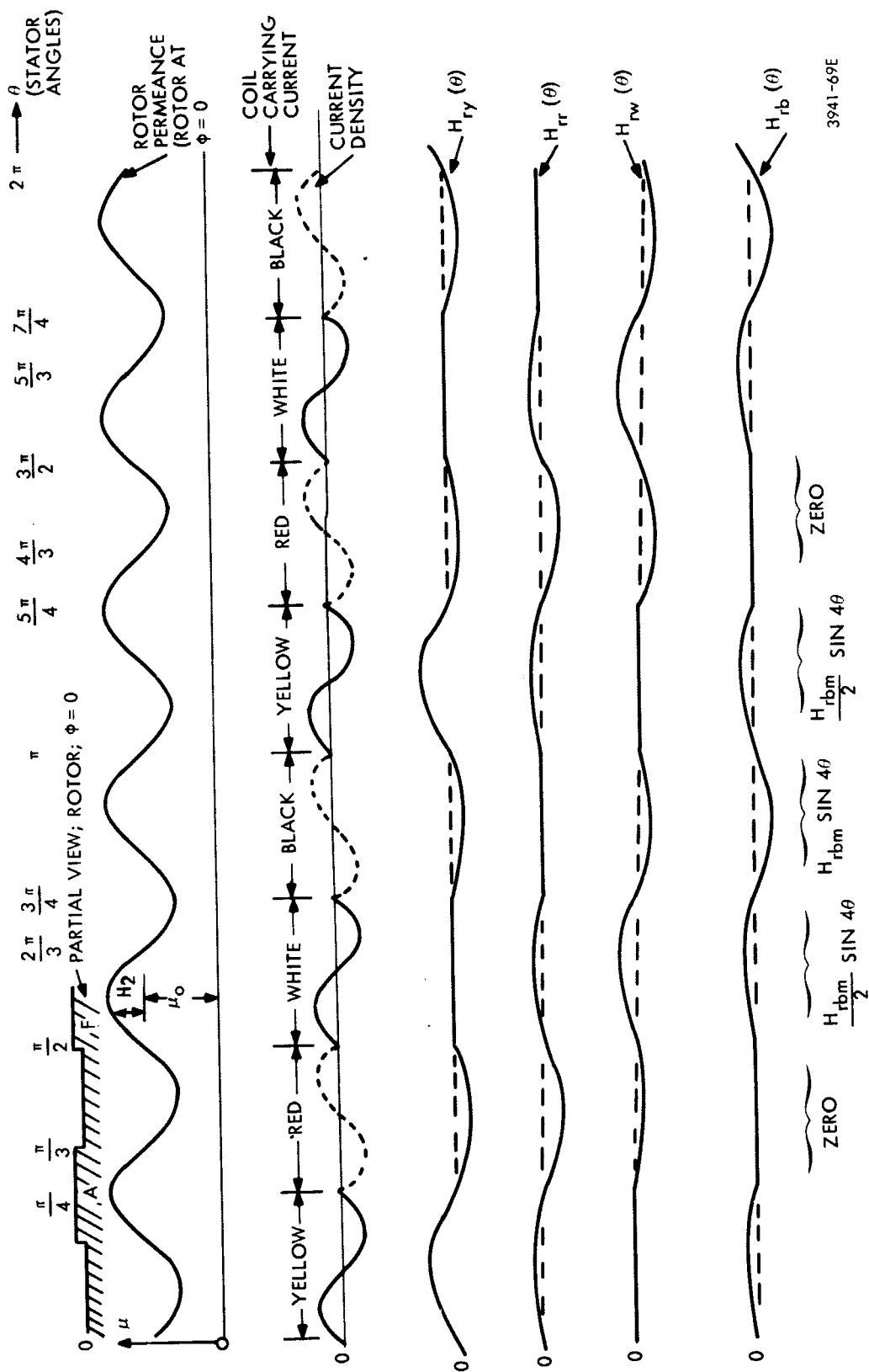


Figure 50. Step Motor Air Gap Variables

$$\begin{aligned}
H_{rym} \sin 4\theta, & \quad \text{for } 0 < \theta < \pi/4 \\
\frac{1}{2} H_{rym} \sin 4\theta, & \quad \pi/4 < \theta < \pi/2 \\
0, & \quad \pi/2 < \theta < 3\pi/4 \\
\frac{1}{2} H_{rym} \sin 4\theta, & \quad 3\pi/4 < \theta < \pi
\end{aligned} \tag{50}$$

The rest of the components of H_{ry} and the components of H_{rr} , H_{rw} , and H_{rb} can be written by inspection. The field intensity and current for the corresponding coil are related by expressions such as

$$H_{rym} = NI_y/g \tag{51}$$

where all coils have N turns and g is the airgap length in meters. The airgap energy can be written as

$$W_m = \int_V H \cdot B dV = \int_0^{\ell} \int_0^g \int_0^{2\pi} H_r B_r (r d\theta) (dx) (dz) \tag{52}$$

where x is the linear distance across the airgap, z is axial length along the stator periphery, and θ is the angle around the stator.

The airgap energy is then

$$W_m = r \ell g \int_0^{2\pi} \mu H_r^2 d\theta \tag{53}$$

where $B_r = \mu H_r$ has been introduced. Substitute for the different components of H in terms of the currents to obtain

$$W_m = \frac{r \ell N^2}{16g} \left[\begin{aligned} & (2I_y + I_r + I_b)^2 \int_0^{\pi/4} + (I_y + 2I_r + I_w)^2 \int_{\pi/4}^{\pi/2} \\ & + (I_r + 2I_w + I_b)^2 \int_{\pi/2}^{3\pi/4} \\ & + (I_y + I_w + 2I_b)^2 \int_{3\pi/4}^{\pi} + (2I_y + I_r + I_b)^2 \int_{\pi}^{5\pi/2} + \dots \end{aligned} \right] \tag{54}$$

where the abbreviation $\int_{\theta_1}^{\theta_2}$ stands for $\int_{\theta_1}^{\theta_2} \mu \sin^4 \theta d\theta$.

The integral is evaluated over the indicated ranges and with all terms collected, the energy can be written as

$$W_m = \left[\begin{aligned} &L_1 (3I_b^2 + 3I_w^2 + 3I_y^2 + 3I_r^2 + 4I_b I_y + 2I_w I_y + 4I_w I_b \\ &\quad + 4I_w I_r + 2I_b I_r + 4I_y I_r) \\ &+ L_2 \cos 6\phi (I_w^2 + I_b^2 - I_y^2 - I_r^2 + 2I_w I_b - 2I_y I_r) \\ &+ L_2 \sin 6\phi (I_w^2 + I_r^2 - I_b^2 - I_y^2 + 2I_w I_r - 2I_y I_b) \end{aligned} \right] \quad (55)$$

where

$$L_1 = \frac{\pi \mu_0 \ell r N^2}{8g} \quad \text{and} \quad L_2 = \frac{8 \mu_2 \ell r N^2}{21g}.$$

The motor torque and induced voltages are found by evaluating the partial derivatives of W_m with respect to ϕ for torque, and with respect to each of the currents I_b , I_w , I_y , and I_r for voltage. The results are as follows:

$$T_m = 6L_2 \left[\begin{aligned} &\sin 6\phi (I_y^2 + I_r^2 - I_w^2 - I_b^2 + 2I_y I_r - 2I_w I_b) \\ &+ \cos 6\phi (I_w^2 + I_r^2 - I_y^2 - I_b^2 + 2I_w I_r - 2I_y I_b) \end{aligned} \right] \quad (56)$$

$$[V] = ([R] + [Z]) \times [I] \quad (57)$$

$$[R] = R_c \times [U] \quad (58)$$

Where T_m is motor torque, $[V]$ is the matrix of coil terminal voltages, R_c is the resistance of each of the four coils, $[U]$ is the unit matrix, $[I]$ is the matrix of coil currents and the matrix order is y, r, w, b.

The $[Z]$ matrix is as follows:

$(6L_1 - 2L_2 \cos 6\phi - 2L_2 \sin 6\phi) p$ $+ 12L_2 (\sin 6\phi - \cos 6\phi) \dot{\phi}$	$(4L_1 - 2L_2 \cos 6\phi) p$ $+ 12L_2 (\sin 6\phi) \dot{\phi}$	$2L_1 p$	$(4L_1 - 2L_2 \sin 6\phi) p$ $- 12L_2 (\cos 6\phi) \dot{\phi}$
$(4L_1 - 2L_2 \cos 6\phi) p$ $+ 12L_2 \sin 6\phi \dot{\phi}$	$(6L_1 + 2L_2 \sin 6\phi - 2L_2 \cos 6\phi) p$ $+ 12L_2 (\sin 6\phi + \cos 6\phi) \dot{\phi}$	$(4L_1 + 2L_2 \sin 6\phi) p$ $+ 12L_2 (\cos 6\phi) \dot{\phi}$	$2L_1 p$
$2L_1 p$	$(4L_1 + 2L_2 \sin 6\phi) p$ $+ 12L_2 (\cos 6\phi) \dot{\phi}$	$(6L_1 + 2L_2 \cos 6\phi + 2L_2 \sin 6\phi) p$ $+ 12L_2 (\cos 6\phi - \sin 6\phi) \dot{\phi}$	$(4L_1 + 2L_2 \cos 6\phi) p$ $- 12L_2 (\sin 6\phi) \dot{\phi}$
$(4L_1 - 2L_2 \sin 6\phi) p$ $- 12L_2 (\cos 6\phi) \dot{\phi}$	$2L_1 p$	$(4L_1 + 2L_2 \cos 6\phi) p$ $- 12L_2 (\sin 6\phi) \dot{\phi}$	$(6L_1 + 2L_2 \cos 6\phi - 2L_2 \sin 6\phi) p$ $- 12L_2 (\sin 6\phi + \cos 6\phi) \dot{\phi}$

3.1.8 Structure Equations

The servomotor torque, Equation (56) is related to the motion of the antenna by equations which take into account the mass, stiffness and damping of the antenna, the gearing and the servometer characteristics. For north-south and rotational axes (ψ_{23} and ψ_{34}) the structure is approximated by a cantilever and mass with the relationships shown diagrammatically in Figure 51. The output angle (θ) shown in the figure may be identified with ψ_{23} or ψ_{34} . The derivation of the structural parameters is covered in the mechanical design section. The diagram for the east-west axis (ψ_{12}) is simpler because the structure is not cantilevered and only rotational terms are needed to describe its behavior. The case where θ is equal to ψ_{12} is obtained from Figure 51 by setting $B_{12} = B_{21} = B_{11} = K_{12} = K_{21} = K_{11} = 0$. The parameters shown on the figure have the following definitions when referred to the motor shaft.

J_m	Inertia of the servomotor rotor about its axis of rotation.
K_g	Stiffness of the gearing (shafts).
B_g	Friction of the gearing (bearings).
J_g	Inertia of output gear and that part of the structure located at the motor end of the cantilever.
$K_{11}, K_{12}, K_{21}, K_{22}$	Structural stiffness constants. See below.
$B_{11}, B_{12}, B_{21}, B_{22}$	Structural damping constants. See below.
J	Inertia of cornucopia, etc. located at outboard end of cantilever.
M	Mass of cornucopia, etc., located at outboard end of cantilever,
ℓ	Length of cantilever.

The variables appearing in the diagram are as follows:

T_m	Motor torque. See Equation (56).
$\dot{\phi}$	Rotor angular velocity.
ϕ	Rotor position.
θ_1	Rotation of motor and of cantilever.
θ_c	Rotation of the cantilever ($\theta_1 - \theta$).
T	Cantilever moment applied to inertia J .
F	Cantilever force applied to mass M .
δ	Displacement of mass M .

The stiffness and damping constants are related to structural coefficients as follows:

$$\begin{aligned} K_{11} &= k_{11} \ell^2; K_{12} = k_{12} \ell = K_{21}; K_{22} = k_{22} \\ B_{11} &= b_{11} \ell^2; B_{12} = b_{12} \ell = B_{21}; B_{22} = b_{22} \end{aligned} \quad (59)$$

The k 's are members of the stiffness matrix $[k]$, which is discussed in Section IV along with the damping coefficient matrices. Figure 51 summarizes the relationships of the structural parameters and the characteristic equations of the motor.

3.1.9 Satellite Motion Equations--Interferometer

In general, the location and orientation of the satellite does not coincide with the satellite inertial reference frame described previously. When the satellite drifts from its nominal location in space the subpoint moves relative to the surface of the earth. It is assumed that this motion is detected from the earth and that the actual subpoint is used to update the coordinates of the beacon and desired pattern center location. Updated beacon offsets are transmitted via uplink to the satellite. Consequently only the effects of satellite body rotation about the "corrected" inertial reference frame will be considered.

The order of rotation of the satellite is taken arbitrarily as y, x, z . The angles of rotation are defined as $\psi_{sy}, \psi_{sx}, \psi_{sz}$ and the rotation matrices are $[R_{sy}], [R_{sx}]$ and $[R_{sz}]$. The beacon coordinates in terms of the satellite body are given by

$$[B_{11}] = [R_{sz}] \times [R_{sx}] \times [R_{sy}] \times \begin{bmatrix} -R_e \cos \lambda_B \sin \phi_B \\ R_e \sin \lambda_B \\ R_s - R_e \cos \lambda_B \cos \phi_B \end{bmatrix} \quad (60)$$

where

R_e, R_s, λ_B , and ϕ_B have been defined earlier. The beacon coordinates in terms of the antenna reference frame are found from $[B_{11}]$ by applying $[R_{12}^a], [R_{23}^a]$, and $[R_{34}^a]$; thus

$$[B_4^a] = [R_{34}^a] \times [R_{23}^a] \times [R_{12}^a] \times [B_{11}] \quad (61)$$

3.1.10 Satellite Motion Equations - Polarization Detector

When the satellite is unperturbed and the gimbal angles are equal to ψ_{12}^d, ψ_{23}^d and ψ_{34}^d the polarization angle, θ_r , mentioned in Section 3.1.5 is caused to be zero by suitable adjustment of the polarization angle of the beacon transmitting antenna. The effect of satellite translation is eliminated by updating the polarization adjustment made on the ground. The effect of satellite rotations and gimbal angles equal to ψ_{12}^a, ψ_{23}^a , and ψ_{34}^a is described in this section. The polarization detector is assumed to sense on the y axis. The electromagnetic field polarization is changed continually. As a result,

the direction of the E-field at the satellite varies. The maximum E-field will lie on a plane which rotates about the line $\overline{OB}_4$ shown in Figure 48. The ground polarization adjustment is made so that the rotating plane includes the y_4 axis (y axis after rotations $\psi_{12}^d, \psi_{23}^d, \psi_{34}^d$; unperturbed satellite) at a reference time. The satellite-borne polarization detection subsystem is provided with an indication of the reference time by a frequency modulation technique to be described in a later section. The E-field plane which rotates around $\overline{OB}_4$ will include the polarization detector (y' axis) at a time different from the reference time if the y' axis is located by satellite rotations different from zero and gimbal angles, ψ^a , different from ψ^d . The time difference between these two events is a direct measure of the angle, θ_r , between the two planes.

The angle, θ_r , can be found if ψ^d, ψ^a , and the beacon location are known. Figure 48 shows the polarization reference plane, P , and the plane P' at an angle θ_r from P . The polarization detector lies along the y' axis of plane P' . The polarization detector at y' will experience a null output at the instant that plane P' includes OY' . The reference time instant corresponds to the instant that plane P includes the line $\overline{OY}$.

In the satellite inertial frame the sensing axis of the detector is

$$[E_1] = [R_{sy}^t] \times [R_{sx}^t] \times [R_{sz}^t] \times [R_{12}^{at}] \times [R_{23}^{at}] \times [R_{34}^{at}] \times [E_4'] \quad (62)$$

where

$$[E_4'] = \begin{bmatrix} 0 \\ 1 \\ 0 \end{bmatrix} \text{ and } [R^t] = [R^{-1}] \quad (63)$$

In the unprimed reference frame,

$$[E_4] = [R_{34}^d] \times [R_{23}^d] \times [R_{12}^d] \times [E_1]. \quad (64)$$

The coordinates of the detector can be found in terms of a reference frame which is beacon oriented with its origin at 0, its z axis along the line $\overline{OB}_4$, and its y axis along the line $\overline{OA}$. The coordinates become

$$[E_5] = [R_\alpha] \times [R_\beta] \times [E_4] \quad (65)$$

where α and β are shown on Figure 48.

Finally the sensing axis of the detector is found in a reference frame determined by rotating about the z axis ($\overline{OB}_4$) an amount θ_r so that the x coordinate of the axis is zero. This is

$$\begin{bmatrix} 0 \\ y_6 \\ z_6 \end{bmatrix} = [R_{\theta_r}] \times [E_5]. \quad (66)$$

From the definition of a counterclockwise rotation about the z axis:

$$x_5 \cos \theta_r + y_5 \sin \theta_r = 0. \quad (67)$$

Thus the expression for the polarization angle becomes:

$$\theta_r = -\tan^{-1} (x_5/y_5). \quad (68)$$

The relation between satellite motion, actual gimbal angles, and the polarization angle is shown in Figure 52.

3.1.11 System Summary

The functional relations derived in this section (3.1) can be summarized by combining the partial block diagrams shown in Figures 47, 51, and 52. The result, shown in Figure 53, places in evidence the closed loop nature of the system and also shows how satellite motions enter the closed loop as a disturbance. The desired pattern location and orientation measured in earth coordinates are converted into beacon coordinates measured relative to the antenna on the satellite by means of a computation carried out on earth. The actual beacon coordinates, relative to the satellite antenna, depend upon the antenna orientation relative to the satellite and the satellite orientation relative to its nominal (or undisturbed) orientation. The form of the relationship is indicated by the system geometry block. The difference between desired and actual beacon coordinates is converted into three signals by the interferometers and polarization detector. These three error signals are inputs to the three gimbal servos. The gimbal servos continue to rotate as long as the error signals exceed the threshold value driving the antenna pattern boresight to the desired location and orientation.

3.2 EQUIPMENT DESCRIPTION AND THEORY OF OPERATION

This section describes the operation of the pointing system in terms of specific equipments. The sensing system is discussed first in terms of the prime mode of operation and then as it functions in the redundant mode. Lastly the steering equipment is described as it operates on the error signals generated by the sensing system in each of the two modes of operation and orients the antenna relative to the satellite.

3.2.1 Sensing Equipment - Prime Mode

The sensing system operation in the prime mode will be described in terms of the equipment shown in Figure 54. This satellite-carried sensing equipment operates in conjunction with reference beacon signals transmitted from a single ground station. The diagram shows the equipment associated with one interferometer axis and the polarization axis.

One pair of interferometer antennas is shown at the upper left of the figure. These receive ground beacon signals in the prime sensing mode and transmit to the ground in the redundant mode. They are mounted in such a way that they are fixed with respect to the high power beam shaping antennas but squinted slightly off the main beam boresight direction.

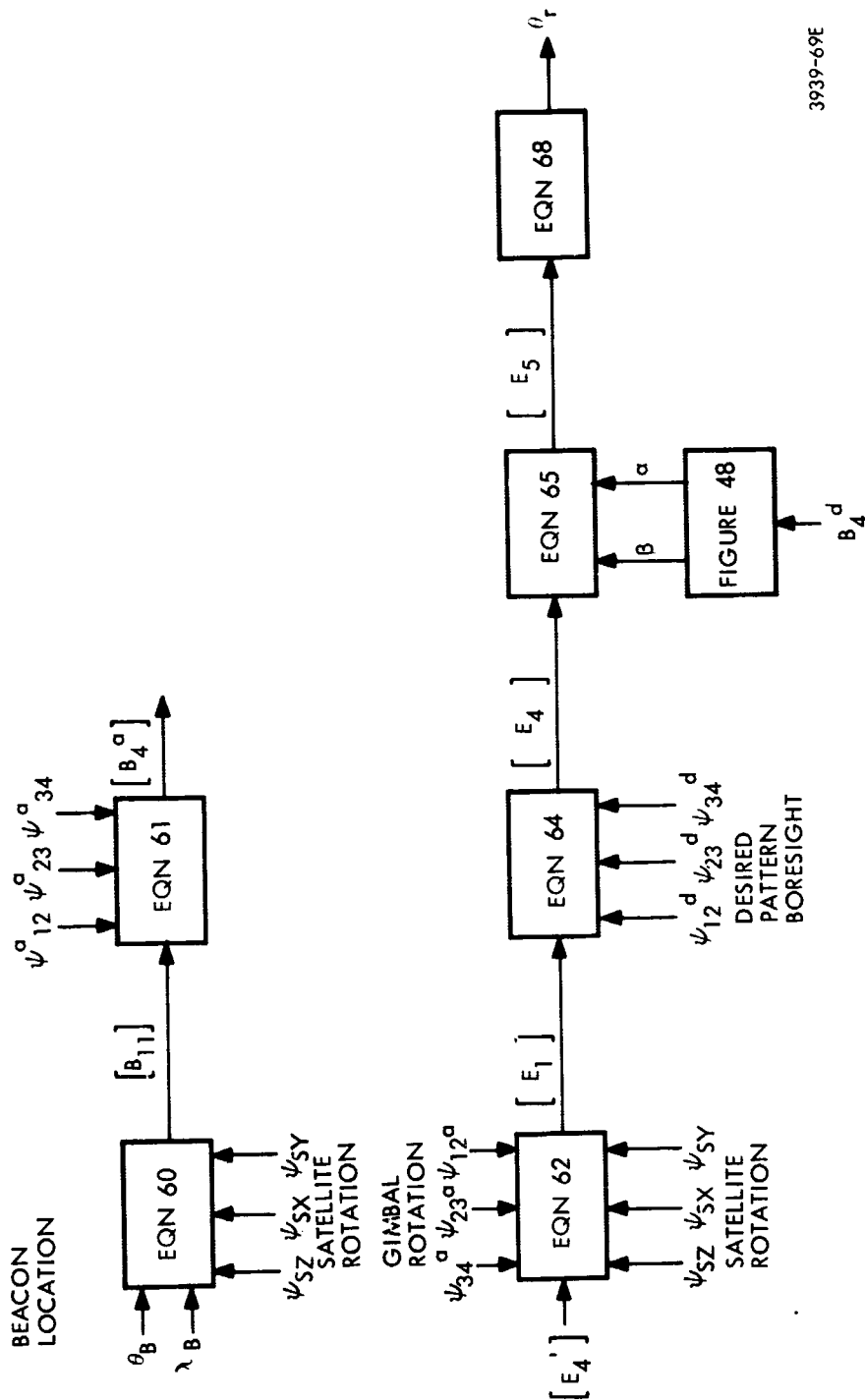
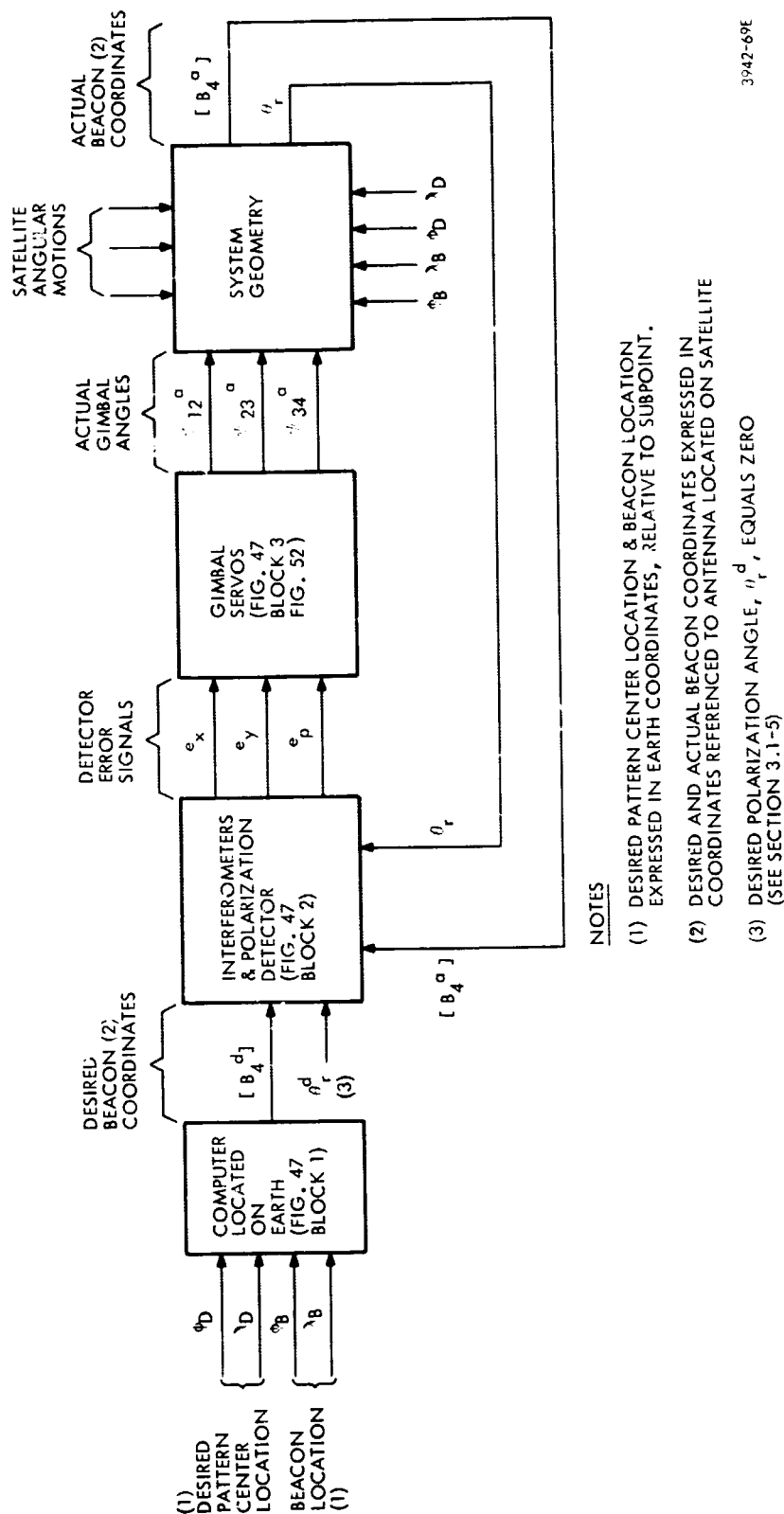


Figure 52. Partial System Block Diagram No. 3 (Satellite Disturbances)



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Figure 53. Overall System Block Diagram

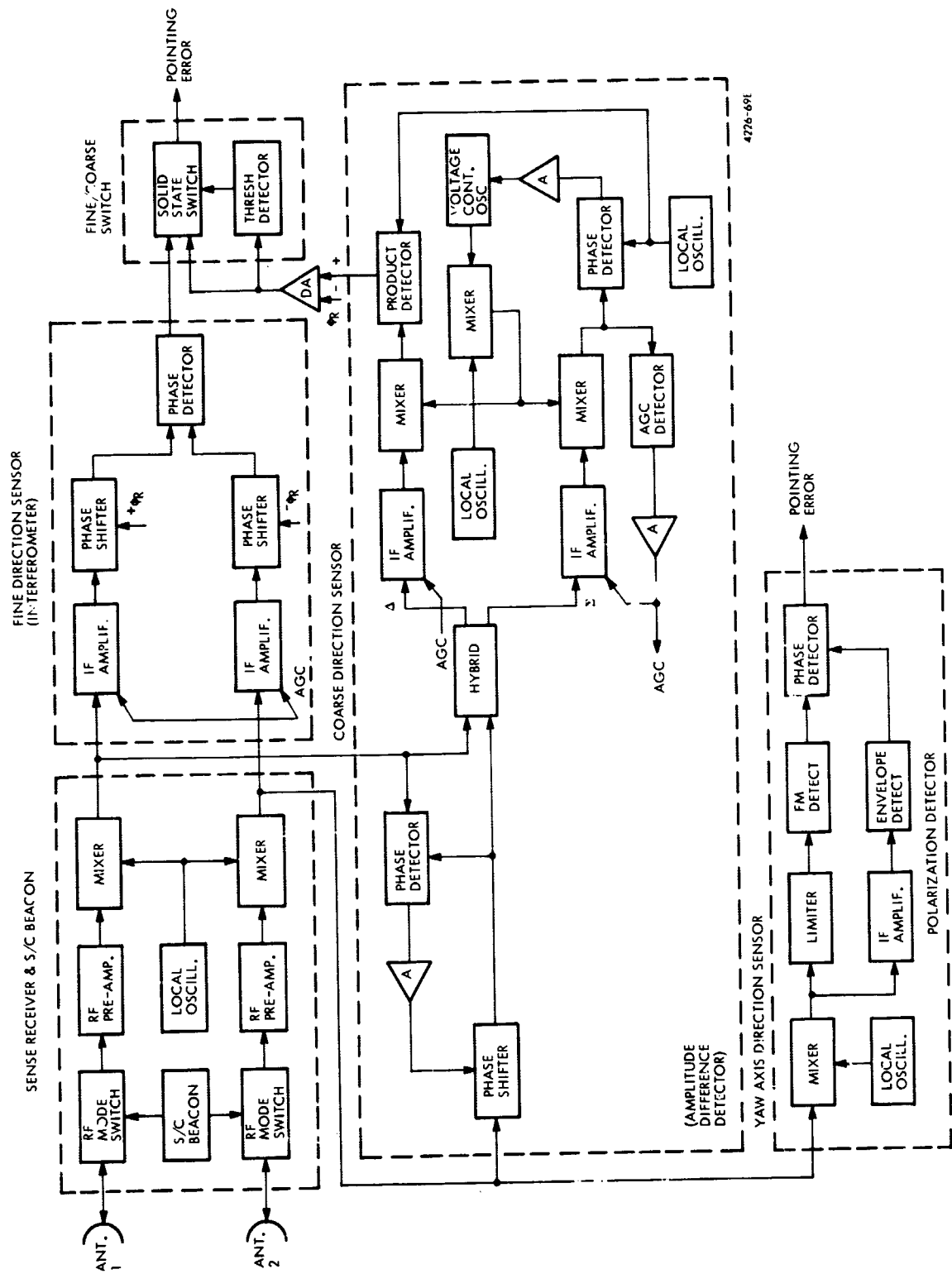


Figure 54. Block Diagram of Primary Sensing System

The signals transmitted to the spacecraft sensing system are: an unmodulated CW beacon signal at frequency $f_1 = 11.5$ GHz with circular polarization and a frequency modulated beacon signal with a center frequency $f_c = f_1 + \epsilon$ transmitted with linear polarization from an antenna rotating about its boresight axis at f_r revolutions per second ($f_r = 30$ Hz). The constant offset ϵ is to be such that f_1 and f_2 can be amplified in the same narrow band receiver channel and yet readily separated by standard frequency selection circuitry. The FM on the polarization beacon signal is to be at the rotational frequency f_r and phased to provide a reference signal for comparison with the polarization induced AM.

