

COMBINED ACQUISITION AND TRACKING RADAR
A PARAMETRIC STUDY

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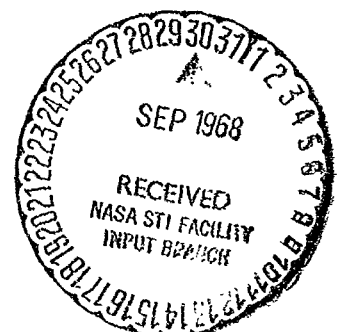
July 1968

Prepared under Contract No. NAS 9-7678 by
TEXAS INSTRUMENTS INCORPORATED
Equipment Group
13500 North Central Expressway
Dallas, Texas 75222

for

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
MANNED SPACECRAFT CENTER

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FOREWORD

SPONSORING ACTIVITY	National Aeronautics & Space Administration Manned Spacecraft Center Houston, Texas
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ABSTRACT

The objective of the "Combined Acquisition and Tracking Radar Parametric Study" is to obtain a complete analysis and investigation of the state-of-the-art optimum design parameters of a space-qualified electronically scanned phased-array radar system. The study is performed in an attempt to support design parameters for development of a phased-array radar system to be used as an experiment on board an Apollo spacecraft.

The study effort concerned investigation into radar subsystem areas which are most affected by the phased-array nature of the antenna. These areas included antenna, array components, signal processing, data processing, thermal control, and reliability.

Candidate system configurations include corporate-fed, active element arrays and space-fed, front- or rear-illuminated arrays. The frequency range of interest is 1.5 GHz to 20 GHz. The nominal tracking problem is to be able to acquire and track a 50 m² target at a range of 100 nm.

Early parts of the study result in a weight limitation on the radar of 350 pounds, a size limitation on the array of a six-foot diameter, and a prime power limitation of 1000 watts to the array. Under these limitations an optimization procedure is performed which maximizes the returned signal-to-noise ratio for each of six discrete frequencies chosen as candidates, and for the three types of arrays. The optimum system under the assumed limitations included a 292-element, 1.5-GHz corporate-fed, active-element array which yielded satisfactory acquisition and tracking performance at a range of 110 nm for a 50 m² target.

TABLE OF CONTENTS

Paragraph	Title	Page
1.0	INTRODUCTION	1-1
2.0	OPERATIONAL REQUIREMENTS	2-1
2.1	General Problem Statement	2-1
2.2	System Constraints	2-1
2.3	Geometrical Constraints	2-1
2.4	Required Operational Capability	2-3
3.0	SUBSYSTEM PARAMETRIC ANALYSES	3-1
3.1	Antenna Analysis	3-1
3.1.1	Introduction	3-1
3.1.1.1	Corporate-Fed Arrays	3-4
3.1.1.1.1	Single Rf-Power Source Array	3-4
3.1.1.1.2	Active-Element Array	3-5
3.1.1.2	Space-Fed Arrays	3-6
3.1.2	Corporate-Fed Arrays	3-7
3.1.2.1	Design Tradeoffs	3-8
3.1.2.2	Properties of Corporate-Fed Arrays	3-12
3.1.3	Space-Fed Arrays	3-20
3.1.3.1	Design Tradeoffs	3-20
3.1.3.1.1	Front Versus Rear Feeding	3-20
3.1.3.1.1.1	Front-Fed Array	3-20
3.1.3.1.1.2	Rear-Fed Array	3-22
3.1.3.1.2	Aperture Geometry and Illumination	3-23
3.1.3.1.3	Power-Handling Requirements	3-24
3.1.3.2	Properties of Space-Fed Arrays	3-28
3.1.3.2.1	Phase Correction and Quantization	3-28
3.1.3.2.2	Mutual Impedance Effects and Compensation	3-30
3.1.3.2.3	Thinning Considerations	3-31
3.1.3.2.4	Feedhorn Design	3-33
3.1.3.2.5	Module Configurations	3-36
3.1.4	Array Techniques	3-41
3.1.4.1	Antenna Element	3-41
3.1.4.2	Optimum Matching	3-42
3.1.4.3	Polarization Requirements and Implementation	3-45
3.1.4.3.1	Reflect-Array Polarization Diversity	3-46
3.1.4.3.2	Rear-Fed Element Polarization Diversity	3-50
3.1.4.4	Array Gain Versus Number of Elements	3-53
3.1.4.5	Gain Variation with Scan	3-53

TABLE OF CONTENTS (Continued)

Paragraph	Title	Page
3.1.4.6	Beamwidth Variation with Scan	3-55
3.1.4.7	Beam Pointing Error	3-55
3.1.5	Plasma Effects	3-58
3.1.5.1	Target Radar Cross-Section Change	3-58
3.1.5.2	Plasma Sheath Effects on Radar Performance	3-60
3.2	Array Components Analysis	3-60
3.2.1	General Array Discussion	3-60
3.2.2	Active Array Components Analysis	3-61
3.2.2.1	Introduction	3-61
3.2.2.2	Microwave Power Transistors	3-70
3.2.2.3	Microwave Diodes	3-78
3.2.2.4	Transmitter Chains	3-83
3.2.2.5	Phase-Shift Networks	3-88
3.2.2.6	Receiver Circuits	3-93
3.2.2.7	Rf Module Synthesis	3-99
3.2.2.8	Rf Module Parametric Results	3-105
3.2.2.9	Bulk and Transit Time Device Considerations	3-111
3.2.2.10	Rf Manifold	3-120
3.2.2.11	Signal Generator	3-135
3.2.3	Space-Fed Array Components Analysis	3-137
3.2.3.1	Introduction	3-137
3.2.3.2	Phase-Shifter Considerations	3-138
3.2.3.3	Waveguide Diode Phase Shifters	3-140
3.2.3.4	Diode Microwave I/C Phase Shifters	3-140
3.2.3.5	Diode Phase-Shift Modules	3-146
3.2.3.6	Ferrite Phase Shifters	3-147
3.2.3.7	Ferrite Phase-Shifter Modules	3-152
3.2.3.8	Transmitter Analysis	3-155
3.2.3.9	Space-Fed Array Analysis	3-162
3.3	Signal Processing Analysis	3-166
3.3.1	Range-Tracking Subsystems	3-166
3.3.1.1	Open-Loop Range Measurement	3-166
3.3.1.2	Split-Gate Range Tracker	3-169
3.3.1.3	Separated Gate Range Tracker	3-178
3.3.2	Angle-Tracking Subsystems	3-180
3.3.2.1	Monopulse	3-181
3.3.2.2	Conical Scan	3-185
3.3.2.3	Conopulse	3-191
3.3.2.3.1	System Description	3-191
3.3.2.3.2	Signal Processing	3-196

TABLE OF CONTENTS (Continued)

Paragraph	Title	Page
3.4	Data Processing Analysis	3-206
3.4.1	Multiple Mode, Multiple Target Capability	3-207
3.4.1.1	Multiple Mode Capability	3-207
3.4.1.2	Multiple Target Capability	3-209
3.4.2	Beam Steering and Stabilization Capability	3-210
3.4.2.1	Phase-Shift Equations for All-Electronic Stabilization and Scanning of Beam Axis	3-211
3.4.2.2	Scan Programmer	3-217
3.4.2.3	Phase-Shift Computer	3-217
3.4.3	Beam-Pattern Scan-Control Capability	3-221
3.4.3.1	Search Time	3-221
3.4.3.2	Angular Rate	3-225
3.4.3.3	Beam-Scan Mode Control	3-227
3.4.4	Data Accumulation	3-229
3.5	Thermal Analysis	3-229
4.0	SYSTEM PERFORMANCE ANALYSIS	4-1
4.1	Introduction	4-1
4.2	System Parametric Analysis	4-1
4.2.1	Target Detection Analysis	4-1
4.2.2	Target Tracking Analysis	4-10
5.0	VEHICLE INTERFACE	5-1
5.1	Types of Radar Data Displays	5-11
5.1.1	Galvanometers	5-11
5.1.2	Servo-Driven Tapes	5-11
5.1.3	Cathode-Ray Tubes	5-12
5.1.4	Light-Emitting Diode Arrays	5-12
5.1.5	Electroluminescent Displays	5-12
5.2	CAT Radar Display Requirements	5-13
6.0	MECHANICAL ANALYSIS AND DESIGN	6-1
7.0	RELIABILITY ANALYSIS	7-1
7.1	Reliability Indices	7-1
7.2	Calculation of Reliability	7-2
7.2.1	Constant Failure Rate Case	7-2
7.2.2	Case of Partial Redundancy	7-2
7.3	System Reliability	7-5
7.3.1	Conditions Assumed	7-5
7.3.2	Power Conditioner	7-5

TABLE OF CONTENTS (Continued)

Paragraph	Title	Page
7.3.3	Phased Array	7-8
7.3.4	Transmitter	7-9
7.3.5	Other Functional Blocks	7-10
7.3.6	Combined System Reliability	7-11
7.4	Environmental Factors	7-12
7.4.1	Temperature	7-12
7.4.2	Radiation	7-12
7.4.3	Shock, Vibration, and Thermal Cycling	7-15
7.4.4	On-Off Cycling	7-16
7.4.5	Meteoroids	7-17
7.4.6	Off-Duty Reliability	7-20
7.5	Reliability Summary	7-20
8.0	SYSTEM SYNTHESIS AND OPTIMIZATION	8-1
8.1	Introduction	8-1
8.2	System Synthesis	8-1
8.2.1	Array Synthesis	8-1
8.2.1.1	Corporate-Fed Array	8-1
8.2.1.2	Space-Fed Array	8-6
8.2.2	Radar System Synthesis	8-15
8.3	System Optimization and Specification	8-18
9.0	RECOMMENDATIONS FOR FURTHER STUDY	9-1
9.1	Higher Duty Cycle Radar Systems	9-1
9.2	Experimental Array Verification Measurements	9-2
9.3	Conopulse Radar Analysis	9-2
9.4	Latching Ferrite Thin-Film Phase Shifters	9-3
9.5	Additional Study with Different Restrictions	9-3
9.6	Surface Wave Device Development	9-3

APPENDIX

APPENDIX A	CHARACTERISTICS OF PHASED ARRAYS
APPENDIX B	MATCHED-FILTER THEORY AND PULSE COMPRESSION
APPENDIX C	EXPERIMENTAL PHASED-ARRAY ANALYSIS

REFERENCES

BIBLIOGRAPHY

LIST OF ILLUSTRATIONS

Figure	Title	Page
2-1	Closing Velocity for Two Circular Orbits	2-2
3-1	Classification of Electronically Scanned Phased-Array Antennas	3-3
3-2	Corporate-Fed Array	3-5
3-3	Front-Fed Array	3-6
3-4	Rear-Fed Array	3-7
3-5	Element Spacing Versus Maximum Scan Angle for Grating-Lobe Prevention in Phased Arrays	3-9
3-6	Rectangular Aperture Tradeoffs	3-10
3-7	Circular-Aperture Tradeoffs	3-11
3-8	Aperture Diameter Versus Total Number of Elements . .	3-13
3-9	Beamwidth Versus Gain for Corporate-Fed and Space- Fed Arrays	3-14
3-10	Number of Elements Versus Gain for Corporate-Fed and Space-Fed Arrays	3-15
3-11	Determination of Beam-Pointing Angle	3-17
3-12	Quantization Phase Error for Linear Delay Across Array	3-18
3-13	Phase-Shifter Quantization Effects	3-19
3-14	Front-Fed Illuminated Array	3-21
3-15	Rear-Fed Illuminated Array	3-22
3-16	Gain Versus Beamwidth	3-25
3-17	Number of Elements Versus Gain	3-26
3-18	Array Center Element Peak-Power Requirements	3-27
3-19a	Plane Phase Front Quantization	3-29
3-19b	Space-Fed Configuration	3-29
3-20	Transmission Loss Versus VSWR	3-32
3-21	Rear-Fed Array Module	3-33
3-22	Basic Feedhorn Types (Waveguide)	3-35
3-23	Feedhorn Considerations and Tradeoffs	3-35
3-24	F/D Versus a/λ for Optimum Monopulse Sensitivity . . .	3-36
3-25	Space-Fed Array Feed Design	3-37
3-26	Pencil-Beam Feedhorn Design Curves	3-38
3-27	10-dB Beamwidths for Open-End Waveguide and Horn Radiators (Average Flare Angle, 20 Degrees)	3-38
3-28	Module Block Diagrams	3-39
3-29	Module Mockup Interior View	3-39
3-30	Example of H-Plane Optimum Impedance Matching	3-44
3-31	30-Degree H-Plane Simulator	3-45
3-32	Polarization Requirements	3-46
3-33	Probe Orientation Versus Transmitted Polarization	3-47

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
3-34	Feed Polarization Circuits	3-48
3-35	Reflectarray Module Polarization Diversity	3-49
3-36	Added Insertion Loss for Polarizer	3-52
3-37	Circular Aperture	3-54
3-38	Gain Variation with Scan	3-56
3-39	Gain and Beamwidth Variation with Scan	3-57
3-40	Linear Array of Dipoles Scanned in the H-Plane	3-58
3-41	Beam-Pointing Error	3-59
3-42	Simplified Block Diagram of Space-Fed and Active- Element Arrays	3-62
3-43	MERA Radar System Block Diagram	3-64
3-44	Two Sides of the MERA Rf Module	3-67
3-45	Rf Module Block Diagram	3-68
3-46	L-49 Transistor Geometry	3-71
3-47	MERA Module Power Amplifier-Multiplier Chain	3-73
3-48	L-158A Geometry	3-74
3-49	L-195 Transistors	3-75
3-50	L-195 Pulsed Amplifier Performance	3-76
3-51	Pulsed L-195 Amplifier with Two Chips	3-77
3-52	Beam-Lead Surface-Oriented Switching Diode	3-79
3-53	Beam-Lead Planar Varactor Diode	3-80
3-54	Schottky Barrier Diode	3-81
3-55	Beam-Lead Surface-Oriented Diode Structure	3-82
3-56	Planar Beam-Lead Varactor Structure	3-84
3-57	Monolithic Structure for Series-Parallel Diode Combina- tions	3-85
3-58	Amplifier-Multiplier Chains for Rf Modules Showing Gain, Power and Efficiency	3-86
3-59	Simplified Block Diagram of 25-Watt L-Band Amplifier	3-88
3-60	L-Band Transistor Amplifier Pulsed Power and Gain Versus Frequency	3-89
3-61	Combination Loaded-Line/Line Length Switch	3-92
3-62	Maximum Available, Unneutralized, Common-Emitter Gain and Optimum Noise Figure	3-94
3-63	Predicted Noise Figure and Gain of L-196 Transistor	3-95
3-64	Receiver Performance for Rf Modules	3-96
3-65	10-110 MHz Video Amplifier	3-97
3-66	Image-Terminated Mixer	3-98

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
3-67	Image-Terminated Mixer If. Noise Figure with a 2.2-dB If. Thin-Film Amplifier	3-99
3-68	1.5-GHz Rf Module	3-100
3-69	2.25-GHz Rf Module	3-100
3-70	3.0-GHz Rf Module	3-101
3-71	6.0-GHz Rf Module	3-101
3-72	9.0-GHz Rf Module	3-102
3-73	18.0-GHz Rf Module	3-102
3-74	Module Input Power Versus Frequency for Fixed Duty Cycle and Module Peak Output Power Versus Frequency	3-109
3-75	Number of Modules Per 1 kW of Prime Power Versus Frequency for Fixed Duty Cycle	3-110
3-76	Peak Power Transmitted Per 1 kW Prime Power Versus Frequency for Fixed Duty Cycle	3-112
3-77	Average Power Transmitted Per 1 kW Prime Power Versus Frequency for Fixed Duty Cycle	3-113
3-78	Transmit Efficiency and Gain Versus Frequency for Fixed Duty Cycle	3-114
3-79	Module Efficiency Versus Frequency for Fixed Duty Cycle	3-115
3-80	Solid-State Device Power to August 1967	3-116
3-81	Schematic Diagram of Injection Phase Locking	3-118
3-82	9.0-GHz Rf Module Using Pulsed Oscillator	3-119
3-83	Rf Manifold Block Diagram	3-122
3-84	S-Band Distribution System	3-123
3-85	If. Distribution and Monopulse Comparator	3-124
3-86	Corporate Feed Structure	3-125
3-87	Rf Manifold Configuration	3-126
3-88	Rf Manifold Top View	3-127
3-89	Rf Manifold Bottom View	3-128
3-90	Rf Manifold Side View	3-129
3-91	Manifold S-Band Amplitude Distribution	3-130
3-92	Manifold S-Band Phase Distribution (1 Degree Increments)	3-131
3-93	Manifold 500-MHz Phase Distribution (1 Degree Increments)	3-132
3-94	Manifold Deviation from Required 500-MHz Amplitude Distribution	3-133

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
3-95	Signal Generator Block Diagram	3-136
3-96	Center Element Peak Power for Space-Fed Array	3-139
3-97	Diode Phase-Shifter Types	3-142
3-98	L-Band Phase-Shift Network	3-143
3-99	Four-Bit S-Band Phase Shifter	3-145
3-100	Diode Phase-Shifter Module Block Diagram	3-146
3-101	Reciprocal Latching Phase Shifter	3-151
3-102	Nonreciprocal Latching Phase Shifter	3-153
3-103	Drive Energy-Transmissive Ferrite Phaser in Microstrip	3-154
3-104	Insertion Loss-Transmissive Ferrite Phaser in Microstrip	3-155
3-105	Reciprocal, Reflective or Transmissive Ferrite Phase- Shifter Module Block Diagram	3-156
3-106	Nonreciprocal, Transmissive Ferrite Phase-Shifter Module Block Diagram	3-156
3-107	Hard Tube Modulator	3-160
3-108	Line Type Modulator Diagram	3-161
3-109	Magnetic Modulator Diagram	3-161
3-110	Single-Pulse Breakdown Power for Air-Filled Standard Waveguides	3-163
3-111	Transmitter and Receiver Circuit Description for Space- Fed Array	3-164
3-112	Invariant Portion of Signal-Processing System	3-167
3-113	Combined Detection and Range Measurement System	3-167
3-114	Split-Gate Range-Tracking System	3-170
3-115	Waveforms of Split-Gate Range-Tracking System	3-170
3-116	Simplified Block Diagram of Split-Gate Range-Tracking System	3-172
3-117	Separated-Gate Range-Tracking System	3-179
3-118	Waveforms of Separated Gate Range-Tracking System	3-179
3-119	Monopulse Rf Receive Circuitry Block Diagram	3-182
3-120	Monopulse Receiving and Processing Circuitry Angle Track Block Diagram	3-183
3-121	Sum-and-Difference Beam Patterns of a Monopulse Tracking System	3-184
3-122	Conical-Scan Radar Angle Track Block Diagram	3-186
3-123	Relationships Between Target, Scanning Beam, and Received Signal for Conical-Scan Tracker	3-187

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
3-124	Angular Error Slope and Power Loss for Conical-Scan Tracker	3-189
3-125	Conopulse Transmission and Reception System	3-192
3-126	Conopulse Scan Coordinate System	3-196
3-127	Power Spectral Density of If. Signal (No Modulation) . . .	3-198
3-128	Power Spectral Density of Detected If. Signal	3-198
3-129	Effective Power Spectral Density of Detected If. Signal .	3-199
3-130	Power Spectral Density of Output of Conopulse Oscillator	3-199
3-131	Power Spectral Density of Output of Phase-Sensitive Detector	3-199
3-132	Angular Error Slope and Power Loss for Conopulse Tracking System	3-204
3-133	Block Diagram for Near- and Far-Range Modes	3-208
3-134	Array Face Coordinates	3-212
3-135	Relation of x, y, z Coordinates to the X, Y, Z Coordinates .	3-212
3-136	Rotation of Axes	3-214
3-137	Definition of α and β	3-214
3-138	Scan Programmer	3-218
3-139	Phase-Shift Computer	3-219
3-140	Phase Control in Rf Module	3-220
3-141	Search Geometry	3-221
3-142	Allowable Search Time Versus Velocity	3-222
3-143	Range for Equal Radar Search Time and Target Time in Search Sector	3-224
3-144	Two-Dimensional Velocity Representation	3-225
3-145	Multibar Raster Scan	3-226
3-146	Search-Sector Reduction	3-226
3-147	Beam-Pattern Scan Controller	3-226
3-148	Basic Output Data Block Diagram	3-228
3-149	Presentation of Operating Envelope for Array Cooled from Surface	3-232
3-150	Estimate of Fluid-Loop Heat-Rejection Performance . . .	3-234
4-1	Probability of Detection (P_d) as a Function of Signal-to-Noise Ratio (S/N) and Probability of a False Alarm (P_n)	4-2
4-2	Search Sector and Search Scan Pattern for CAT Radar . .	4-4
4-3	Filter Mismatch Loss Relative to a Matched Filter	4-6
4-4	Post Detection Integration Loss Relative to Coherent Integration	4-7