The RF mode switches are remotely controllable T/R switches which can handle the S/C beacon power required in the redundant mode. In the prime mode they connect the incoming beacon RF to the receiver front end. These each consist of a preamplifier and mixer. One local oscillator (LO) serves all four channels. The LO frequency is chosen to provide a convenient intermediate frequency (IF). At this point input signals for the coarse sensor are picked off. In the fine sensor each channel has an IF amplifier with AGC (only phase information is important in an interferometer). The IF amplifiers are sufficiently narrow band to block the polarization tracking signal.

The receiver channels from the four interferometer antennas through the IF amplifiers have very nearly identical phase characteristics so as to keep instrumentation errors acceptably small. This may be possible to achieve by careful matching of the electrical, thermal and temporal characteristics of the components. However, closed-loop control of the channel phase characteristics can be implemented if necessary.

Electrically controlled variable phase shifters are included in each channel. These introduce a net phase offset between the channels of each interferometer pair such that when that pair is physically oriented in the desired manner the phase difference detected by the phase detector is zero. Any error in the boresight pointing direction will cause a nonzero phase detector output which can actuate the servo system to correct the error.

The phase shifter offset inputs are computed on the ground on the basis of the known earth-satellite geometry and the desired location of the satellite transmission beam (See Section 3.1.4). The offset is transmitted to the satellite via the command uplink.

The coarse direction sensor is a simple amplitude monopulse receiver which uses the squinted interferometer antenna amplitude responses to resolve the interferometer ambiguities. The need for a coarse sensor is discussed in Section 3.3.1. The output of one sense receiver mixer is phase shifted by the phase detector-amplifier-phase shifter combination functioning in a closed loop. The purpose of this is to make the two inputs to the hybrid have the same phase. The hybrid outputs will then be the sum and difference of the two first IF amplitudes. These are amplified and mixed with a signal which phase-locks the sum signal second IF to a local oscillator by means of the phase detector and voltage controlled oscillator. The difference signal second IF will have an amplitude proportional to the absolute value of the received signal amplitude difference normalized to the amplitude sum by the AGC system. The difference will either be in-phase or out-of-phase with the local oscillator to which the sum is

phase-locked. This local oscillator may be used, therefore, as a reference signal for a product detector which will give a bipolar DC output which is a measure of the offset angle between the ground beacon direction and the line which bisects the individual interferometer antenna boresight directions. This is compared in a differential amplifier (DA) with the commanded offset angle to obtain a coarse pointing error.

The automatic gain control (AGC) system used the detected sum second IF to control the gain of all IF amplifiers. The detected difference output is thus normalized to the sum and relatively independent of received signal strength. The AGC system in the fine sensor is not critical to its operation, since the information in this channel is conveyed by the phase rather than the amplitude of the IF signals.

The fine/coarse switch functions to select the error output which drives the servo system. A threshold detector monitors the coarse pointing error and controls the solid-state switch. When the coarse error is less than approximately 1 degree, the fine error is used in the servo; otherwise the coarse is used.

The polarization detector uses the first IF from one of the interferometer antennas. This is converted to a convenient second IF and filtered to pass only the polarization beacon signal. The polarization signal is frequency modulated at the beacon with a 30-Hz phase reference and transmitted from the linearly polarized ground antenna. The polarization is caused to rotate about the boresight axis at a 30-Hz rate. The equipment used is described in Section 3.3.4. The second IF signal is also passed through a limiter which removes the AM. It is then FM detected to extract the phase reference signals. The two detected 30-Hz signals are fed to a phase detector whose output is used as a servo error.

The phase of the 30-Hz reference is adjusted on the ground so that the polarization detector will have zero output when the spacecraft antennas are correctly oriented. This phase angle is computed by using known geometric data, the characteristics of the hardware, and the desired pattern orientation.

3.2.2 Sensing Equipment - Redundant Mode

The redundant sensing system configuration is shown in block diagram form in Figure 55. The spacecraft antennas are the same for both the primary and redundant sensing modes. In the redundant mode the onboard transmitter is sequentially switched to the four interferometer antennas. The switching sequence is coded so that it may be determined on the ground which of the four spacecraft antennas is in use at any instant of time.

The ground receiver has a mixer-IF stage following the single antenna. The decommutator switches the received pulses into four channels each of which carries signals from one particular S/C antenna. Pulses from each interferometer pair are AM detected. The detector outputs are sampled, held, and compared in amplitude. These differences depend in a known way on the interferometer squint angles and are coarse measures of the S/C antenna boresight direction relative to the line-of-sight to the S/C. They are used in the ground computer to resolve the interferometer pattern ambiguities.

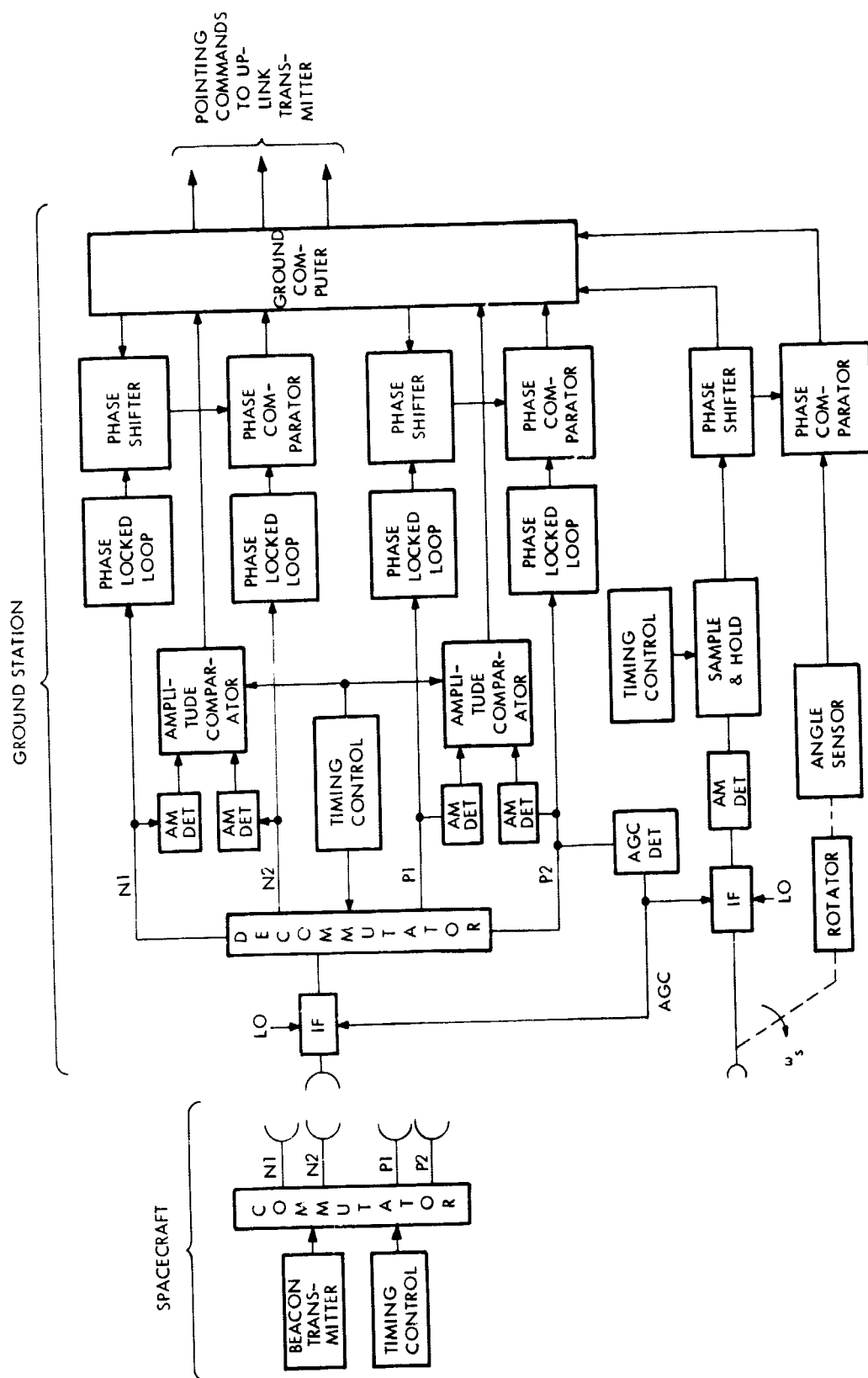


Figure 55. Block Diagram of Redundant Sensing System

The decommutated IF pulses are also fed to four phase-locked loops. These are designed so that during the time an input pulse is present they become entrained to it by means of a phase detector and VCO, while during the interpulse period the VCO frequency is held at the entrained value. By suitably designing the S/C beacon pulse train timing and the phase-locked loop, it is possible to convert the low duty cycle decommutated IF pulse into continuous signals with extremely good phase coherence relative to the incoming pulses.

The output signals from the phase-locked loops are phase-compared in pairs in the same way as in the prime system, to extract the interferometric pointing measurement. The phase shifter inputs come directly from the computer which automatically adjusts them to maintain null outputs from the phase comparators.

The redundant system polarization detector utilizes the same antenna and associated equipment that is used to transmit in the prime sensing system (See Section 3.3.4). The received pulse train is converted to IF and amplitude detected. The sample and hold is strobed by the timing control to select detected pulses from only one S/C antenna. The effective PRF must be large compared with the mechanical scan rate of the antenna in order to reconstruct the polarization induced AM accurately. The polarization angle detection technique is described in Section 3.3.4.

In effect then the redundant system is simply the prime system working in reverse. The same physical antennas are used both on the S/C and on the ground. Because the interferometer array is transmitting in the redundant mode, the ground receiver must be somewhat more complex than the S/C receiver. This is not considered a problem, however, since the ground equipment is not subject to stringent constraints on size, weight and power consumption.

3.2.3 Steering Equipment

Each of the three antenna axes is provided with analog to digital error conversion and servo control equipment. The timing and control logic shown in Figure 56 is common to all three axes, thus, the operation of only one axis is described.

Analog input (pointing error) is converted to parallel digital data by means of a differential amplifier (DA), threshold detector, gate (G), oscillator, up-down converter, and digital-to-analog converter. During each conversion time interval the gate, G, is enabled by the convert command. If the threshold detector level exceeds a predetermined amount, local-oscillator (OSC) pulses are entered into the up-down counter. The counter output is converted to analog signals by the DA converter. The counter output increases until the analog output equals the analog input (pointing error plus bias). The bias level is set at a value corresponding to one half the full plus-to-minus range of signal. Thus, when the analog pointing error signal is zero the counter will be half full; a totally filled counter corresponds to full clockwise error; a totally empty counter corresponds to full counterclockwise error. Accordingly, the most significant counter bit (0 or 1) corresponds to clockwise or counterclockwise error; that is, the most significant digit is an error sign indication.

Once the conversion interval is complete, the servo motor control circuits become active. A set of step command pulses is issued by the timing logic. Each pulse

Figure 56. Detailed Block Diagram-One Axis

counts the counter down and enables the gate at the counter output. The counter output pulse is delivered to the proper (clockwise or counterclockwise) channel of the motor driver circuits by means of the forward-reverse logic circuits. The error sign bit (UDC MSD) controls the forward-reverse logic. The motor driver circuit advances the motor excitation by one increment and causes the motor to step in the proper direction. This action continues upon receipt of successive step commands until the up-down counter is empty. Further step commands are gated out and do not cause any motion. The counter remains empty until the next convert command is received at which time the conversion process is repeated.

The uplink decoder is used to convert digital signals received via uplink to analog offset commands which are used by the antenna orientation sensing system as reference signals.

Operation of the system in the secondary mode ($\bar{M}$) is indicated on the same figure. The major difference is that the servo command signal in the up-down counter is equal to a value received from the ground via uplink. The command signal is parallel loaded into the up-down counter. Servo motion occurs between the parallel load time intervals.

The three servos are stepped in sequence; the interlacing is controlled by the three step commands furnished by the timing and control logic (step 1, step 2, and step 3). The control logic also furnishes beacon timing signals for use in the secondary mode ($\bar{M}$). Shift pulses control loading of the uplink decoder input register. The uplink receiver digital logic circuits are synchronized with the timing logic shown to ensure that the uplink decoder register is not loaded at the same time the uplink receiver output registers are changing.

3.3 EQUIPMENT DESIGN PARAMETERS

This section defines in detail the specific equipment characteristics that have been selected for the design and presents some of the reasons for the selection. Alternate equipments and techniques that were considered are subsequently discussed in Section 3.4.

3.3.1 Interferometer Sensors

The interferometer in its simplest form is illustrated in Figure 48. It consists of two antennas separated by a distance and each feeding a phase meter. The phase difference $\Delta\theta$ in the received signals is related to the angular direction of the source θ . The differential length between the two reception paths Δl is given by

$$\Delta l = d \sin \theta. \quad (69)$$

If the received signals have a wavelength λ , the relative phase between them will be 2π radians for each wavelength of path difference. Thus,

$$\Delta\phi = \frac{2\pi}{\lambda} \Delta l = \frac{2\pi d}{\lambda} \sin \theta. \quad (70)$$

A sensitivity function S that relates the incremental change in measured phase shift per unit change in angle of arrival θ can be defined as

$$S = \frac{d(\Delta\phi)}{d\theta} = \frac{2\pi d}{\lambda} \cos \theta. \quad (71)$$

Sensing a change of 1 degree in measured phase shift is well within the state-of-the-art.

If the one degree phase uncertainty is to correspond to an angular error of 0.01-degree at the antenna the sensitivity must have a value given by

$$S = 1^\circ / 0.01^\circ = 100 \quad (72)$$

with this value of sensitivity the spacing to wavelength ratio can be found and is given by

$$\frac{d}{\lambda} = \frac{S}{2\pi \cos \theta} \leq \frac{100}{2 \cos 4.3^\circ} = 16 \quad (73)$$

where $\cos \theta$ has been evaluated at the extreme of the pattern coverage region. Wavelength λ is related to frequency f by

$$\lambda \text{ (ft.)} = 0.984/f \text{ (GHz)}. \quad (74)$$

Thus, if we choose a beacon frequency in the vicinity of 12 GHz, the wavelength will be

$$\lambda = 0.984/12 = 0.082 \text{ ft.} \quad (75)$$

$$\text{and } d = 16 \lambda = 1.3 \text{ ft. (39.7 cm)}. \quad (76)$$

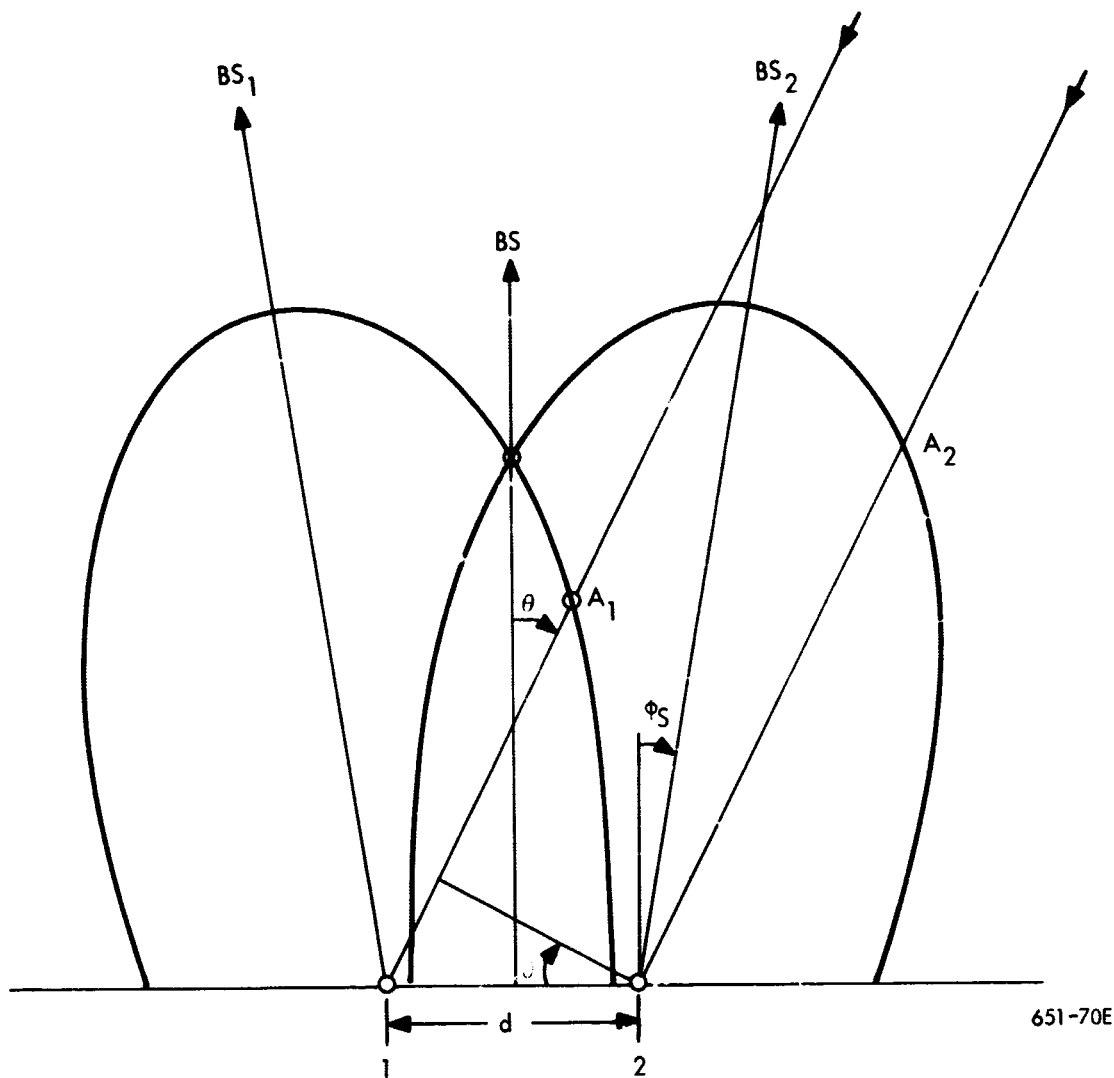
While the interferometer will thus yield measurements of angle of arrival with good accuracy, these measurements are ambiguous. To obtain the required angular coverage and sensitivity, the spacing must be chosen so large that $\Delta\theta$ goes through several cycles of 2π radians. In particular for $d/\lambda = 16$ and $\theta_{\max} = 4.3^\circ$

$$\Delta\theta_{\max} = \frac{2\pi d}{\lambda} \sin \theta_{\max} = 2\pi (1.2) \text{ rad.} \quad (77)$$

Thus, the total phase shift is more than $\pm \pi$ radians and provision must be made in the sensing system to resolve this ambiguity. This is done by squinting the basic interferometer apertures and making use of simple amplitude monopulse measurements.

The antenna pattern geometry required to implement the squinting technique is illustrated in Figure 57. The individual antenna patterns have boresight directions BS_1 and BS_2 which are tilted to either side of the composite boresight direction BS by the squint angle θ_s . Beacon signals arriving from the direction θ will be received with amplitude A_1 at antenna 1 and amplitude A_2 at antenna 2. Comparison of these amplitudes yields a coarse measurement of θ .

To meet the sensing system requirements the individual interferometer antennas should have sufficient beamwidth so that adequate signal strength is maintained for



651-70E

Figure 57. Interferometer Antenna Boresight Squinting

angles of arrival between ± 4.3 degrees. A beamwidth specification of 10 degrees is adequate for this purpose assuming that relatively small squint angles can be used.

To check this assumption the amplitude monopulse patterns corresponding to different squint angles were computed based on a simple sinusoidal primary antenna response. The results indicated that a 10-degree bandwidth and a squint angle in the range of from 1 to 3 degrees will yield coarse sensor performance adequate to resolve the interferometer ambiguities. A standard and relatively simple monopulse receiver can be used to process the signals and generate coarse errors for the servos.

3.3.2 Polarization Angle Sensors

The simplest imaginable technique for sensing polarization consists of aiming a linearly polarized antenna towards the source of received signal and then rotating the antenna about its boresight direction until either a maximum or a minimum is observed in the received signal strength. The antenna output depends on the cosine of the angle ρ between the direction of the incident polarization vector and the receiving antenna sensitive direction. Thus the received signal e may be represented as

$$e = A \cos \rho. \quad (78)$$

If the orientation of the polarization vector is fixed with respect to the earth then the measurement of e onboard the spacecraft can be interpreted as a measurement of the spacecraft antenna rotation orientation with respect to the reference earth-fixed direction. There are, however, difficulties associated with this simple scheme. For one thing the system is sensitive to low frequency variations in propagation losses, i.e., to the slowly varying component of fading. For another the measurement sensitivity which may be defined as

$$S_m = \frac{de}{d\rho} = -A \sin \rho \quad (79)$$

varies in magnitude from zero to A , the maximum value occurring when $\sin \rho = \pm 1$, i.e., when $\rho = \pm 90$. Therefore, to optimize the system sensitivity, the response null should be tracked. This means that either the spacecraft or the ground beacon antennas should be rotated to maintain the desired null condition. Clearly, the steering of a spacecraft polarization tracking antenna, independent of the main pattern, is much more difficult to implement than rotation of the ground beacon antenna. Thus, it is proposed to orient the polarization beacon antenna to provide at the spacecraft a polarization vector aligned such that a null response corresponds to the correct rotational position of the main pattern. There are still problems with this simple system in obtaining adequate sensor accuracy. If the system noise level near the response null amounts to as much as 1 percent of full scale, then there will be uncertainty in the null location as measured in the receiver of $\Delta e \approx 0.01A$. For a system sensitivity of A the equivalent uncertainty in polarization vector angle $\Delta \rho$ is

$$\Delta \rho = e/S_m = 0.01A/A = 0.01 \text{ rad} = 0.57^\circ. \quad (80)$$

It can be seen from this result that, even if the measurement noise is less than the assumed value, the angular accuracy is marginal in relation to the allowable error, as discussed in the error analysis section.

These difficulties are alleviated by making the following modification. First, continuously rotate the ground beacon antenna at an angular frequency which is high compared with the atmospheric fading spectral components and yet low enough not to present mechanical problems. In order to avoid the requirements for physically rotating the antenna feed itself, it is proposed to use a rotary half-wave plate as described in Section 3.3.4. The effect on the transmitted signal is the same as though the feed were rotated. A frequency of 30 RPS is a suitable value. The polarization receiver output will be amplitude modulated at this frequency. The receiver then will include narrow band filtering and thus the system noise level significantly lowered. To improve the sensitivity, a 30-Hz reference signal is transmitted to the spacecraft using FM. This reference signal is FM detected and compared in phase with the AM. The phase difference is a direct measure of the main pattern rotational alignment relative to the FM reference. Commanded offset angles are introduced into this system on the ground by controlling the phase of the FM.

The system described above will provide an order-of-magnitude improvement in error performance over the direct null sensing technique partly because of the noise reduction resulting from narrow-banding, and partly because a phase null can be measured more accurately than an amplitude null. With this technique pattern rotation sensing errors can be kept below a few tenths of a degree. This level is adequate to meet overall system specifications.

The redundant mode sensing of rotational motion of the spacecraft antenna about its boresight axis can be accomplished on the ground by sensing the plane of polarization of a linearly polarized beacon signal transmitted from the spacecraft. By rotating the half-wave plate as described in Section 3.3.4, the measurement sensitivity can be enhanced just as in the primary mode using the same antenna and feed system. The polarization tracking ground receiver must decommutate the time multiplexed satellite signal to extract transmission from one particular spacecraft antenna.

3.3.3 Servo Control Equipment

The servo motor output shafts are coupled to the antenna by means of gearing which is described in detail elsewhere. The choice of step-motor size and gear ratio is made on the basis of a compromise between: motor capabilities and size, gear ratio, gear complexity, the requirements imposed by the specification (pattern slew rate), and the load (antenna inertia). The numerical value of the gear ratio is a convenient parameter on which to base the following trade-offs. The larger the gear ratio the faster the motor must turn for a specified antenna slew (maximum) rate. The larger the gear ratio the smaller the reflected inertia. Stepper motor speed capability increases with decreasing reflected load. Any given stepper motor has a maximum rate at which it can turn in the total absence of reflected load. The larger the gear ratio the smaller the reflected load torque and, therefore, larger ratios result in the use of small, high speed motors. A small motor requires less power from the motor driver circuit than a larger motor.

Because of the low slew rate requirements of the specification, any one of several motor-gear combinations is satisfactory. The specification requires that the target be moved from any point in the coverage area to any other point in 15 minutes or less. This corresponds to a gimbal angle change of 9 degrees or less. In addition,

the gimbal rate must be capable of compensating for 1 degree per second satellite drift rate. The total slew rate requirement is approximately equal to 1.7 degree per second.

The velocity constant of the system described is equal to the reciprocal of the conversion period. This assertion may be verified by referring to Figure 58. In the figure, the responses to two input ramps are shown. The larger ramp is one which the system is just capable of following by stepping continually between conversion times. The maximum rate is R_m , the corresponding error is E_m , and the two are related by $R_m T_c = E_m$. For the smaller input, say at $R_m/2$, the error is reduced by the same factor (at least). The ratio of rate to error is the velocity constant, K .

It is seen that $K = \frac{R_m/2}{E_m/2}$

Therefore, $K = R_m/E_m$. But this is numerically equal to $1/T_c$.

In conclusion, $K = 1/T_c$. (81)

The threshold detector is set to operate at a magnitude corresponding to one step. Thus, if the error at any conversion time is incrementally less than one step, no corrective antenna motion will occur.