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
4-5	Post Detection Integration Loss Relative to Coherent Integration	4-7
4-6	Collapsing Loss	4-8
4-7	Effective Integrated If. Signal-to-Noise Ratio for System Parameters Defined in Table IV-3 Eight Pulses Integrated.	4-16
4-8	Single Pulse If. Signal-to-Noise Ratio for System Parameters Defined in Table IV-3	4-18
4-9	Rms Thermal-Noise Error in Open-Loop Range Mea- surement for System Parameters Defined in Table IV-3 Pulses Integrated	4-19
4-10	Rms Thermal-Noise Error in Split-Gate Range Tracker for System Parameter Defined in Table IV-3	4-20
4-11	Rms Thermal-Noise Error in Separated Gate Tracker for System Parameters Defined in Table IV-3	4-21
4-12	Comparison of Thermal-Noise Response of Range Trackers for System Parameters Defined in Table IV-3	4-22
4-13	Comparison of Maximum Trackable Target Velocities for Range Trackers with System Parameters Defined in Table IV-3	4-23
4-14	Comparison of Lag Errors of Range Trackers for Sys- tem Parameters Defined in Table IV-3	4-24
4-15	Rms Thermal-Noise Error in Angle Tracking Loops for System Parameters Defined in Table IV-3	4-25
5-1	MERA-Type Corporate-Fed Array	5-2
5-2	Conformal Array	5-3
5-3	Electronically Scanned, Rear-Fed Array with Tube-Type Transmitter	5-4
5-4	TRC Basic Power Circuit	5-5
5-5	Efficiency and Regulation Time Ratio Control	5-7
5-6	Output Power (kW) at 28 Volts Dc	5-8
5-7	Battery Power Storage Capacity (Kilowatt-Hour)	5-9
5-8	Battery Power Storage Capacity (Kilowatt-Hour)	5-10
6-1	Six-Foot Octagon-Shaped Array Mounted on Service Module	6-2
6-2	1.5-GHz Module	6-4
6-3	2.25-GHz Module	6-4
6-4	3.0-GHz Module	6-5

LIST OF ILLUSTRATIONS (Continued)

Figure	Title	Page
6-5	6.0-GHz Module	6-5
6-6	9.0-GHz Module	6-6
6-7	18.0-GHz Module	6-6
6-8	Module Weight Analysis	6-7
6-9	Weight Estimate of 31.5-Ft ² Aperture (Approx. 6-Ft Octagon)	6-9
6-10	Space-Fed Module Weight Analysis	6-11
6-11	Weight Estimate for 31.5-Ft ² Transmissive Space-Fed Array	6-12
6-12	Space-Fed Array Weight Analysis	6-13
6-13	Weight as a Function of Frequency for a Space-Fed Array	6-14
6-14	Weight as a Function of Frequency for an Active-Element Array	6-15
6-15	1-kW System Weight Power and Volume Estimate	6-16
7-1	Basic System Functional Block Diagram	7-6
7-2	Power Conditioner	7-6
7-3	Array Survival Probability Versus Time for J Allowed Failures	7-7
7-4	Combined System Reliability	7-11
7-5	Average Component Failure Rate	7-13
7-6	Module Failure Rate Due to Meteoroids	7-19
8-1	Array Limitation Dependence on Pulselength	8-4
8-2	Optimization Flow Chart	8-7
8-3	Range Dependence on Σ , α , and L	8-8
8-4	Optimum Corporate-Fed System Performance	8-9
8-5	Average Tx Power-Gain ² Product Versus Number of Modules for Space-Fed Array	8-13
8-6	Peak-Power and Pulselength Tradeoff for Space-Fed Array	8-14
8-7	CAT Corporate-Fed Radar System	8-19

LIST OF TABLES

Table	Title	Page
III-1	Phased-Array Tradeoffs	3-2
III-2	Classification of Phase Shifters	3-16
III-3	Thinning Considerations	3-33
III-4	Element Grouping Considerations	3-40
III-5	Radiating Element Type	3-42
III-6	Maximum Scan Angles of Waveguide Array	3-44
III-7	Possible Polarization Configurations	3-51
III-8	Power Amplifier Output Transistor Parameters	3-72
III-9	MERA Power Amplifier Stages	3-72
III-10	Varactor Diode and Step Elastance Diode Characteristics	3-83
III-11	Power Requirements for Diode Phase-Shift Networks	3-93
III-12	Rf-Module Parameter Results	3-108
III-13	Phase-Shifter Characteristics	3-140
III-14	Modules for Space-Fed Arrays with Diode Microstrip Phase Shifters	3-147
III-15	Examples of Ferrite Phaser Designs	3-149
III-16	Transmitter Tubes for Space-Fed Array	3-158
III-17	Solid-State Magnetic Modulator	3-162
III-18	Transmitter/Receiver Contributions for Space-Fed Array	3-165
IV-1	Summary of Performance Equation of Range-Tracking Systems	4-11
IV-2	Summary of Performance Equations of Angle-Tracking Systems	4-13
IV-3	CAT Radar Parameters	4-15
VII-1	Antenna Module Failure Rates	7-8
VII-2	Solid-State/Vacuum-Tube Power-Chain Comparison	7-11
VII-3	Functional Block Failure Rates	7-11
VII-4	Effects of Excessive Radiation	7-14
VIII-1	Solid-State Module Parametric Dependence on Frequency	8-2
VIII-2	Tabulation of Optimum Σ for Each Frequency	8-10
VIII-3	Space-Fed Array Parameters	8-16
VIII-4	Tabulation of Optimum Σ for Each Frequency for the Reflective and Transmissive Arrays	8-17
VIII-5	CAT Radar System Specifications	8-20

1.0 INTRODUCTION

This report constitutes the end result of a study, "Combined Acquisition and Tracking Radar: A Parametric Study," performed under Contract No. NAS 9-7678 and covering the period between 14 March 1968 and 14 June 1968. The study was performed for the Manned Spacecraft Center of the National Aeronautics and Space Administration.

The Combined Acquisition and Tracking (CAT) Radar is expected to be the first noncooperative radar which NASA will develop for spaceborne application. This study was conducted in order to support design parameters for the development of a phased-array radar system to be used experimentally on board an Apollo spacecraft. The primary objective of the study is to obtain a complete analysis and investigation of the "state-of-the-art," optimum design parameters of a space qualified, electronically scanned, phased-array radar system.

The study effort consisted of four phases. Phase One consisted of defining the operational requirements of the CAT Radar System. This was done so that boundaries could be placed around the problem in order that a significant contribution could be achieved in the relatively short period of time allocated for completion of the study.

Phase Two involved not only the compilation of data concerning the subsystem areas within a phased array radar but also the analysis of this data. The objective of Phase Two is a complete space qualified radar.

Phase Three consisted of the synthesis of the best possible corporate-fed array radar and the best possible space-fed array radar systems. This was done under the conditions of an upper limit on size, weight, and power consumption.

Phase Four saw the comparison of corporate-fed and space-fed array radar. These two systems were compared for recommendation as to the preferred design.

The study was conducted under this general guideline: the most important output of the study would be the parametric trade relationships that were uncovered so that the study results would still be valid and useful even though various limitations which were used in deriving a preferred system design would change in the future. Of secondary importance is the preferred system description. This is strictly valid only under identical assumptions concerning size, weight, and power consumption limitations.

A useful algorithm is included which, for conditions similar to those described in this study, permits one to optimize many parameters of a phased-array radar under his own conditions of size, weight, and power.

The importance attached to the parametric study portion of the effort accounts for the relative weighting of later sections, having more pages devoted to the studies than to the system synthesis itself.

A brief word is in order at this point concerning the general area of acquisition and tracking radars for noncooperative targets. NASA space-borne radars in the past have been cooperative, meaning that the target houses a transponder which detects the transmitted pulse, improves the signal-to-noise ratio, shifts modulation to a different frequency, and transmits a pulse back to the master radar which computes range, etc., by standard methods. This scheme has many advantages over noncooperative radar systems. This accounts for the historical fact that cooperative space radars have been developed and noncooperative space radars have not. However, all these advantages are worthless if the objective is to detect and track targets which have no working transponders on board. The need to track new kinds of targets initiated the proposed development of the CAT radar. Possible applications for the CAT radar are discussed in the next section.

The CAT radar is expected to have a search capability, a tracking capability, and a capability which permits tracking one target on a time-shared basis while continuing a search pattern. The study is largely concerned with possible ways of using phased arrays to detect targets and derive range and angle error signals. This is necessary in closing the loop so that the radar system can track a target.

2.0 OPERATIONAL REQUIREMENTS

2.1 General Problem Statement

The CAT radar is intended to be the first NASA radar to function in a space environment with noncooperative targets. Possible applications of CAT radar include:

- Rendezvous with space objects, including vehicles in distress, inactive satellites, space stations, unknown vehicles, and experimental probes

- Compile a catalog of radar signatures from test targets which are ejected from Apollo vehicles

- Locate and record data concerning asteroids.

The purpose of CAT radar is to acquire a target with a nominal cross-section of 50 met² at ranges up to 200 nm and track this target as long as it remains within the field of view. Still another purpose of the CAT radar study is to determine the requirements on various radar subsystems which allow the radar system to meet this objective.

2.2 System Constraints

The permissible type of radar system is constrained to be an electronically scanned phased-array radar. Possible array implementations which will be considered include:

- Corporate-fed array—single power source

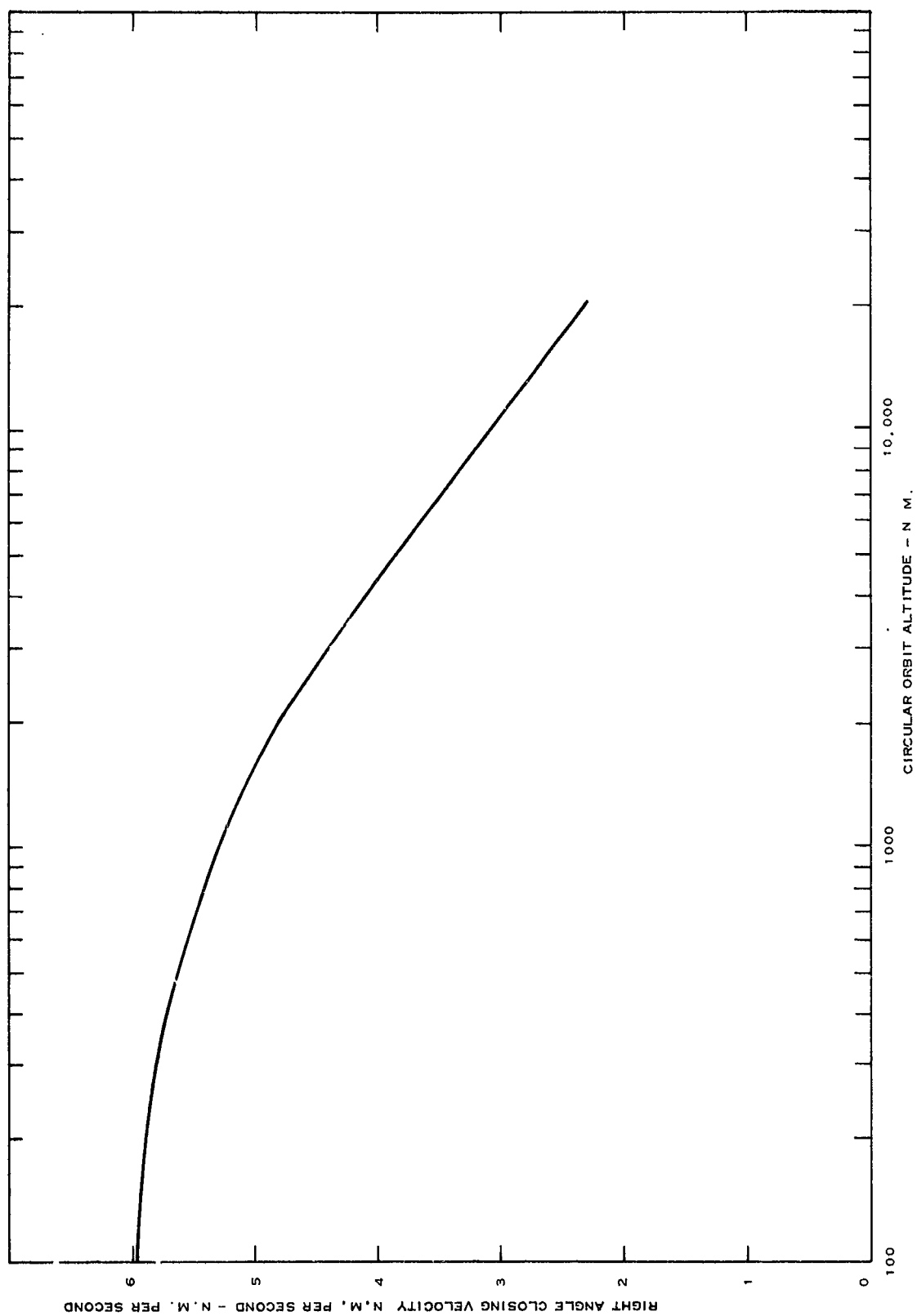
- Corporate-fed array—elemental power sources

- Space-fed array—rear illumination (lens)

- Space-fed array—front illumination (reflective).

2.3 Geometrical Constraints

In considering the range of velocities which are possible between two objects in near earth orbits, the problem is essentially reduced to one of matter of angle between the two velocity vectors. As a worst case condition for study purposes, the spacecraft carrying the CAT radar is assumed to encounter the approaching target at a right angle with respect to the spacecraft flight path. Assuming that both the spacecraft and the target are in circular earth orbits of the same radius, the relative velocity between the two objects is 1.414 times that of each object. Because orbital velocity is greatest at lower altitudes, the worst case will be assumed to occur at an altitude of 100 nm. At this altitude the circular orbit velocity is 4.2 nm per second and the right angle approach velocity is about 6 nm per second. Figure 2-1 shows how this right angle closing velocity depends on altitude for two objects in circular orbits.



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Figure 2-1. Closing Velocity for Two Circular Orbits

Another type of rendezvous situation occurs when both the spacecraft carrying CAT radar and the target are in coplanar orbits. This situation requires use of the CAT radar for terminal guidance information necessary for a successful rendezvous. Means other than CAT radar have been used to establish the coplanar orbits much like the Gemini and Agena target orbits. The problem is to accommodate the residual error velocity between the two space objects with the CAT radar tracker.

For purposes of this study, the maximum velocity capability of the Gemini radar will be the goal for the rendezvous mode of the CAT radar. This maximum velocity capability can be seen in a picture of the Gemini radar display on page 103 of the 11 October 1965 issue of Product Engineering magazine and is 500 feet per second.

2.4 Required Operational Capability

This section will include a summary of the radar parameters which have been assumed necessary in attaining the desired operational capability.

Acquisition—The CAT radar shall be able to acquire a non-cooperative target of 50 met² cross section at ranges between 5 and 200 nm

Tracking—The CAT radar shall be able to track a noncooperative target of 50 met² cross section at ranges between 5 and 200 nm

Polarization Diversity—On the basis of target detection and tracking, it is difficult to justify a requirement for polarization diversity since linear polarization performs adequately in all cases except the one where a pencil-like target is oriented perpendicular to the polarization. Since this is an unlikely event for space targets, it is believed that a firm requirement exists only for linear polarization. In the interest of completeness, it is desirable that the CAT radar be capable of polarization diversity for achievement of more effective use of the space experiment environment and time. A goal for the study is to define a series of possible radar system implementations. The implementations range from one linear polarization only at the bottom to the one at the top which permits transmitting either sense circular polarization, while having separate processing channels for each of the two linear polarizations received

Range Interval and Accuracy—The CAT radar shall have an effective range interval of 5 to 200 nm. On this interval the desired 3σ range accuracy is 0.25 percent

Range Rate Accuracy—The desired 3σ range rate accuracy is 0.25 percent or 1 fps, whichever is greater

Field of View—The field of view shall, at a minimum, encompass ± 45 degrees in elevation, and ± 45 degrees in azimuth

Angular Locational Accuracy—The target azimuth and elevation angles shall be resolved within 0.5 degrees

Detection and False Alarm Probabilities—The desired detection and false alarm probabilities are 0.9 and 10^{-6} respectively

Search Time—The CAT radar shall be able to search the entire field of view and acquire a target within 10 seconds

Growth Potential—The CAT radar system design shall be of a type which readily incorporates advances in the state-of-the-art for an improved system. Also, the radar system performance shall be expected to improve correspondingly as more resources become available like prime power, space, weight, cooling capacity, and money.

Interface—It is expected that the spacecraft housing the CAT radar will not only be occupied and controlled by astronauts, but also will include a guidance computer capable of high-speed data exchange with the radar system. When necessary, spacecraft orientation data is supplied to the radar and target data is supplied to the guidance computer/astronaut in detection and tracking situations.

3.0 SUBSYSTEM PARAMETRIC ANALYSIS

In order to synthesize a radar system which is capable of meeting the operational requirements discussed in Section 2.0, data must be gathered and analytical studies performed within each of the subsystem areas which constitute a radar system. These trade studies are the most important part of this study report. They represent the compilation and analysis of much data concerning the applicability of electronically scanned phased-array radars to the problem of detecting targets in the space environment. Given the physical constraints of a spacecraft, this problem is a formidable one indeed.

The subsystem areas covered in this section include antenna, array components, signal processing, data processing, and thermal control. In the next section, following the analytical results of this section, there will be a system performance analysis. In this analysis, the detection and tracking capability of the radar system is derived, based on the performance potential of each subsystem as discussed here.

3.1 Antenna Analysis

Because of the additional costs of phased arrays, careful array analysis should be planned in the design of the systems. Each array should have optimum capability for its particular mission to achieve maximum cost effectiveness. This section will cover many of the pertinent antenna factors to be considered in the design of phased arrays for spacecraft radar. Trade-off data will be presented to compare space-fed arrays with corporate-fed arrays. Design tradeoffs peculiar to each array type will be presented as well as design criteria common to both array types. Table III-1 illustrates some of the tradeoffs between the array types.

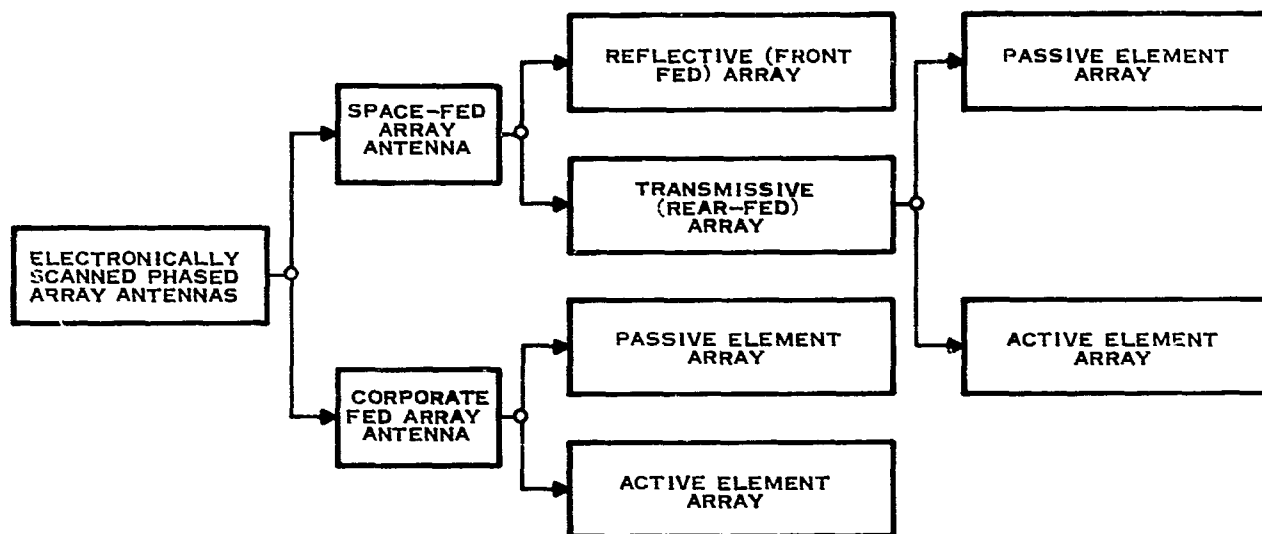
The first paragraph, which will give an introduction to the basic array types, will be followed by a paragraph on corporate-fed arrays. Next will follow a paragraph presenting design tradeoffs and properties of space-fed arrays. This will be followed by a discussion of various array techniques common to both array types. The antenna analysis section will be concluded with a paragraph on plasma effects. Appendix A will present detail information on phased-array parameters, such as mutual coupling and radiating element design.

3.1.1 Introduction

Phased arrays often are classified by the manner in which they are fed as illustrated in Figure 3-1. In general, arrays are space-fed or corporate-fed.

Table III-1. Phased-Array Tradeoffs

SPACE-FED (TRANSMISSIVE)	SPACE-FED (REFLECTIVE)	CORPORATE-FED (ACTIVE ELEMENT)
Capable of high rf power	Capable of high rf power	All solid-state
Good sidelobe performance	Aperture blockage causes higher sidelobes	Adapts to any aperture shape
Requires reciprocal phase shifters or phase switching between transmit and receive	Minimum weight	Highest aperture efficiency
Multiple Feedhorns can offer varied aperture distributions	Good dc to rf efficiency	Permits implementation of compact system
High-power phase shifters require significantly more dc power	Array complexity is reduced making this system less expensive	Highest reliability
	Overall system efficiency is best at higher frequencies	No high rf peak power to control
		Overall system efficiency is best at lower frequencies



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Figure 3-1. Classification of Electronically Scanned Phased-Array Antennas

A space-fed (or illuminated) array consists of radiating elements which are fed optically from a source outside of the array (otherwise, in ways much like the usual parabolic reflector). The two types of space-fed arrays are the front-fed array and rear-fed array. The front-fed array consists of a feed device, normally a horn radiator, which illuminates the elements of the array from the front. This energy is received by the individual array elements, given the proper elemental phase shift, and reradiated through the same set of array elements. The horn used to illuminate this type of array will present some aperture blockage to the array, the degree of blockage depending on the size of the array.

The rear-fed array employs a transmissive system similar to a dielectric lens. The array is, again, externally illuminated by a feed (normally a horn), but from the rear. The energy is received by antenna elements on the rear of the array, transmitted through phase shifters, and then reradiated from the front side by a second set of antenna elements.

A corporate-fed array consists of an array of radiating elements which are fed through a tree-like power-distribution network. The two types of

corporate-fed arrays are (1) the single-source array where the rf power is centrally generated and then divided and fed to the elements through the distribution tree, and (2) an active element array in which the power amplification is achieved in the modules.

The single-source array uses a single rf source and a passive power-distribution network. The single rf-source method has many of the major disadvantages of the other methods and few of the major advantages of either for most space applications.

The active element array is similar in general concept to any other type of corporate-fed array. Its unique feature is its modular molecular electronic construction and the generation of rf power in the individual modules. The basic technology is direct outgrowth of the Molecular Electronics for Radar Applications (MERA) program Contract AF 33(615)-1993.

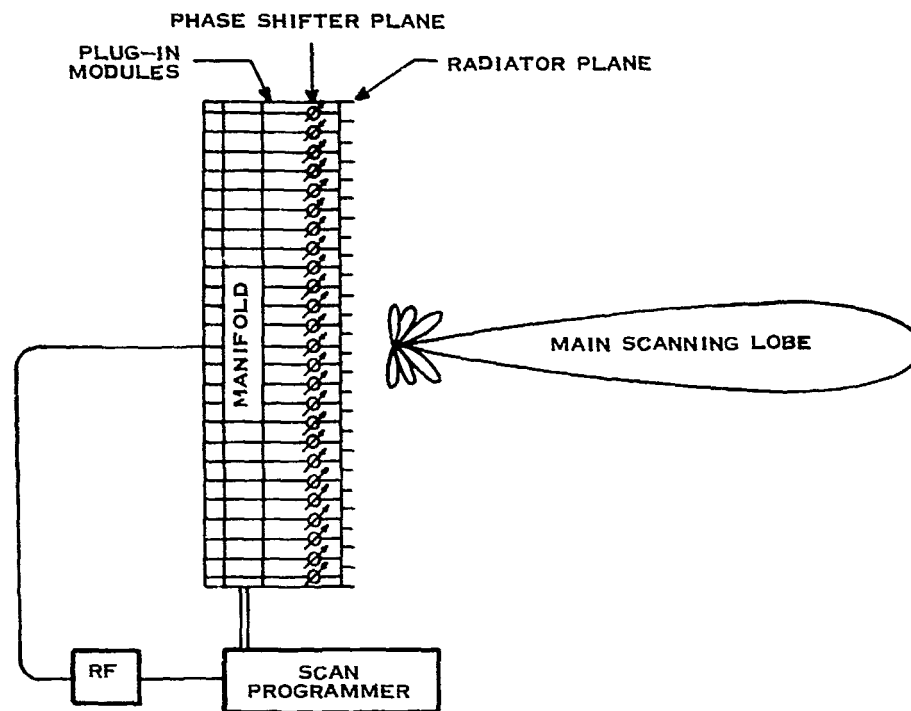
3.1.1.1 Corporate-Fed Arrays

The corporate-fed array may consist of (1) a single high-power rf source feeding an array of passive elements, or (2) a low-power source feeding an array of active elements.

3.1.1.1.1 Single Rf-Power Source Array

The single high-power, rf source, corporate-fed array has the same advantage in dc-to-rf conversion efficiency as does the illuminated array when compared to a system utilizing active elements. However, this type of array differs from the space-fed system in that it employs a passive feed structure to achieve power division rather than by optical illumination. Figure 3-2 illustrates the corporate-fed array. With most radar systems, there is a relatively high rf peak power distribution to the phased-array elements, and the corporate feed structure is necessarily heavy in order to prevent excessive losses in the high-power rf distribution system.

The phase shifters are located in the array element modules, and a dc manifold is integrated with the rf corporate feed network to provide phase shifting commands to the phase-shift driver circuits. A characteristic of this type of corporate-fed array is that the phase shifters must operate in a relatively high-power environment as compared to the active-element array discussed in the next paragraph. In general, this type of array is considered to be less satisfactory than the others described. The primary disadvantages are weight and complexity introduced by the high-power rf distribution system, eliminating it as a candidate for the space radar application.



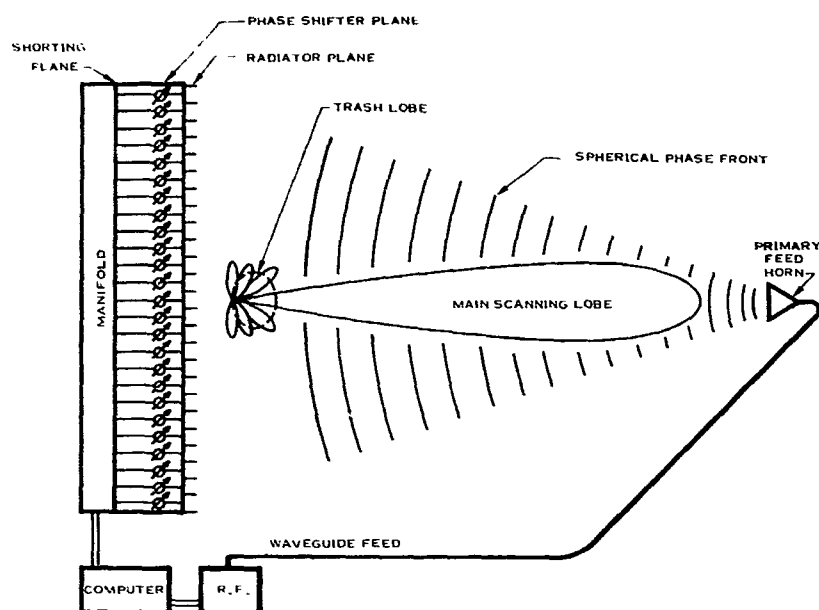
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Figure 3-2. Corporate-Fed Array

3.1.1.1.2 Active-Element Array

The corporate-fed phased-array system and the space-fed system, previously discussed, utilize high-power tube-type transmitters feeding arrays of passive elements. However, recent advances in microwave integrated circuits permit the mechanization of a solid-state system, replacing the high-power tube transmitter with a low-power solid-state unit. Power amplification and phase shifting are then achieved in microwave integrated circuits located in the individual array elements. These elements (or modules) may also include receiver circuitry that provide outputs at intermediate frequencies. The individual modules are interconnected through a distribution manifold which provides low-level rf, if., dc and phase-control distribution for the system.

The most significant advantage of this active-element approach is the inherent reliability that it implies. This type of system is also more compatible with space and weight restrictions encountered in space applications. In addition, the radiation efficiency of a corporate-fed array is greater than the space-fed array because there is no edge spillover from a primary feed-horn such as exists with the latter system.



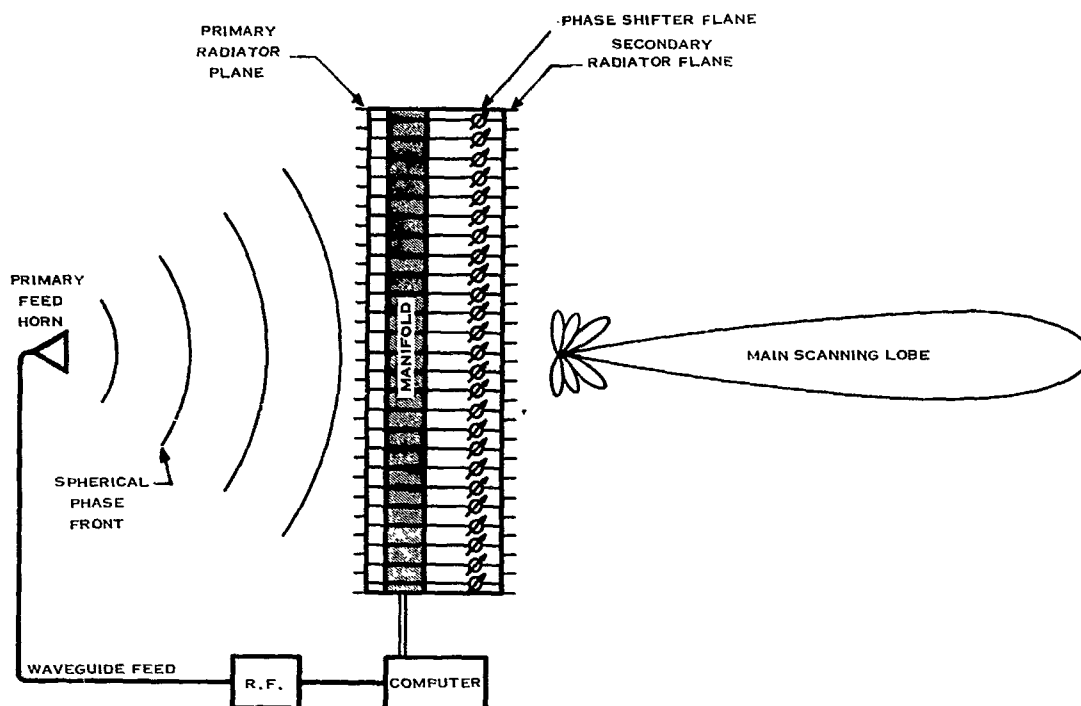
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Figure 3-3. Front-Fed Array

3.1.1.2 Space-Fed Arrays

A space-fed array may either be front-fed as in Figure 3-3 or rear-fed as in Figure 3-4. As shown, the illuminated array consists of one or more primary feedhorns, a number of antenna elements with associated phase shifters, and a manifold which connects the phase shifters of each element to a central scan computer. Choice of feed method usually depends on system requirements such as antenna volume, sidelobe level, and weight.

Space-fed or optically-fed arrays can provide weight-saving advantages over some corporate-fed phased arrays. Instead of using a strip-line or waveguide manifold feed system to distribute the power to the radiating array elements, the illuminated (or space-fed) array makes use of optical methods in feeding an array composed of many antenna elements. As with standard parabolic reflector antennas, the optically fed array is illuminated by a primary feedhorn. The amplitude distribution imposed on the array depends on the inverse square law for energy division. [The amplitude distribution usually will follow some form of $\cos^n(\theta)$ -on-a-pedestal.] The aperture phase distribution from the primary feedhorn will take the form of a spherical



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Figure 3-4. Rear-Fed Array

wavefront impinging on a plane surface. However, this spherical phasefront can be corrected to a plane phasefront by appropriately adjusting the phase-shifter setting of each individual element. Some phase error will be experienced with digital phase shifters, but the error due to the imperfect correction of the spherical phase may be used to advantage to reduce quantization side-lobes as discussed later in this section. The tradeoff parameters such as gain, beamwidth and power handling requirements will be discussed in Subsection 3.1.3.1.2.

3.1.2 Corporate-Fed Arrays

A corporate-fed array is quite attractive for space applications because of its improved aperture efficiency and compact packaging. These features imply increased antenna gain for a given package size. There are several design tradeoffs to consider in a corporate-fed array, and these tradeoffs are generally straightforward antenna aperture theory. This section will discuss these aperture parameters and will also present some of the properties common to the corporate-fed array. These properties include phase errors and quantization, mutual impedance, thinning and module configurations.

3.1.2.1 Design Tradeoffs

The tradeoffs associated with the corporate-fed aperture differ from the space-fed aperture in that there are no spillover losses for the corporate-fed system. The corporate-fed array can be uniformly illuminated by distributing equal power to all elements, whereas the space-fed array illumination depends on the feed horn pattern and location which implies an amplitude taper on the array.

A number of design curves are given in the following pages which relate the aperture size, beamwidth, gain, sidelobe level and number of array elements.

The element spacing in a phased array is a function of the maximum required scan angle and, assuming uniform rectangular spacing, is given by the following equation.

$$\frac{d}{\lambda} = \frac{1 - \frac{1}{n}}{1 + \sin \phi}, \quad (3-1)$$

where

d = spacing between elements

λ = wavelength

n = number of elements along the scan plane

ϕ = maximum scan angle.

Grating lobes will form in the principal scan planes if the array is scanned beyond the maximum scan angle given above. However, because of the severe impedance changes experienced at scan angles near grating-lobe formation in visible space, it may be very desirable to make the spacing even smaller than indicated in Equation (3-1). As illustrated in Figure 3-5, an element-to-element spacing of one-half wavelength will give good coverage for large arrays out to a 60-degree scan angle.

The array beamwidth is determined by the aperture size, shape and amplitude distribution. Figure 3-6 shows the beamwidth versus rectangular aperture width for various amplitude distributions. The same quantities for a circular aperture are shown in Figure 3-7. Figure 3-7 shows an example for a 10λ diameter aperture illustrating the range of beamwidths available from the different amplitude distributions. The sidelobe level for the uniformly illuminated circular aperture is lower than a rectangular aperture. This is because of the apparent amplitude taper of the circular aperture achieved by removing the corners of the rectangular aperture. The theoretical sidelobe