Three "designs" are compared in Table 11 as a function of the gear ratio. Two motors are considered; both produce 15-degree steps at the motor shaft each time the excitation is advanced.

It can be seen that a size 15 motor with 5000 to 1 gear ratio is capable of meeting the slew specification ($3 > 1.7$); operates well within the manufacturer's specified maximum motor rate at the load of 23 G-cm² ($16.67 < 300$); and has velocity and quantization error of acceptable magnitude (0.01 degree and 0.003 degree). The same comments are true for the size 11 motor at 5000:1 ratio. The size 15 motor at 2500:1 ratio has a greater slew capability margin ($6 > 1.7$) but also has a greater quantization error (0.006 instead of 0.003). The lower gear ratio would require one less mesh and thus provide a slight improvement in reliability. Weight and size differences are negligible.

Since the optimization process is relatively insensitive to parameter choice in the range considered, no further investigation was pursued and somewhat arbitrarily the size 15 motor and 2500:1 ratio was chosen. This particular motor ratio choice was studied by means of digital computer simulation of the motor's describing equations and the equations corresponding to a rigidly-coupled 90 G-cm load. The simulated motor was capable of 100 step/second motion, consistent with manufacturers specifications.

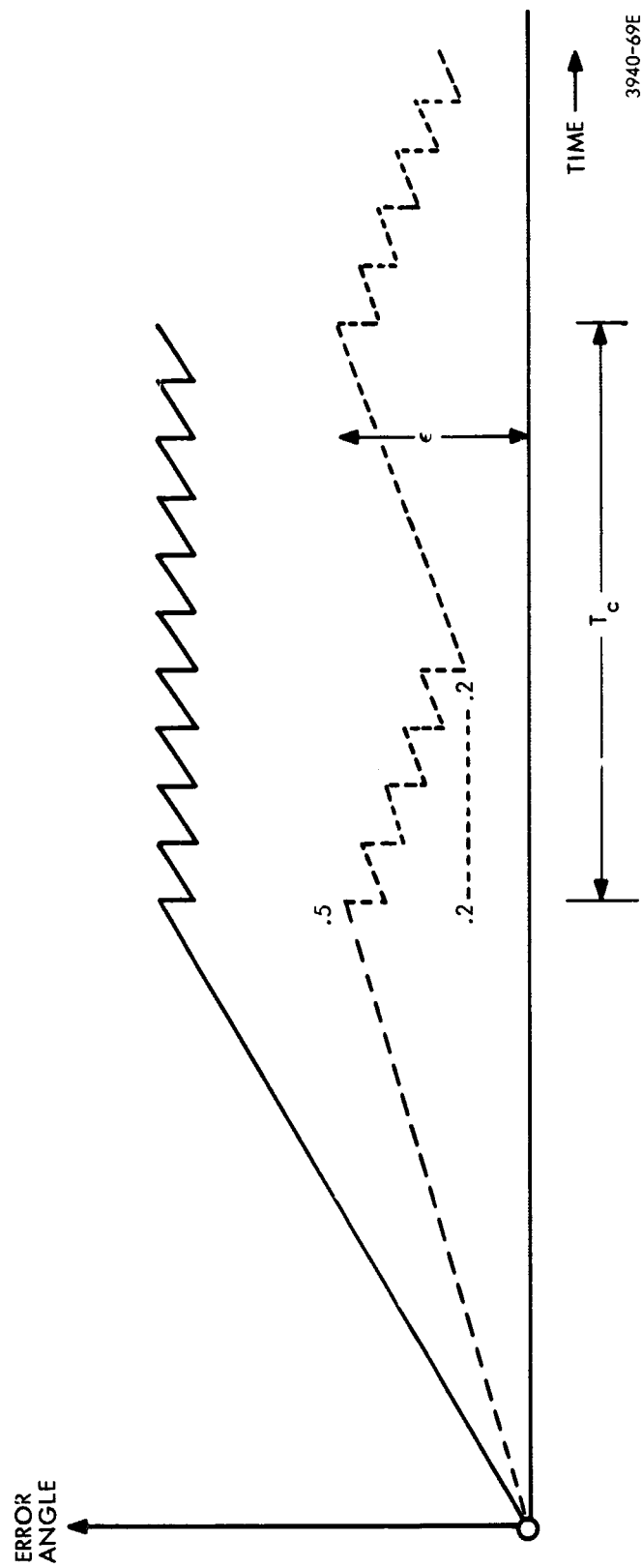


Figure 58. Response to a Ramp Input

TABLE 11
SERVO EQUIPMENT PARAMETRIC DESIGN

GEAR REDUCTION	2500	5000	7500
STEP SIZE (degrees)	0.006	0.003	0.002
STEP FREQUENCY (steps/second)	16.67	16.67	16.67
STEP FREQUENCY (steps/minute)	1000	1000	1000
MAXIMUM RATE (degree/minute)	6	3	2
STEP PERIOD (seconds)	0.06	0.06	0.06
CONVERSION PERIOD (seconds)	0.6	0.6	0.6
STEPS/CONVERSION	10	10	10
VELOCITY CONSTANT (1/seconds)	1.67	1.67	1.67
VELOCITY ERROR (degrees)	0.01	0.01	0.01
QUANTIZATION ERROR (degrees)	0.006	0.003	0.002
REFLECTED LOAD INERTIA (G-cm ²)	92	23	10.2
SIZE 15 MOTOR CAPABILITY (steps/second)	150	300	400
SIZE 11 MOTOR CAPABILITY (steps/second)	will not operate	100	150

3.3.4 Beacon Terminal Equipment

3.3.4.1 Link Computations

The interferometers (NS and EW) and the polarization rotation sensor perform their function by comparing the phase between two signals. In the case of an interferometer in the prime mode, the phase measurement is made between the signals that arrive at the two horns. In the redundant mode, phase measurement is made at the ground terminal site between signals that are alternately radiated from the two horn antennas forming the interferometer on the satellite.

To make the phase comparison with an accuracy compatible with the system pointing requirements, a comparatively high signal to noise ratio is required. Specifically, the error in measuring phase is given by: ³

$$\delta\phi = \frac{1}{(2 C/N)^{1/2}} \quad (82)$$

where

$\delta\phi$ = phase error in degrees

C/N = carrier to noise ratio.

If a C/N ratio of 30 db is used, the error in the phase measurement is 0.0224 degrees. Since the interferometer has a sensitivity of 100, this error corresponds to a pointing error contribution of 0.0224×10^{-2} degrees. This value when RSS'd with the other system errors does not contribute significantly to the pointing error. Thus, a carrier to noise ratio of 30 db is adequate for the interferometer in terms of the 0.1-degree pointing requirement. Similarly, for the polarization rotation sensor, the 30 db C/N ratio is adequate for the spacecraft rotation measurement.

Since the downlink is generally the limiting portion of the link, the redundant mode, wherein the interferometer transmits to the terrestrial beacon site, was considered first in defining terrestrial equipment. Furthermore, a 0 dbW transmitter power level at the satellite interferometer was chosen as a reasonable compromise between satellite and earth terminal equipment complexity. A center frequency of 11.5 GHz was tentatively picked in the analysis to minimize antenna equipment size. In an actual system, a somewhat higher frequency would most likely be used to minimize interaction between the interferometers and the communications channels.

At the ground terminal, an uncooled parametric receiver has been assumed to minimize complexity of the terminal. The receiver will have a noise temperature of approximately 470°K. For the NS and EW interferometers the ground terminal antenna will be circularly polarized and the interferometer horns in the satellite will be linearly polarized. While this imposes a 3-db loss, it permits the simultaneous use of one of the interferometer horns in the rotation correction function where polarization angle is measured. In the interferometers the circular polarization minimizes signal variation due to spacecraft rotation. The ground terminal equipment also incorporates a rotating linearly polarized transmit and receive capability which is used in the rotation correction

polarization angle measurement. Operation of the NS-EW interferometers is at f_0 (nominally 11.5 GHz), while the polarization angle measurement function is at $f_0 + \delta f$. δf is required to be sufficiently large to enable RF diplexing of f_0 and $f_0 + \delta f$, while yet small enough so that a single receiver in the satellite can be used to process both the interferometer and polarization angle signals.

The downlink computations are summarized in Table 12 for the NS-EW interferometers. A ground terminal gain of 49 db is required. At 12.0 GHz, this corresponds to an antenna diameter of 10 feet. The equipment characteristics outlined in Table 12 are also adequate for the polarization angle measurement necessary for the redundant rotation correction function. The operation of the latter system is described in Section 3.3.4.3.

Operation of the beacon terminal in the prime mode (uplink) requires the radiation of appropriate CW and AM/FM signals to the satellite for the interferometer and polarization angle measurements, respectively. The design of the beacon terminal was such that the same ground antenna was used for the uplink or prime mode as in the redundant mode. In other words, the available gain for the terrestrial antenna in the prime mode was set at nominally 49 db.

The receiving equipment in the satellite for the prime mode uses a receiver with a 6 db noise figure front end. Effective bandwidth in the receiver was set at 1 kHz. Losses in the feedline and switching network (1.1 db) together with the fact that the satellite antennas view the earth, result in a degradation in the receiver noise temperature to approximately 1458°K. Use of these satellite receiver characteristics results in the uplink summary given in Table 13. Requirement of a 30 db C/N ratio to obtain the necessary system accuracy sets the beacon transmitter power at 8 dbW.

3.3.4.2 Beacon Terminal RF Equipment

The beacon terminal equipment block diagram is shown in Figure 59. The basic characteristics of the RF equipment are summarized in Table 14. In the prime mode for the NS-EW interferometer, the transmit signal at f_0 is routed into a polarizing quadrature hybrid as shown in Figure 59. Outputs of the hybrid pass through diplexers to an orthomode transducer at the feed. At this point, the necessary space and phase quadrature between signals is present for circular polarization. This signal is used by the spacecraft interferometer as described in Section 3.3.1. The diplexers permit simultaneous use of the feed for the polarization angle measurement.

In the redundant interferometer mode signals arriving from the individual spacecraft interferometer antennas are routed to the interferometer receiver by correct positioning of switch SW. Operation of the receiver is given in Section 3.2.2.

The polarization antenna measurement capability is incorporated into the feed by providing a separate elliptically polarized signal which is amplitude-modulated by rotation of the polarization ellipse. The elliptically polarized signal is provided by a polarizing section as shown in Figure 59, which is mechanically rotated and coupled to the transmitter in the prime mode by a rotary joint. Output of the polarizing section is fed into an orthomode transducer. Output from this transducer goes through a diplexing network which enables connection to the feed orthomode transducer.

TABLE 12

INTERFEROMETER DOWNLINK SUMMARY - REDUNDANT MODE

Free Space Path Loss (12.0 GHz)		-206 db
Rain Attenuation		-6 db
Total Path Loss		-212 db
Satellite Transmitter Power		0 dbW
Satellite Antenna Gain	24 db	
Pattern Loss ($\pm 10^\circ$ Satellite Motion)	-3 db	
Polarization Loss	-3 db	
Time Loss	-0.5 db	
Switch Loss	<u>-1.8 db</u>	
	+15.7 db	
Equipment Gain (less ground terminal gain)		+15.7 dbW
Required C/N Ratio		30.0 db
Allowable Effective Receiver Noise Power		-226.3 dbW
Effective Receiver Noise Density (470°K)		-202 db/Hz
Effective Receiver Bandwidth (300 Hz)		24.8 db
Ground Terminal Antenna Gain, G_R		49 db
$(G_R - 226.3 + 202 = 24.8)$		

TABLE 13

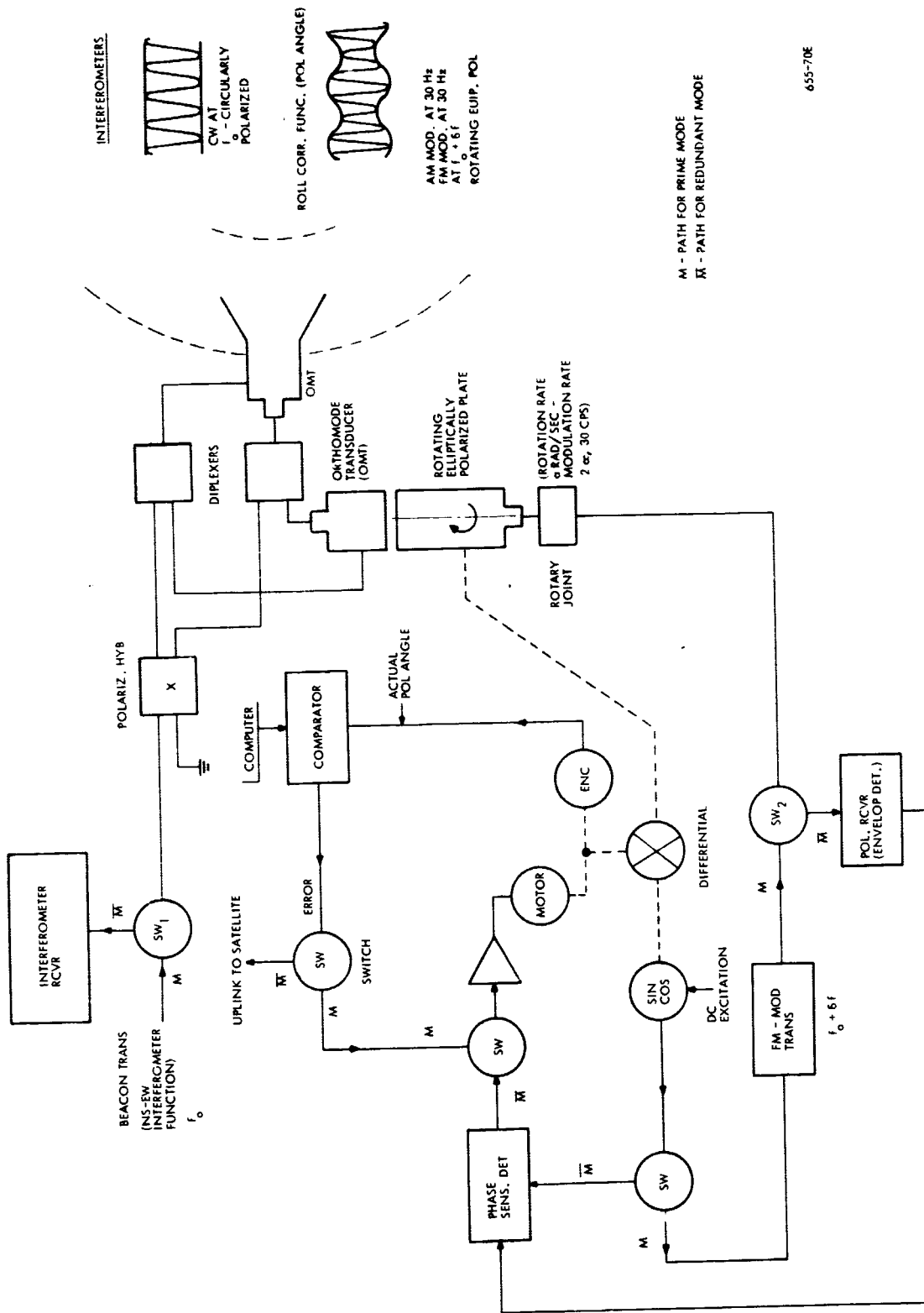
INTERFEROMETER UPLINK SUMMARY - PRIME MODE

Free Space Path Loss (12.0 GHz)		-206 db
Rain Attenuation		-6 db
Total Path Loss		-212 db
Equipment Gain (less transmitter power)		
Satellite Antenna Gain	24 db	
Pattern Loss ($\pm 10^\circ$ satellite motion)	-3 db	
Polarization Loss	-3 db	
Ground Terminal Antenna Gain	<u>49 db</u>	
	+67 db	67 dbW
Required C/N Ratio		30 db
Allowable Effective Receiver Noise Power		-175 dbW
Effective Receiver Noise Density (1458°K)		-197 db/Hz
Effective Receiver Bandwidth (1000 Hz)		30 db
Ground Terminal Transmitter Power, P_T		8 dbW
$(P_T + 197 - 175 = 30)$		

TABLE 14

BEACON TERMINAL CHARACTERISTICS

	PRIME MODE	REDUNDANT MODE
FREQUENCY	≈ 12.0 GHz	≈ 12.0 GHz
ANTENNA DIAMETER	10 FT. (3.04 m)	10 FT. (3.04 m)
ANTENNA GAIN	49 db	49 db
POLARIZATION		
NS-EW INTERFEROMETER FUNCTION	CIRCULAR	CIRCULAR
ROTATION CORRECTION FUNCTION	ROTATING LINEAR	ROTATING LINEAR
GROUND RECEIVING SYSTEM NOISE TEMP	-	470°K
GROUND TERMINAL TRANSMITTER POWER	8 dbW	-
SATELLITE TRANSMITTER POWER	-	0 dbW



655-70E

Figure 59. Beacon Terminal Equipment

simultaneously with the interferometer signals. Rotation of the polarizing section at a rate of α radians per second causes the radiated signal as viewed with a linearly polarized receiver to be amplitude modulated at 2α radians per second. In the system operation a phase comparison is made between the amplitude modulated signal caused by mechanical rotation and the impressed FM modulation on the transmitter operating at $f_0 + \delta f$. Detail of this polarization measurement used for satellite roll correction is given in Section 3.3.4.3.

3.3.4.3 Beacon Terminal Polarization Angle Measurement Equipment

The polarized angle system includes electromechanical elements with function in conjunction with the RF elements in the manner described next. The same electromechanical elements are used in both prime and redundant modes with suitable reconnection as described.

In the prime mode, the elliptical plate assembly rotates at 15 cps, as described in Section 3.3.4.2 and shown in Figure 59. This shaft rotation is coupled to a sin-cosine transducer via a 2:1 gearbox and a mechanical differential. The sin-cosine transducer is excited with DC. Its output voltage is a sine-wave at 30 cps. The phase of the 30-cps sine wave can be adjusted relative to the "phase" of the elliptical wave by rotating the second differential input shaft. The 30 cps sine wave is used to control the frequency deviation of the output of a frequency modulator. In this manner the FM signal shown as $f_0 + \delta f$ on Figure 59 is produced. The phase shift input to the differential is servo controlled in the prime mode so that it is numerically equal to the value of phase shift determined by the ground computer. The computer output and the output of an encoder on the phase-shifter shaft are compared to derive the servo error signal. The computer output is the value of phase shift which must be introduced on the ground in order that the satellite polarization axis servo (yaw) always seeks a null between the AM and FM detected signals at the polarization receiver.

In the redundant mode, the input to the phase-shifter servo is the phase difference between the envelope detected signal obtained from the ground receiver (whose phase is a measure of the satellite antenna polarization angle) and the output of the sin-cosine transducer. The servo acts to null the difference. The amount of motion required at the second differential input shaft is equal to the actual polarization angle. In the redundant mode this motion, suitably encoded, is compared to the computer output which is the same value used in the prime mode. Any difference is the error signal for the airborne yaw axis servo and is sent to the servo via uplink. Switch-over from prime to redundant mode should result in no motion of the phase-shifter servo or the yaw-axis servo (except for problems occurring in the receiver due to the time required to overcome switching transients in band-pass or smoothing filters).

3.3.5 Ground Station Computation

The ground station computation requirements for support of the pointing system in the prime mode of operation consists of solving the coordinate transformation equations to determine the desired offset angles for transmission to the satellite. Corrections for satellite orbital drift and other predictable perturbations can be computed if necessary. Under normal circumstances the required computations will need

to be updated at relatively infrequent intervals. No problems are anticipated in satisfying either the solution time or the accuracy requirements with a modest computer.

The requirements for ground computation in the redundant sensing mode are essentially identical to those of the prime mode. The major difference is that the error computations must now be made on a real time basis consistent with the allowable pointing errors and the maximum satellite drift rate since the computer is now part of the closed position loop.

3.4 ALTERNATE COMPONENTS AND TECHNIQUES

This section describes the alternate components and techniques associated with the pointing system that received significant attention during the design process.

3.4.1 EW and NS Sensing Techniques

Several alternative sensing techniques were considered during the design. Simple direction finding systems using a highly directive antenna and both amplitude and phase monopulse systems were studied analytically. None of these methods was capable of sufficient sensitivity to meet the specified pointing accuracy requirements. Thus, the interferometric technique was selected without a competitive alternative.

The use of one interferometer antenna as a common member of each crossarm pair, and thereby reducing the number of antennas required from four to three, was considered. Difficulties are inherent in this configuration since the coarse sensing system requires the squinting of the common antenna in two orthogonal directions. The resulting asymmetrical beamwidth and gain requirements were found to involve more complications in the antenna and signal processing systems than could be reasonably justified.

Two alternate techniques were considered for resolving the interferometer ambiguities. One involved the use of additional interferometer apertures with reduced spacing to provide a coarse but unambiguous measurement of angle of arrival. The other means involved obtaining an independent measure of pointing direction at some initial time, perhaps using the redundant sensing system, and then keeping running track of subsequent motions relative to the initial direction.

The first of these techniques has been used successfully with very long baseline ($d/\lambda > 50$) ground-based interferometers where quite good accuracy was needed to resolve multiple closely-spaced ambiguities. It has the disadvantage, however, of requiring additional antennas on the spacecraft. The tracking method has the serious drawback that a complete reacquisition procedure is needed in the event of a momentary loss of track. Upon careful consideration of these possibilities it was concluded that squinting of the interferometer antennas represented the best solution to the ambiguity problem. The increase in the receiver requirements was modest in comparison with the mechanical problems of mounting additional antennas on the spacecraft or the functional problems of the tracking method.

3.4.2 Rotation Sensing Techniques

An alternate rotation sensing technique that was considered utilizes two ground beacons at locations with maximum separation consistent with the 25 degree elevation requirement. This approach has been shown to provide adequate EW and NS measurement sensitivity. However, a study of the geometry for sensing rotation about the boresight direction revealed problems. For optimum sensitivity this beacon should be as far off to the side as possible (ideally, 90 degrees onboard another satellite). However, to satisfy the 25-degree elevation requirement this beacon can be no more than 7.9 degrees from the subpoint direction. Thus, the rotation measurement component must be obtained from beacons located between 7.9 and 15.8 degrees of the boresight direction. This information is cross-coupled to the pattern translation measurements, thus requiring additional on-board computation to extract the rotational measurement components. These components occur at much lower sensitivities and are subject to further degradation by the computations. Another drawback of this rotation-sensing technique is that it requires sensing antennas with broader beamwidths than if only one beacon location were used, thus, further reducing the measurement sensitivity.

3.4.3 Steering System Alternates

Alternates that were considered include gimbal sequence, position encoders and motors. In each of these three cases, system considerations were paramount in making the choice of type of component.

The three gimbals are referred to as inner, middle and outer. The inner set is closest to the main body of the satellite. The three motions to be accomplished with respect to pattern motion on the earth are east-west, north-south and pattern rotation. The six combinations are summarized in Table 15. Combinations 1 and 2 are not discrete choices since the middle and inner gimbal function is interchanged throughout the day. This choice is mechanically possible but would require additional complexity to resolve the pointing errors into steering commands. Combinations 3 and 5 give rise to gimbal lock every twelve hours when two sets of gimbals become parallel. Combinations 4 and 6 are both feasible, in that neither requires error resolution nor suffers from gimbal lock. Combination 6 was chosen because it appeared to be more straightforward to describe analytically.

The choice of using position encoders depends upon the desirability of closing a position feedback loop around the antenna servo on the satellite. If this were done the pointing errors would have to be converted into position rate signals by means of an integrator on the satellite in the prime mode or in the earth-borne computer in the redundant mode. Since the system, when operating normally, will not experience repeated pattern location command changes and since the platform disturbances are characterized by low-frequency drift phenomena, the additional stabilization achieved by the position loop does not appear to be warranted from a system standpoint. Eliminating the loop avoids the use of an encoder and an integrator, thus simplifying the resultant system.

The choice of a continuous-motion or a discrete-motion motor depends upon the kind of servo used. Since a digital control system is comprised of two-state devices,

TABLE 15

ALTERNATE GIMBAL ARRANGEMENTS

Number	Outer Gimbal	Middle Gimbal	Inner Gimbal	Comments
1	East-West	North-South	Rotation	Role of middle and inner gimbals is interchanged every six hours
2		Rotation	North-South	
3	North-South	East-West	Rotation	Gimbal lock occurs every twelve hours
4		Rotation	East-West	
5	Rotation	East-West	North-South	Gimbal lock occurs every twelve hours
6		Rotation	East-West	

it is relatively unaffected by variations in component performance during the lifetime of the system. In this sense it is more reliable than an analog system. In addition, the use of devices in either a saturated-on or a totally-unconductive-off state often leads to a more efficient system with lower heat dissipation. Since the servo errors will be predominantly those due to low frequency disturbances (satellite orientation drift or diurnal east-west rotation) the use of digital signals will not adversely affect the ability of the system to point with desired accuracy. A stepping-motor is a logical choice for use with digital servos since it is basically a two-state device (as far as any given winding is concerned). The specific choice of motor between the major types (variable reluctance or permanent magnet) does not affect the preliminary design carried out in the present study. Both types are available with stator windings only (avoiding the use of brushes, slip-rings or solid-state commutating devices). Accordingly, a reluctance type stepper motor was chosen.

3.5 ERROR ANALYSIS

Pointing angle errors arise due to errors in the three gimbal angles, caused, in turn, by errors due to atmospheric anomalies, receiver equipment inaccuracies, servo equipment errors, and thermal distortion. The three gimbal axis errors combine to produce a pointing error as shown in Figure 60. The expression for pointing error is

$$d\theta_p = [d\psi_{12}^2 + (d\psi_{23} + 0.039 d\psi_{34})^2]^{1/2} \quad (83)$$

Because of the similarity of the servo and receiver equipment it was assumed that $d\psi_{12} \cong d\psi_{23}$. On the other hand, the polarization detector associated with the ψ_{34} axis is not expected to have commensurate error magnitudes to those of an interferometer. The trade-off curve shown in Figure 61 is based on the error relationships shown in Figure 60. It was used as a guide in apportioning sensing system errors between the interferometer and the polarization angle detectors. The various error sources are discussed in the following sections and then summarized in a combined error tabulation.

3.5.1 Atmospheric Propagation Errors

One of the sources of error in the sensing system is the turbulence and inhomogeneities in the atmosphere which cause a variety of effects in an electromagnetic wave propagating through it.

3.5.1.1 Interferometer Sensing Errors

For the interferometer measurements, only phase and direction of arrival fluctuations are of interest, assuming, of course, that the signal amplitude can be maintained well above the ambient noise. The interferometer receives two signals from a distant single source. The absolute phase shift over each of the paths is subject to random fluctuation about some mean value. The difference in these mean values is what is measured. If the two signals were to follow identical paths, they would be subject to the same phase fluctuations and the random disturbances would cancel in the phase comparison. However, as the receiving antennas are separated, the two paths are separated and the correlation between the phase fluctuations in the two received

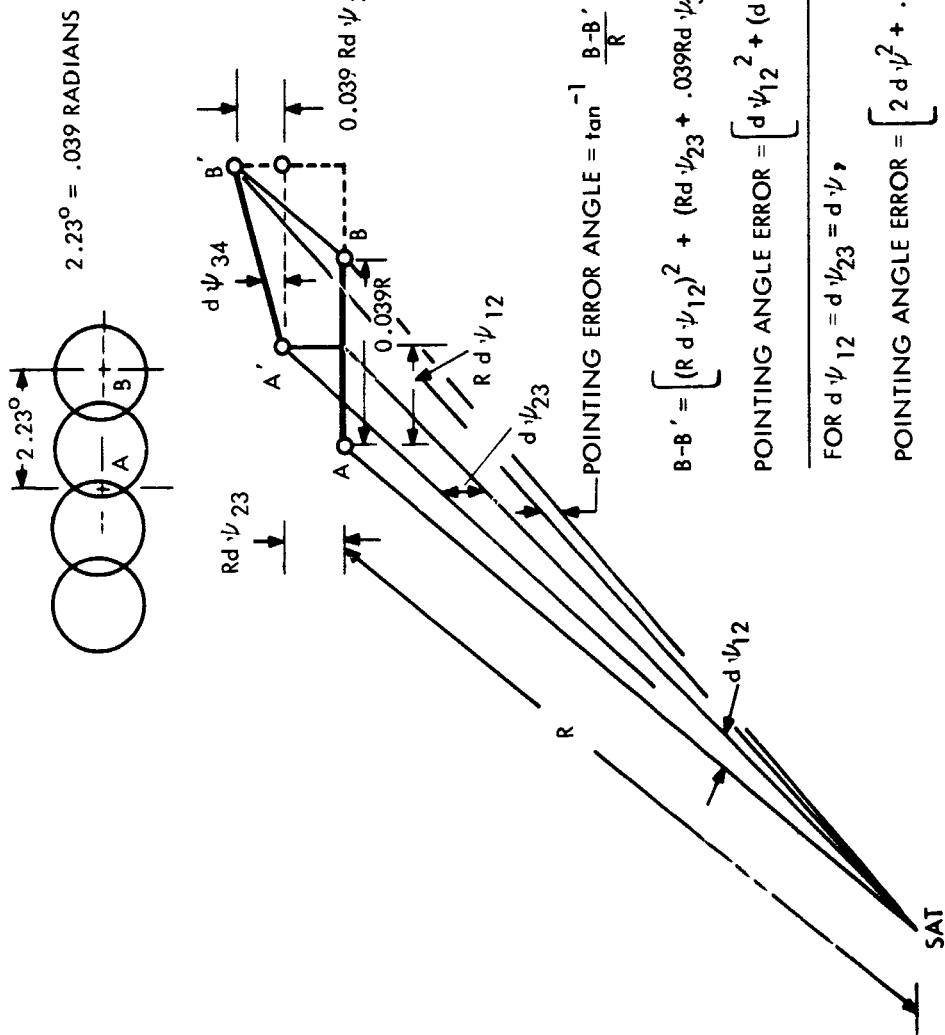


Figure 60. Geometry of Pointing Angle Errors

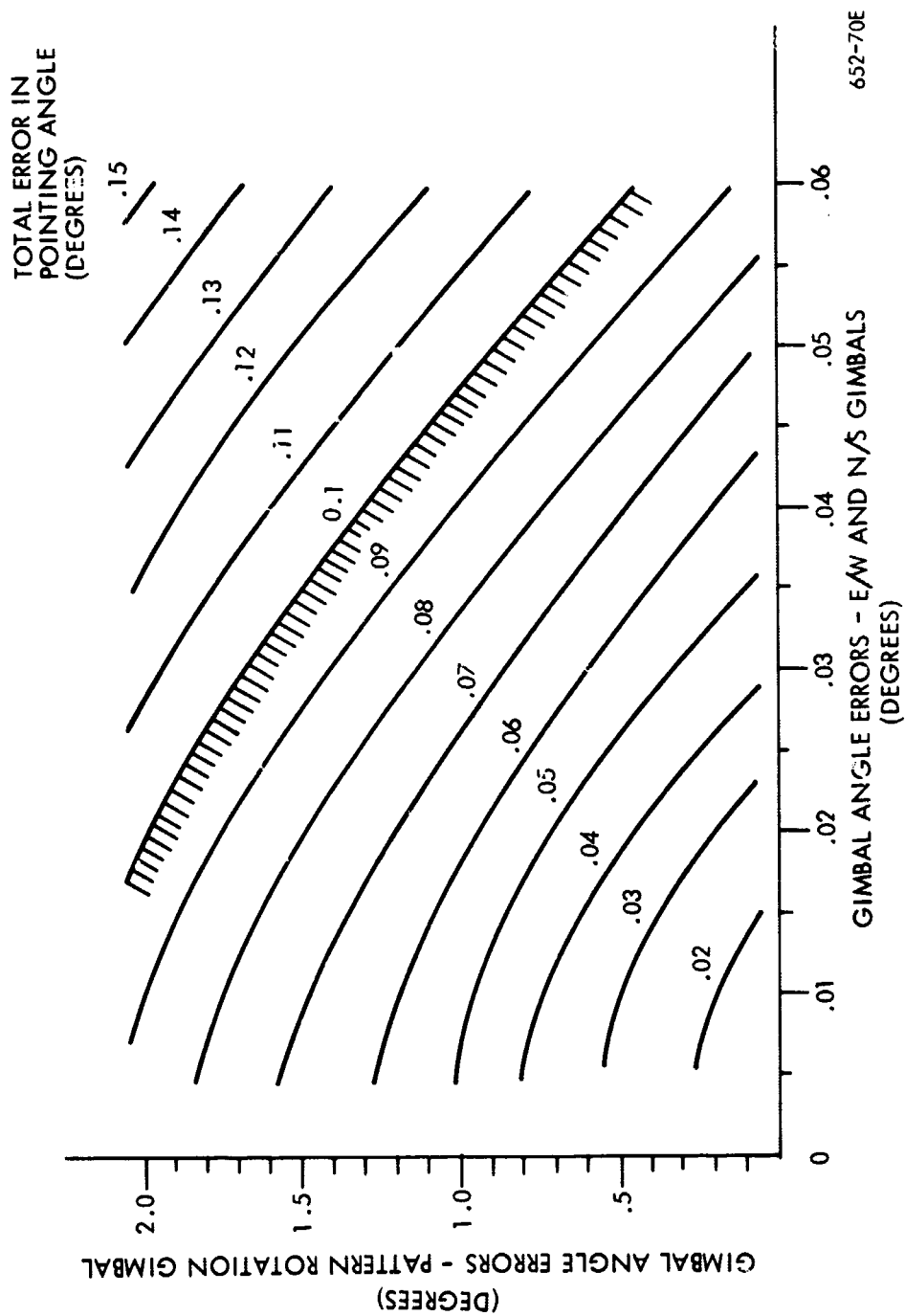


Figure 61. Pointing Error Due to Gimbal Angle Errors

signals will decrease. Since the interferometer is used to measure the angle of arrival of the beam signal, it is useful to relate random fluctuations in angle of arrival θ to those in the phase angle of the received signal. To a good approximation, this relationship can be expressed as

$$\sigma_{\theta} = \frac{\lambda}{2\pi d} [2 \sigma_{\phi}^2 (1 - c(d))]^{1/2} \quad (84)$$

where

- σ_{θ} = std. deviation in angle of arrival
- σ_{ϕ} = std. deviation in phase ϕ of received signal
- λ = wavelength of beacon signal
- d = interferometer aperture spacing
- $c(d)$ = correlation function of random phase fluctuations in two received signals.

A number of experiments have been performed to determine empirically atmospheric effects on interferometer performance. Data taken at 10 GHz over a near-horizontal 15-mile path in Hawaii⁴ has been summarized in Figure 62. Unfortunately, no source of experimental interferometer measurement noise is available for high elevation paths through the entire atmosphere, and there is no clear-cut theoretical basis for relating the information which is available to the problem of interest. It is possible, however, to draw some conclusions based on the following observations:

1. The data presented in Figure 62 was taken over a nearly horizontal path through tropical air.
2. Based on the density vs. altitude profile of the atmosphere, 95 percent of the atmosphere is below a height of 15 miles.
3. Available radar refraction data⁵ show that for elevations above 60 degrees, essentially all refraction takes place below 15 miles altitude and that the magnitude of such refractions is less than 25 percent of the refraction occurring over 15-mile near-horizontal paths.
4. For a beacon at the center of the required coverage area, the elevation to the satellite is approximately 73 degrees.

Considering these observations together it seems reasonable to expect that the interferometer measurement noise introduced by the atmosphere will not be more than that plotted in Figure 62. For a value of d/λ of 16 and a frequency near 12 GHz, it is predicted that wide-band RMS measurement noise will be less than 0.0015 degrees. This conservatively pessimistic prediction is small compared to the other errors arising in the sensing system.

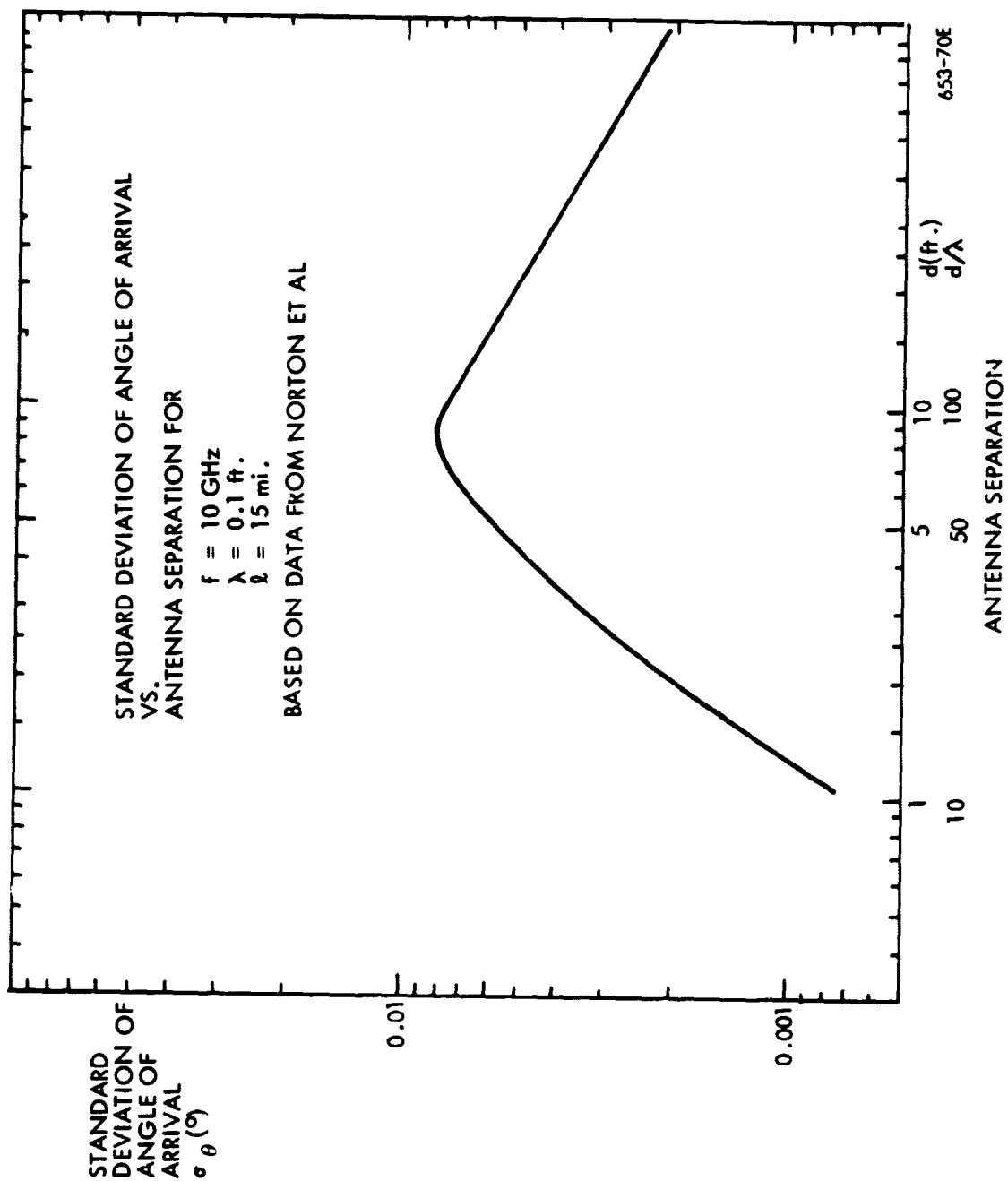


Figure 62. Standard Derivation of Angle of Arrival vs Antenna Separation

Fluctuation in the angle of arrival of the beacon signal at the interferometer receiving antennas is caused by atmospheric refraction. This effect has a DC or mean value which is a well-known function of range and elevation. For elevations above 60 degrees, refraction is less than 0.01 degree for any range. This component of error which is already quite small can be calibrated out of the system, since the beacon-satellite geometry is essentially invariant. Rapid random fluctuations in refraction about the mean do occur and cause the image of a source to quiver. However, the magnitude of this effect is small compared with the mean value and may therefore be neglected so far as the interferometer system is concerned.

3.5.1.2 Polarization Sensing Errors

Errors in the polarization tracking system are introduced as a result of rotation of the wave polarization vector in the ionosphere. Physically, this rotation is the result of interaction between the wave and the charged particles in the atmosphere in the presence of the earth's magnetic field, and is an effect which is inversely proportional to the square of the frequency. According to generally accepted theories⁶, the one-way angular rotation of the polarization vector is given by the expression:

$$\Omega(h) = \frac{4.36 \times 10^4}{f^2} \int_{h_1}^{h_2} g(h) H \cos \theta N_e dh \quad (85)$$

where

f = frequency in Hz

$$g(h) = \frac{r_e + h}{[(r_e + h)^2 - (r_e \cos E)^2]^{1/2}}$$

r_e = radius of earth

h = vertical height variable

E = elevation angle

H = earth's magnetic field strength

θ = angle between H and direction of propagation

N_e = electron density.

If we assume a transmission path from a beacon on the earth near the center of the coverage area to the satellite, then the elevation angle E will be large ($\approx 70^\circ$), the magnetic field strength H near its equatorial value (0.31 gauss) and the angle θ large enough ($< 89^\circ$) for quasi-longitudinal conditions to prevail. For these conditions the predicted polarization rotation through the atmosphere is less than 0.1 degree at a frequency of 12 GHz. This prediction is based on conservative assumptions with the

actual value being far less under nighttime conditions. It should also be pointed out that this predicted rotation is the sum of a relatively large but slowly varying component and a very small random component. The former component can be predicted and, therefore, largely corrected if more stringent design requirements were warranted.

3.5.1.3 Redundant Mode Errors

Errors caused by atmospheric effects when in the redundant sensing mode do not differ significantly from those expected in the primary system. This is due to the basic similarity of the two systems in the primary system, including the use of the same beacon frequencies, ground based equipments and ground station location.

3.5.2 Receiver/Detector Errors

Receiver and detector errors are evaluated in both the prime and redundant mode for one interferometer channel and the polarization sensing channel. The two interferometer channels are assumed to contribute identical components to the total pointing error.

3.5.2.1 Interferometer Sensing Errors - Prime Mode

The interferometer receiver has two separate channels which ideally should have identical phase characteristics. While they can be aligned and calibrated initially to a very high accuracy, temperature variations and aging will inevitably produce differential phase shifts in the two channels. The magnitude of this effect depends on exactly how the receiver is designed. At the very least, the two channels must be matched electrically and thermally. If necessary, it would be possible to introduce a pilot signal into the front end of both channels and match their phase characteristics under closed-loop control using the pilot signal to sense any differential phase shift. Pending a detailed receiver design analysis, the differential phase shift error has been estimated at 0.1 degree. Because the design sensitivity of the interferometer is 100, this corresponds to a cross pointing error of 0.001 degree.

It is proposed that two electronically controlled phase shifters be used in the interferometer channel operating in a push-pull configuration to compensate for non-linearities. Latching ferrite phase-shifters are capable of producing continuous phase variations of sufficient magnitude to cover the required range. The IF frequency can be selected to best match the phase shifter characteristics. Residual non-linearities can be calibrated out to a considerable degree. Random phase shift errors can be kept under 1 degree with state-of-the-art devices. This equates to 0.01 degrees of pointing error.

Because of the introduction of the commanded offset angle as a calibrated phase shift, the nominal phase detector output is zero, i.e., it operates as a null detector. This type of operation reduces the effect of inaccuracies to a minimum and it is estimated that phase errors due to this source can be kept below 0.1 degree which is equivalent to 0.001 degrees pointing error.

3.5.2.2 Polarization Angle Sensing - Prime Mode

In the polarization sensor, the measurement information is contained in the phase of the 30-Hz frequency and amplitude modulated signals. Because of the low frequency involved, it is easier to obtain good phase matching in the two polarization channels than in the interferometer. This is due to the fundamental limitation on time delay stability. The same delay errors in the receiver have much less effect at 30 Hz than at the much higher IF of the interferometer receiver. No special circuitry is needed therefore, in order to obtain a differential phase error in the two channels of less than 0.1 degree. For the polarization channel, all phase errors are directly related to pointing error.

For the reasons presented in the previous paragraph, no particular difficulty is expected in designing a 30-Hz phase shifter having an operating range of ± 25 degrees to an accuracy of 0.2 degree. Note that this component is ground based and may be subjected to close environmental control and recalibration.

The same error contribution has been assigned to the polarization angle phase detector as to the one used in the interferometer receiver. Inaccuracies of less than 0.1 degree are well within the state-of-the-art at the frequency of interest.

3.5.2.3 Interferometer Sensing Errors - Redundant Mode

The redundant sensing system is very similar in operation to the primary system. The main difference between the two is the addition of time multiplexing in the redundant system interferometer channels. As a consequence of this difference the redundant system hardware requirements are more complicated than those of the primary system and thus there are sources of error not present in the primary system. However, since the receiving portion of the necessary hardware is ground-based, its performance can be optimized without imposing spacecraft size, weight and power consumption constraints.

The spacecraft beacon transmits sequentially via each of the four interferometer antennas. Ideally, the phase shift through each of these four channels should be equal and fixed. Since each channel consists only of a switch, a length of cable and an antenna, the initial matching of phase characteristics should be very stable with time. The error due to differential phase shift in the transmitter channels has been estimated at 0.001-degree angular cross pointing error (0.1-degree phase error at the beacon frequency).

The ground receiver channel up to the input of the phase-locked loop consists of an antenna, a decommutator and an IF amplifier. Any variation in the differential phase shift between paired channels will cause an error in the sensed angle of arrival. Various techniques including automatic phase compensation based on a pilot signal could be used to achieve minimum differential phase shift characteristics of the basic receiving channels. However, the time multiplexing of these channels necessitates operation in a sampled data mode which will reduce the accuracy in comparison to a continuous system such as the primary sensing system. This source of error is allocated 0.6 degree in phase or 0.006 degree in pointing angle.

The function of the phase locked loop is to reconstruct the phase information contained in a sequence of received RF pulses. It operates in effect as a sample and hold circuit. By using a highly stable VCO and by carefully designing the feedback loop it is possible to realize phase-locked loop performance such that system angular error due to this cause will be less than 0.006 degree (0.6 degree phase error at the phase-locked loop operating frequency).

The same type of phase shifter can be used in both the primary and the redundant sensing systems. Because it is readily accessible for recalibration, the redundant system phase shifter error has been estimated at one half that of the primary system. With a phase shifter error under 0.5 degree in phase, its contribution to the system angular error will be less than 0.005 degree.

The same error allocation has been made for the redundant phase comparator as for the primary system. The phase comparator is always operated in a narrow range about its zero output point. System angular errors from this source can readily be held below 0.001 degree.

It is assumed that in the digital processing of sensing system measurements and in the digital transmission of computed uplink commands no significant errors are introduced. For this assumption to be valid the data channel must have a minimum capability both in terms of precision (bits/word) and also capacity (words/sec). These requirements are well within the state-of-the-art and no difficulty is anticipated in satisfying them. This applies to both the interferometer and the polarization channels of the sensing system.

3.5.2.4 Polarization Angle Sensing - Redundant Mode

The polarization receiver for the redundant sensing system is less complex than that required for the primary system. Consequently, its contribution to the system error is estimated at about half the value for the comparable portion of the primary system thus introducing an angular error of 0.05 degree.

The polarization receiver phase shifter and phase comparator in the redundant sensing system can be identical to the primary system phase shifter and phase comparator. The angular errors associated with these two subsystems are therefore 0.20 and 0.10 degrees, respectively.

3.5.3 Servo Errors

There are three errors which are attributable to the servo system and they are identical in each axis. The errors are quantization, dynamic or velocity lag, and bias drift.

The quantization error arises due to the dead band of the threshold detector and is equivalent to the step size of the motor reflected thru the gearing to the sensing axis. As described in Section 3.3.3, the step size has been set at 0.006 degree.

The velocity constant has been set equal to 1.7. Thus, at the maximum satellite drift rate of 1 degree/minute or 0.0167 degree/sec, the velocity-lag error is 0.01 degree.

Bias signals and DA converter signals are derived from the same reference source. Bias or calibration errors arise from differential drifts, due to aging, in the values of the precision resistors which are used in the bias and DA networks. The maximum error which is to be converted is 10 steps in magnitude and thus a one per-cent drift in resistor value corresponds to less than 0.001 degree error.

3.5.4 Thermal Distortion Errors

Thermal distortion errors have been evaluated at both symmetric (12 noon and 12 midnight) and non-symmetric (6 AM and 6 PM) sun angles. The error sources and methods of analysis are detailed in Section 4.4. The worst case results of this analysis are included in the pointing error summary.

3.5.5 Pointing Error Summary

Except for the components of the thermal distortion errors, all errors are considered to be statistically uncorrelated. Accordingly RSS errors in the three gimbal axes are obtained and these errors are then combined in accordance with Figure 60 to obtain an overall pointing error. The non-symmetric sun angle thermal errors are slightly greater than the symmetric sun angle errors derived in Section 4 and are used in the tabulations shown in Table 16 for the primary sensing system and Table 17 for the redundant sensing system. The principal source of error is thermal distortion about the east-west (pitch) axis. All other errors are nearly negligible, in comparison. The total error is within the allowable error of 0.1 degree.

TABLE 16
POINTING ERROR SUMMARY
(Primary Sensing System)

<u>ERROR SOURCE</u>	$\frac{\Delta\psi_{12}}{(\text{E-W})}$	$\frac{\Delta\psi_{23}}{(\text{N-S})}$	$\frac{\Delta\psi_{34}}{(\text{Rotation})}$
Propagation	0.0015°	0.0015°	0.1°
<u>Receiver/Detector</u>			
Receiver Channel	0.001	0.001	0.1
Phase Shifter	0.01	0.01	0.2
Phase Detector	<u>0.001</u>	<u>0.001</u>	<u>0.1</u>
RSS Total	0.011	0.011	0.245
<u>Serv</u>			
Quantization	0.006	0.006	0.006
Velocity Lag	0.01	0.01	0.01
Bias Drift	<u>0.001</u>	<u>0.001</u>	<u>0.001</u>
RSS Total	0.012	0.012	0.012
<u>Thermal</u>			
Interferometer	0.040	0.0004	-
Cornucopia	<u>0.038</u>	-	-
Arithmetic Total	<u>0.078</u>	<u>0.0004</u>	-
RSS GRAND TOTAL	0.080°	0.016°	0.264°

$$\text{TOTAL POINTING ERROR} = ([0.080]^2 + [0.016 + (0.039)(0.264)]^2)^{1/2} = 0.084^\circ$$

TABLE 17
POINTING ERROR SUMMARY
(Redundant Sensing System)

ERROR SOURCE

	$\frac{\Delta\psi_{12}}{(E-W)}$	$\frac{\Delta\psi_{23}}{(N-S)}$	$\frac{\Delta\psi_{34}}{(Rotation)}$
Propagation	0.0015°	0.0015°	0.1°
<u>Transmitter/Receiver/Detector</u>			
Transmit Channel	0.001	0.001	-
Receive Channel	0.006	0.006	0.05
Phase Lock Loop	0.006	0.006	-
Phase Shifter	0.005	0.005	0.2
Phase Comparator	<u>0.001</u>	<u>0.001</u>	<u>0.1</u>
RSS Total	0.010	0.010	0.23
<u>Servo</u>			
Quantization	0.006	0.006	0.006
Velocity Lag	0.01	0.01	0.01
Bias Drift	<u>0.001</u>	<u>0.001</u>	<u>0.001</u>
RSS Total	0.012	0.012	0.012
<u>Thermal</u>			
Interferometer	0.040	0.0004	-
Cornucopia	<u>0.038</u>	-	-
Arithmetic Total	<u>0.078</u>	<u>0.0004</u>	-
RSS GRAND TOTAL	0.080°	0.016°	0.25°

TOTAL POINTING ERROR = 0.084°

SECTION IV

MECHANICAL DESIGN

The mechanical design of the antenna includes three significant aspects: hardware, structural, and thermal. Hardware design is concerned with the physical layout of components and the associated design of bearings, gearing, lubrication, friction, leakage rates and partial pressures. The structural design sizes the various structural members to survive the worst case dynamic loadings and interacts with the physical design concept to assure that structural weight is being used efficiently. The thermal design establishes a controlled temperature environment for the antenna components such that materials are not degraded and thermal distortions do not impair such parameters as antenna pattern sidelobes and pointing direction or the operation of the drive system. Temperature control is established by specifying thermal coatings, insulators, conductive path, radiating surfaces, or by varying component location.

The three aspects of the mechanical design, while discussed independently, must in reality interact simultaneously during the design stages to assure that the final concept is an optimum solution to the design constraints and objectives.

4.1 HARDWARE DESIGN AND ANALYSIS

The configuration of the final design shown in Figure 63 and 64 is the result of several iterations and represents an optimization of the various design parameters within the given constraints and objectives. The mechanical characteristics of the parts associated with the communication (RF) subsystem are described first, followed by a description of the components associated with the pointing system.

4.1.1 Communication System Components

All communications components are shown in Figure 63. In this design, the cornucopias are oriented such that the lower two have their polarizers facing up while the top three have theirs facing down. This configuration results in minimum length waveguide runs, thus minimizing the RF losses, while still complying with the constraint of the desired launch volume. This configuration also lends itself to efficient structural design, where the axes of the tubular supports roughly intersect the cornucopias' center of gravity, thus reducing the torsion loading on the tubes. The center of gravity support is a desirable technique for supporting the cornucopias since the most rigid part of the antenna (the horn) is used as the main support. Alignment of the cornucopias is easily effected at the flange connecting the antenna to the tubular support structure. Ample room is provided in the center of the cornucopia support structure for electronics and RF components such as diplexers and RF switches. The four-channel limited-motion rotary joint is implemented as flex waveguide in the center of the yaw and roll axes platforms while the 360-degree rotary joint is contained within the pitch axis shaft.