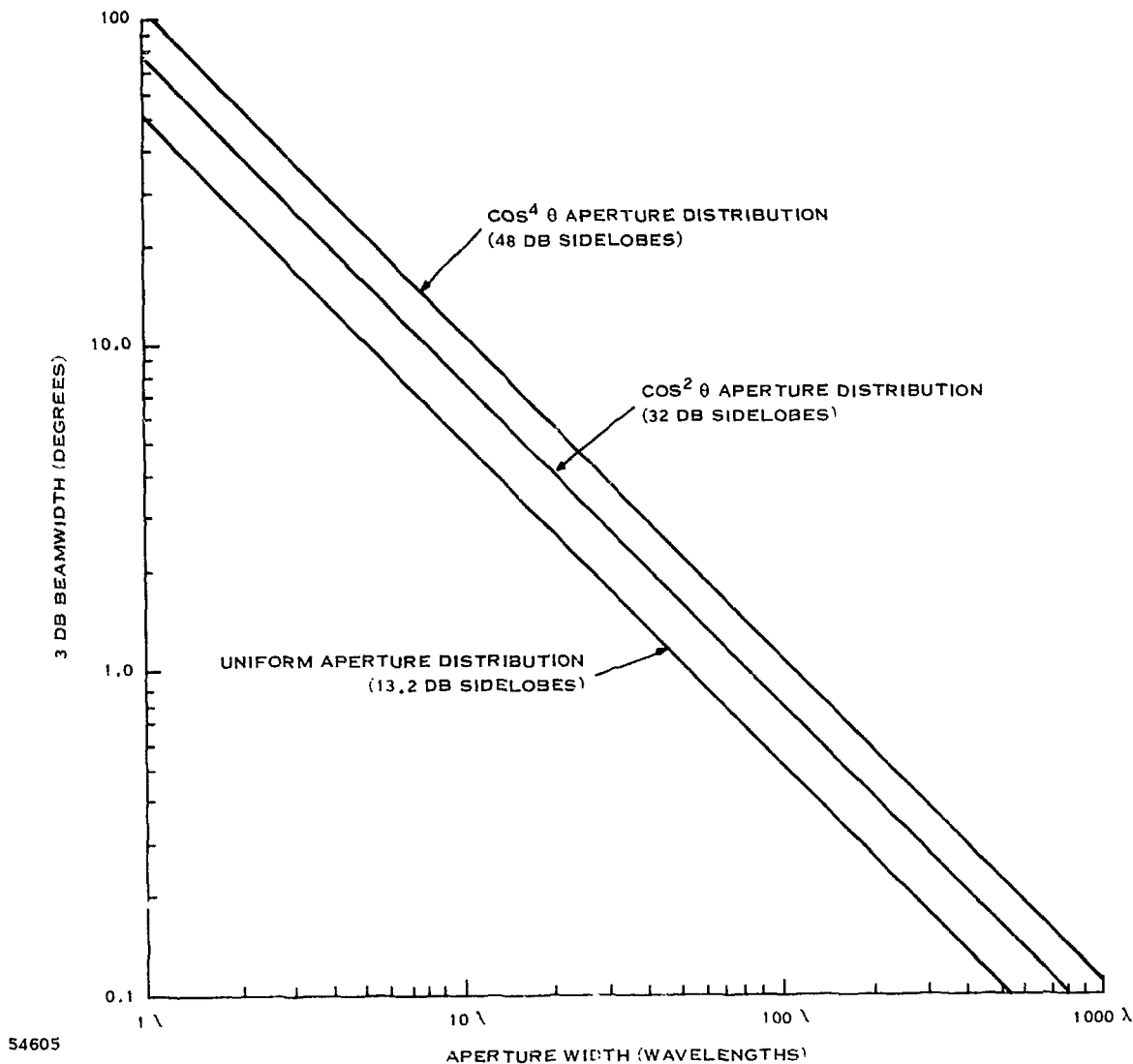
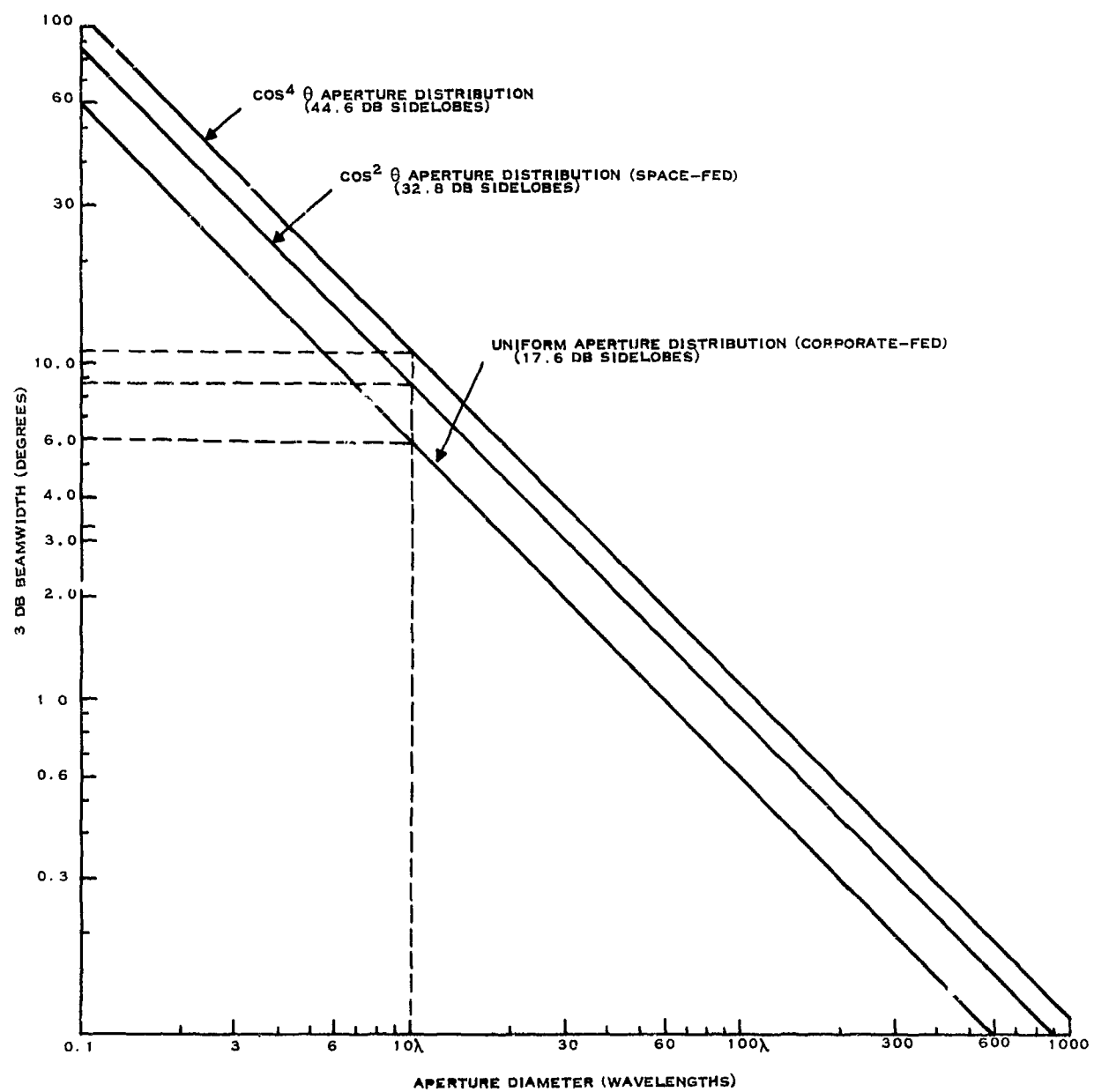


Figure 3-6. Rectangular-Aperture Tradeoffs

levels shown are not usually obtained in practice because of the effect of errors. In phased arrays such things as phase quantization, phase and amplitude errors have a large effect on the final sidelobe level. In which case, the array is usually designed with a more severe taper than the ideal theoretical taper in order to ensure that the desired sidelobe level is achieved. Errors are far more important when sidelobes below 20 or 25 dB are desired and careful consideration must be given at each step in the design. Probably the most efficient method to use in the determination of expected sidelobe levels is to program the array on a digital computer.



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Figure 3-7. Circular-Aperture Tradeoffs

To achieve improved sidelobe control in an active element corporate-fed array, the amplitude is controlled on both transmit and receive. The transmit distribution is generally uniform because maximum power is transmitted via amplifiers operating in saturation. Therefore most of the sidelobe control must be achieved by both aperture shape and tapering on the receive mode. The receive if. manifold is heavily amplitude tapered on the receive mode. The receive if. manifold is heavily amplitude tapered to give lower "one-way" sidelobe levels. Thus sidelobe control differs in the corporate-fed array from the space-fed in that the former requires attenuation in the modules or amplitude tapering in the manifold, whereas, the space-fed system amplitude taper is controlled with the feedhorn.

The amplitude tapering for the corporate-fed array may also be achieved with element thinning or space tapering.^{1, 2} This technique involves the removal of elements in such a fashion as to approximate an amplitude taper. However, space tapering is not generally performed on a solid-state active array, usually because the maximum aperture size is critical and maximum radiated power is desired from the array. Hence, the maximum number of array modules are generally used to completely fill the aperture.

Figure 3-8 shows the number of array elements for a given aperture diameter as a function of square element spacing. This illustrates also the resulting beamwidth, if uniform amplitude illumination is used. Triangular element spacing, discussed in greater detail in Appendix A, is useful if conical scan volumes are used. In this case, the triangular element spacing reduces the number of elements by 13 percent when compared to square spacing.³ A disadvantage of triangular spacing is that the number of columns (or rows) in the array doubles, thus changing the scan computer requirements. Square element spacing is generally easier to fabricate, therefore, several of the following curves will assume square spacing.

The boresight gain of the array is related to the beamwidth and aperture efficiency. Off-boresight gain is usually a function of the scan angle and will be discussed in greater detail in paragraph 3.1.4. Assuming a circular aperture, the beamwidths will be symmetrical and the gain may be related as shown in Figure 3-9. The aperture efficiencies shown are approximate values and are a function of the particular aperture configuration. Figure 3-10 relates the gain to the number of array elements assuming a $\lambda/2$ square grid.

3.1.2.2 Properties of Corporate-Fed Arrays

Many of the properties of corporate-fed arrays are similar to those of space-fed arrays and will be discussed in general in both paragraph 3.1.4 and Appendix A. Lengthy material, such as mutual coupling, will be

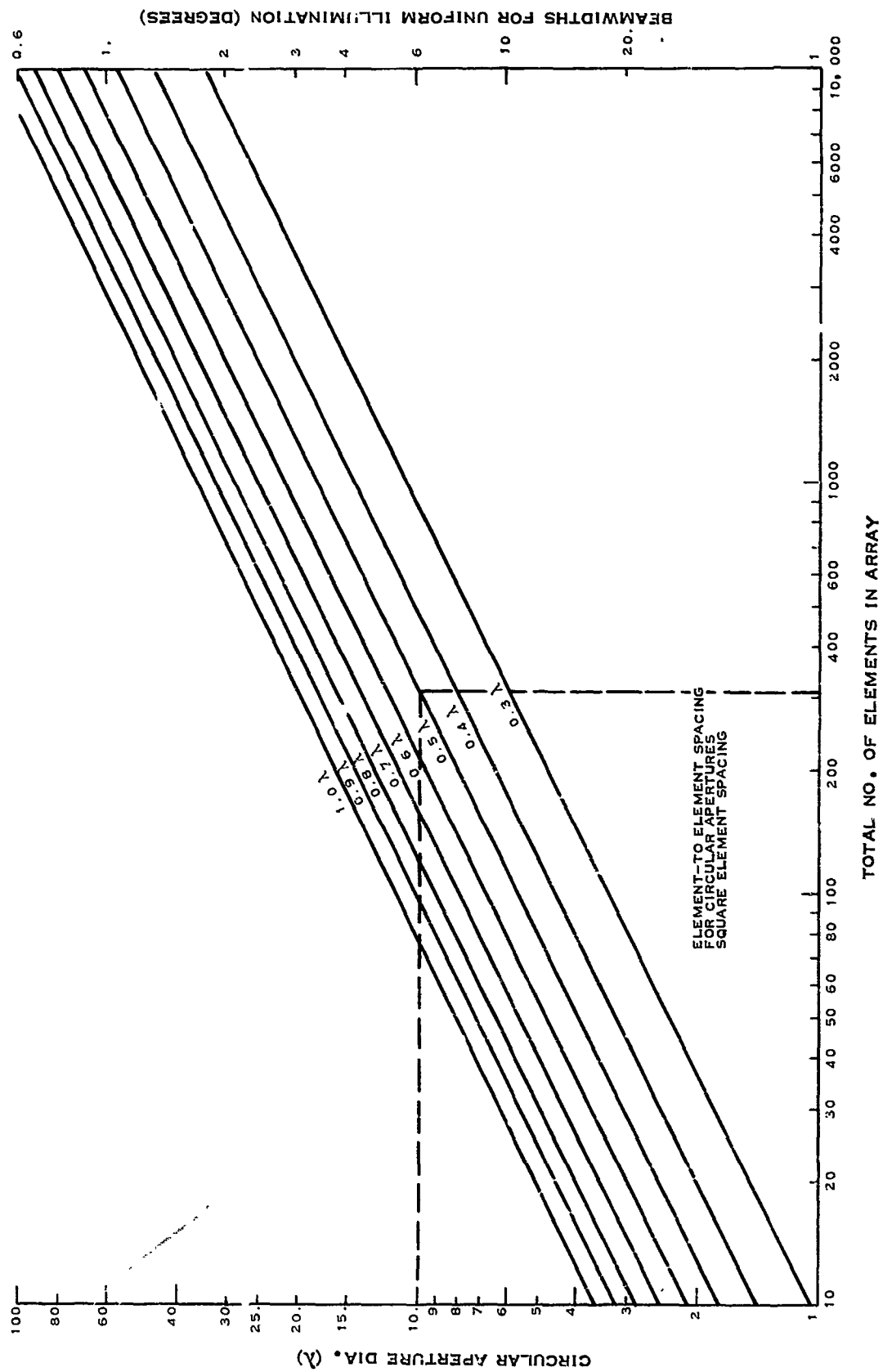


Figure 3-8. Aperture Diameter Versus Total Number of Elements

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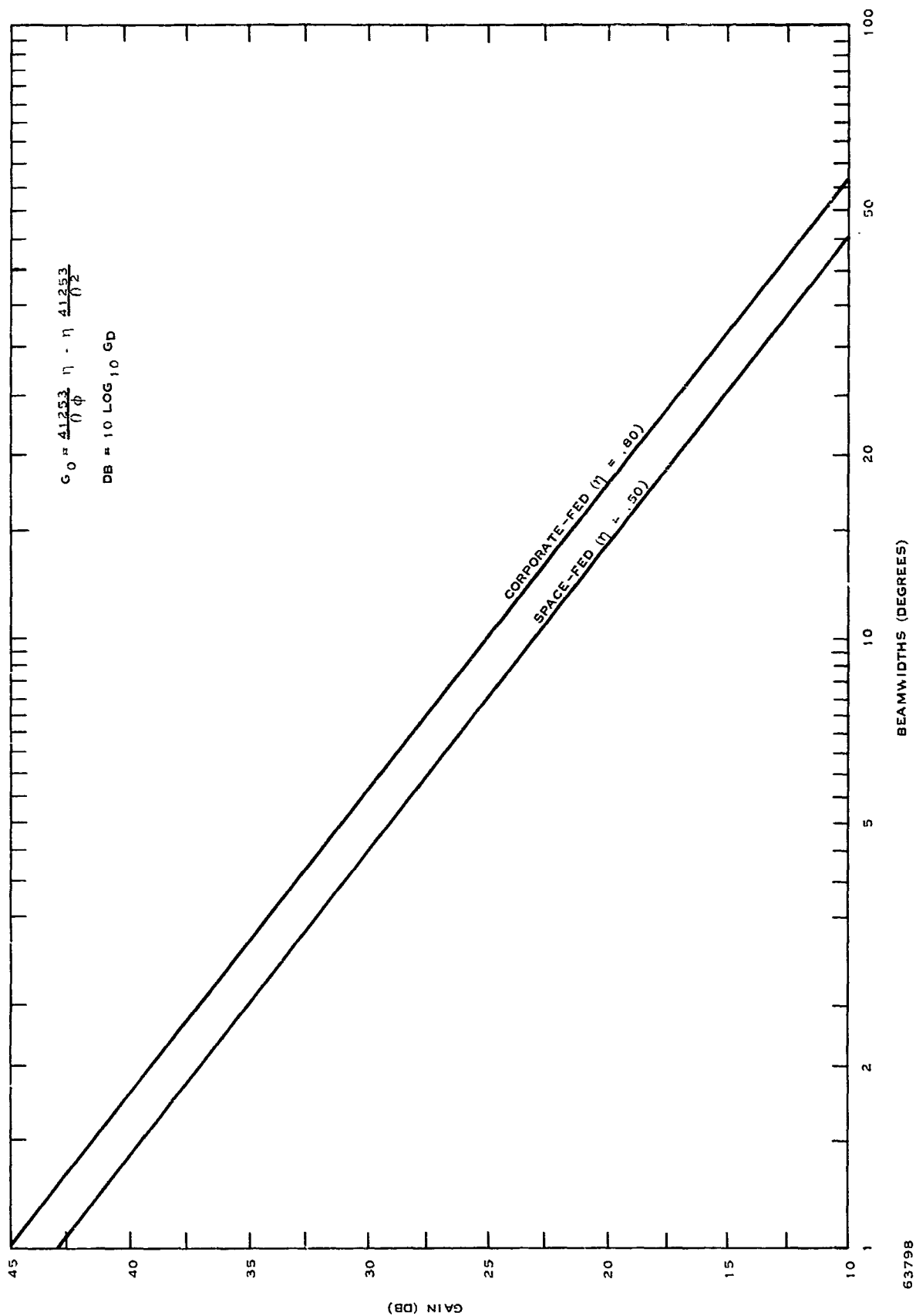
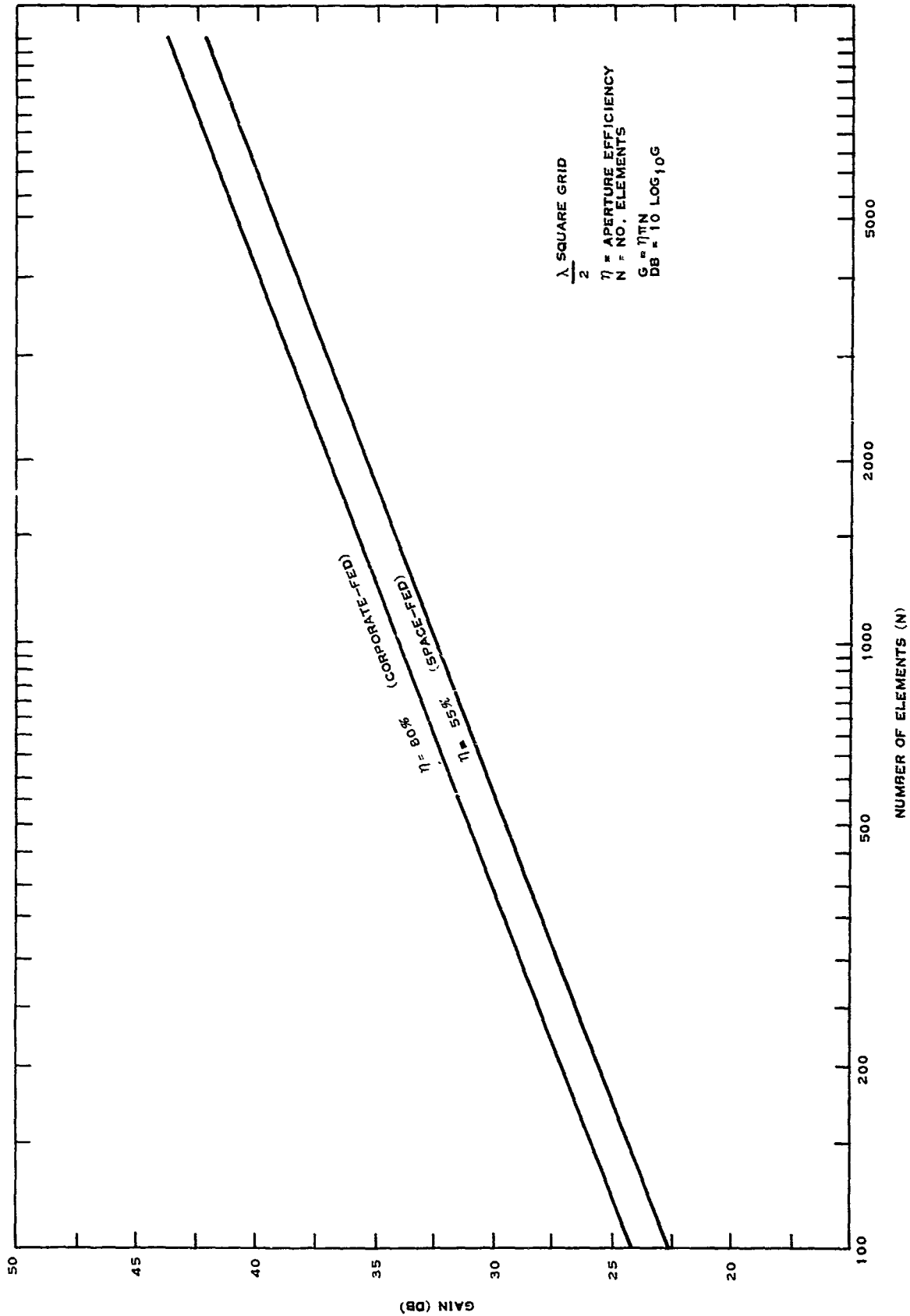


Figure 3-9. Beamwidth Versus Gain for Corporate-Fed and Space-Fed Arrays



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Figure 3-10. Number of Elements Versus Gain for Corporate-Fed and Space-Fed Arrays

covered in Appendix A. Subsection 3.2 will not only discuss the differences between the actual array modules, but also will present some very useful design data for both array modules and array module types. Paragraph 3.1.3.2.5 will cover some of the module-grouping considerations (number of array elements per module). Beam-pointing, phase-quantization, and phase-error effects as discussed in this paragraph will apply both to corporate-fed and space-fed arrays. They will be discussed additionally in paragraph 3.1.3.2.1, illustrating the influence of space feeding.

Beam pointing and shaping in a phased array is accomplished by the phase setting of phase shifters in series with each radiating element. A beam-scanning computer calculates the necessary phase for each element in order to collimate the beam at some position in space. The scan computer is a digital device, and the variable phase shifter is either a digital device or a digitally controlled analog device. Because of their digital nature, phase shifters can be classified as shown in Table III-2.

Table III-2. Classification of Phase Shifters

Phase Shifter	Minimum Discrete Phase Shift	Maximum Quantization Error
1-bit	180 degrees	90 degrees
2-bit	90 degrees	45 degrees
3-bit	45 degrees	22.5 degrees
4-bit	22.5 degrees	11.25 degrees
5-bit	11.25 degrees	5.625 degrees

Although the exact phasing required for a particular beam position can be easily calculated, quantization of this phase value by the scan computer to conform to the available shifter phase value necessitates some tradeoff or compromise between beam-pointing accuracy, sidelobe level, cost per element, prime power requirements, and scan computer complexity.