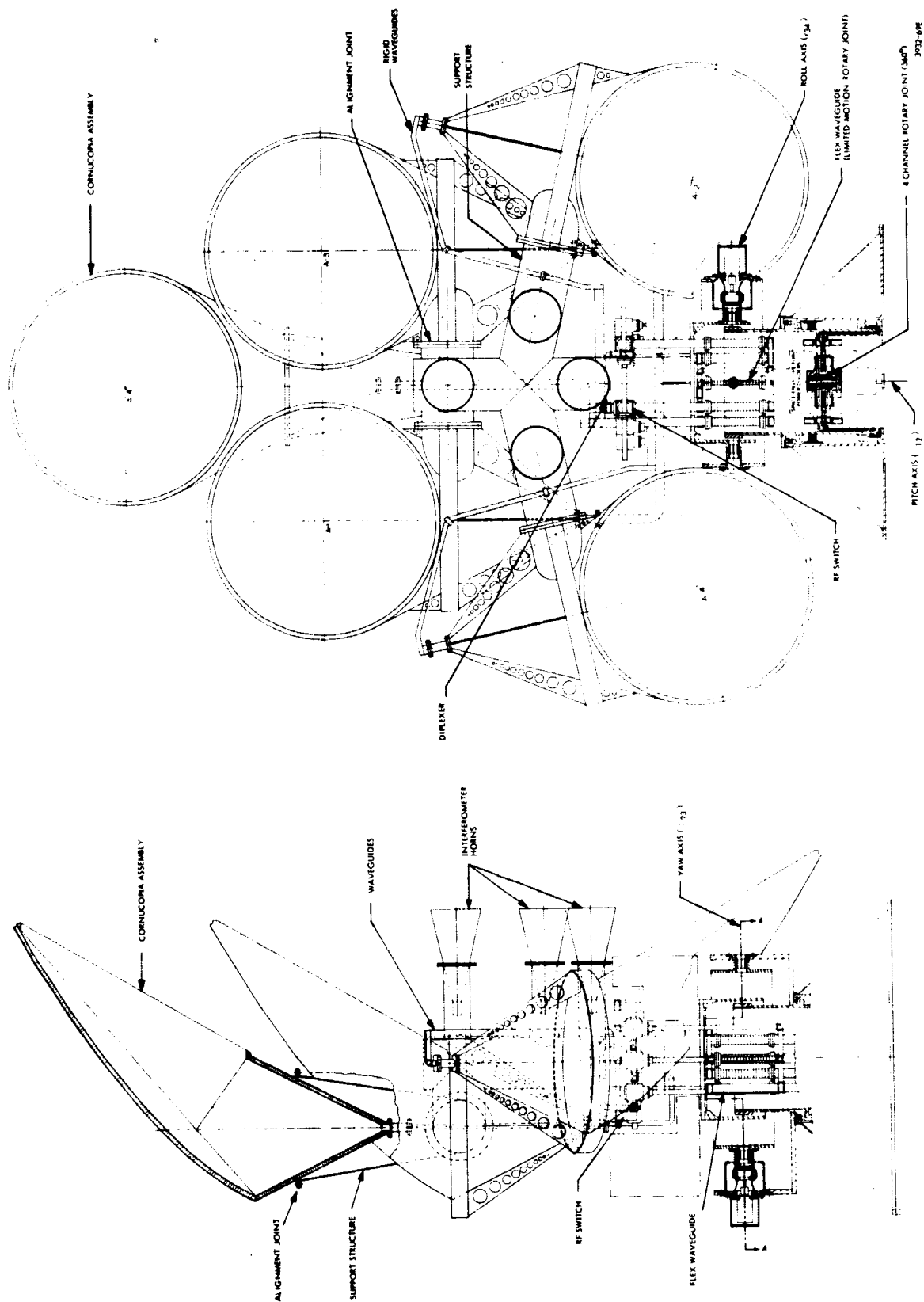


Figure 63. Antenna Assembly

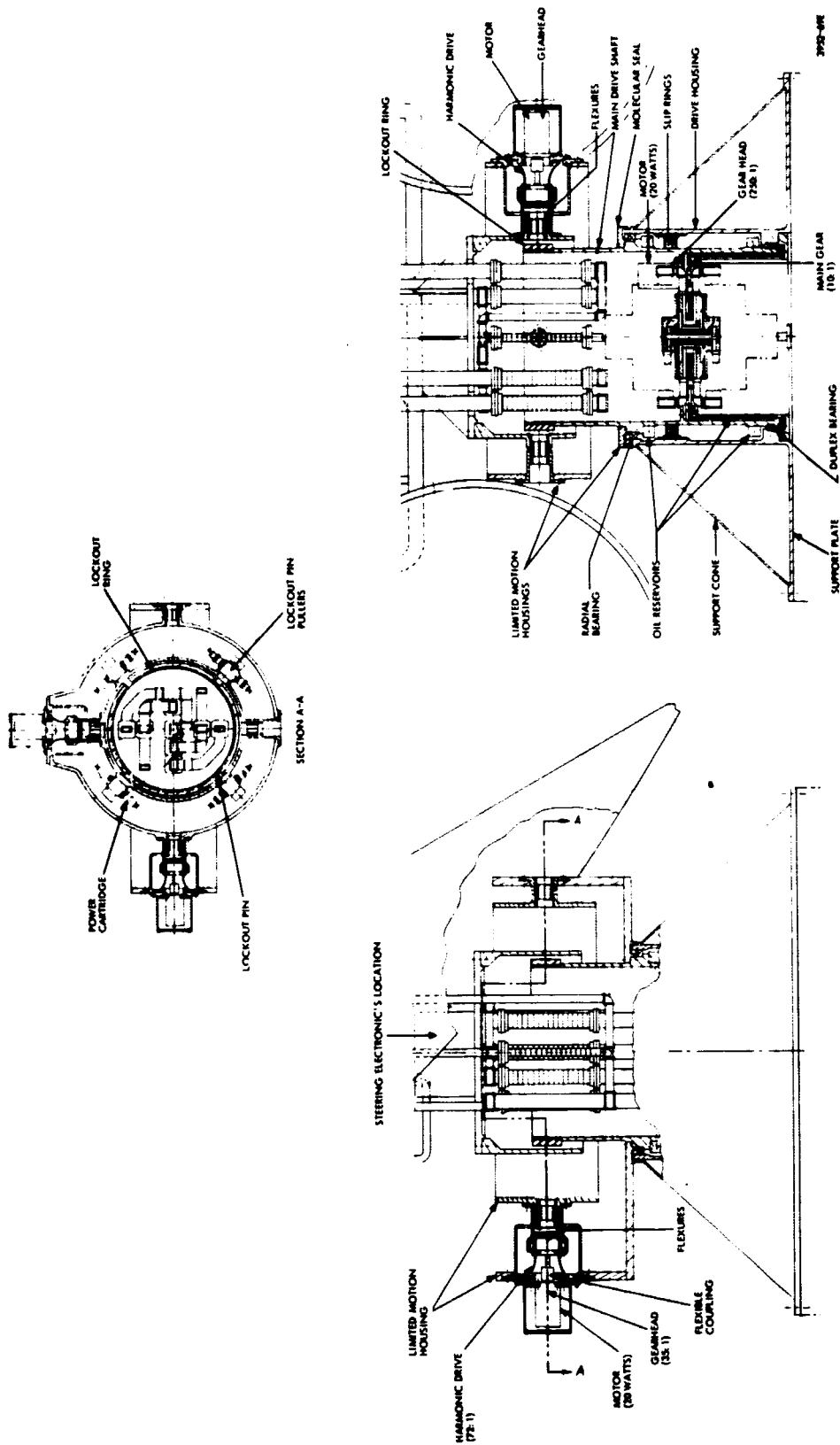


Figure 64. Antenna Assembly Pointing System Details

4.1.1.1 Cornucopias

A cornucopia is composed of three geometrical shapes: a cornical horn, a cylindrical section, and a reflector that is a portion of a paraboloid. Loads are transferred from the cornucopia's horn to the tubular antenna support structure through a ring and rib each having an "I" shaped cross section. The reflector and cylinder are supported from the horn and not tied to other portions of the antenna system. Aluminum honeycomb was the material specified for the cornucopias. It was selected for its high thermal conductivity and its high ratio of moment of inertia to weight. A high thermal conductivity is needed to minimize the thermal distortions which cause reduced antenna gain, increased sidelobes and pointing errors. An efficient minimum weight structure is required not only because the cornucopias have significant weights by themselves but also because their weight affects the weight of the structure required to support them.

4.1.1.2 Limited Motion Rotary Joint

Flexible waveguide is used for limited motion rotary joints to achieve small angle rotation about two orthogonal axes. In the design of the limited motion rotary joint, it is essential that the length of the flexguide be minimized so that the RF loss will not be excessive and the resulting power dissipation will not cause too high a bulk temperature. The selected design achieves rotation about two orthogonal axes by deflecting two sections of flexguide as cantilever beams. The flexguide must be properly located in relation to the axis of rotation in order to minimize the moments and bending stress for a given angular deflection and thus minimize the required length. To determine the minimum length, the flexural rigidity of a section of flexguide was first determined experimentally. From beam theory, the resisting torque imposed by the flexguide on the gimbal system was computed for a 6-inch (15.25 cm) length and 10-degree rotation and was found to be minimal. The minimum radius of curvature was also computed, and was found to be at least 15 times the minimum bend radius for the flexguide, specified by the manufacturers. It was further assumed that the minimum bend radius does not consider repeated flexing which is conservative. The flexguide manufacturers recommend that beryllium copper be used for flexguide that will be repeatedly stressed. Considering that beryllium copper has a ratio of endurance limit to yield stress of about 4/15, a 6-inch (15.25 cm) length of flexguide continually rotating ± 10 degrees has a minimum safety factor of four. A typical commercial type of flexguide that would be satisfactory is Airtron Co. sideseam flexguide, type AS, which is designed for repeated flexing and vibration and has no seams that have relative motion between metal parts.

In order to achieve the four-channel capability required of the limited motion joint, the cantilevered sections of flexguide are located side by side as shown in Figure 64. To reduce launch vibration stresses and act as a thermal conductive path, the rigid waveguides that connect the flexguides of each axis are supported by the yaw axis platform through cutouts in the roll axis platform.

4.1.1.3 Diplexer

The diplexers which are only used in the auxiliary mode of operation, are mounted at the base of the support structure as shown in Figure 63. Since diplexers

each dissipate about 256 watts of power, good conductive paths are provided to the surfaces that radiate to free space, thus, maintaining the diplexer temperature and the temperature of the adjacent electronics packages at reasonable values. This component was accorded a detailed thermal design since it dissipates more power than any other component and its RF performance is sensitive to temperature variations.

4.1.1.4 RF Switch

To achieve the RF requirements of low loss and good channel isolation, it is necessary to use an electrically actuated mechanical switch having a waveguide rotor that inter-connects different stator waveguides as it rotates. The rotor waveguides interface with non-contacting RF chokes. The bearing lubrication technique is the same as is used on the other components in the system that require lubrication, i.e., Ball Brother's Corp. Vac Kote lubricant impregnated in the balls and spacers of the bearings plus oil reservoirs and molecular seals. The RF switch is located in the same area as the diplexers.

4.1.1.5 360-Degree Rotary Joint

The four-channel rotary joint design with a capability of continuous rotation is shown in detail in Figure 65 and 66. The rotary to stationary interface is in the coaxial tube region shown in cross-section in the second figure. The RF choke joints separating the rotating and stationary parts have large clearances that assure that there will be no rubbing contact under all operating extremes. The remainder of the rotary joint consists of power dividers that feed the coaxial sections. The inner channel is fed by one port and thus does not have a power divider. The adjacent channel is fed by two ports and requires one divider that is contained within the plates that house the coaxial sections. The third channel is fed by four ports that require three power dividers, two of which are within the plate assembly and one of which is in waveguide. The outer channel is fed by eight ports and thus requires seven power dividers, four of which are in the plate assembly leaving three in waveguide. Figure 65 shows the waveguide power divider configuration while Figure 66 shows cross sections of the plate assembly power dividers. The entire assembly has been sized to fit within a shaft that is eleven inches in diameter. Although the total RF loss in the 360-degree rotary joint is not particularly high, the coaxial cylinders that form the RF transmission lines have large thermal resistances. To prevent excessive hot spot temperatures in these cylinders, a concerted effort was made to minimize the length of the coaxial section of the rotary joint. Temperature differentials along the cylinders in this section are dependent on the square of their length since RF dissipation is directly proportional to length. In the Thermal Analysis Section, it is shown that the temperature differentials along the cylinders are reasonable for the design shown, the maximum being 32.2°F. The joint also has a "counterflow" feature where alternate cylinders are attached to the rotor and stator. This increases radial radiation heat transfer since the hot end of one cylinder radiates to the cold end of an adjacent cylinder. This design also evenly distributes the thermal power flowing into the rotor and stator so it can be more effectively conducted to a surface that radiates to free space and thus maintain a lower bulk temperature of the joint. In the Thermal Analysis Section it is shown that bulk temperature of the joint will not exceed 149°F.

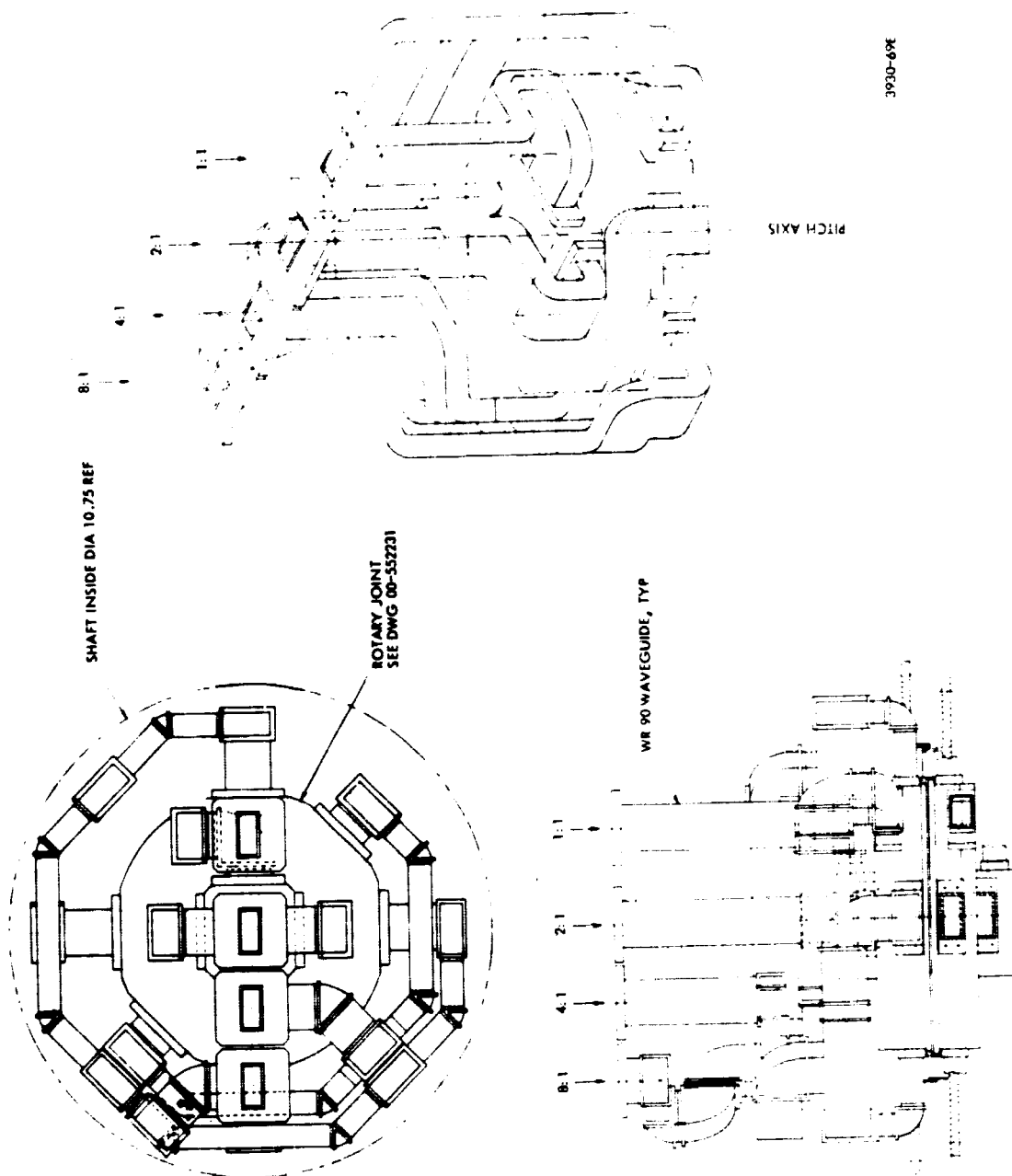
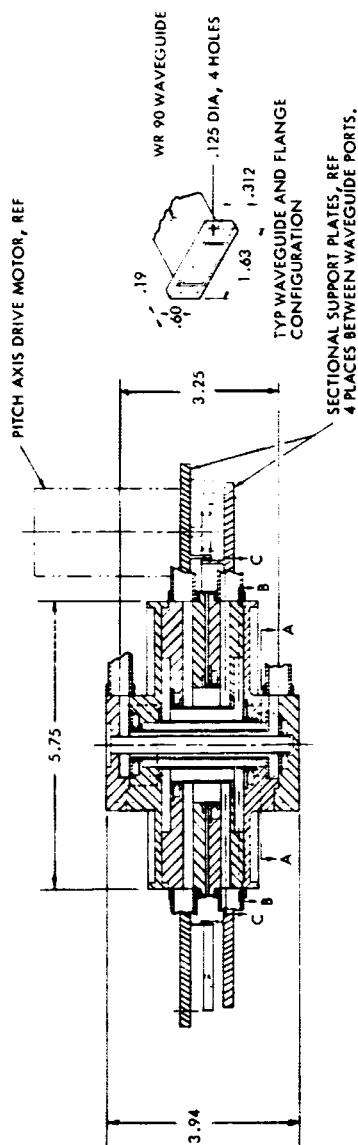
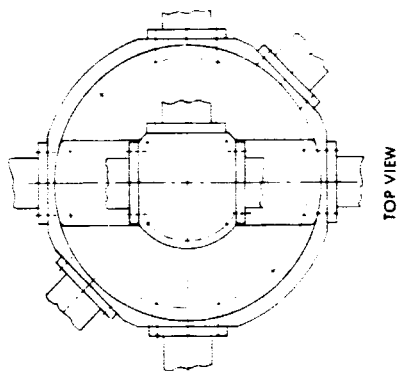
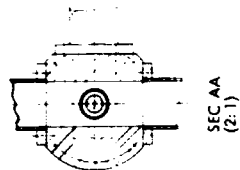
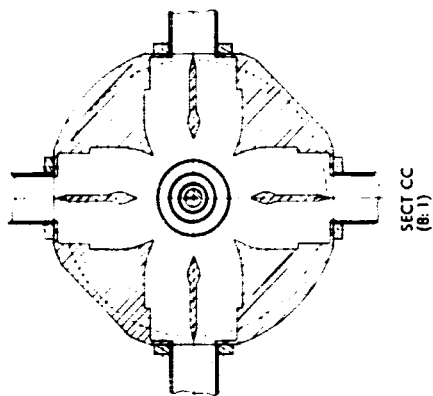


Figure 65. Four-Channel Rotary Joint Assembly



NOTES: 1. FOR RF CAVITY CONFIGURATION SEE DWG 00-522210
SECTION A-A, B-B, C-C

SECT BB
(4:1)

3931-69E

Figure 66. Four-Channel Rotary Joint Assembly (Power Dividers Not Shown)

Minimizing the lengths of the cylinders in the joint is also structurally advantageous since maximum stress is a function of the square of the cylinder's length. For the design shown stress and deflection calculations show that there is no structural problem.

Calculations of maximum partial pressures in the rotary joint and waveguide have been made to aid in the evaluation of high voltage breakdown phenomena. The first approach assumed there were no exit orifices and therefore the pressure in each component would build up to a steady state value. Using this assumption, heating of a component by the dissipated RF power would have no effect on the steady state pressure since although the outgassing rates would initially be higher as a result of the temperature rise, the total mass of gas introduced into the component would be the same after infinite time. The steady state pressure was calculated by integrating curves of the outgassing mass flow rate vs. time, obtained by an analytical curve fit to experimental data¹, and then applying the ideal gas equation of state. The curve fit was assumed an exponential, which is the correct analytical expression for molecular flow through a long tube of small cross sectional area. The resulting steady state pressures calculated for an aluminum rotary joint and waveguide are $4.28\text{E-}2$ Torr and $1.37\text{E-}2$ Torr, respectively.

The second approach was a closer approximation to the real situation and considered molecular flow through the rotary joint and waveguide to space. Using molecular flow equations from S. Dushman², both the orifice pressure drops, and those along the rotary joint and waveguide were determined. At the maximum flow rate, which occurs on entry into orbit (time = 0) the orifice pressure drop from the end of the waveguide to space is $2.77\text{E-}5$ Torr. The pressure distribution versus length along the rotary joint and waveguide is parabolic since the outgassing mass flow rate is proportional to surface area, and molecular flow through a constant area duct is proportional to pressure gradient. The maximum pressure drop between the rotary joint and end of the waveguide is $4.94\text{E-}3$ Torr. Since the pressure of space at synchronous altitude is less than 10^{-12} Torr, the maximum pressure in the joint is $4.97\text{E-}3$ Torr at time zero.

Heating by dissipated RF energy will cause a slight increase in this pressure, due to increased outgassing rate. Data for aluminum indicates, however, that the outgassing rate increases by only an order of magnitude for a temperature rise of 776°F above room temperature. Since the hottest bulk temperature of the rotary joint is 149°F , temperature rise will not cause significant pressure increases. It is also possible for microcracks in the rotary joint to trap air and cause slightly higher local pressures. The effect of these pressures can only be determined experimentally, however it is felt the safety factor on voltage breakdown (see RF section) is high enough to account for this effect.

4.1.1.6 Support Structure

The support structure for the cornucopias is made of thin wall tubes, which are efficient structural members in bending, torsion, and axial loading. Since launch vibration loading can be along any axis, the symmetry of the thin wall tube has a further advantage in that it can transfer loads in any axis. The wall thickness and diameters of the tubes are designed to roughly the same buckling and yield safety factors in the

cases where a minimum gage condition does not govern. The material chosen for the structure is 6061-T6 aluminum, an alloy that can be easily machined and welded, resulting in weight saving since welded joints are generally lighter than bolted flanged joints. Aluminum also has a high stiffness-to-weight as well as strength-to-weight ratio (in comparison to fiberglass) which is an important consideration since in thin walled tubes, buckling can be the governing failure mode. The high thermal conductivity of aluminum also minimizes temperature differentials and the resulting structural distortions.

4.1.2 Pointing System Components

The pointing system includes the interferometer horn assembly, shown in Figure 63, and the sensing and steering electronics, motors, and mechanical drive assembly, shown in Figure 64. The interferometer sensing horns are located on the antenna support structure close to, but separate from the cornucopias. This configuration minimizes thermal pointing errors between the horns and cornucopias, yet uncouples the interferometers and their sensitivity to translation between phase centers from the cornucopias which are only sensitive to rotation. The gimbal sequence is such that the 360 degree axis is closest to the satellite and thus, only prime power rather than motor control signals pass through the slip rings required for the 360-degree axis. Control signal paths to the motors cross only the limited motion axes, where flex cables, rather than slip rings are used. The most compact design of the pointing system locates the four-channel rotary joint inside the pitch axis shaft. Since for a fixed load capacity, bearing friction generally increases with diameter, the clearance diameters of the rotary joints are minimized so that the 360-degree shaft diameter is not excessive.

4.1.2.1 Interferometer Horns

The interferometer horns are located in the center of the support structure as shown in Figure 63. They are located adjacent to the cornucopias in order to minimize thermal distortions between the horns and cornucopias. The horns are supported by thin walled cylinders and a flange that may be shimmed for convenient alignment. The horns are machined from 6061-T6 aluminum to a 0.030-inch wall thickness with a resulting weight of one lb. per horn. The use of more sophisticated fabrication techniques would have a negligible effect on weight but would materially reduce the thermal conductivity and increase distortions, thus adversely affecting the thermal pointing errors.

4.1.2.2 Sensing and Steering Electronics

The system's electronics are located at the base of the support structure, as shown in Figure 64. The configuration of the support structure and cornucopias provides ample room for the electronics in this location. The electronics have not been treated below the black box level in the mechanical design. Weight, volume, and heat dissipation figures, have been included in the weight estimates, structural analysis, and thermal analysis. Due to the large amount of heat generated by the electronics, and the small temperature variations usually required by these components, a detailed thermal design will be required in this area.

4.1.2.3 Slip Rings

Due to the gimbal sequence and the location of the electronics on the support structure, the slip rings, which are required on the pitch or 360-degree axis shaft, only have to carry prime power to the electronics. The slip rings take advantage of the same lubrication system that is used for the radial and duplex bearings on the pitch axis shaft. Locating slip rings on such a large diameter (11-inch, 28 cm) shaft is usually considered undesirable, due to the increased circumferential distance travelled per minute for a constant RPM. This is not a consideration in this design, since the shaft's angular velocity is so small (one revolution per day), that it only accumulates a total of 1825 revolutions for a 5-year design life. Ball Brother's Research Corp. has data⁵ that shows no degradation of properly lubricated slip rings operating in a space environment for over 10^6 revolutions. Mounting the slip rings on the main shaft leaves the space inside available for the RF rotary joint, thus simplifying the total design.

4.1.2.4 Limited Motion Drive System

Only limited rotation of ± 10 degrees is required about the roll and yaw axes. The limited motion drive, shown in Figure 64, consists of a hermetically sealed motor and gearhead, harmonic drive, and low speed shaft supported by flexural pivot bearings. The only part of this system that is exposed to vacuum and requires special lubrication is one mesh of the harmonic drive.

To achieve the required angular resolution about the three axes using a motor of reasonable size, a speed reduction from the motors to the shafts is necessary. The most critical gear mesh in the speed reducer is at the low speed end where backlash must be very small. To achieve low backlash in the limited motion axes, where the motor and speed reducers may mount on axis, a harmonic drive is used for the final reduction. Briefly, harmonic drives, such as those manufactured by USM Corp., work as follows. An elliptical cam, called a wave generator, deflects a flexible spline. This spline has teeth which are cut to the same circular pitch and mesh of a circular spline, which in our design is attached to the output of the speed reducer. The flexspline has fewer teeth than the circular spline, the difference being a multiple of 2. Therefore, the flexspline has a smaller diameter than the circular spline. Rotation of the wave generator by the high RPM motor causes the flexspline to deflect and "ride around" the circular spline, the only teeth meshing are those at the ends of the elliptical cam's major axis. For one revolution of the wave generator, the flexspline moves relative to the circular spline a circumferential distance corresponding to two teeth. For a large number of teeth, then, a large speed reduction may be obtained between the wave generator and the circular spline. As shown in Figure 64, the flexspline is "fixed" and could be thought of as part of the housing for the limited motion axes.

In addition to low backlash, the harmonic drive hermetically seals the limited motion motors and gearheads from the vacuum environment and minimizes moving parts required for the gear reduction. Since flexures rather than ball bearings are used to support the low speed limited motion shaft, the only moving part exposed to vacuum and thus requiring special lubrication is the mesh between the flexspline and circular spline of the harmonic drive. This mesh will be lubricated by impregnating Ball Brother's Corp. Vac Kote lubricant in the teeth for primary lubrication, and providing oil reservoirs and molecular seals to assure a lubricant supply lasting for a five-year life.

4.1.2.5 Limited Motion Bearing System

Flexures are used as the limited motion bearing system. The advantage of these bearings is that small angle rotation of a shaft relative to a housing is obtained by deflecting a series of structural elements, thereby eliminating friction contacts and the associated lubrication requirements.

The two-strip type flexure that is recommended is commercially available from Bendix Corp. Data are available on load capacity and torsional spring constant, which are calculated from beam theory (knowing Young's modulus, the endurance limit for the flexure's material and the angle of rotation required). The two-strip flexure rotates about the intersection between the two strips in the untorqued position. For relatively large angles of rotation, the instant center of rotation of the flexures deviates slightly from the zero rotation nominal, as is discussed by Weinstein³ and Troeger⁴. The center shift for this design at ± 10 -degree rotation is given by Bendix Corp.⁵ as 0.003 inches (0.076 mm). This motion will not affect antenna pointing accuracy, and motion relative to the harmonic drive is compensated for by a flexible coupling.

4.1.2.6 360-Degree Drive System

The 360-degree mechanical drive consists of a motor and gearhead with a pinion driving a large internal gear that is attached to the housing. The pitch axis shaft is supported by duplex and radial bearings. As was the case with the limited motion axes, a gear reduction is required, and backlash must be minimized. Since the mechanical drive cannot be mounted on axis, due to the space occupied by the four-channel rotary joint, a large diameter internal gear was chosen for the final reduction. For a fixed machining tolerance, backlash is minimized in a large diameter gear since backlash in radians is roughly equal to tolerance divided by gear radius. The moving parts in the motor and gear train are lubricated by Vac Kote process including oil reservoirs and molecular seals.

4.1.2.7 360 Degree Bearing System

The 360 degree shaft is supported by a pair of 12-inch duplex bearings mounted back-to-back and a 13-inch radial bearing that floats in the axial direction to allow for thermal expansion. The balls and spacers of all bearings are impregnated with Ball Brother's Corp. Vac Kote lubricant for primary lubrication. In addition, oil reservoirs are located close to the bearings, so the lubricant will evaporate in a controlled manner, being contained by molecular seals that reduce molecular flow to space. This lubrication technique has been proven in life tests conducted at Ball Brothers Research Corp. and has been used successfully on several satellites. Beryllium liners are provided at the bearings to assure compatibility of thermal expansion and thus minimize any increase in bearing friction due to thermal effects.

4.1.2.8 Support Structure

Launch loads are transferred from the cornucopias and their supporting structure through the pointing subsystem's shafts, housings, and drives to the main body of the satellite in the following manner. The limited motion axes are locked out by four retractable pins, and thus the loads are transferred directly from the main tube of the

antenna support structure into the pitch axis shaft. This shaft must have a large diameter to accommodate the four-channel 360-degree rotary joint and the limited motion rotary joints that fit inside. Therefore, with any reasonable wall thickness, the shaft can withstand very high loading and easily support the reactions from the pins. The large diameter bearings supporting the pitch axis shaft also can withstand high radial and axial loads and do not require locking out. Radial loads from the duplex and radial bearings, and axial loads from the duplex bearings are transferred through the pitch axis housing and main support cone and plate to the 3-foot diameter satellite interface ring. The material selected for all the housings, shafts, and main support cone and plate, is 6061-T6 aluminum. It was selected for its good strength and stiffness, ease of fabrication, and high thermal conductivity.