Figure 3-11 illustrates the relationship between beam-pointing angle and the required phase-shifter setting (Ψ_n) which is given by

$$\Psi_n = n \Psi_1 \text{ (modulo } 2\pi) \quad (3-2)$$

where

$$\Psi_1 = d/\lambda (2\pi) \sin \theta$$

d = equal spacing between radiators

λ = wavelength (same units as d)

θ = scan angle.

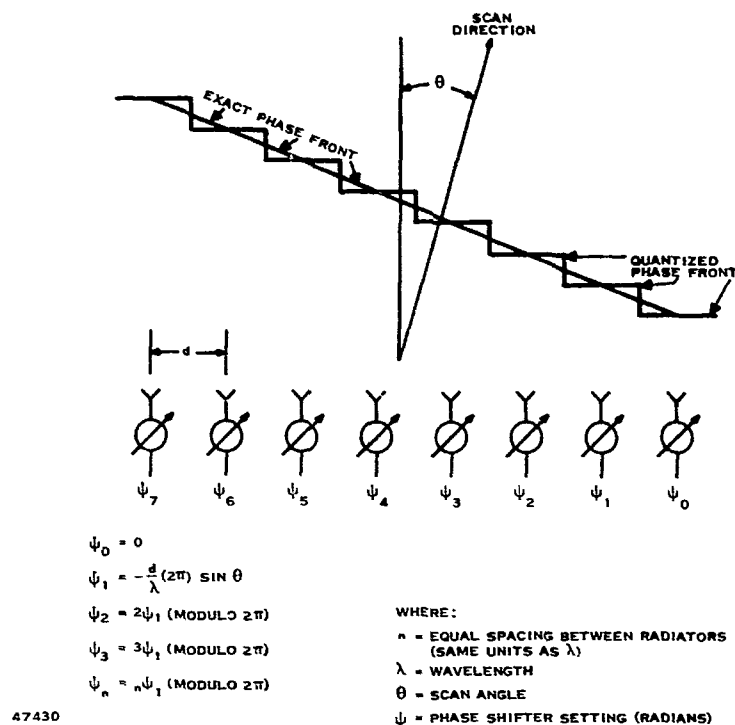


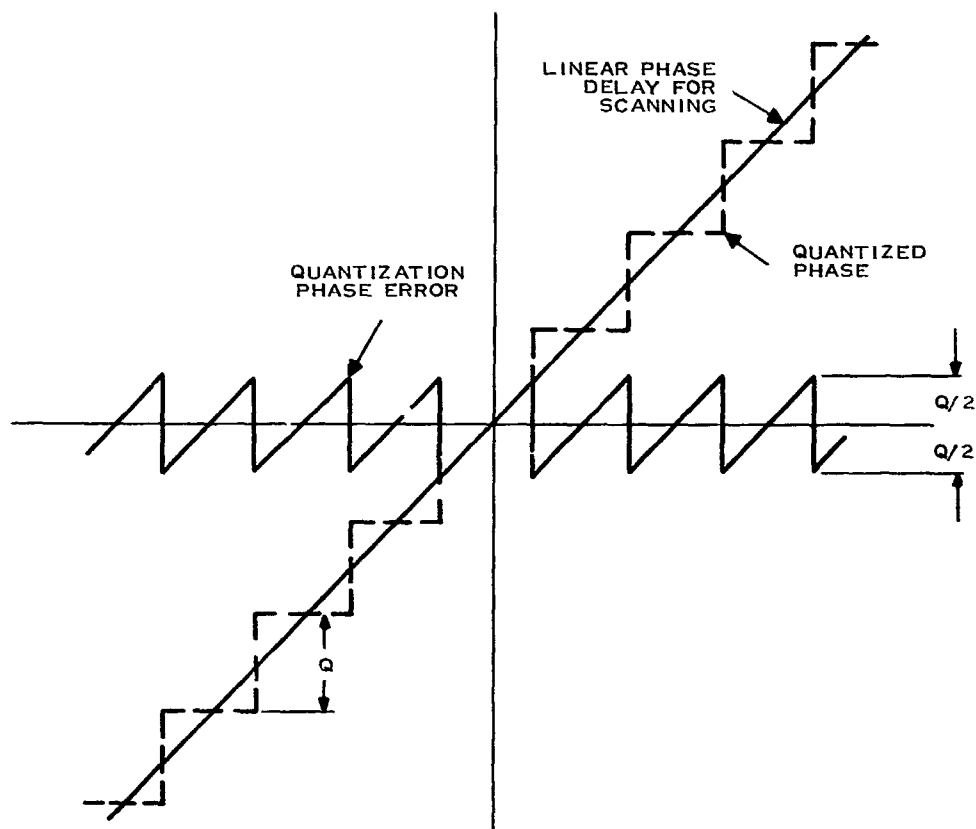
Figure 3-11. Determination of Beam-Pointing Angle

The phase delay required for scanning the beam is a linearly increasing function across the array. This must be approximated as closely as possible by the digital phase shifters through quantization of the desired linear phase distribution. Figure 3-12 shows the effect of quantizing this linear phase distribution. Q is the quantization increment which, for binary quantization with B bits, is equal to

$$Q = \frac{2\pi}{2^B} \text{ radians.} \quad (3-3)$$

Figure 3-12 shows that the error for a quantized linear phase delay is a saw-tooth function with peak values to $\pm Q/2$ with equal or regular spacing of the discontinuities.

For beam-steering angles very near broadside, the quantization of the phase shifters will cause large subsections of the phased array to have the same value of phase shift. The phase center spacings of these "subarrays" will generate grating lobes which will not be attenuated by the subarray pattern. These subarray grating lobes caused by the phase quantization for small scan angles will appear as large sidelobes with amplitudes greater than the other sidelobes which result from the aperture amplitude distribution. Of



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Figure 3-12. Quantization Phase Error for Linear Delay Across Array

course, with more phase "bits" (phase increments) per phase shifter, the size of the subarrays resulting from phase quantization will be reduced, and the magnitude of phase quantization sidelobes may be reduced to a value below the normal sidelobe level. Figure 3-13 illustrates the appearance of quantization sidelobes at small scan angles. At larger scan angles the phase-quantized subarrays will be spaced closer, and the quantization sidelobes will not be as significant as they are at small scan angles.

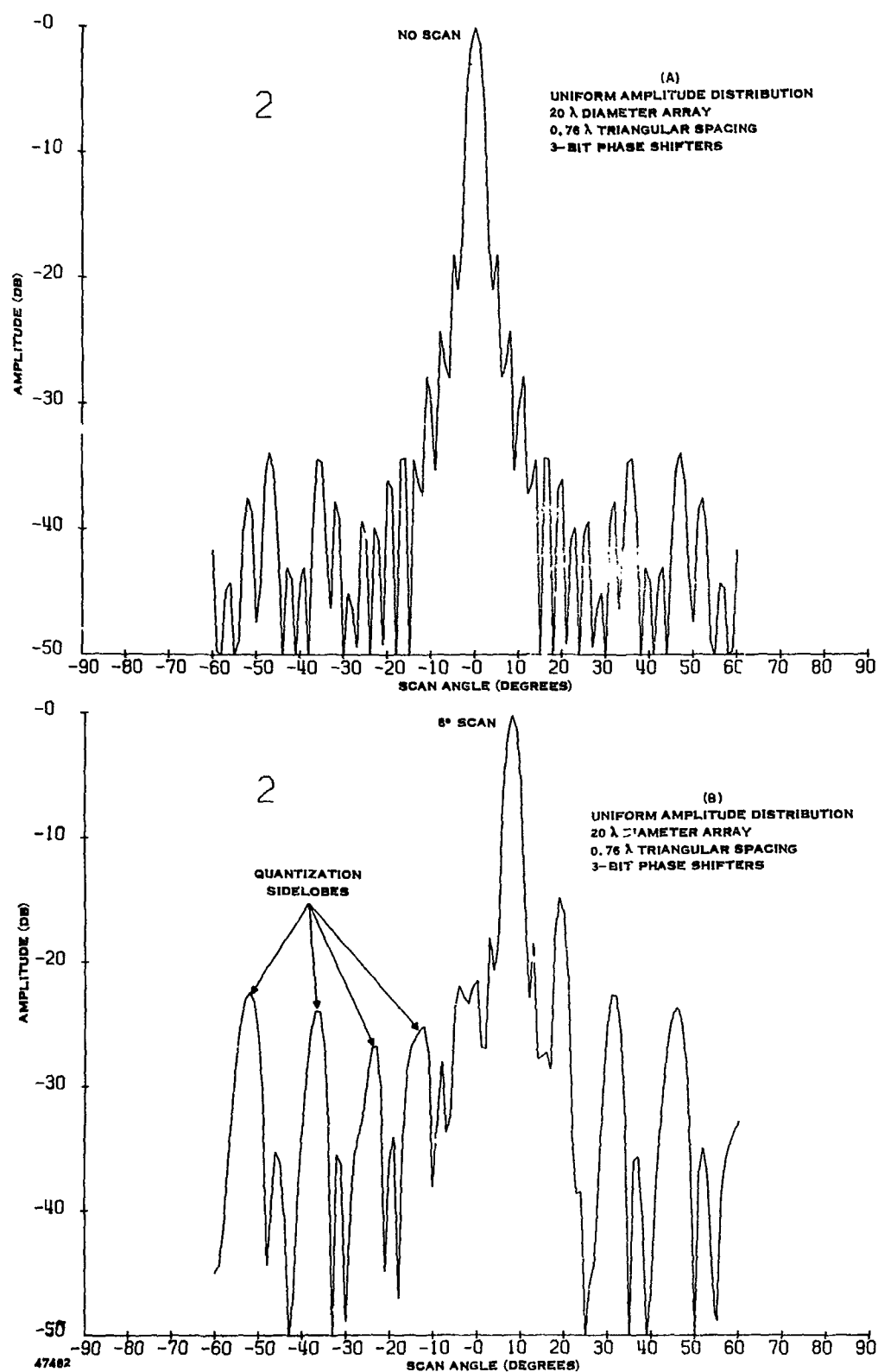


Figure 3-13. Phase-Shifter Quantization Effects

3.1.3 Space-Fed Arrays

Some of the basic characteristics of the space-fed array were discussed in the introduction (paragraph 3.1.1). Although this information will not be repeated exactly, some of the data will be mentioned again in the following paragraphs. Emphasis will be placed on the comparison between space-fed array types as well as between space-fed and corporate-fed arrays.

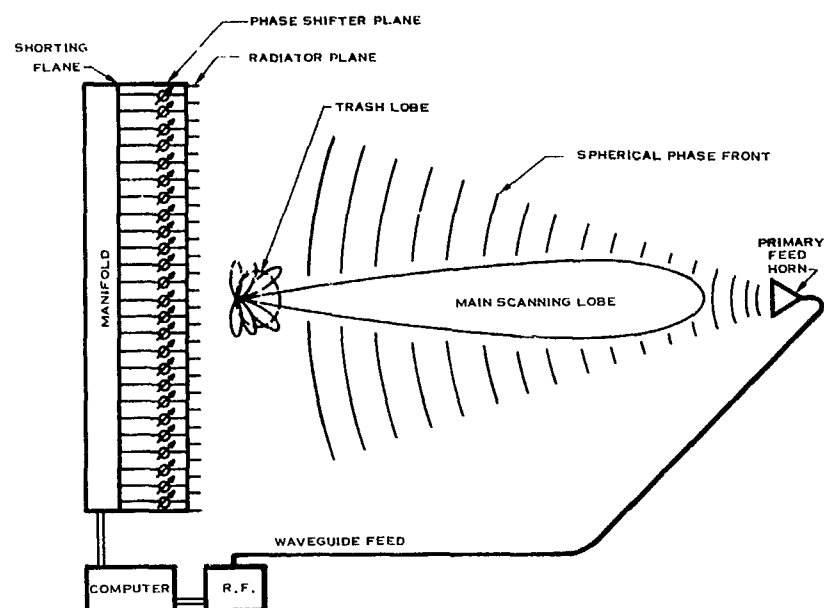
3.1.3.1 Design Tradeoffs

With space feeding there is the choice of either front or rear feeding. The aperture geometry is similar to that of corporate-fed arrays, previously covered in paragraph 3.1.2.1, the main differences being lower aperture efficiency and automatic edge tapering for space-fed systems.

3.1.3.1.1 Front Feeding Versus Rear Feeding

3.1.3.1.1.1 For the front-fed array, or "reflectarray" (Figure 3-14), the modules consist of a radiator element, a phase-shifter section, and a shorting point. The wavefront from a primary feedhorn is received by the radiator element and transferred to the phase-shifter section. The scan computer corrects for the spherical phasefront across the aperture, determines the phase necessary for the required scan direction, and quantizes this amount to the closest phase-shifter increment. In the front-fed array, the energy is coupled through the phase shifter twice because it reflects from the shorting point. The wave then is retransmitted from the single radiator element.

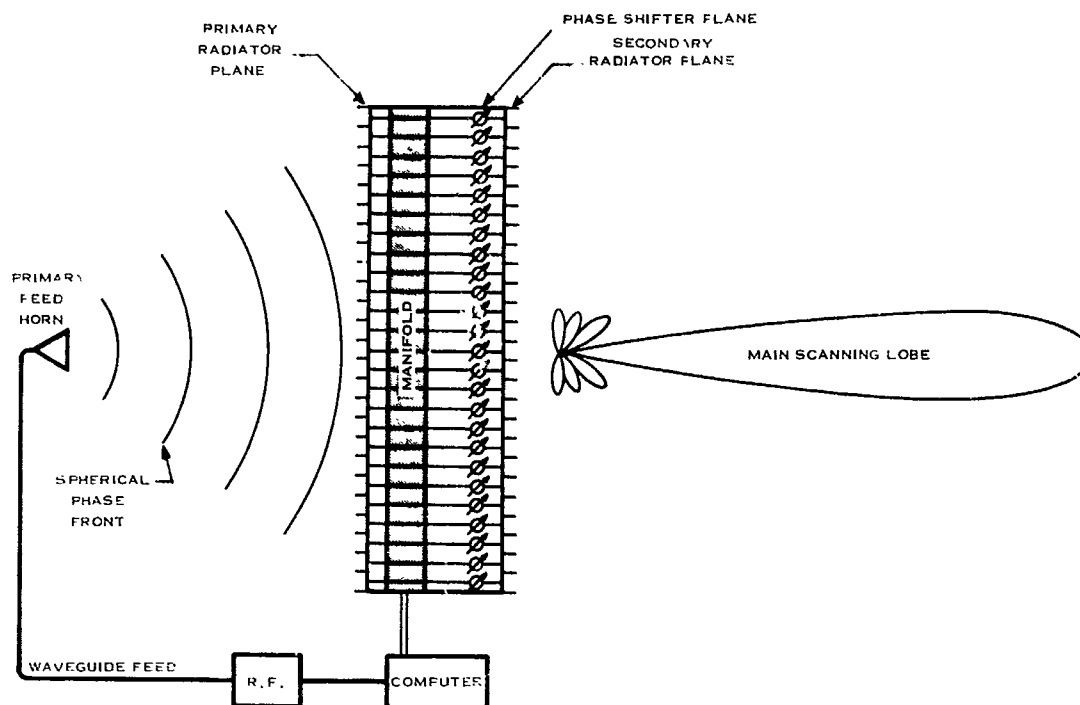
Aperture blockage and trash-lobe level in front-fed arrays are functions of array size and F/D ratios. [An F/D ratio is the ratio of the distance between the feedhorn and the array face (F) to the array face diameter (D).] The trash lobe (Figure 3-14) results from the specular or mirror reflection of the primary wavefront from the array radiator plane. The angle of reflection will equal the angle of incidence and is determined by the location of the primary feedhorn with respect to the array face. If the primary feed is located along the centerline of the array, the incident angle and reflection angle will lie along the mechanical boresight axis. The effect of the trash lobe on system operation depends on the trash-lobe amplitude as compared to the gain of the main beam. Usually, the trash lobe will appear as a low-amplitude, nonscanning sidelobe. For instance, consider the case of a reflector antenna with 27-dB gain. If the primary feedhorn had a gain of 10 dB and perfect specular reflection were assumed (no edge spillover) from a ground plane, the trash lobe would appear as a -17-dB sidelobe. Of course, a front-fed illuminated array would not have perfect specular reflection, and the trash lobe normally would be on the order of -25 dB referenced to the main scanning beam.



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Figure 3-14. Front-Fed Illuminated Array

The effect of aperture blockage on illuminated arrays is related to aperture size. With small arrays (less than $D = 10\lambda$), the aperture blockage of the primary feedhorn can cause excessively high sidelobes and severe beam deformation at certain scan angles. However, for larger apertures (greater than $D = 20\lambda$) only a small percentage of the aperture would be blocked and the sidelobe levels would not suffer appreciably. When the primary feedhorn is along the array centerline, maximum aperture blockage will result. To reduce aperture blockage, the primary feedhorn must be placed off boresight, which places the trash lobe to one side. Because of the nature of illuminated phased arrays, the antenna designer has great flexibility in optimizing an antenna system for each specific application. The choice of feedhorn placement may depend upon the system pointing angle requirements. For instance, the feedhorn would be offset toward the top of an aperture that scans "down" for a particular system requirement. The feed blockage shadow will then move toward the top of the aperture as the beam is scanned below boresight. The effect of aperture blockage for various beam pointing angles may be analyzed via the use of a computer program.



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Figure 3-15. Rear-Fed Illuminated Array

3.1.3.1.1.2 Rear-Fed Array

In many respects, the rear-fed illuminated array (Figure 3-15) is similar to the front-fed array. A significant difference between the two is elimination of the trash-lobe and feed aperture blockage of the front-fed array. Holding other parameters constant, the elimination of the aperture blockage of the front-fed array always will cause a reduction in the array sidelobe level. The extent of sidelobe improvement of the rear-fed array compared to the front-fed array is a function of the sidelobe level available without blocking as well as the percentage of aperture blocking. As the array size increases, the percentage of aperture blocking decreases. This difference implies that for large arrays the sidelobe improvement expected for rear feeding may not increase significantly.

The use of a rear-fed array becomes more desirable when the feed structure is complex. Some multipurpose systems require several different feedhorns for optimum completion of the mission. These feedhorns may be arrayed to provide the correct illumination function to the phased array. Information providing the location of their phase center may be entered in the scan computer memory to properly collimate the otherwise skewed beams. For

instance, if a twelve-horn monopulse primary feedhorn is used for improved monopulse performance,^{4, 5} the rear-fed array becomes more desirable than the front-fed array.

Another difference between the two space-fed array configurations is the phase-shift element. The phase-shift element for the reflective (front-fed) array requires an rf short as well as a two-way path through the phase shifter. The transmissive (rear-fed) array requires only a single path through the phase shifter for a single transmit or receive mode. Phase-shifter losses are about the same for both array types.

The phase-shifter control manifold is another system tradeoff between the two array types. The manifold for the front-fed array is much easier to construct as compared to the transmissive configuration. For the transmissive, or rear-fed array, the manifold must fit between the radiating elements. This represents a difficult requirement for close-spaced phased arrays that must scan over large angles.

Polarization requirements also enter the tradeoff analysis between the front- and rear-fed arrays, depending upon the type of phase shifter selected. If a multipolarized system is required, it is sometimes advantageous to perform the necessary polarization switching in the array element itself (rather than in the feedhorn structure). For instance, a switching circulator may be used in the transmissive array element to perform the polarization switching with a minimum of rf loss. This eliminates the requirement for high-power devices in the feed. This also can be accomplished in front-fed arrays with some types of phase shifters.

In summary, the major tradeoff considerations relevant to the application of space-fed arrays are the effects of aperture blockage and complexity of array fabrication. Volume and weight are also important considerations. Exact module configuration of the array should also be considered and will be discussed in paragraph 3.2.

3.1.3.1.2 Aperture Geometry and Illumination

Aperture geometry refers to the shape of the aperture and method of rf power distribution. Considerations of aperture geometry and illumination apply equally to the design of space-fed arrays and horn-fed parabolas. The aperture efficiency of space-fed antennas will generally be less than enclosed-fed antennas where the efficiency can range as high as ($\eta = 85$ percent). The efficiency of the space-fed antenna is affected by edge illumination, spillover, and aperture blockage and will generally be on the order of ($\eta = 55$ percent).