Although the 360-degree shaft, bearings, and housing have inherent high load capability, the two limited motion axes drives can not be highly loaded especially in torsion. In order for the gears in these drives to withstand torsion about either of the two limited motion axes, their size and weight would be prohibitively large. The lock-out technique shown in Figure 64 uses four retractable pins to lock the cornucopia support structure directly to the pitch axis shaft. This configuration provides load paths that bypass the two limited motion drives thus structurally isolating the components. The loads will be carried mainly by the relatively stiff pins, rather than the more flexible limited motion shafts and housings. The pins are located in the plane of the limited motion axes, thus maximizing the clearance between the lockout ring and the antenna support structure during the ± 10 -degree rotations. The pin is retracted by gas pressure from a low impact power cartridge acting on a piston which forms the end of the pin. The pin is then held in the retracted position by holding clips. Two redundant power cartridges are provided for reliability. These devices have been successfully used in many space applications.

4.1.3 Weight

A weight breakdown is shown in Table 18. The total system weight is calculated to be 237 lbs. (107.5 kg) which is below the 250-lb (113.5 kg) limit in the design requirements. Although the system's overall design concepts are optimized for minimum weight, choices of materials considered other properties as well as weight. The stress calculations that these weights are based upon are also somewhat conservative since only worst case sections were considered. The weight tabulations are therefore without the benefit of refinement through the use of more exotic materials, and more detailed dynamic and stress analysis. If such an effort were undertaken, it is estimated that the total system weight could be reduced by 10 to 20 percent.

4.2 STRUCTURAL ANALYSIS

The structural analysis effort is responsible for providing an efficient structural concept that minimizes weight while providing adequate margins against failure during ground test, launch, and in orbit operation.

4.2.1 Design Criteria

The criteria for sizing structural members is defined by the factors of safety established against yield, ultimate, and buckling strengths of the appropriate material

TABLE 18
ANTENNA WEIGHT ESTIMATE

STEERING SUBSYSTEM				SENSING SUBSYSTEM		
Item No.	Part	Qty/ Assy	Total Weight	Item No.	Part	Total Weight (lbs.)
1	Housing	1	19.3	1	Interferometer Horn	4 2.1
2	Shaft	1	15.7	2	Interfer. Horn Support	4 1.5
3	Shaft Top Ring	1	1.8	3	Sensing Electronics	AR 37.0
4	Base Cone	1	6.5	TOTAL WEIGHT		
5	Gimbal Yoke	1	9.0			40.6
6	Roll Platform	1	9.1	RF SUBSYSTEM		
7	Yaw Platform	1	6.8			
8	Antenna Mtg. Structure	1	18.1			
9	Flexture Brg	4	1.0			
10	Tob Ball Brg	1	2.6	1	Cornucopia	5 45.0
11	Duplex Ball Brg	1	2.5	2	RF Switch	5 6.3
12	Duplex Brg Lock Ring	1	0.6	3	Diplexer	1 0.3
13	Oil Reservoirs	2	1.3	4	Rotary Joint	1 3.0
14	Squib Assy	4	4.0	5	2:1 Power Divider	1 0.3
15	Pitch Axis Bull Gear	1	2.0	6	4:1 Power Divider	1 0.5
16	Pitch Axis Drive	1	3.0	7	8:1 Power Divider	1 1.0
17	Roll Axis Drive	1	5.0	8	Rigid WR 90 W/G	176 IN 2.5
18	Yaw Axis Drive	1	5.0	9	Flex WR 90 W/G	48 IN 2.1
19	Steering Electronics	AR	9.0	10	Polarizer	5 0.4
20	Hardware	AR	5.0	11	W/G Flanges	37 1.2
TOTAL WEIGHT			126.3	12	RJ Stator Support	1 6.0
				13	RJ Stator Support	1 0.5
				14	W/J Mtg. Clamps	6 0.1
				15	Switch & Diplexer Brackets	2 0.4
				16	Flex W/G Support	2 0.2
				TOTAL WEIGHT		
						69.8
GRAND TOTAL				GRAND TOTAL		
						236.7 lbs.

when applying the most severe or governing loading condition. In analyzing the recommended design and sizing the members, the minimum factors of safety for aluminum were established at 1.5 on yield and 2.0 on the less exact failure mode of buckling. In many cases, minimum gage material conditions occur with a resulting increase in the factors of safety. This design criteria will establish a conservative preliminary design that still falls well within the weight goal of 250 lbs. (113.5 kg).

4.2.2 Governing Loading Conditions

The four structural loading conditions of shock, vibration, steady state acceleration, and acoustic noise are specified in the design requirements. Acoustic noise was found to be less severe than shock and vibration. Steady state acceleration is also less severe than shock and vibration except for structural resonant frequencies less than 10 Hz where there is no vibration input and considerable shock isolation. As will be seen in later sections, the minimum resonant frequencies are above 10 Hz and thus, the steady state acceleration levels will not govern.

Comparisons between the severity of the shock and vibration loading were made on the basis of a single degree of freedom model. For such a system, subjected to base excitation by a rectangular shock pulse, maximum amplification is obtained if the natural period of the structure divided by the period of the shock pulse is less than two, and the critical damping ratio is zero. The 30-g shock pulse with a period of 0.008 second excites resonant frequencies greater than 62.5 Hz at maximum amplification. The peak vibration inputs are 1.4 g from 10 to 60 Hz and 2.82 g from 60 to 2000 Hz. For the single degree of freedom model, the maximum vibration response is at its natural frequency and is damping controlled. Two cases (for structure's with first resonances greater and less than 60 Hz) are now considered to arrive at a design dynamic load that can be used in the stress analysis.

Structures with resonances above 60 Hz will see both the most severe shock and vibration loading, since maximum shock amplification will occur and vibration inputs are higher. Thus maximum responses for a single degree of freedom system subjected to both specified inputs have been plotted versus the damping ratio. The curves cross at a peak acceleration value of 55 g and a damping ratio of 0.026. This critical damping ratio corresponds to a resonant vibration amplification factor, Q of 19.2, a reasonable value for preliminary structural design. Thus, at resonant frequencies above 60 Hz, the effects of shock and vibration are roughly equal, and the maximum acceleration response is 55 g.

If a structure's first resonance is less than 60 Hz, both the shock and vibration maximum response will reduce, the shock by lower amplifications and the vibration by the lower input of 1.41 g instead of 2.84 g. Assuming the same critical damping ratio as in the previous case, the vibration maximum response of a single degree of freedom system is about 30 g. From shock transmissibility curves it is found that this level is greater than the shock response for structural natural frequencies less than 21.2 Hz. Therefore, from 10 to 21.2 Hz, vibration response of 30 g is most severe, from 21.2 to 60 Hz, shock governs with a peak response between 30 g and 60 g, and above 60 Hz, vibration and shock loading are both roughly equal at 60 g.

4.2.3 Stress Analysis

For purposes of a preliminary stress analysis, all structures except the antenna support structure, were assumed to have fundamental resonances above 60 Hz, and thus would be subjected to the 60 g dynamic load. The first resonance of the antenna support structure is assumed to be less than 21.2 Hz, which will be verified in the Vibration Analysis Section, where the 30 g dynamic load applies. The member sizings for the critical structural areas, as determined by the stress calculations, are shown in Figure 67.

In analyzing the cornucopia, the reflector was considered as a cantilevered panel without benefit of any support offered by the cylinder. Therefore, the section of the honeycomb sandwich reflector that is closest to and supported by the horn is designed with thicker skins than the rest of the panel. The transition between the reflector and closed horn is perhaps the most critical section of the cornucopia since it must withstand the bending moment from the reflector and yet is an open semi-circular cross section having a minimal flexural rigidity. The closed conical horn cross sections are good structural shapes having large moments of inertia. The most critical section is the joint between the antenna support structure and horn, which is reinforced by a rib and ring.

The most critical sections of the side and top tubes of the support structure occur at the intersection with main support tube where moments are highest and both buckling and yielding govern the selection of wall thickness and diameter. The base of the main support tube must transfer the highest moments in the antenna support structure, and again both buckling and yielding govern the design. To reduce weight, the wall thickness of this tube is tapered to 0.030 inches (0.076 cm) which is assumed to be a minimum gage thickness for these 6-inch (15.25 cm) diameter aluminum tubes.

Loads are transferred from the antenna support structure into the pitch axis shaft through the retractable pins. Since the pins unsupported length is small compared with their diameter, these loads are predominantly carried in shear. These pins are designed with a large safety factor since their weight is small compared with the total system weight, and yet a pin failure would be catastrophic to the system's operation. The largest moment in the pitch axis shaft is at the top bearing. The wall thickness of the shaft considers stress concentrations due to shoulders, and grooves. The critical section of the main support cone and plate is at the interface between the antenna system and satellite.

4.2.4 Vibration Analysis

The transverse frequencies of the antenna system were calculated to check that the first resonance was less than the 21.2 Hz assumed when determining the dynamic loads on the antenna support structure. The longitudinal frequencies, i.e., along the pitch axis, were not computed since the transverse loads govern the design of the antenna support structure and mechanical drive components by producing the largest moments in these structures. The antenna was modeled with four degrees of freedom. The mechanical drive and gimbal system area was assumed to be a mass and inertia. This rigid mass and inertia was attached to a rigid base (assumed to be the satellite), by two springs, which represent the main support cone and plate. The mass and

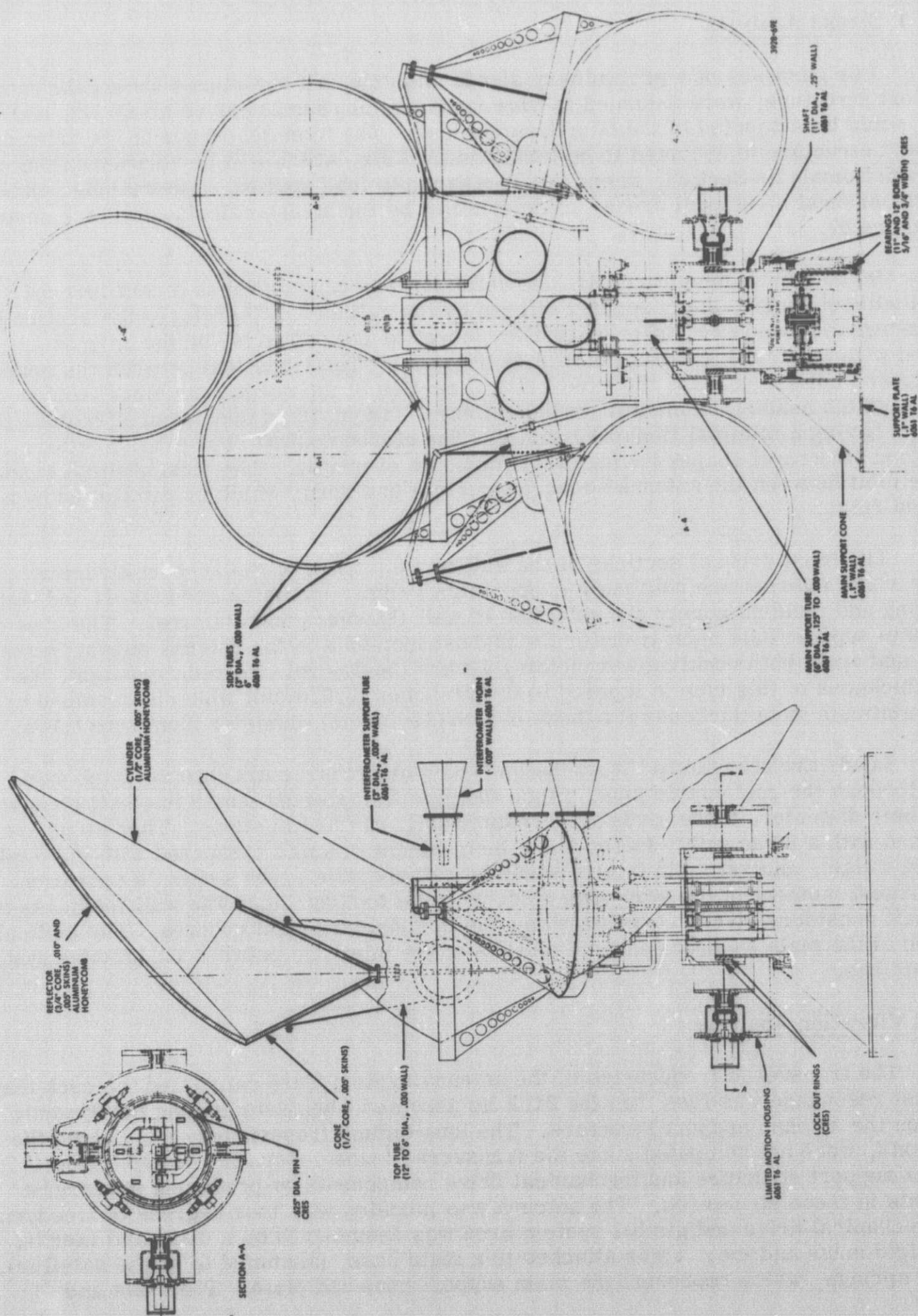


Figure 67. Critical Structure Areas

inertia of the cornucopias and tubular support structure was represented by another mass and inertia which was assumed connected to the gimbal area mass and inertia by elements in the stiffness matrix of the cantilever beam that represents the antenna support structure. Using Lagrange's equations, the equations of motion for the model were determined and checked by equilibrium of spring and inertia forces and moments. The entries in the mass and stiffness matrices were calculated, and the Eigen value problem solved by computer. In addition to the four degree of freedom model, a one and two degree of freedom free vibration problem was solved to check the effect on the lowest resonant frequency as the structure was approximated more closely. For the one degree of freedom case, the flexibility of the drive area, and the inertia of the cornucopia structures were neglected. For the two degree of freedom case, only the flexibility of the drive area was neglected. The resonant frequencies resulting from the three calculations are presented in Table 19.

TABLE 19
RESONANT FREQUENCY

Case	f_1	f_2	f_3	f_4
1 d-o-f, (weightless cantilever beam with end mass)	29.3 Hz	-	-	-
2 d-o-f, (weightless cantilever beam with end mass and inertia)	20.0 Hz	68.3 Hz	-	-
4 d-o-f	15.9 Hz	59.7 Hz	365 Hz	774 Hz

It can be extrapolated from these frequencies that if a more detailed analysis were performed including more degrees of freedom, the resulting first mode natural frequency would be less than 15.9 Hz, and the prior assumption that the first mode is less than 21.2 Hz is valid. By analyzing the Eigen vectors of the model matrix for the four degree of freedom case, it was found that the first two resonances are primarily motions of the lumped mass and inertia of the cantilever beam, which represents the support structure for the cornucopias. The two higher modes are primarily motions of the lumped mass and inertia of the mechanical drive area.

4.2.5 Structural Model for Control Systems Analysis

A two degree of freedom model of the antenna structure has been developed for incorporation into the limited motion axis model for the control system dynamic analysis. The structure was approximated as a lumped mass and inertia on the end of a weightless cantilever beam since modes of these masses will be most closely coupled to the control system transient response. The springs that attach the antenna system to the satellite, are assumed infinitely stiff. The mass and stiffness matrices for the two degree of freedom model are of the form:

$$\begin{bmatrix} m & 0 \\ 0 & J \end{bmatrix} \quad \text{and} \quad \begin{bmatrix} k_{11} & k_{12} \\ k_{12} & k_{22} \end{bmatrix}$$

To determine the structural damping matrix for use in the control system transient analysis, a modal damping technique was used. Using the modal matrix obtained from the vibration analysis for the two degree of freedom approximation, the equations of motion were transformed to modal coordinates. The equations were normalized by equating the mass matrix to the identity matrix, and then damping was included. The resulting expression was

$$[I] \{\ddot{Z}\} + [\bar{B}] \{\dot{Z}\} + [\bar{K}] \{Z\} = 0 \quad (86)$$

where:

- $[I]$ = identity matrix
- $\{Z\}$ = modal coordinate vector
- $[\bar{K}]$ = modal stiffness matrix
- $[\bar{B}]$ = modal damping matrix

The modal damping matrix is then given by

$$[\bar{B}] = \begin{bmatrix} \frac{\omega_1}{Q_1} & 0 \\ 0 & \frac{\omega_2}{Q_2} \end{bmatrix} \quad (87)$$

where:

- ω = circular frequency for the first and second modes
- Q = quality factor for the first and second modes

The damping matrix $[B]$ was computed by assuming $Q_1 = Q_2 = 20$,

and recalling that $[\bar{B}] = [\theta]^T [B] [\theta]$, where $[\theta]$ is the modal matrix.

4.3 THERMAL ANALYSIS

The goal of the thermal design of a space antenna system is to maintain temperatures of the components in the system as close to room temperature, and each other, as possible. The temperature control must be maintained to prevent excessive material degradation, thermal stress, and thermal distortions. Material degradation will result from either high or low temperatures compared with room temperature

and when coupled with thermal stress can cause material yielding. Thermal distortions may cause excessive bearing friction, pointing error, or RF degradation.

The design techniques available to achieve temperature control may be classified as active or passive. Passive techniques are more desirable since they require no mechanical moving parts or control. Examples of passive systems are thermal coatings, blankets, isolators, conductive paths, or sunshades. Examples of active systems are servo controlled heaters and louvers. To achieve thermal control of the antenna system, only passive systems have been employed, and it is concluded from the results of the bulk temperature and thermal distortion analyses that temperature control of the antenna system is such that thermal effects do not degrade performance beyond the limits of the design specification.

The components in the antenna system that are critical in the thermal design are listed in Table 20 along with the possible problems that could occur if thermal control were not achieved.

TABLE 20
THERMAL DESIGN COMPONENTS

COMPONENT		POSSIBLE PROBLEMS
(1)	Cornucopias	RF degradation, pointing errors
(2)	Antenna Support Structure	pointing errors
(3)	Sensing Horns	pointing errors
(4)	Electronics	material degradation
(5)	Diplexers	RF degradation, thermal stress, material degradation
(6)	Flexguide	RF degradation, material degradation
(7)	Rotary Joint	RF degradation, thermal stress, material degradation
(8)	Shafts	Bearing Friction

These critical areas will be investigated in the analyses that follow:

4.3.1 Bulk Temperature Analysis

A 20 node bulk temperature analysis of the antenna system was conducted, using the thermal model shown in Figure 68. The sun was assumed to be in the earth's equatorial plane, even though in summer and winter, the earth-sun line is inclined 23.5 degrees to this plane. This change in angle has a small multiplicative effect on the solar radiation heat inputs to the antenna system of $\cos 23.5$ degrees or 0.917. The bulk temperature analysis was treated as a transient problem throughout the day. However, steady state from day to day, i.e., solar inputs and node temperatures, continually change with time during the day; but a plot of temperature vs. time for each node is

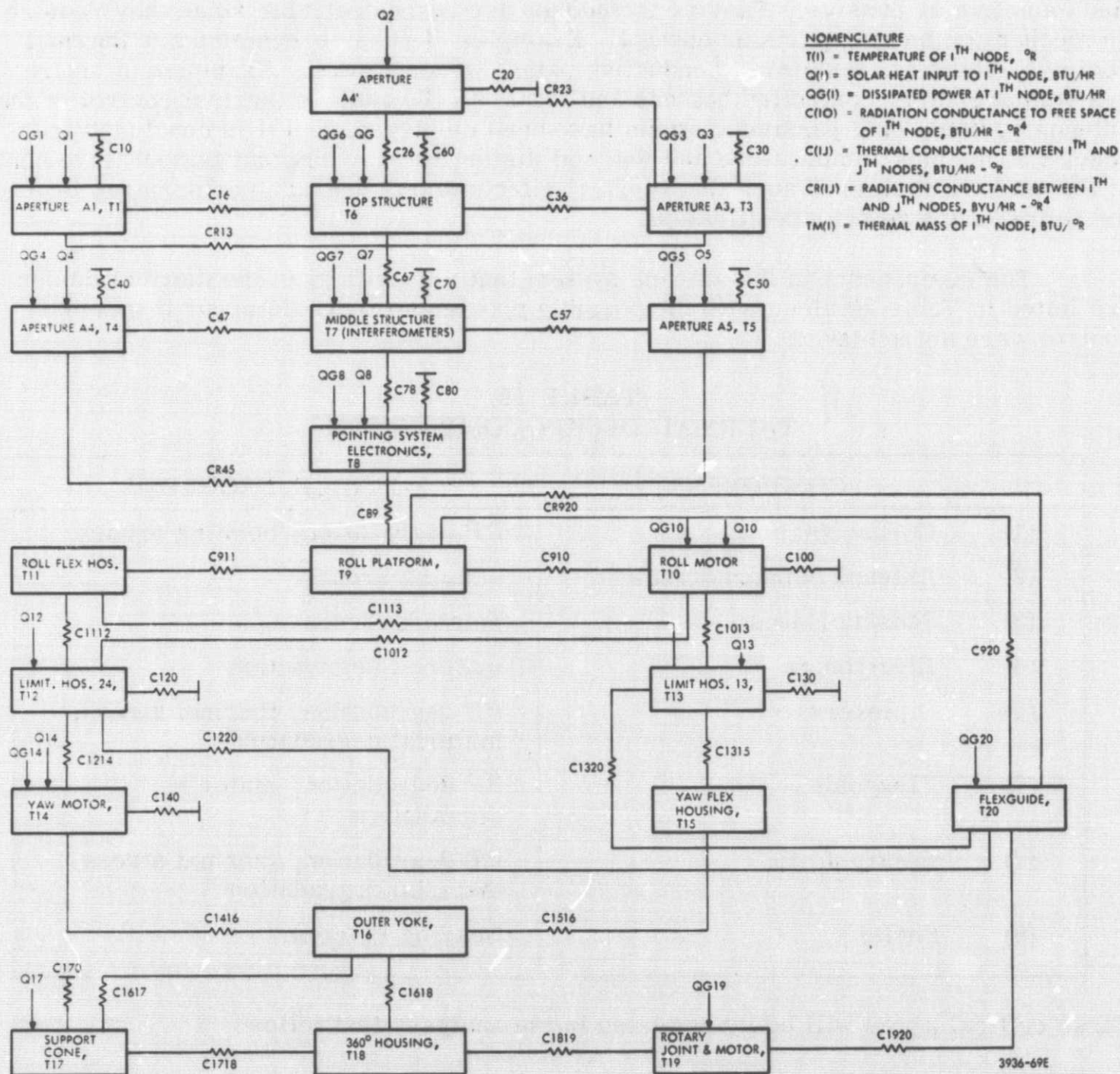


Figure 68. Bulk Temperature Model

identical from day to day, after initial conditions die out. This approach neglects seasonal changes, which were shown to have a minor effect on solar power inputs. In the thermal model, radiation between nodes was neglected except in the most significant cases such as between the flexguide and the roll axis platform, and between cornucopias. This is a conservative assumption since if more radiation paths were included, temperature differentials between nodes would be reduced. The interface between the satellite and the antenna is assumed to be insulated.

Due to the high RF power of the system, and the resulting energy dissipation in the RF and pointing system components, the system has many significant heat sources besides solar radiation. These on board sources are listed in Table 21 for both the normal and multichannel modes of operation. While of significance during the day, these sources are reduced to zero during the eclipse when the solar panels provide no power.

TABLE 21
HEAT SOURCES

<u>Heat Source</u>	<u>Dissipation</u>	
	<u>Normal Mode</u>	<u>Multichannel Mode</u>
(RF Components)		
Rigid Waveguide	139 Watts	139 Watts
Flexguide Waveguide	59	59
Rotary Joint (Coax Section)	78	78
Rotary Joint (Power Dividers)	100	100
RF Switches	74	74
Diplexers	<u>0</u>	<u>516</u>
	450 Watts	966 Watts
(Pointing System Components)		
Sensing Electronics	70	70
Steering Electronics	10	10
Motors	<u>60</u>	<u>60</u>
	140 Watts	140 Watts
Total	590 Watts	1106 Watts

The thermal power sources are lumped at the most appropriate nodes in Figure 68. The solar heat inputs, also indicated, depend on sun angle and the thermal coatings applied to the structure.

Both radiative and conductive thermal conductances are considered in the analysis. Radiative conductance is defined as the product of the shape factor, the area of the surface upon which the shape factor is based, and the Stefan-Boltzmann constant.

Thermal power radiated from one surface to another is obtained by multiplying this conductance by the difference of the fourth powers of the surface temperatures. If the surface is radiating to free space, the second "surface" temperature is absolute zero, and the shape factor is the one between the radiating surface and free space. Thermal power conducted from one node to another is obtained by multiplying the conductance by the temperature difference between nodes.

Since the analysis is transient, thermal masses are required for each node and are defined as the product of the weight lumped at the node and the specific heat of the material. The nomenclature used for the heat sources, conductances, and thermal masses is given in Figure 68.

From the 20 node thermal model, the twenty thermal power balance equations are written. For example, the equation for Node 1, which represents cornucopia aperture No. 1 is:

$$Q_1 - TM_1 \frac{dT_1}{d\theta} + C_{16} (T_6 - T_1) = C_{10} (T_1)^4 \quad (88)$$

These differential equations are non-linear, and are solved by computer using a third order Runge-Kutta method yielding the 20 node temperatures as a function of time. The initial conditions used for the program were the steady state temperature at the sun angle with the smallest rates of change of temperature with respect to time (12:00 noon). These steady state temperatures are simply the solution to the 20 non-linear simultaneous thermal power balance equations that result if all of the derivatives of temperature with respect to time are equal to zero. The Runge-Kutta method provides a convenient way to solve these simultaneous equations by assuming arbitrary initial conditions and running the program until a steady state solution is reached. The only change that results from assuming different sets of initial conditions is in the convergence time. Using as initial conditions the 12:00 noon temperatures, the program was run for five cycles (5 days "computer time"). Differences between temperatures of the same node in the fourth and fifth day were less than 1°F thus indicating that the day steady state solution has been reached.

Of the various sets of thermal coatings tried, the following set was considered most satisfactory, based on the criteria of minimal temperature extremes and differentials between nodes. All exterior surfaces are painted white except the cornucopias which are coated with Rinsed Aluminum paint (solar absorptivity = 0.33, room temperature emissivity = 0.24). All interior surfaces are black. Plots of temperature versus time for the critical nodes are shown in Figure 69 and 70. Components critical to the pointing error are shown in Figure 69 for the normal mode only. Those critical from a thermal stress and/or material degradation point of view are shown in Figure 70 for the normal or auxiliary mode depending on which is most severe. A discussion of these temperatures follows.

The cornucopias have a relatively large variation in bulk temperature with changing sun angles, the extremes being 149°F and -54°F. The large drop in temperature during eclipse is caused by the light-weight design of the cornucopia's structure, resulting in small thermal time constants. The effect of these variations on pointing accuracy is discussed in the next section.