The aperture shape is related to the aperture illumination in that both are important considerations when designing the antenna for particular sidelobe criteria. For instance, keeping the aperture width and height constant and using a uniform amplitude illumination, the theoretical sidelobe levels of rectangular and circular apertures are 13.2 dB and 17.6 dB,

respectively. The advantage of using a circular aperture is apparent. Determining the phase and amplitude distributions required for a particular beam shape is an application of conventional aperture theory. The exact configuration required to achieve the necessary aperture illumination is obtained after considering system performance tradeoffs.

The aperture size, beamwidth, and amplitude illumination of a circular aperture are related as illustrated in Figure 3-7. The maximum aperture diameter establishes the smallest attainable beamwidth for a pencil-beam mode. Using a 10λ diameter array as an example the beamwidths range from 6 to 11 degrees for various amplitude distributions. A corporate-fed array can provide a uniform amplitude distribution which results in maximum gain and minimum beamwidth from a given aperture. On the other hand, a horn-fed parabola or space-fed phased array inherently provides a tapered amplitude distribution and lower sidelobes can result. The corporate-fed array may also be illuminated with a low-sidelobe amplitude taper. The theoretical sidelobe levels indicated in Figure 3-7 are shown as a guide and these levels will not necessarily be obtained with a particular antenna design. Aperture blockage, surface errors or feedhorn defocusing, and phase shifter phase and amplitude (insertion loss) errors will contribute to the general degradation of the theoretical sidelobe level of antenna systems.

The gain of an antenna is related to its aperture efficiency and shape of far-field pattern. Gain of a typical space-fed array, assuming 55-percent aperture efficiency and pencil beams may be determined from Figure 3-16. If square half-wavelength spacing is assumed the antenna gain may be related to the number of array elements as shown in Figure 3-17.

3.1.3.1.3 Power-Handling Requirements

The rf power-handling requirement of a space-fed array element is a function of the aperture amplitude distribution. This aperture distribution will usually be a tapered distribution with the array edge elements receiving from -8 dB to -20 dB less power than the array center elements. The exact aperture distribution is a function of feedhorn beamshape and gain and distance between the feedhorn and array. If an amplitude distribution is used that gives on the order of -25-dB Sidelobes (without aperture blockage), then Figure 3-18 will illustrate the expected maximum power per element. As an example, a 1000-element circular array with a 25-watt (pulse-peak power) primary feed would require array elements in the aperture center rated at approximately 70 watts (pulse-peak power).

This power-handling requirement for space-fed arrays will affect the choice of phasor, rf power source, feedhorn design and various radar parameters such as pulsewidth and duty cycle. For instance high prf radar systems will, in general, have much lower peak powers; and phase shifter power-handling requirements will be greatly reduced.

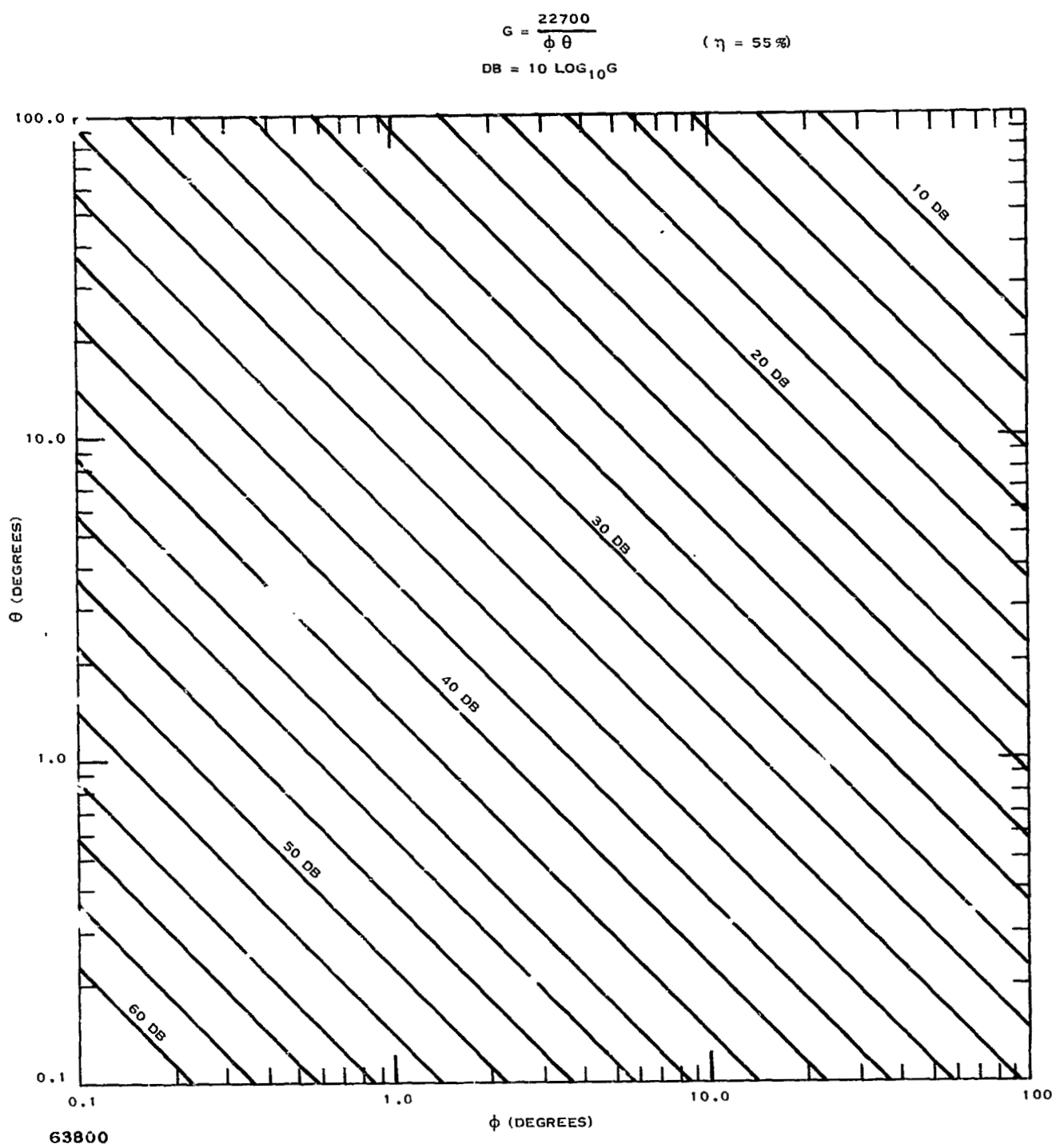
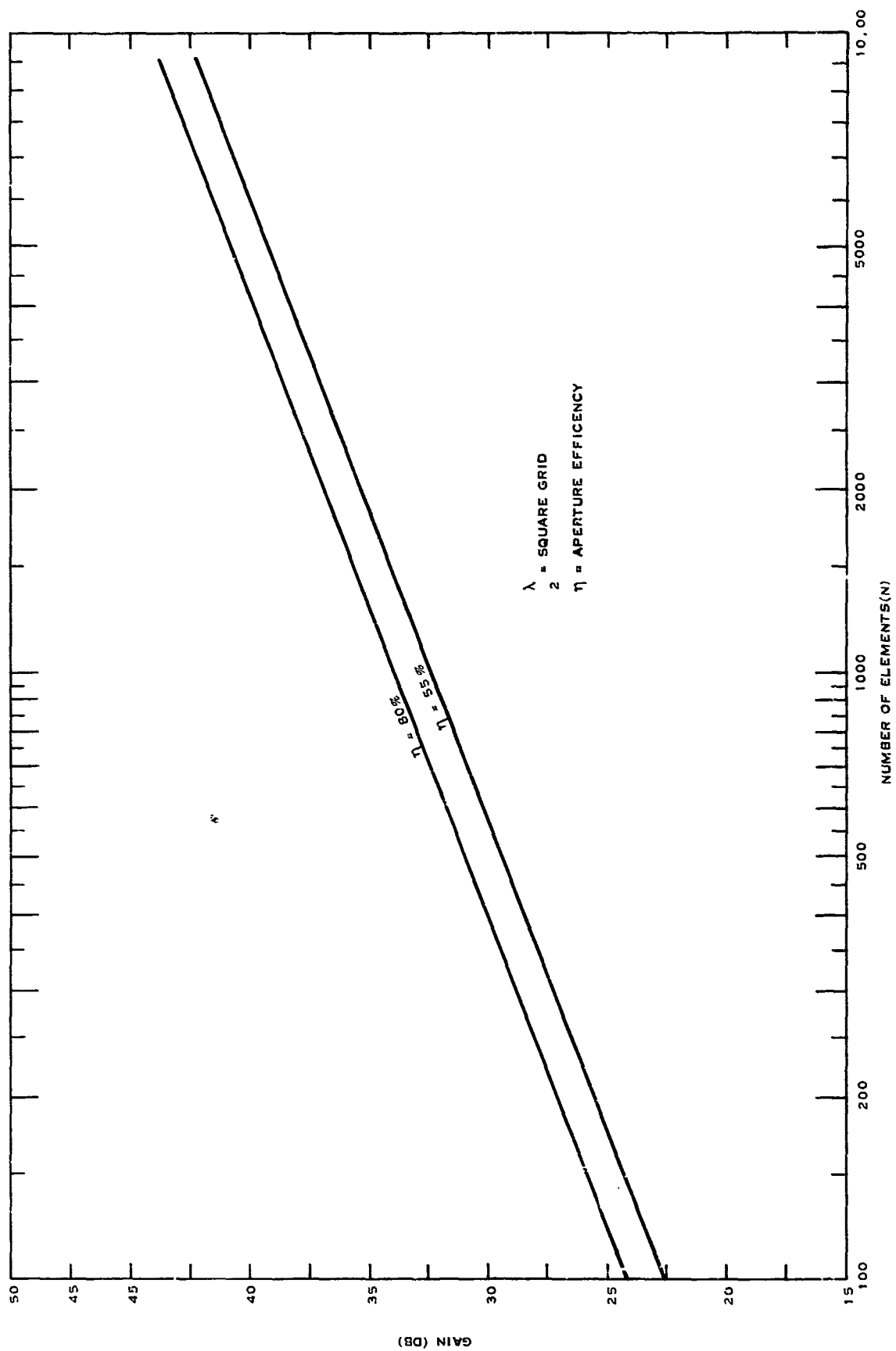


Figure 3-16. C in Versus Beamwidth for Space-Fed Arrays



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Figure 3-17. Number of Elements Versus Gain

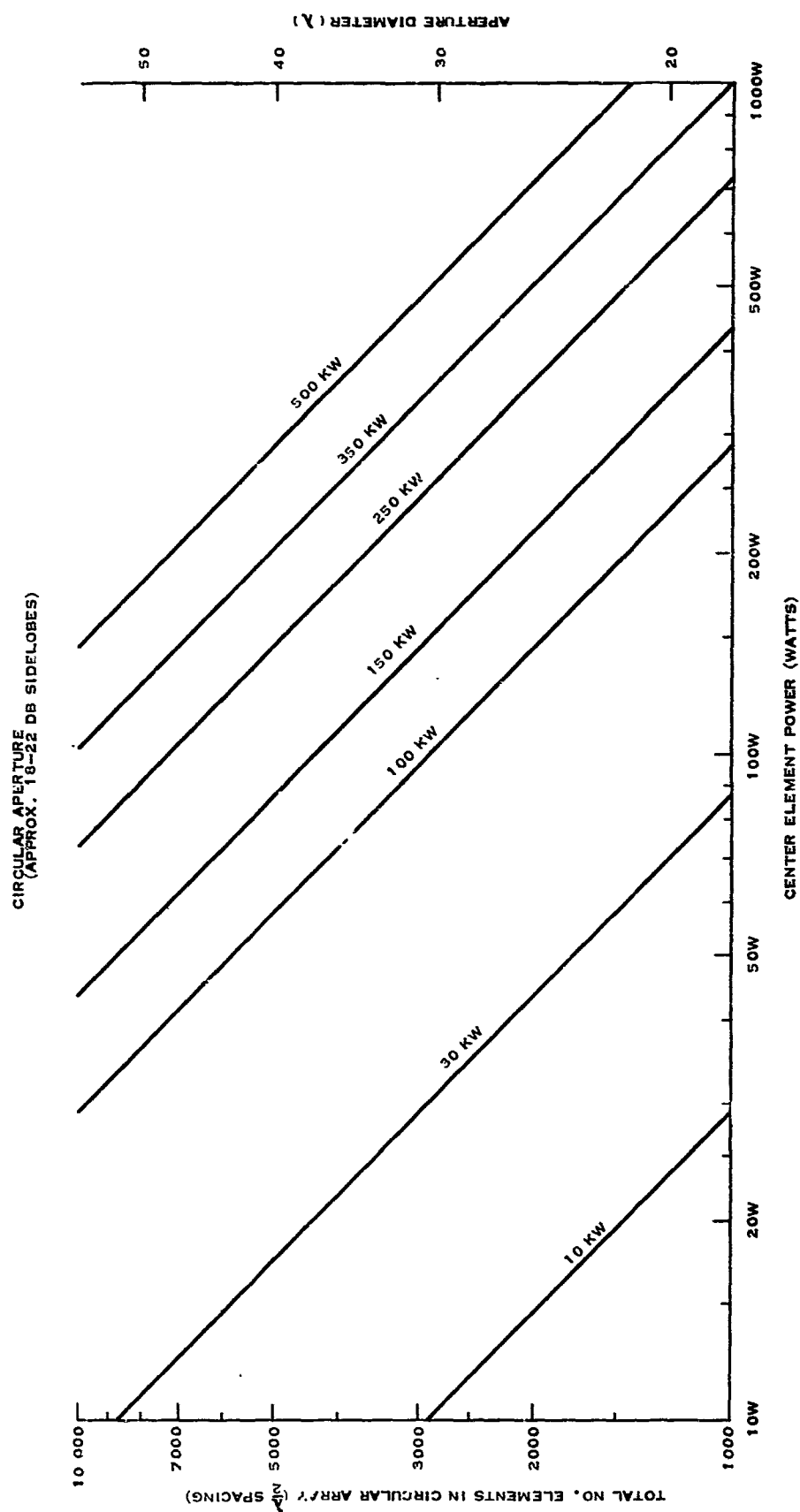


Figure 3-18. Array Center Element Peak-Power Requirements

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3.1.3.2 Properties of Space-Fed Arrays

Most space-fed arrays have a number of properties that are characteristic to all phased arrays. Included are phase quantization, mutual coupling, and beam variation with scan. Although many of these properties have been discussed elsewhere, the characteristics particular to space feeding an array will be covered in this subsection.

3.1.3.2.1 Phase Correction and Quantization

Proper phasing of the array elements determines the beam shape and beam-pointing characteristics of the array. The phasing required for a particular scan position or beam shape is computed by the scan computer which controls the phasor. Usually, digital control is used which results in quantization of the exact computed phase value. At certain scan positions, usually near boresight, the quantized phase values can add in a coherent fashion and tend to generate high sidelobes. These sidelobes, termed quantization sidelobes, are a function of the phase-shifter phase increment (number of bits), array size, array element pattern, array amplitude distribution, and scan position. When the phasors are used to collimate the spherical phase front from the feed structure of a space-fed array, the quantized phase distribution across the array face is randomized to some extent and the quantization sidelobes are reduced. Figure 3-19 illustrates the difference between quantization phase errors generated from plane phase fronts and spherical phase fronts.

The phase correction for the spherical phase front can be separated into the sum of a parabolic phase correction and a small remainder term. The parabolic part for each antenna element turns out to be the sum of two terms which depend, respectively, on the row and column location of that element; thus, the correction is naturally performed by a row and column steering computer. The remainder term increases with increasing curvature of the spherical phase front which occurs as the feed point is moved closer to the array of elements. If the feed point distance (F) is much closer than half of the array diameter (D), (if $F/D = 0.5$), the error resulting from the omission of the remainder term will be parabolic in shape which will tend to broaden the beam. In addition, if the feed point distance is significantly greater than the array diameter (if $F/D = 1$), the advantage of the nonregular quantization phase error is reduced and quantization sidelobes will tend to increase.

In summary, space feeding with ($0.5 \leq F/D \leq 1$) can offer advantages in reduction of quantization sidelobes. The exact quantization sidelobe level expected with space-fed arrays is also a function of array scan computer design and instrumentation as well as array characteristics such as amplitude taper and expected range of random phase and amplitude errors. The accurate prediction of quantization sidelobe levels may be best determined by a pattern analysis on a digital computer.

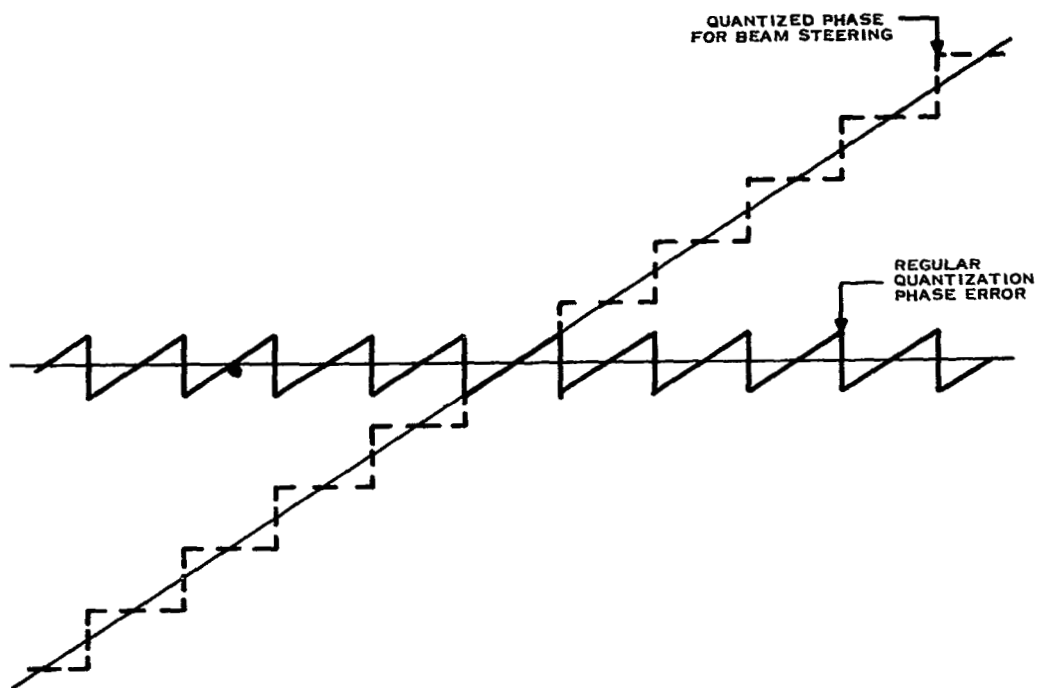


Figure 3-19. Plane Phase Front Quantization (a)

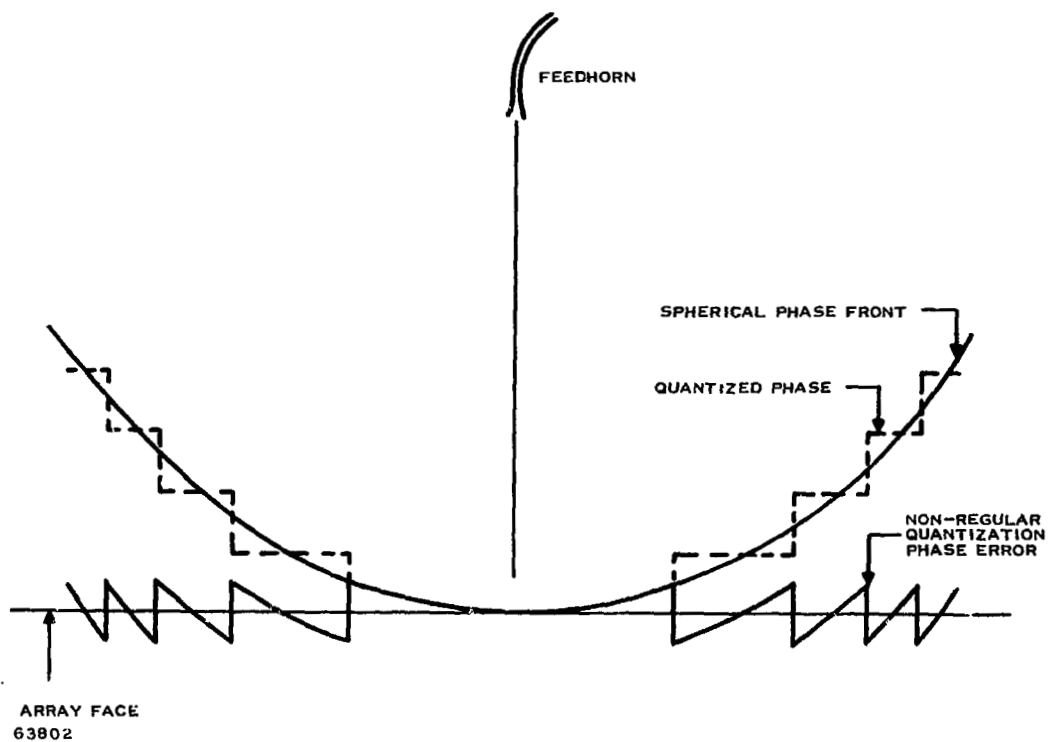


Figure 3-19. Space-Fed Configuration (b)

3.1.3.2.2 Mutual Impedance Effects and Compensation

Mutual coupling in phased-array antennas has been discussed extensively^{6, 7} and is covered in greater detail in Appendix A. Mutual coupling is caused by interaction of the rf fields of the phased-array radiation elements, and the analysis of this effect has been emphasized because of its importance in array design. The presence of mutual coupling in a phased array causes a gain variation, as the beam is electronically steered, as well as increased sidelobes and impedance mismatch of the radiating elements.