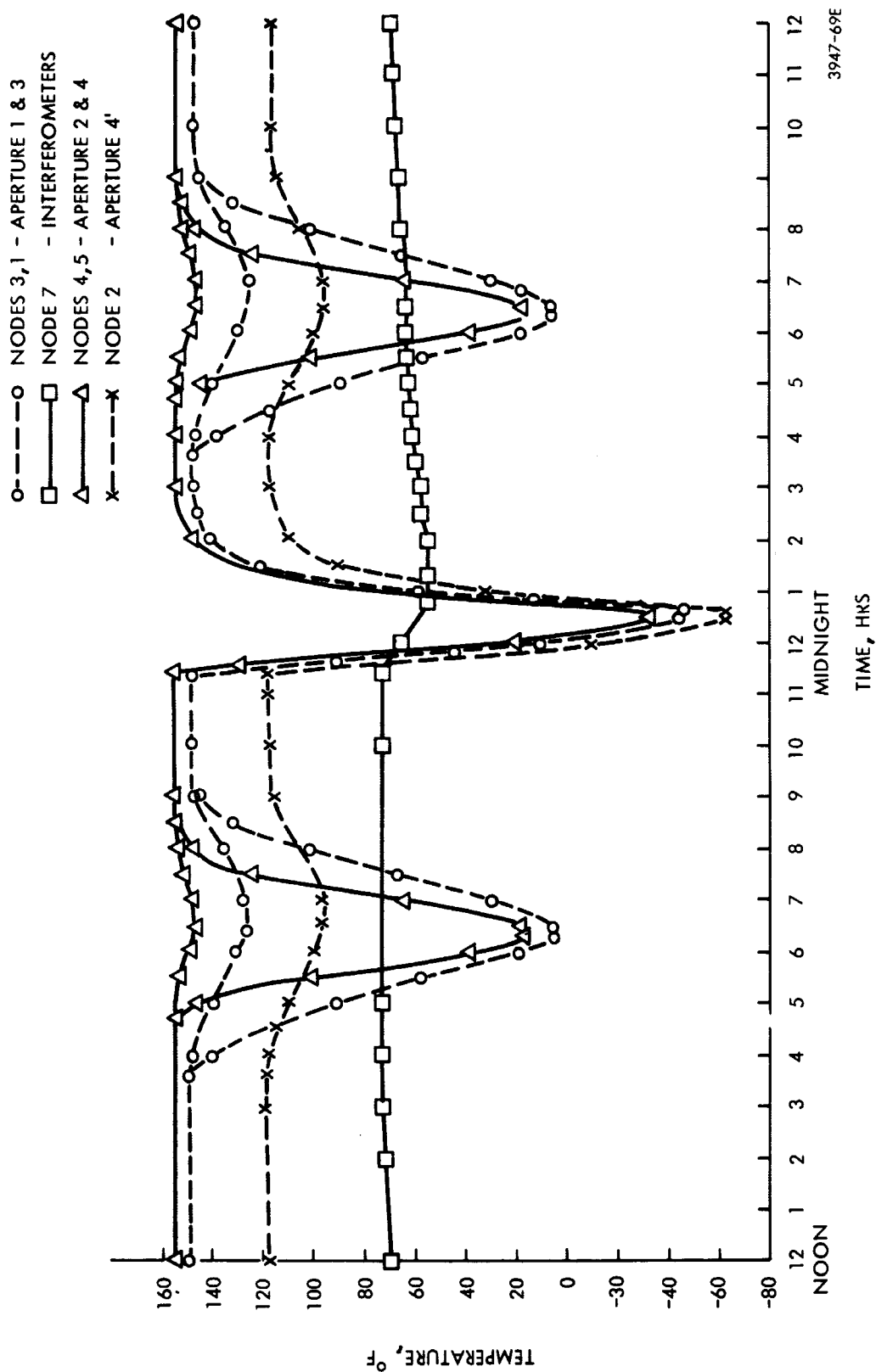


Figure 69. Bulk Temperature Profile (Components Affecting Pointing Error)

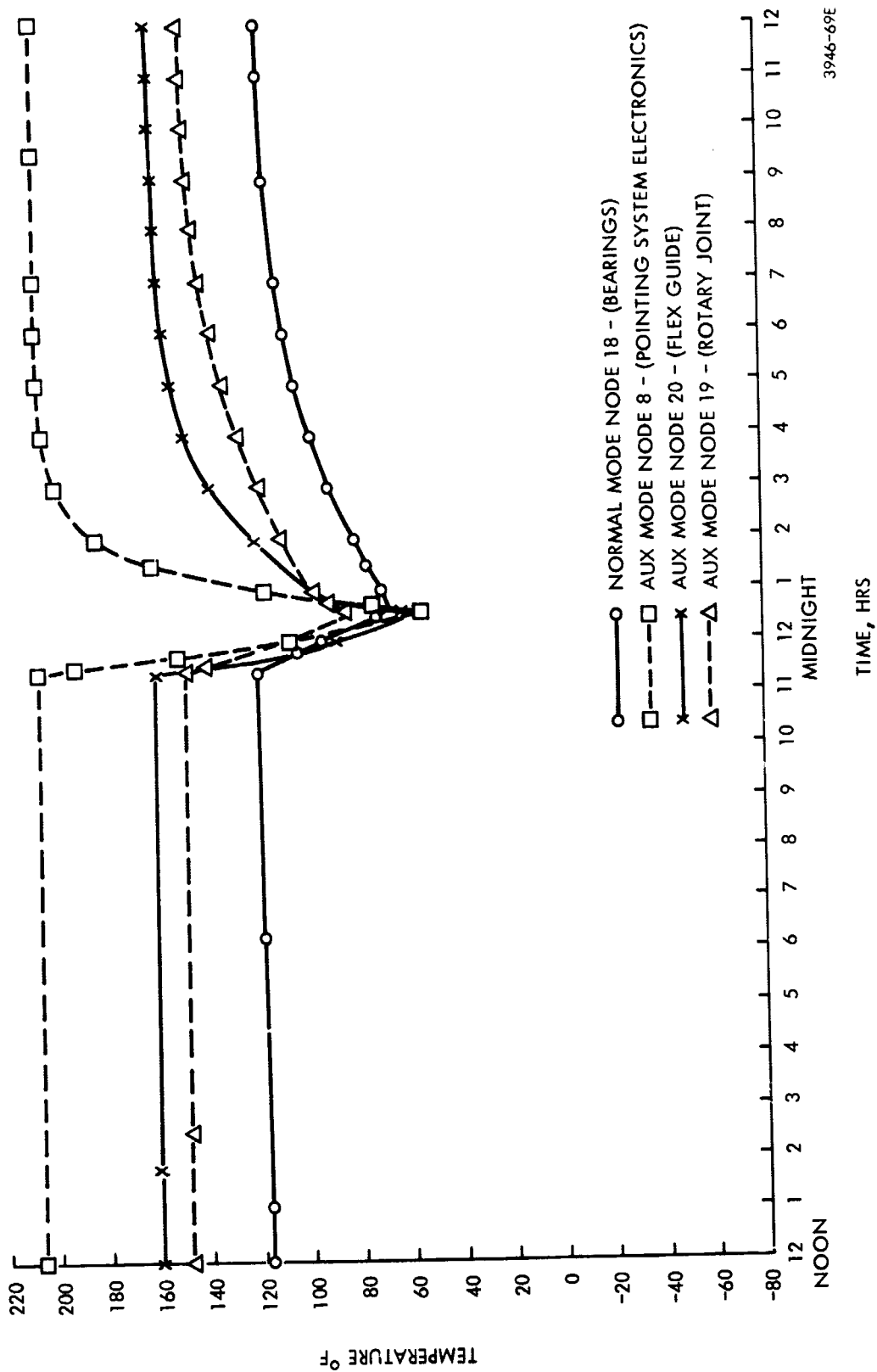


Figure 70. Bulk Temperature Profile (Thermally Sensitive Components)

The antenna support structure and interferometer horn structure have bulk temperatures varying from 120°F to -60°F just prior to leaving eclipse. These temperatures are of concern since deflections of the tubular beams that support the cornucopias and horns will cause pointing errors. The effect of bulk temperature on the circumferential temperature differential in the tubular beams is small since conduction is the predominant mode of heat transfer. It is desirable to have the structure warm, however, since radiation inside the tubes will increase with temperature thus reducing the differential. In the operational modes bulk temperature is close to room temperature.

Bulk temperatures of the electronics and diplexer node vary from 208°F to -1°F . This is the hottest area in the antenna system since both the electronics and diplexers (in the auxiliary mode) dissipate a large amount of power. Although this peak temperature appears manageable, care will have to be taken in the design of the electronics to minimize thermal resistances to surfaces that radiate to free space.

The flexible waveguide has one of the higher bulk temperatures (162°F in the auxiliary mode) due to its heat dissipation and poor thermal conductance. This temperature has been minimized by using minimal lengths of guide to accomplish the two axis flexing, and by providing a conductive path between the waveguide and yaw axis platform. This path also acts as a structural support for the flexguide during launch.

The hot spot temperature inside the four-channel rotary joint is derived in part from the bulk temperature analysis. The maximum bulk temperature of the rotary joint is 149°F . The predominant thermal resistances in the rotary joint are the cantilevered tubes that form the coax lines. These are also the surfaces where significant RF power dissipation occurs. Knowing the dissipated power in each cylinder, the parabolic temperature distribution along each cylinder was calculated. The maximum temperature differential for the worst case cylinder is 32.2°F . Adding this differential to the highest bulk temperature of the rotary joint yields a hot spot temperature of 181°F . This temperature will not cause excessive thermal stress, material degradation, or thermal distortions which would result in degradation of the RF performance.

The 360 degree axis shaft is aluminum with beryllium liners for thermal expansion compatibility with the steel bearings. The housing is aluminum which has a coefficient of thermal expansion roughly twice that of steel or beryllium. Therefore, the bulk temperature of the shaft and housing area should always be above room temperature so that bearing friction will not increase. The minimum temperature for these nodes is shown in Figure 70 to be 69°F . In addition to this requirement, the temperature differential between the shaft and housing should be minimal.

4.3.2 Pointing Error Analysis

The thermal distortions that contribute to the system pointing error are discussed in this section. The distortions of significance are those that shift the interferometer sensing direction relative to the cornucopia boresight direction. The interferometer sensing direction is altered by differential translations of the interferometer horns in the boresight direction but is not affected by rotations about their phase centers. Differential translations that alter the horn spacing introduce secondary errors that appear when the interferometer is offset from its null by a phase bias which is introduced as a pointing command. The cornucopias boresight direction is sensitive

to rotations about any axis perpendicular to the boresight but is not affected by translation in any direction or by rotation about the boresight.

The basic pointing error equation developed in the pointing system section is:

$$\text{Total error} = \left[(\Delta \psi_{12})^2 + (\Delta \psi_{23} + 0.039 \Delta \psi_{34})^2 \right]^{1/2} \quad (89)$$

where

$\Delta \psi_{12}$ = total change in boresight east-west pointing angle

$\Delta \psi_{23}$ = total change in boresight north-south pointing angle

$\Delta \psi_{34}$ = total change in the boresight rotation angle.

The east-west and north-south boresight rotation errors in the above equation are root sum squares of all of the system errors including the thermal errors $\Delta \psi_{12}^T$ and $\Delta \psi_{23}^T$. Thermal errors do not contribute to rotation of the four apertures as a rigid body about the pattern boresight ($\Delta \psi_{34}^T$).

The thermal errors were examined for the symmetric sun angles (12 noon and 12 midnight) and for the nonsymmetric sun angles (6 AM and 6 PM) since the worst case thermal errors will occur at one of these conditions. The components that contribute to the above mentioned errors are limited to the interferometer horns, the cornucopias, and the structure between the two. The manner in which the errors occur for each of the sun angles considered is summarized in Table 22.

In the distortion calculations that follow, only normal mode temperatures are considered, since the error specification does not apply to the multi-channel mode of operation. Also, the distortions of the top cornucopia, aperture A4', are not considered since this aperture is only used in the multi-channel mode. The method used in calculating these errors will now be described.

4.3.2.1 Symmetric Sun Angle Thermal Errors

The change in distance between the phase centers of the interferometer horns in the roll and pitch axes causes east-west and north-south rotation errors. The relationship between this distance and rotation error has been determined in the Pointing System Section as

$$\text{rotation error} = 5.73 \frac{dD}{D} \quad (90)$$

where

dD = change in phase center separation

D = initial phase center separation.

TABLE 22
SUN ANGLE ERRORS

Sun Angle	Component	$\Delta\psi_{12}^T$	$\Delta\psi_{23}^T$
symmetric	interferometer	change in distance between the phase centers in the roll axis	change in distance between the phase centers in the pitch axis
	cornucopia	rigid body rotation of cornucopia about the pitch axis due to support structure rotations	rotation of the cornucopia's reflector about the roll axis
non-symmetric	interferometer	(1) change in distance between the phase centers in the roll axis plus (2) unequal translation of the phase centers in the yaw axis due to non-symmetric heating	change in the distance between the phase centers in the pitch axis
	cornucopia	east-west component of rotation of cornucopia's reflector	north-south component of rotation of cornucopia's reflector

For the symmetric sun angle conditions, these translations are the sum of the change in length of the antenna support structure between the horns due to its bulk temperature differing from room temperature, and the change in distance between the horn's phase centers due to rotation of the antenna support structure. The change in distance between interferometer horn phase centers was determined for the interferometer bulk temperature furthest from room temperature (56°F). Substituting in the previous equation, the contributions to $\Delta \psi_{12}^T$ and $\Delta \psi_{23}^T$ were computed.

North-south rotation errors are caused by cornucopia reflector rotation due to temperature gradients normal to the cornucopia reflector surface. The prime mode of heat transfer normal to the honeycomb structure was found to be conduction, and the differential between skins was 2.5°F. The slope of the linear strain distribution produced by this gradient is equal to 2.5 times the coefficient of thermal expansion divided by the reflector thickness. From the geometry of a beam in pure bending, the rotation of the beam in radians is equal to this slope times the length of the beam.

East-west rotation errors are caused by rigid body rotations of the cornucopias relative to the interferometers; this is due to rotations of the tubular members in the antenna support structure that connects the interferometers to the cornucopias. To evaluate the rotation of the tube, a detailed circumferential temperature distribution was determined by a 20 node two dimensional analysis that considered solar inputs, radiation to free space, radiation between nodes, and conduction between nodes with additional uniform heating added to satisfy the previously calculated bulk temperature. This resulted in a non-linear temperature distribution across the tube with extreme values differing by 13.3°F. From this temperature profile, a linear strain distribution was found, using equations developed by Timoshenko⁶, that satisfied the condition that the tube be free from external forces during rotation i. e., the cantilevered tube is not prevented from rotating by any external constraints. The rotation of the tubular beams was then computed from the linear strain distribution.

4.3.2.2 Non-Symmetric Sun Angle Thermal Errors

For the non-symmetric sun angles, east-west rotation errors due to interferometer horn distortions are caused by changes in the distance between the horn phase centers due to bulk temperature changes, and by unequal translation of the phase centers in an axis parallel to the yaw axis due to non-symmetric heating. The first error was computed from the non-symmetric bulk temperature furthest from room temperature, (64°F). The second was calculated by conservatively assuming the temperature difference between the interferometer horns was the same as the maximum differential between cornucopias (132°F). Knowing the difference in translation of the phase centers, the rotation error in radians is equal to this value divided by the horn separation distance. North-south rotation errors are also caused by changes in distance between interferometer horn phase centers in the non-symmetric condition. This rotation error is numerically the same as the first of the east-west rotation errors.

In addition there is a temperature differential across the cornucopia's reflector that causes a rigid body rotation of the elliptical reflector about its major axis, with a significant component in the east-west directions. The temperature differential was assumed to be the same as the maximum between cornucopias (132°F). The rotation vector for the reflector was determined and its east-west component found.

4.3.3 Thermal Pointing Error Summary

The thermal pointing errors are summarized in Table 23 for both the symmetric and non-symmetric sun angles.

TABLE 23
THERMAL POINTING ERROR SUMMARY

Sun Angle	Component	$\Delta \psi_{12}$ (East-West)	$\Delta \psi_{23}$ (North-South)
symmetric	interferometer	.00216°	.00216°
	cornucopia	.0242°	.0637°
SUM		.0264°	.0659°
non-symmetric	interferometer	.0399°	.0004°
	cornucopia	.038°	-
SUM		.078°	.0004°

The total thermal error about any one axis is the arithmetic sum of the components rather than a RSS total since the errors combine in a deterministic manner. The total thermal errors (ψ^T) are then combined in an RSS manner with the other system errors and introduced into the overall system error equation to generate a total pointing error of 0.08 degrees for symmetric sun angles and 0.08 degrees for non-symmetric sun angles.

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SECTION V

DESIGN VERIFICATION RECOMMENDATIONS

This section recommends the various tests and development of critical components necessary to verify the design concept proposed in the main body of this report. The recommendations are presented in sections as applicable to the various subsystems.

5.1 RECOMMENDED RF TESTS

5.1.1 Horn-Reflector Evaluation

The preliminary experimental results summarized in Section 2.2.1 have verified the feasibility of using a multi-mode horn-reflector antenna to reduce sidelobes. Experimental work is necessary in work of this nature since analysis is limited by the second order effects which at sidelobe levels of 30 db are significant. One example of this is the horn-reflector radiation pattern in the longitudinal plane. Here, radiation at some aspect angles is the sum of radiation from the horn section directly, and that from the parabolic surface. Complexity of this condition forces analytical treatment to neglect the horn component.

In view of the shortcomings of analytical treatment, an experimental program to determine the detailed performance characteristics of the multi-mode horn-reflector antenna is required. Particular emphasis should be placed on demonstrating the limit of low sidelobe capability. The major parameters that should be investigated are: horn flare angle, straight or folded (Cass-horn) horn configuration, mode exciter form, and horn to aperture interface configuration. The characteristics that would be measured are: General radiation pattern properties, sidelobe level (in the principal and several diagonal planes), bandwidth, cross polarization levels, and input VSWR.

As presently envisioned, evaluation of horn-reflectors with flare angles of 14, 20 and 26.5 degrees is necessary. These antennas would be compared to determine the degree to which sidelobe control is dependent on flare angle. The antennas with an α of 14 and 20 degrees would also be evaluated in folded form as illustrated in Figure 71. This configuration enables the use of small flare angle designs without their attendant large physical size.

5.1.2 Horn-Reflector Thermal Sensitivity Tests

The multi-mode horn-reflector communications antenna chosen in this conceptual design study is subjected to a variety of thermal environments. This changing environment is due largely to the variations in solar illumination with orbital motion and season change. Also, the power dissipated in the RF transmission components is radiated in part to the horn-reflectors thereby significantly influencing their thermal environment.

The thermal environment about the antennas will influence their pointing direction and sidelobe level since it causes distortion in the reflector and horn sections. The amount of radiation pattern degradation that occurs in a horn reflector when

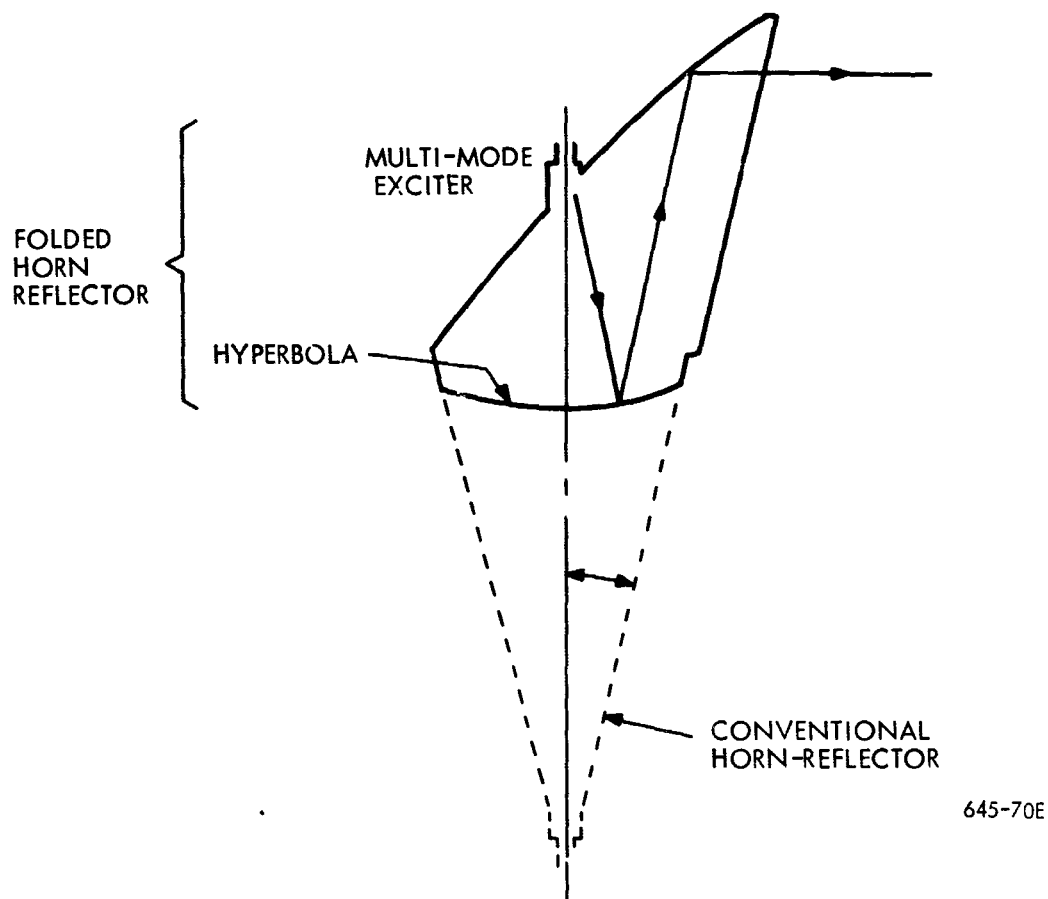


Figure 71. Folded Horn-Reflector Antenna

physically distorted can be determined analytically. The analytical procedure is a complex one, however, since the horn reflector is a non-symmetrical structure, and the analysis requires three successive steps which leads to compounding of errors. Temperature distributions must be determined by a three dimensional radiation-conduction analysis. Given the temperature profiles, distortions can be approximately determined using a finite element structural analysis. Finally, these distortions can be transformed to radiation pattern performance.

In view of this difficulty it is recommended that a combined analytical and experimental program be implemented to assess the thermal sensitivity of the horn-reflector with respect to sidelobe level and pointing direction. This program would consist of the following: First, the worst case thermal environments for the antenna would be analytically determined. From this data the temperature profiles of the antenna structure would be computed.

Next, this temperature distribution would be simulated on a full-scale horn-reflector constructed in a manner suitable for space use. This simulation would be conducted in a microwave anechoic chamber. Beam pointing errors and sidelobe degradation would be measured in a conventional manner and compared to the baseline characteristics measured in a normal laboratory environment.

This combined analytical and experimental work would show conclusively the operational performance of the conceptual design.

5.1.3 Rotary Joint Development

The rotary joint necessary in the system is of complex design. The complexity arises from the high average power (four kW total), the number of channels and the high isolation requirement. The first factor necessitates the use of large conductors. This together with the four channel capability results in transmission sections that can propagate higher order modes. To insure suppression of these modes, multi-point excitation of the coaxial transmission lines is required. Again because of the high average power, waveguide excitation of the coaxial rotary sections is necessary. Thus, the waveguide power dividers used in feeding the joint add considerable mechanical and electrical complexity to the device.

While all of the techniques employed in the rotary joint are proven, such a joint operating at four kW total average power in X-band has never been built. Hence, the quality of performance (VSWR, isolation, etc.,) of such a device has not been demonstrated. More important is the lack of demonstrated evidence that such a joint can safely handle the one kW average power per channel. In the preliminary design of this study, losses in the joint coaxial line, losses in transition regions, and in the waveguide feed network have been computed using conventional transmission line attenuation formulas. Because the joint is comparatively small in length in terms of wavelengths and since considerable transitions and physical discontinuities exist in the joint, high localized losses, not predicted by the former attenuation formulas, may exist within the joint. High localized losses may inhibit the power capability to below the required four kW average total level.

Development and test of a 4-channel rotary joint is recommended. This effort would verify the rotary joint performance capabilities, particularly its average power limitation. Testing of the joint would be for the standard electrical parameters, i.e., insertion loss, VSWR, isolation, wow, and power handling. The latter test would require a sufficiently high average power source such that an inner temperature profile of the joint could be made to predict the maximum average power capability.

5.1.4 RF Power Handling Tests

The power handling tests that are recommended for the various systems components are summarized in Table 24. Test of the antennas is not necessary in that their large size precludes power handling failure in any of the three possible mode, i.e., multipaction, ionization and average power heating. In conducting the tests on the various components the appropriate pressure and thermal environments must be simulated. The environments of the various components would be estimated by computation and simulated heaters, shrouds, etc. in the conventional manner.

The performance of RF power tests on broadcast satellite components will be hindered by the general unavailability of high-power sources at 12.0 GHz. A survey of power sources has indicated that no magnetrons or pulsed klystrons that provide one kW or more of peak power at 12.0 GHz are available. This is unfortunate in that a comparatively expensive tube (magnetron) cannot be used to determine the multipaction or ionization breakdown capability of a component.

Some CW sources are available in the upper portion of X-band. These TWT sources generally fall in power ranges either below 200 watts or above eight kilowatts. The latter source would provide full simulation of four kilowatt channels simultaneously and peak power test capability to eight kilowatts on individual components. While this peak power may not be sufficiently large to cause breakdown (by any mechanism), it would prove substantial power handling safety factors in the components tested.

The CW TWT high power source is limited in range to within 3 db of its rated output. To provide a variable power source, a variable attenuator will be required between the source and component under test. This will enable definition of the exact level at which breakdown occurs provided it is below the maximum available power. While it is desirable that the power limits of all components be determined, this will probably not be possible because of the limited peak power available (8 to 10 kW). The available peak level is, however, large enough to demonstrate a sufficient safety factor relative to the one kW operational power level.

5.2 POINTING SYSTEM VERIFICATION

5.2.1 Sensing System Verification

The purpose of the test is to verify that the error budget assigned to the sensing system is realistic by testing a prototype of the sensing system electronics. The prototype shall include those functional elements of the satellite borne system required to perform tests on one axis of the interferometer system and the polarization angle detector. In addition, the equipment shall be sufficient to test the operation of the interferometer front end in the redundant mode. Component choice and packaging shall be

TABLE 24. RECOMMENDED RF COMPONENT POWER TESTS

Component	Multipaction	Ionization	Average Power	Remarks
WR-90 Waveguide			x	
WR-90 Waveguide diplexer	x	x	x	
Polarizer			x	
RF Electromechanical	x	x	x	Should be tested with lubricant outgassing products present
WR-90 Flexible Waveguide			x	
Four Channel Rotary Joint	x	x	x	Should be tested with lubricant outgassing product pre- sent. Average power tests should be conducted on the four channels simulta- neously.

consistent with the design objectives for the flight equipment, and shall be such that the test results shall be applicable to the flight equipment. A functional description of the sensing system can be found in Section 3 of this report. In addition, the sources of system error have been described in Section 3.5.

The following test descriptions cover the prime mode interferometer, prime mode polarization angle detector, and redundant mode interferometer. It is not proposed to carry out any separate tests on the redundant mode polarization angle detector since its operation is a subset of prime mode operation.

5.2.1.1 Prime Mode Interferometer

This test is applicable to that part of the receiver which is common to both prime and redundant modes. Errors to be evaluated include the receiver channel phase error, phase shifter error, and the phase detector error.

The source and magnitude of receiver channel phase errors will be investigated by introducing test signals at the input of the two channels. The input generator will consist of an oscillator of appropriate frequency and a calibrated phase shifter. Phase difference measurement will be made at the outputs of the RF preamplifier, mixer, and IF amplifier. Sets of measurements will be made under different environmental conditions to simulate the variations to be encountered during flight. A test program shall be carried out to determine the drift in phase shift over an extended period of time under simulated space environment.

Additional tests to determine phase shifter errors may be combined with the tests made to determine receiver channel phase errors. The phase difference at the output of the two phase shifters is determined with the two input IF signals in phase for different values of the phase shift command signal. A calibration curve is prepared. Tests shall be performed to determine how the calibration curve is affected by environmental changes and by long-term drift.

Further tests to determine the magnitude of phase detector errors may also be combined with the previous tests. The calibration curve for the phase detector is determined by varying the phase between the two input signals. The measurements should be performed in a cyclic manner, i. e., starting from zero phase shift, increase the phase difference in one sense to a maximum; then reduce the phase shift to zero by increments. Continue with phase difference of the opposite sense to a maximum and then back to zero. The resultant response should be examined for curvature and double valued response. The phase shifter tests shall be made so as to determine the effects of environmental change and long-term drift.

5.2.1.2 Prime Mode Polarization Detector

Errors to be evaluated include the receiver channel error, phase shifter error, and phase detector error. Receiver channel errors are those which produce a phase shift between the FM modulation signal and the AM modulation signal. Because of the different characteristic of the modulation it is not feasible to compare signals at the inputs to the two detectors (FM detector and envelope detector). Hence the phase shift will be measured at the output of the detectors and compared to the phase shift introduced between the two modulations impressed on the RF input to the mixer.

Phase shifter and phase detector performance will be tested in a manner similar to that described for the interferometer.