Exact analysis of coupling is difficult because of the number of variables that must be considered and the complex relationship between the variables. Fortunately, various approximations can be used when making analytical studies of the problem. The coupling analysis can be experimentally verified by measurements in a phased array, but this requires a complete active array which could be very costly. Waveguide simulators are useful tools for experimentally verifying coupling analysis because only a few array elements are required.

Mutual coupling is caused by an element producing fields on the surface of adjacent array elements. These electromagnetic fields on the surface of an element cause a change in the element impedance which is dependent on the scan position of the array. For an infinite array the mutual coupling causes only a variation of impedance with scan position. However, coupling in a finite array not only causes a variation of impedance with scan, but also alters the array pattern and produces edge effects. Edge effects appear as an impedance variation from one element to another across the array. The largest impedance variations occur near the edge of the array and for that reason are called edge effects. The added influence of coupling in a finite array is caused by the unsymmetrical environment of the elements in the array. Practical radar arrays consist of many elements so there is some merit to examining mutual coupling of infinite arrays as well as finite arrays.

The use of an array simulator⁸ for radiating element design has become popular in recent years. The waveguide simulator is based on the principle of multiple images and simulates the impedance function (including mutual coupling) of an infinite array. This approximation holds well for large finite arrays as long as simulation of the edge elements is not desired. Although simulation may be performed at only certain discrete scan angles for particular polarizations, the simulation technique is still quite useful for array design.

Mutual coupling considerations for space-fed arrays may be separated into rear-fed and front-fed arrays. Since a radiating element may be easily matched to a particular scan angle, the "pick-up" radiators on the rear

of a rear-fed module may be exactly matched to the feedhorn. Since this "rear" beam of the array does not scan, an ideal load may be provided to one port of the transmissive phase shifter via proper element impedance matching. On the other hand, the radiator on the "front" of the module must cover a much larger scan volume—the scan volume of the scanning radar beam. Although this radiator may be impedance matched to cover the required scan volume with a minimum overall mismatch (paragraph 3.1.4 discusses this optimum matching technique), there will still be a considerable impedance mismatch which will seriously affect the operation of the module. A VSWR of 2 or 3 is not uncommon for wide scanning phased arrays and this mismatch will degrade the array performance. Figure 3-20 illustrates the expected transmission loss resulting from high values of VSWR.

Another effect of high values of VSWR is the possible formation of reflection sidelobes caused by multiple reflections through the feed lines and phasors.⁹ These reflection sidelobes are formed with certain types of feed systems and are more of a problem with reciprocal phase shifters than with nonreciprocal phasors. The early work on reflection sidelobes¹⁰ was with enclosed-fed feed systems with both isolated and unisolated ports. However, the same effect would apply to space-fed arrays (perhaps on a limited basis). For the case of front-fed (reflective) arrays, the type of phase shifter (i.e., reciprocal or nonreciprocal) should make little difference. However, rear-fed arrays offer the option of using a nonreciprocal circuit element to load the reflected power due to element mismatch. For instance consider the use of a switching circulator as shown in Figure 3-21. By using a thin-film or microstrip configuration circulator it would be feasible to switch the device on a pulse-to-pulse basis in order to prevent power reflection problems in the transmit mode. Disadvantages of this scheme include an extra component for each element as well as added insertion loss. However if multipolarization requirements specify the use of nonreciprocal elements it may be feasible to consider such a scheme. Polarization requirements and possible module layouts will be discussed in greater detail in paragraph 3.1.4

3.1.3.2.3 Thinning Considerations

There are several tradeoffs to consider when aperture thinning is used in a phased array. The cost and weight of the array are reduced, but at the same time the radar gain is decreased. Aperture thinning may also be used to reduce the sidelobe level if the aperture amplitude distribution is relatively flat (not tapered) and a space-tapering criterion is used. Here, space tapering implies that the elements are selectively thinned (removed) in such a manner so as to result in an effective amplitude tapering of the aperture. Space tapering should be performed only after careful analysis to ensure that secondary sidelobes will not be generated as the array is scanned.

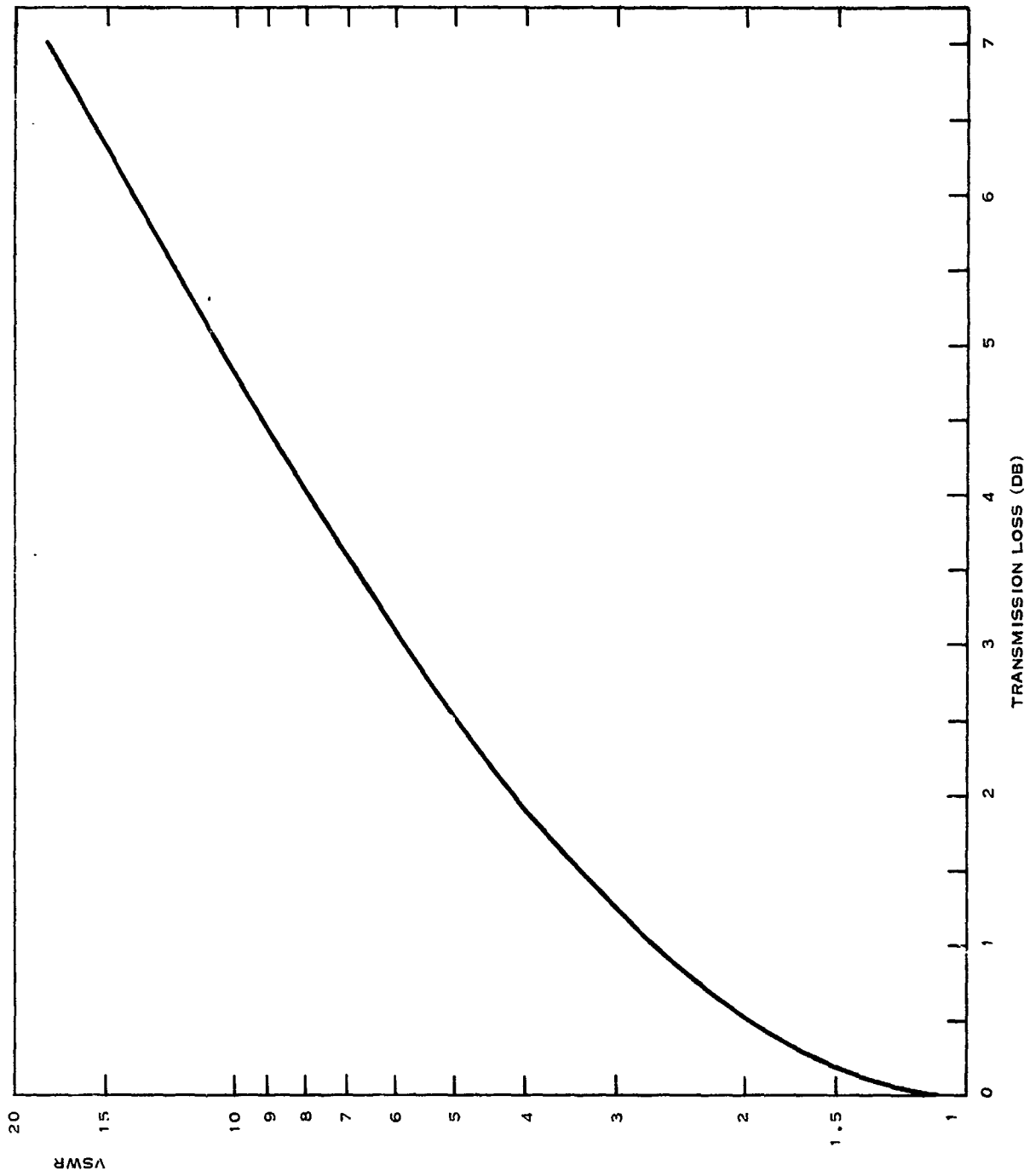


Figure 3-20. Transmission Loss Versus VSWR

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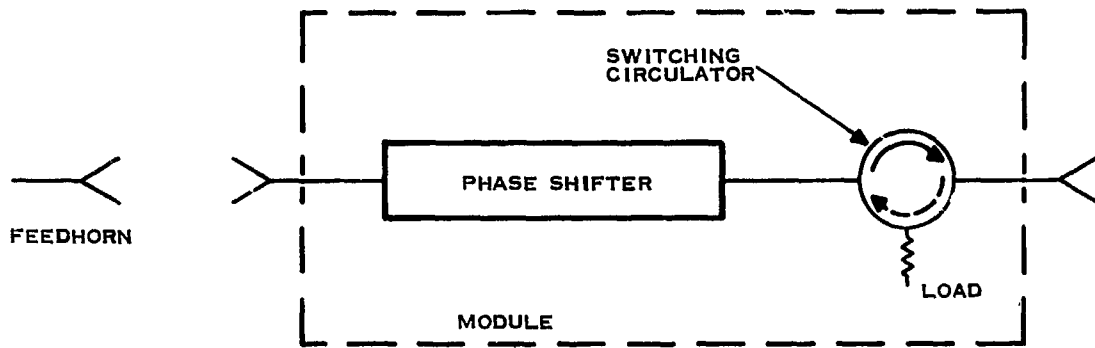


Figure 3-21. Rear-Fed Array Module

Thinning may be compared to the basic shapes of square and round apertures. It is obvious that a round aperture is simply a "corner-thinned" square array of the same width. With uniform amplitude distributions, the "thinned" round aperture generates -17.6 -dB sidelobes which may be compared to the -13.2 -dB sidelobes available from the rectangular (or linear) aperture. By thinning the filled circular array, a further sidelobe reduction is possible, when using uniform amplitude distributions. However, the exact amount of thinning for an optimum configuration is a function of aperture size, required beam coverage angles, phase quantization and amplitude distribution. Thinning for sidelobe control is more of an advantage for corporate-fed arrays using uniform amplitude distributions for improved transmit efficiency. However, when designing large space-fed arrays (i.e., greater than 2000 elements), aperture thinning should be considered for reduction of array cost and weight. If the application of the array requires high reliability and the array elements are a critical item, then thinning should not be considered. Array thinning considerations are summarized in Table III-3.

Table III-3. Thinning Considerations

DISADVANTAGES	ADVANTAGES
Gain loss	Reduced array cost and weight
Array reliability reduced	Sidelobe reduction in some cases via space tapering

3.1.3.2.4 Feedhorn Design

Figure 3-22 illustrates some of the basic waveguide feedhorn configurations used for space-fed arrays. These feedhorns are generally used

for frequencies 6.0 GHz and above, although they may be used for lower frequencies as well. Dipole feeds have a weight advantage for the lower frequencies and should be considered in the system design.

The location of the feedhorn is an important tradeoff consideration in the design of a space-fed phased array. Both the monopulse patterns and the effects of quantized phase-front correction must be considered. Such things as number of phase shifter bits and type of scan computer instrumentation affect the overall system tradeoffs. Figure 3-23 lists some of the basic feedhorn considerations of a space-fed phased array. Much of the feedhorn design criteria for space-fed arrays is similar to focal-point parabola design. For instance, if the array modules are used for polarization switching, a vertical polarized feedhorn may be used for a 4-lobe monopulse system. Since only vertical polarization is required in combination with 4-lobe monopulse processing, the optimum feedhorn is a dual horn/multimode design.¹¹ The array is illuminated with greater efficiency in the azimuth plane (H-plane) by the use of a multimode feedhorn, resulting in less spillover loss. Conventional sum and difference hybrids are used for the E-plane, but different waveguide modes are used to generate the H-plane sum and difference signals, offering approximately 25 percent greater error sensitivity than the conventional 4-horn monopulse feed.¹² Since monopulse feedhorn design has been covered in the literature for focal point parabolas, design details will not be covered extensively in this report.

Figure 3-24 illustrates the effect of feedhorn size on the F/D ratio of the parabola.¹³ This curve represents a "compromise" optimization between the antenna pattern angular sensitivity and the range sensitivity. These sensitivity functions (which are a function of the F/D ratio and feedhorn aperture) are important in monopulse antenna design because the antenna may be optimized for maximum tracking accuracy performance. These functions have been developed in the literature for both focal point parabolas and Cassegrainian systems.¹⁴⁻¹⁷

The edge taper on a space-fed array is similar in design to a dish antenna, except that the space attenuation factor at the edge of the aperture is modified because the array is a planar surface rather than a parabola. Figures 3-25, 3-26, and 3-27 illustrate typical design curves for the determination of the primary feedhorn size. Using Figure 3-25 as an example, let us choose a focal length/diameter ratio of 0.5 ($F/D = 0.5$), the subtended angle of the feed is 90 degrees, which gives a space attenuation factor of 3 dB. As shown in Figure 3-26, this gives a required feedhorn 10-dB beamwidth of 82 degrees $\left(\frac{90}{1.1}\right)$, for a 15-dB (12 + 3 dB) edge taper. The required E- and

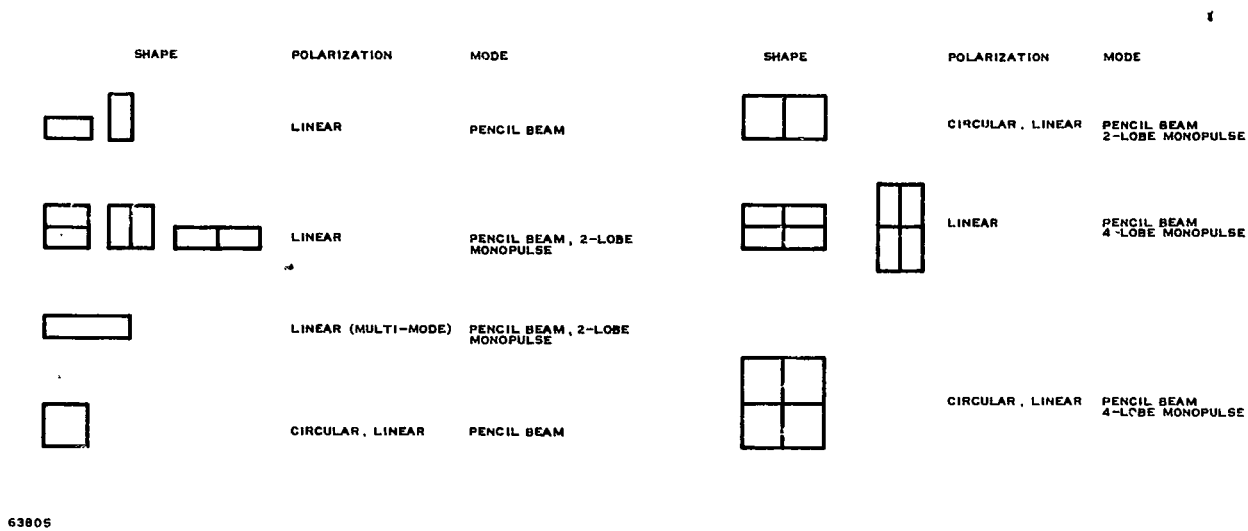
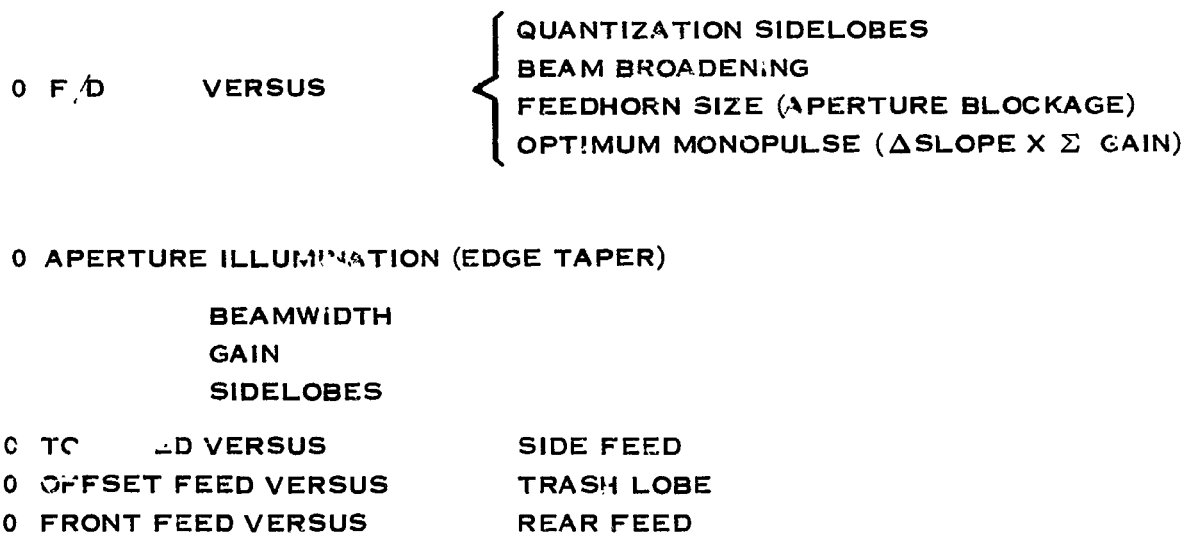


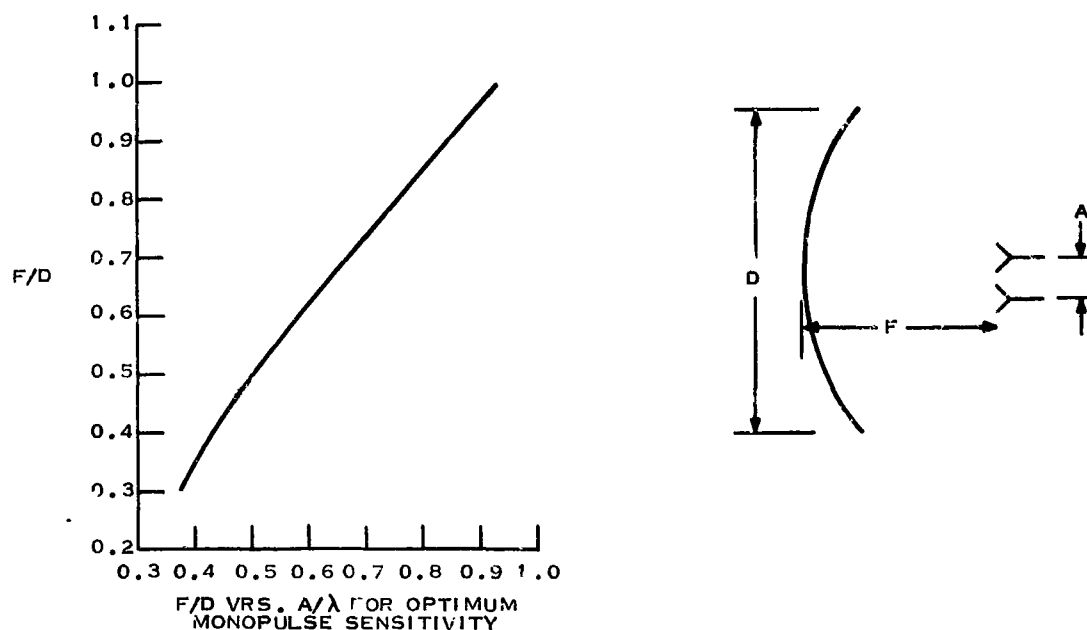
Figure 3-22. Basic Feedhorn Types (Waveguide)

FEEDHORN CONSIDERATIONS AND TRADEOFFS



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Figure 3-23. Feedhorn Considerations and Tradeoffs



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Figure 3-24. F/D Versus a/λ for Optimum Monopulse Sensitivity

H-plane apertures from Figure 3-27 are therefore about $1.2\lambda \times 1.4\lambda$, respectively, for a symmetrical (i. e., round) aperture illumination. It should be pointed out that these design curves are just approximations and do not take into account the feed pattern from complex feed structures or feed structures with flanges and dielectric pressure seals. The final design of the feed structure will usually require empirical design for optimum array performance.