5.2.1.3 Redundant Mode Interferometer

Additional tests will be carried out to determine the magnitude of the transmission channel phase error and the phase lock loop error which are in addition to those occurring in the prime mode equipment.

Transmitter channel errors are those which result in a different phase between the sequential segments of signal which the commutator applies to the two transmitting horns comprising one axis of the interferometer. The commutated signals must be compared to the uncommutated carrier and any difference in phase shift between the commutated segments must be measured. Measurements are to be made under varying environmental conditions and over an extended period of time sufficient to determine the effect of drift on the phase difference.

The IF amplifier at the receiver input is common to both channels and is not expected to contribute phase error. Additional phase shift error introduced by the phase-locked loops will be determined by measuring the relative phase between the phase lock loop outputs when the two input sampled signals are in phase. The effect of varying environmental conditions and equipment aging will be determined.

5.2.2 Steering System Analytical Verification

A dynamic analysis of the closed loop performance of the steering system is recommended. The analytical model of the system shall include the following component characteristics. The sensing system model will take into account gain variation with pointing error. The servo motor model will include non-linear torque and speed relationships. The load will include a two degree of freedom model of the antenna and the effects of coupling dynamics such as gearing inertia, friction, and backlash. Geometrical factors such as cross coupling and gain variations derived from the equations relating gimbal motion to interferometer beacon coordinates as measured by the interferometers and the equation relating gimbal angles to polarization angle will also be modeled. Other factors to be considered include inputs caused by satellite motion about its inertial reference frame and redundant mode transportation lag effects due to uplink transmission time.

The purpose of the analysis will be to verify by computer analysis that stable system operation can be achieved over the entire range of pattern coverage with a velocity constant of the value described in the report with the effects described in the preceding paragraph taken into account.

5.3 MECHANICAL VERIFICATION

5.3.1 Structural Model Tests

Although detailed vibration analysis can predict natural frequencies and mode shapes with good accuracy, steady state response at resonant frequencies and transient response to a shock pulse is more difficult to evaluate. This is due to the uncertainty

of energy dissipation or damping in the structure, the amount of which dictates response to a given input. Therefore, maximum stress in each component is strongly dependent on the amount of damping present in the structure. Although methods are available for predicting damping in very simple configurations, calculating the amount of damping in a complex structure such as an antenna system is too involved to be useful in design, and accuracy would be questionable. The most efficient procedure is to assume structural damping factors based on test experience, compute response and stress, and arrive at a design. Then a structural model that dynamically simulates the antenna system is constructed and tested. The results of the test either confirm the original design, or suggest changes.

The goal in the design of a structural model is to dynamically reproduce the actual system, and include in detail all the critical areas of the structure so that strain gage measurements can be made on these components. In the present antenna system, the components that should be reproduced in reasonable detail are the cornucopias, tubular support structure, lockout pins, 360-degree shaft and housing, main support cone, and main support plate. The waveguide system, diplexers, RF switches, electronics, rotary joints, and limited motion drives could be simulated by masses placed at the appropriate locations. These items are excluded since they are not essential to the structural integrity of the antenna system. It would be desirable to mount accelerometers at these component locations on the structure, and use these data as vibration specifications for the components. A structural model can be fabricated with far less effort than an actual system, and also provides an opportunity for development of fabrication techniques.

The structural model would be subjected to vibration and shock in three orthogonal axes to the specified levels. Strain gages, placed at the critical stress areas, provide the ultimate information on component safety factors. Accelerometer data are useful in determining damping factor by working backwards through the dynamic analysis, which can then be used to evaluate the effect of design changes on the antenna's structural integrity.

5.3.2 Thermal Vacuum Test

The thermal analysis of the antenna system can be expected to predict bulk temperatures quite accurately since the thermal analysis is based on well known values for key parameters such as thermal conductivity, specific heat, and mass. The purpose of the thermal vacuum test is to verify operation under simulated space environments and thus determine any adverse effects due to thermal gradients, local hot spot temperatures, or thermal distortions since in many cases, these effects cannot be calculated with a high degree of confidence.

The complete antenna system should be used for the thermal vacuum tests since hot spot temperatures of the components such as diplexers, electronics, and rotary joints are of major interest. There is enough of a problem simulating heat loads, and if the structure was also simulated, the test results would be questionable. Individual components such as a cornucopia or rotary joint could be tested separately from the antenna by simulating the appropriate boundary conditions, but any test of the complete system should be performed on the actual hardware.

The thermal vacuum test of a complete antenna system is a major undertaking. The thermal environment must be simulated by high intensity lamps to provide equivalent solar radiation and a low temperature (liquid nitrogen) shroud around the periphery of the chamber to simulate radiation to outer space. In addition, the satellite interface temperatures must be modeled using hot plates or cold plates and the antenna dissipative power must be simulated. Instrumentation must be provided to operate the antenna and to verify that it is operating properly. Actual test operation should occur with the chamber pressure less than 10^{-6} Torr so that convective heat transfer is minimized and the proper temperature profile will be achieved.

A thermal vacuum test of an antenna differs from the testing of other satellite components in that it is not possible to verify all of the RF radiation characteristics while in the chamber due to interactions with the chamber walls. Thus, sufficient instrumentation must be provided to determine the thermal profile of the RF apertures, interferometer horns and the structure between. Subsequently, the thermal profile may be simulated in an anechoic chamber where it is possible to measure degradation in RF performance in terms of gain, sidelobe levels, and thermal pointing errors.

In planning a thermo vacuum test on an antenna of this size, sufficient lead time must be provided to procure a chamber that will hold the 6 ft. by 8.5 ft. by 3.5 ft. (182 cm by 260 cm by 107 cm) antenna plus the environmental simulation equipment and instrumentation.

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SECTION VI DESIGN VARIATIONS

This section assesses the effects of five design variations selected to evaluate the versatility of the design concept and to provide data for tradeoff studies with other satellite subsystems.

6.1 CHANNEL SEPARATION VS. CHANNEL ISOLATION

This variation considers effects caused by reducing the channel separation from 2.0 channel widths to 1.2 channel widths while at the same time reducing the channel isolation requirement from 40 db to 30 db.

In the prime mode, reduction of the channel separation from 2.0 channel widths (channel width = 50 MHz) to 1.2 channel widths does not influence any of the RF components with the exception of the 360 degrees/day rotary joint and the waveguide electro-mechanical switches. Reduction in the channel separation reduces the bandwidth over which the chokes in these devices must operate. Because of this, the attainment of high channel isolation should be somewhat simplified.

Operation in the auxiliary mode, where the diplexer is used to combine two signal channels into one antenna terminal, is influenced by the reduced channel separation. If channel separation is reduced to 1.2 channel widths the diplexing device is not physically realizable unless isolation between the two diplexed channels is reduced to 20 db. For the latter channel separation and isolation, the insertion loss of the diplexer will remain at 0.6 db.

Again, in the auxiliary mode, if the channel separation is retained at the 2.0 channel widths and the isolation requirement is reduced to 30 and 20 db, the insertion loss will be reduced to levels of 0.4 db and 0.3 db, respectively.

In summary, the smaller channel spacing and reduced isolation variations do not markedly influence the component design in the prime mode. For the auxiliary mode however, the variations affect the diplexing component. Reduction in channel spacing while requiring the high isolation (30-40 db) is unrealizable in the diplexer. Reduction in diplexer isolation for the 2.0 channel width spacing benefits the system in terms of reduced loss. Size and weight reduction associated with the less severe isolation requirements in the diplexer are insignificant to the system.

6.2 INCREASED RF POWER WITH FOUR CHANNELS

This variation increases the RF Power from 1 kW per channel to 2.5 kW per channel with the number of simultaneous channels remaining fixed at four.

An evaluation of the peak power handling capability of the RF components shows the limiting component in the prime mode of operation to be the 360-degree rotary joint. The rotary joint is limited by multipaction breakdown in the choke sections at the 15 kW level thus providing a safety factor of six on peak power.

An evaluation of the average power handling capability must consider that this power level represents a power dissipation increase in the RF components of a factor of 2.5. This was incorporated into the bulk temperature analysis computer program and the 12:00 noon sun angle condition was run as a steady state calculation. The effect of the power increase on temperatures is more severe at this sun angle, since 12:00 noon is the hottest condition, and it has been demonstrated in previous calculations that this particular sun angle can be approximated as steady state. Results of this analysis show that the temperatures are excessive in the mechanical drive area, especially in the 360-degree and limited motion rotary joints, where temperatures reach 342°F and 268°F, respectively. These high temperatures would cause problems in thermal stress, material degradation, and thermal distortion. It should also be realized that the computed temperatures are bulk temperatures of the components, and that hot spot temperatures will be considerably higher, due to large thermal resistances.

Changing thermal coatings would not have a significant effect on the rotary joint temperatures, since except for the cornucopias the antenna is assumed to be coated with white paint. If the cornucopias were painted white and the heat was effectively transferred from the rotary joints to the antennas, the cornucopias would be too cold during eclipse, when all the heat sources are shut off. Therefore, the only feasible techniques for removing the heat are to either use cooling devices or to provide paths that transfer the heat to the exterior of the satellite where it can be radiated to free space. This second approach is more desirable since it is a static system which is more reliable.

It is concluded that this variation would require extensive thermal redesign in the area of the 360-degree rotary joint to reduce the higher temperatures caused by the increased average power dissipation. Peak power breakdown is not a problem.

6.3 INCREASED RF POWER WITH TWO CHANNELS

This variation increases the RF power from 1 kW per channel to 6 kW per channel but reduces the number of channels from four to two. Thus the total power is increased from 4 kW to 12 kW.

The peak power handling limitation is again determined by multipaction in the choke sections of the 360-degree rotary joint at 15 kW per channel. Operation at 6 kW per channel results in a safety factor of 2-1/2 on peak power breakdown.

Assessment of the average power handling capability of the system starts by noting that while total power has increased by a factor of three, the power dissipated has not increased in proportion. This is due to the reduction in the required number of channels and results in replacing the 4-channel 360-degree rotary joint with a 2-channel rotary joint which has one-third the dissipation at the same total power input.

A worst case bulk temperature analysis was run with the power dissipations adjusted to reflect the increased system power as well as the improved rotary joint efficiency. The result of this analysis showed that the bulk temperature of the 360-degree 2-channel rotary joint would not exceed 160°F and that the flexguide bulk temperatures would not exceed 256°F. In addition, since the 2-channel rotary joint design is less complex than the 4-channel design, it achieves better heat paths thus reducing the temperature gradients within the joint.

It is concluded that this variation is feasible. The thermally sensitive components adjacent to the rotary joint will not be exposed to excessive temperatures. The flexguide, while quite warm, is not adjacent to sensitive components and can itself operate at the elevated temperature without deterioration.

6.4 IMPROVED SATELLITE ATTITUDE CONTROL

This variation assesses effects on the antenna system that would result from reducing the maximum satellite attitude position error from 5 degrees to 0.1 degree and reducing the maximum rate error from 1 degree/min. to 0.1 degree/min.

The system described in the body of this report includes a 3-axis antenna mounted on a stabilized satellite. The three axes are required to move the pattern center throughout the coverage area and orient the pattern axis properly. The solar panels are required to move about one axis for a limited travel to compensate for equator/ecliptic angle. If this concept is retained, an order of magnitude improvement in satellite attitude control has no substantial effect on the size, weight, and complexity of antenna steering equipment required. The decreased satellite drift rate could result in a lower requirement on the velocity constant and a lower slew rate requirement. The former reduction would have no impact on equipment size and weight, the latter could result in a smaller motor with a potential aggregate saving of one or two pounds of weight for the system. In view of the total weight of the antenna system (more than 200 pounds) this possible decrease is not significant. Additional improvements of lesser impact include a reduction in the magnitude of the required steering range and a reduction in the error cross-coupling due to the reduced steering range.

An alternative approach would be to eliminate two of the axis of rotation and the associated drives and use the satellite attitude control to position the pattern to the desired location. One axis of rotation must be retained to permit motion relative to the oriented satellite. Using the satellite attitude control with an accuracy of 0.1 degree in two axes and the previously determined steering system accuracy in the third axis results in a slightly degraded overall pointing accuracy of 0.14 degree.

This approach has two serious drawbacks associated with equipments other than the antenna. First, the satellite attitude control must operate continuously when pointing at other than the subpoint even if the pattern location is not changed. This is due to the continual change in antenna pointing direction relative to the solar oriented satellite. Thus such a configuration places inordinate demands on the satellite attitude thruster propellant.

The second drawback of this configuration concerns the solar panels. By steering the satellite to provide the requisite north-south coverage, an additional 7-degree solar panel misalignment may be introduced resulting in a worst case misalignment of 30 degrees. This much misalignment reduces the power output of the solar panels by approximately 15 percent.

It is concluded that this single axis configuration has more disadvantages than advantages.

6.5 EARTH ORIENTED SATELLITE WITH IMPROVED ATTITUDE CONTROL

This variation considers an earth oriented satellite with its longitudinal axis parallel to the equatorial plane and having an improved attitude control system with a maximum angular position error of 0.1 degree and a maximum rate error of 0.1 degree/min.

The significant change resulting from this variation is that only one limited motion steering axis is required. This axis is parallel to the satellite longitudinal axis and provides capability for north-south coverage without satellite rotation. Such rotation is not desirable since it positions the solar panels at oblique angles to the sun such that performance is seriously degraded. East-west coverage is obtained by rotating the satellite about an axis that does not affect the solar panels.

The improvement in the attitude control system is essential to eliminating two of the antenna steering axis. Without such an improvement, the pointing accuracy requirements would have to be met by antenna steering about three mutually orthogonal axis. The improved attitude control also causes other less significant improvements by reducing steering range, error cross-coupling, velocity constants, and motor size.

A configuration to implement this design variation is shown in Figure 72. It was achieved by minor modifications to the prime system concept described in the main body of this report. The required changes were only in the drive area where the previous 3-axis drive design was replaced by a 1-axis limited motion drive about the axis that previously was capable of 360 degrees of rotation. The previously developed components for the limited motion axis were again implemented into the design.

Flexures are used for the bearings thus eliminating the need for lubrication. The possibility of using flexures without lockouts was considered. A custom flexure could be designed to withstand the launch loads, however, the device would have an unreasonably high resisting torque when rotated five degrees. Thus the recommended configuration utilizes pin pullers and lockouts and small flexures. The locking out during launch is achieved by four pins in the same configuration that was used in the present design. Bending moments and shear and axial loads are transferred from the antenna support tube, through the four pins that are 90 degrees apart, and into the housing. The housing is tied to the satellite in the same manner as the present design by utilizing the main support tube and plate. Due to the lockouts, the launch loads bypass the bearings and thus low resisting torque flexures can be used. The rotating portions of the flexures are tied together by four small diameter tubes, 90 degrees apart. The stationary portions are tied to the housing by four tubes 90 degrees apart, but out of phase from the first set by 45 degrees to allow for the limited rotation. This configuration, though not able to withstand the launch loads with the pins retracted, can support the system in a one g environment to allow prelaunch testing.

The design utilizes a harmonic drive located on axis below the lower flexure. This drive provides low backlash and a hermetically sealed motor and gear head. The mesh between the flex and circular spline is the only point exposed to a vacuum and requiring special lubrication.

Flex waveguide is again used for the limited motion rotary joint since it provides minimum RF loss and has excellent mechanical characteristics. The flexguide sections are mounted as cantilever beams located such that the axis of rotation bisects the length of the beam so the stresses developed during rotation are minimized.

The configuration to implement this design variation as shown in Figure 72 has many advantages. The weight is reduced 16 percent to less than 200 pounds. RF losses are reduced to 0.3 db in the prime mode by eliminating the 360-degrees rotary joint and one limited motion rotary joint. Peak power is still limited at 15 kW per channel by the choke joints in the RF switches. Average power heating due to the RF losses is reduced by 50 percent from that of the prime configuration. Thus the previously determined thermal profile could be expected at 8 kW total RF power and temperatures should be manageable up to 12 kW. The design is such that there are fewer hot spots and generally lower thermal gradients. The pointing accuracy obtained when using the sensing system described in Section 3 and satellite steering about two axes is computed at 0.13 degrees. Satellite attitude control is not required continuously as it was in the variation described in Section 6.5 and the solar panel operation is not degraded by altering the sun angle providing that the pattern configuration is not oriented to the local vertical when steered away from the subpoint.

This configuration takes full advantage of the components developed for the prime concept and provides many additional attractive features.

SECTION VII

PERFORMANCE SUMMARY AND CONCLUSIONS

This section summarizes and evaluates the calculated performance characteristics of a recommended conceptual design and compares them with the specified design requirements. Outstanding features of the design are discussed and conclusions are presented.

7.1 PERFORMANCE SUMMARY

Performance against the design specification is summarized in Tables 25, 26 and 27. In all cases the design requirements have been met or bettered.

Table 25 summarized the RF system performance. Channel width is limited to twice the specified value only by the 360-degree rotary joint. Alternate configurations that do not require this component could have a much larger channel width.

The design provides a safety factor of 15 against peak power breakdown at the specified levels and a thermal design that maintains reasonable temperatures at continuous average power levels 150 percent of the 4-kW specified.

Circular polarization is achieved with an all-aluminum periodic loaded pin polarizer located at the input to the horn feed. No dielectric material is required and thus the component will handle power levels greatly in excess of that specified.

The 0.62 db attenuation, while attributable in part to each component, is primarily caused by the 360-degree rotary joint. The calculated phase delay of 0.07 degrees is due to the length of the waveguide.

The polarization axial ratio is affected only by the aperture and the polarizer. Individually these components have axial ratios of 0.48 db and 0.15 db, respectively. However, the polarizer can be rotationally positioned relative to the aperture in such a manner that cancellation occurs, thus reducing the overall axial ratio well below the specified level of 0.42 db.

The pattern configuration is changed by the actuation of a mechanically operated RF transfer switch and is readily completed in well under ten seconds.

A scale model of the recommended cornucopia apertures was used to obtain actual pattern measurements. Although the model was not constructed to precision tolerances, the near sidelobes were measured at -25 db in one principal plane and -29 db in the other. The backlobes were less than could be measured by the 40 db dynamic range of the instrumentation. The measured sidelobe levels were obtained without resorting to auxiliary radiators or other blockage compensation techniques.

The only components that can affect channel isolation are the choke joints of the 360-degree rotary joint. It is not possible to make meaningful calculations in this area; however, measurements on similar components indicate the desired level of isolation can be achieved by the recommended design.

TABLE 25. RF SYSTEM DESIGN REQUIREMENTS COMPLIANCE

<u>Parameter</u>	<u>Specified Value</u>	<u>Achieved Value</u>
Center Frequency	12.2 GHz	As Specified
Channel Width	50 MHz	> 100 MHz
Channel Separation	100 MHz	As Specified
Number of Changes	4	As Specified
Power per Channel	1 kW (CW)	6 kW
Total Power	4 kW	6 kW
Operation	Continuous	As Specified
Polarization	Circular	As Specified
Beamwidth	2.7° HPBW	As Specified
Beam Separation	0.55 and 0.72 HPBW	As Specified
Beam Configurations	2	As Specified
Attenuation	1 db	0.62 db
Phase Delay	1 degree	0.07 degree
Polarization Axial Ratio	0.42 db	0.33 db
Channels per Beam (Multichannel Mode)	2	As Specified
Pattern Configuration Change Time	10 Minutes max.	< 10 seconds
Near Sidelobes	25 db	25 db
Front-to-Back Ratio	40 db	> 40 db
Channel Isolation	40 db	> 40 db
Channel Mixing Products	40 db	> 40 db

TABLE 26. POINTING SYSTEM DESIGN REQUIREMENTS COMPLIANCE

<u>Parameter</u>	<u>Specified Value</u>	<u>Achieved Value</u>
Pointing Error	±0.1 degree (max.)	±0.086 degree
Spacecraft Angular Error	±5 degrees (3 axis)	As Specified
Spacecraft Angular Rate	±1 degree/minute	As Specified
Spacecraft Station Error	±0.5 degree (Lat & Long)	> Specified
Spacecraft Station Rate	0.1 degree/day	> Specified
Pattern Steering Range (Latitude)	45°N to 10°S	> Specified
Pattern Steering Range (Longitude)	20°E to 20°W	> Specified
Pattern Steering Time	15 minutes (max.)	< Specified
Prime Direction Sensing System	Spacecraft Sensing	As Specified
Complementary Sensing System	Earth Sensing	As Specified
Sensing System Power Level	ERP of 50 DBW (max.)	< Specified
Ground Station Location	S/C View > 25° Above Horizon	Up to 90° Acceptable

TABLE 27. MECHANICAL DESIGN REQUIREMENTS COMPLIANCE

<u>Parameter</u>	<u>Specified Value</u>	<u>Achieved Value</u>
Spacecraft Attitude	Long. Axis Normal to Equatorial Plane	As Specified
Antenna Deployment	Solar Oriented	As Specified
Antenna Orientation	36-inch (91 cm) dia. Mount on Long. End	As Specified
Launch Volume	Sidelooking	As Specified
Launch Weight	370 Cubic feet (10.478 m ³)	220 cubic feet
	250 pounds (113.4 kg)	6.230 m ³
Launch Environment		237 pounds (107.5 kg)
Shock	30G at 8 MS	As Specified
Vibration	1 G _{rms} at 10-60 Hz	As Specified
	2 G _{rms} at 60-2000 Hz	As Specified
Acceleration	6.8 G max.	> Specified
Acoustic Excitation	136 db/Octave (50-1000 Hz)	> Specified
Shroud Temperature	120°F	> Specified
Space Environment	Synchronous Orbit	As Specified

Channel mixing products are neither generated nor attenuated by any of the antenna components. Thus whatever is introduced at the input to the antenna will be radiated from the apertures.

Table 26 summarizes the pointing system performance. The major considerations affecting pointing accuracy are thermal distortions. These have been evaluated at the limiting sun angles, with the worst-case values introduced into the overall pointing error of 0.086 degree.

The allowable spacecraft angular orientation error of ± 5 degrees constrains the proposed design by adding to the required steering range of the limited motion axes. The north-south pattern steering requirement of 45 degrees N to 10 degrees S latitude translates into a steering angle of +7 degrees and -1.75 degrees, which can be considered as ± 4.38 degrees. Adding this value to the satellite motion of ± 5 degrees results in a total N-S steering range requirement of ± 9.38 degrees. The need for ± 1.8 degrees of rotation about the RF boresight is established by the requirement to orient the pattern grouping to the local latitude circle at the extreme corner of the coverage area. Adding this to the satellite motion of ± 5 degrees results in a steering range requirement of ± 6.8 degrees in the other limited motion axis. Both limited motion axes provide ± 10 degrees of motion, thus exceeding requirements. The E-W coverage calls for ± 3.5 degrees motion about the axis providing full 360-degrees rotation to compensate for the solar orientation of the satellite.

The spacecraft maximum angular rate of one degree per minute combined with the steering range stated previously and with the requirement for relocating the pattern to any point within the steering range within 15 minutes, results in an easily achieved maximum speed requirement of 1.27 degrees per minute. The spacecraft angular rate also enters into the determination of the velocity constant required to keep the velocity error at an acceptably small value.

The spacecraft station error of 0.5 degrees in latitude and longitude and the station error rate of 0.1 degree/day are second order contributors to the requirements for steering range, maximum velocity and velocity constant. Thus these design requirements may be increased significantly without affecting the design.

The sensing system requirements are met by an interferometer system for measuring north-south or east-west pattern pointing error, together with a polarization angle sensing system to measure pattern rotation about the boresight axis. Neither system uses the communication signal; therefore, the ERP requirement does not apply. Only one earth station is required and it may be located anywhere within the coverage area.

Table 27 summarizes compliance to the mechanical design requirements. A detailed weight computation of the design shows a total weight of 237 pounds (107.5 kgs). This weight, although subject to reduction by more refined structural analysis, still achieves by a comfortable margin the 250 pounds (113.5 kg) limit of the specification.

The shock and vibration requirements provided the governing dynamic loading conditions to which the structural members were sized with nominal safety margins. Gross margins exist over the acoustic excitation and steady state acceleration requirements.

The antenna thermal environment is determined by the shroud temperature during launch and by the space environment during synchronous orbit. It is found that the in-orbit environment is the more severe and, therefore, governs thermal design. A passive thermal control system has been evolved that limits RF degradation, pointing error and material degradation sufficiently that all design requirements are met or bettered.

7.2 SIGNIFICANT DESIGN FEATURES

The communication system apertures of the recommended design exhibit none of the blockage or spillover losses that are inherent in a parabolic aperture and thus provide the maximum possible gain for a given beamwidth. Blockage compensation devices or other auxiliary radiators are not needed to achieve the required performance. These devices, which under ideal conditions can enhance the performance of any antenna, are susceptible to performance degradation under varying thermal environments and under some circumstances will result in reduced gain and increased sidelobes. Their absence is a desirable feature.

The recommended cornucopia (horn-reflector) aperture is readily adaptable to two beams with small angular separation. For the beam separations called for, it is found that lower sidelobes can be obtained by using separate apertures; however, at reduced separations this would not be the case. The configuration becomes especially attractive at lower frequencies either as presently used or as a focal point feed for a parabolic reflector.

The recommended aperture feed is so located that it forms an integral part of the reflector structure. Its location also makes it possible to restrict temperature excursions to a minimum. These factors minimize thermal displacements of the feed relative to the reflector surface and consequently sustain RF pattern quality and limit thermal pointing errors. The feed and associated polarizer are designed without dielectric material of any kind and are inherently capable of handling high RF power levels.

The earth terminal recommended for use with both the prime and redundant direction-sensing systems can be implemented in one antenna with its associated feed, receiver, and transmitter. It may be located anywhere within the coverage area. Alternative systems utilizing two or more beacon locations are more costly in both equipment and real estate. Implementation of such alternate schemes can pose political problems if beacons are required in adjacent countries not in the area of coverage, or technical performance problems if bodies of water limit the placement of side terminals. In a system employing only one beacon, as recommended, the sensing antennas on the satellite may use narrower beamwidths and, thus, higher gains. Such a system also requires minimum signal processing on the satellite.

The direction-sensing antennas on the satellite are distinct from the communications apertures but closely connected to them structurally in order that relative displacements caused by thermal distortions of the intermediate structure be held to a minimum. This use of separate antennas reduces the coupling of the high power communication signal into the sensing channels and avoids degradation of the sidelobe level of the communication signal that would be incurred if the sensing feed were part of a common aperture. The separation also makes it possible to accommodate the interferometer sensitivity to linear thermal distortions and the communication aperture sensitivity to rotational distortion independently and thus permits a design with reduced thermal pointing error.

The steering system is implemented with digital components driving a brushless stepper motor. This system is advantageous from the standpoint of both reliability and power consumption. Power is dissipated only when there is pointing error.

The mechanical design has been evolved with special attention given to the manner of providing for relative motion of moving parts. Elastic deformation is used in place of rubbing or rolling contact where possible (flexures and flexguide). Where bearings and gears cannot be avoided, they are either hermetically sealed or lubricated with proven space lubrication techniques involving reservoirs, molecular seals, and controlled boil-off.

Thermal control has been achieved by the appropriate specification of thermal conductances and coatings. No active devices are required or used. Although thermal distortions are the major contributor to the total pointing error, the pointing error is within the specification. This is a consequence of a thermally insensitive design configuration, that still needs only conventional fabrication processes and standard, easily fabricated materials. Further improvement in pointing accuracy could be obtained at some increase in cost by utilizing more sophisticated materials and fabrication processes.

The recommended design comprises an interconnected set of components that have been carefully selected and designed to be optimum for high power narrow beam spacecraft antennas. These individual components are versatile in that they may be frequency-scaled or reconfigured to satisfy the antenna requirements of a variety of other high power space communication missions.

7.3 CONCLUSIONS

The most significant conclusion to be reached as a result of this study is that high power narrow beam satellite antennas are feasible and can be implemented with components of present day state-of-the-art technology. This conclusion is clear and unqualified; the antenna design recommended in this report has been shown to meet or exceed all of the design specifications without exception.

It is further concluded that immediate future effort can most effectively be directed toward experimental verification of the RF parameters that are not amenable to precise analytical calculations. The parameters of primary concern include side-lobes, pointing direction and high power breakdown. These parameters should be investigated over extremes of their space environment examining the effects of temperature, vacuum and contamination.

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