3.1.3.2.5 Module Configurations

Figure 3-28 illustrates the basic block diagram for both front and rear-fed phased array modules. Each module has a minimum requirement of a phase shifter and driver circuitry which generally consists of both logic and phase shifter driver. If there is a multipolarization requirement for the module some additional circuitry will be required for polarization switching. This additional circuitry may include polarization logic, an extra phase shifter, microwave hybrids or ferrite circulators.

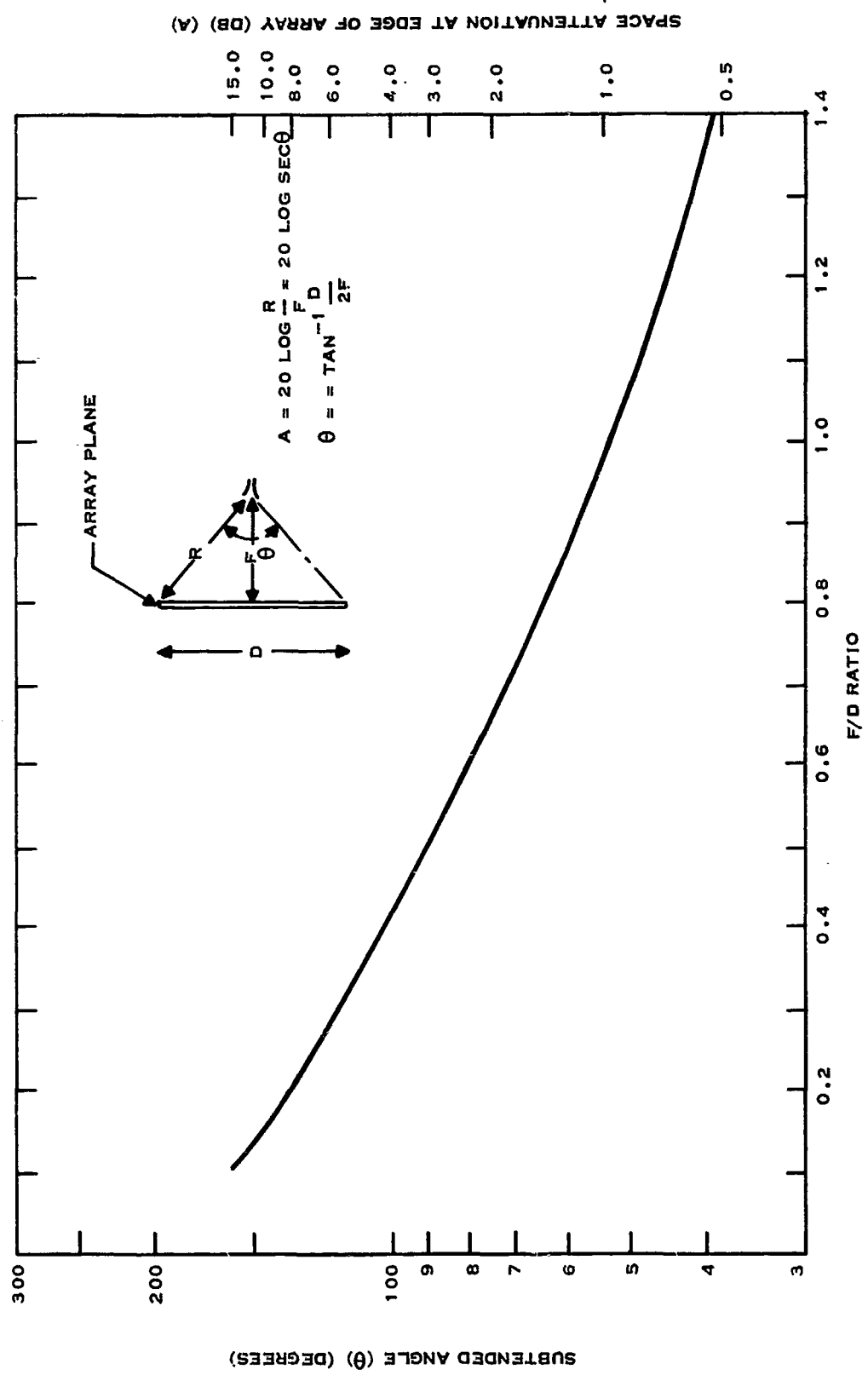
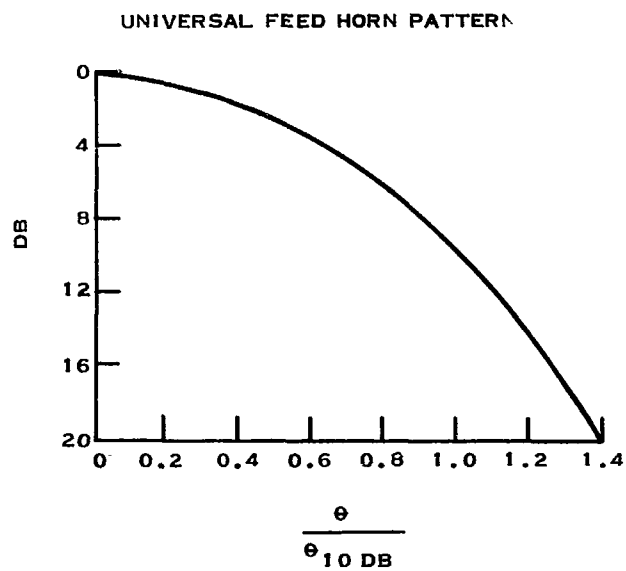


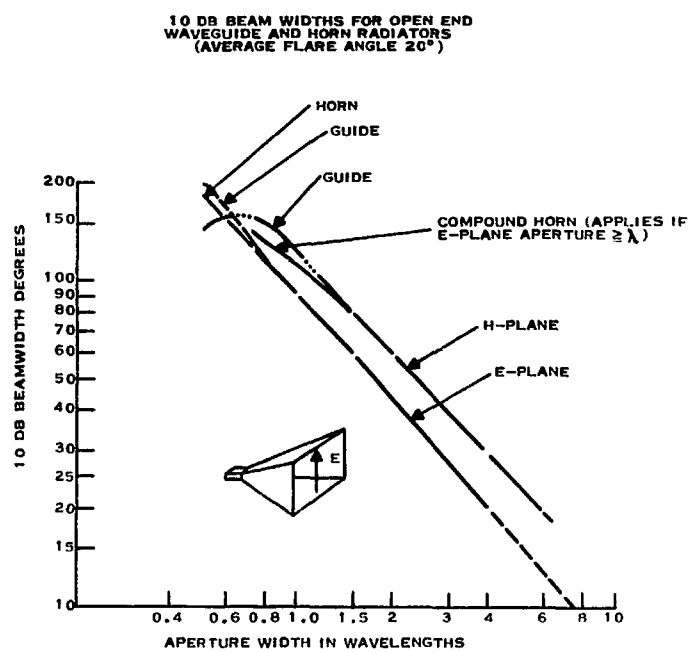
Figure 3-25. Space-Fed Array Feed Design

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Figure 3-26. Pencil-Beam Feedhorn Design Curves



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Figure 3-27. 10-dB Beamwidths for Open-End Waveguide and Horn Radiators
(Average Flare Angle, 20 Degrees)

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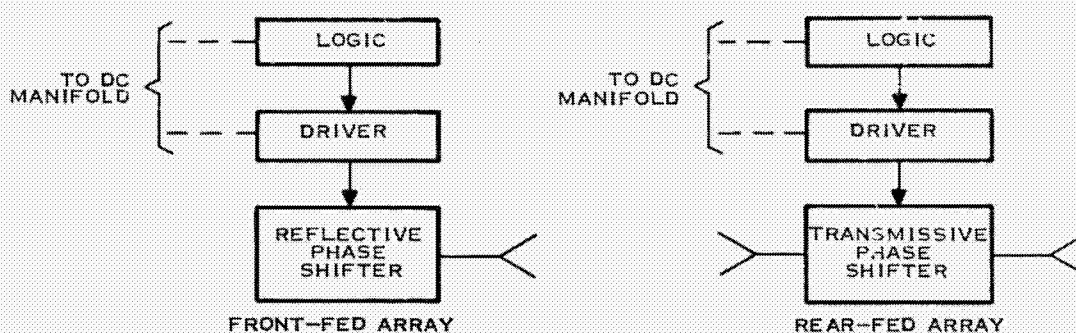


Figure 3-28. Module Block Diagrams

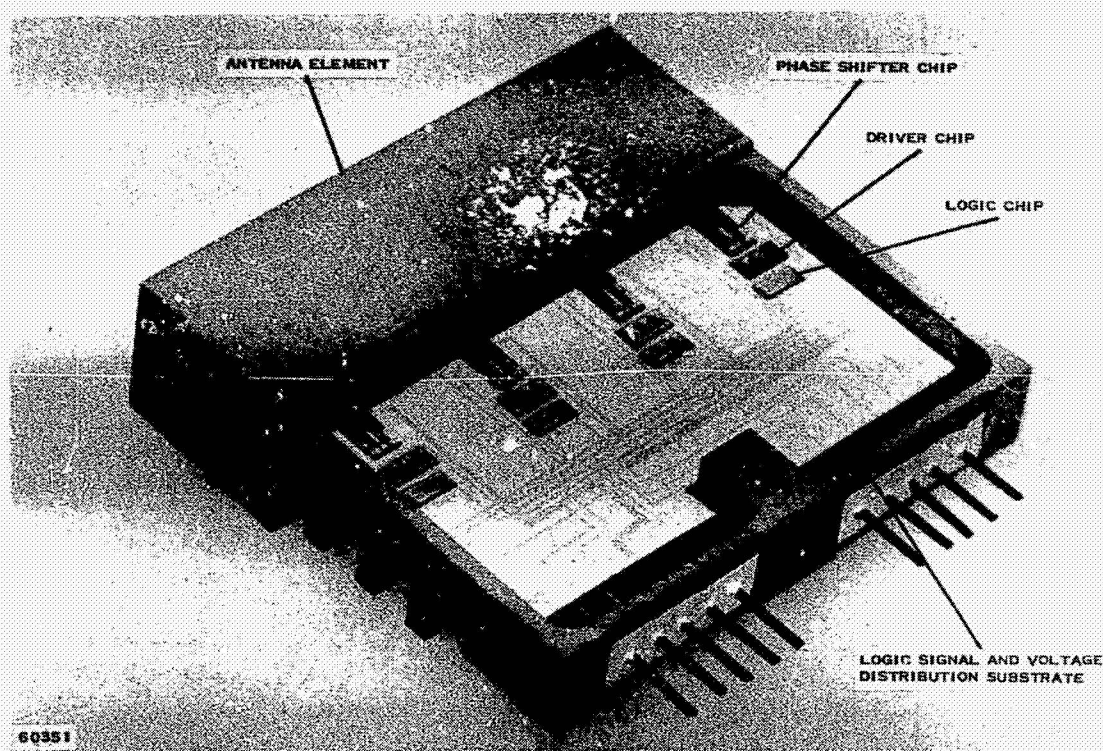


Figure 3-29. Module Mockup Interior View

Figure 3-29 illustrates a typical module for a linear polarized reflective space-fed array. This module, for a K_u -band array, consists of four sets of phase shifters and drivers and a common dc distribution network. By grouping the array elements in groups of four per module, an optimum assembly configuration was achieved for the particular array. The optimum number of elements per array module for a particular array configuration may be determined based on size and assembly requirements.

For instance, as the wavelength approaches one or two inches, items such as construction tolerances, connector sizes, and assembly techniques become important factors in the determination of the optimum package configuration. At the lower frequencies (that is, below X-band), single element modules will probably offer the optimum configuration.

As an example, Table III-4 illustrates tradeoffs considered in the choice of configuration for the module illustrated in Figure 3-29. Choice of antenna element, phasor, polarization, and manifold construction

Table III-4. Element Grouping Considerations

Grouping	Advantages	Disadvantages
	1. Does not restrict shape 2. Best for lower frequencies	1. Large number of plug-in elements 2. Large number of dc connections required per module
	1. Does not restrict shape	1. Large number of plug-in elements 2. Large number of dc connections required (5/element)
	1. Reasonable size for handling 2. Simplified dc distribution (7/4 elements)	1. Slightly restricts antenna shape 2. Manufacturing more complicated than in-line module
	1. Reasonable size for handling 2. Simplified dc distribution (8/4 elements) 3. Easiest to manufacture	1. Slightly restricts antenna shape
 or 	1. Fewest manifold connections	1. Modules cannot be widely used 2. More than one size will usually be required in array.

techniques are all important factors which will influence module configuration. An examination of the number of leads into a module for signal inputs, power, etc., and the physical size of the module itself is in order. The leads required into each module are: (1) digital input for column position, (2) digital input for row position, (3) reset signal, (4) +5 volts for logic supply, (5) +30 volts for driver supply, following Figure 3-29 module as an example. The ground return is assumed to be through the mechanical mounting of the element. These five supplies were required for each element and each element was 0.35×0.35 inch in size. Notice that of the total number of input pins, only two are unique to each module and the other three would be common.

There are two overriding considerations in the grouping of the module. One consideration is to reduce as much as possible the number of connections to ease the manifolding and dc distribution problem. A second and more important consideration is that a grouping should not be so large that either the size and shape of the array is affected or that more than one size group is required to produce an array of the desired shape. The number of connections influences the reliability of the system and the weight of the power distribution system. The principal choice is then the selection of the largest size group that can be used and not significantly affect the antenna design of even small arrays. A grouping of four fulfilled this criteria for the example module. The linear array of four elements has particular advantage in fabrication. All the phase-shift networks and driver circuits are on the same plane and can be bonded into place by one setting of the equipment and one orientation of the module in the fabrication fixture. The module can then be hermetically sealed by simply soldering on one cap as compared to two, the minimum requirement for the square arrangement. Due primarily to the economic advantages afforded, the linear array of four elements was chosen for the example space-fed module.

3.1.4 Array Techniques

3.1.4.1 Antenna Element

The choice of an antenna element for a phased array is a function of polarization, weight, frequency and form factor. In general, the dipole or loaded slot type radiator is more suitable for the lower radar frequencies whereas the waveguide radiator is usually the optimum element for frequencies above about 10 GHz. Table III-5 lists the basic element types for the radar frequencies discussed in this report and indicates the suitable elements for a particular polarization requirement. A more detailed description of these array radiators is given in Appendix A.

Table III-5. Radiating Element Type

Frequency (GHz)	Linear Pol.	Circular Pol.	Multi. Pol.
1.5	Dipole, slot	Crossed dipole, spiral, helix	Crossed dipole, crossed slots
2.25	Dipole, slot	Spiral, helix, crossed dipole	Crossed dipole, crossed slots
6.0	Dipole, slot	Spiral, helix, crossed dipole, slot-dipole	Crossed dipole, waveguide, crossed slots
9.0	Dipole, waveguide, slot	Spiral helix, slot-dipole, waveguide, crossed dipole	Crossed dipole, waveguide, crossed slots
18.0	Waveguide, slot	Waveguide	Waveguide, crossed slots

Application of these radiators is straightforward for both corporate-fed arrays and reflectarrays. However the transmissive space-fed array has a special requirement, that is, the rf power must pass between the dc control manifold. If the transmissive array is linearly polarized, then slots or dipoles are generally used.

When circular or multipolarization requirements are considered, it is obvious that radiators using almost all of the area of a single-array element must be used. In this case, it will often be easier to feed each transmissive element with a linearly polarized element. This case will occupy less area on the rear of the phase-shifter array, therefore, leaving more space available for the dc manifold and assembly structure. This area per element consideration is particularly a problem at C-band and above. At L- or S-band, the problem is slightly reduced and multipolarized radiators on both ends of the transmissive array element become more feasible.

3.1.4.2 Optimum Matching

Optimum impedance matching is an essential factor in designing phased arrays required to scan far from broadside.¹⁸ In such cases, an isolated element match or even a broadside match can result in an intolerable VSWR at large scan angles. In order to choose an optimum matching impedance, one must analytically predict the impedance versus scan for the central

element of the array. For large arrays, the impedance of the central element can be approximated by that of an infinite array element. Hence, the waveguide simulator can be used as an experimental tool for verification of the optimum match, and in certain cases it can be used in the actual matching of the element. An example will be given to illustrate the technique of optimum matching.

Many factors affect the choice of the optimum matching impedance of a particular array system. The main factor is the maximum allowable VSWR at the maximum scan angle. Consider a phased array with goals of scanning ± 25 -degree E-plane and ± 55 -degree H-plane. In order to scan 60-degree H-plane without incurring excessive VSWR, a judicious choice of matching impedance is necessary. The array will be made up of dual ridged open-end waveguides located on a square grid with half wavelength spacings. In order to choose an optimum matching impedance, the element impedance is assumed to be represented by a dipole in free space (exact dual of a slot). Allen has shown that this assumption is valid for the case of unloaded open-end waveguide elements.¹⁹ Using the impedance variation of the dipole, the optimum impedance was chosen using the technique illustrated in Figure 3-30. The diameters of the constant VSWR circles are determined by two points:

$$[R_g / \text{VSWR} - jX_g] \text{ and } [R_g (\text{VSWR}) - jX_g],$$

where R_g is the generator resistance and X_g is the reactance. Using these simple relations, straightforward procedure is used to determine an optimum matching impedance for a particular system.

First, consider the case in which the array elements are matched in the broadside condition. In this case, a constant VSWR circle is plotted with its diameter determined by the two points; $Z_1 = 76 / \text{VSWR} + j33$, and $Z_2 = 76 (\text{VSWR}) + j33$. Similarly, a constant VSWR circle is shown in Figure 3-30 for a 30-degree H-plane match. For the example of the vertically polarized array, the 30-degree H-plane match may be chosen as optimum for the required maximum scan positions. The following table, III-6, illustrates the maximum scans with the associated maximum mismatches for broadside matched elements, isolated matched elements, and 30-degree H-plane matched elements.

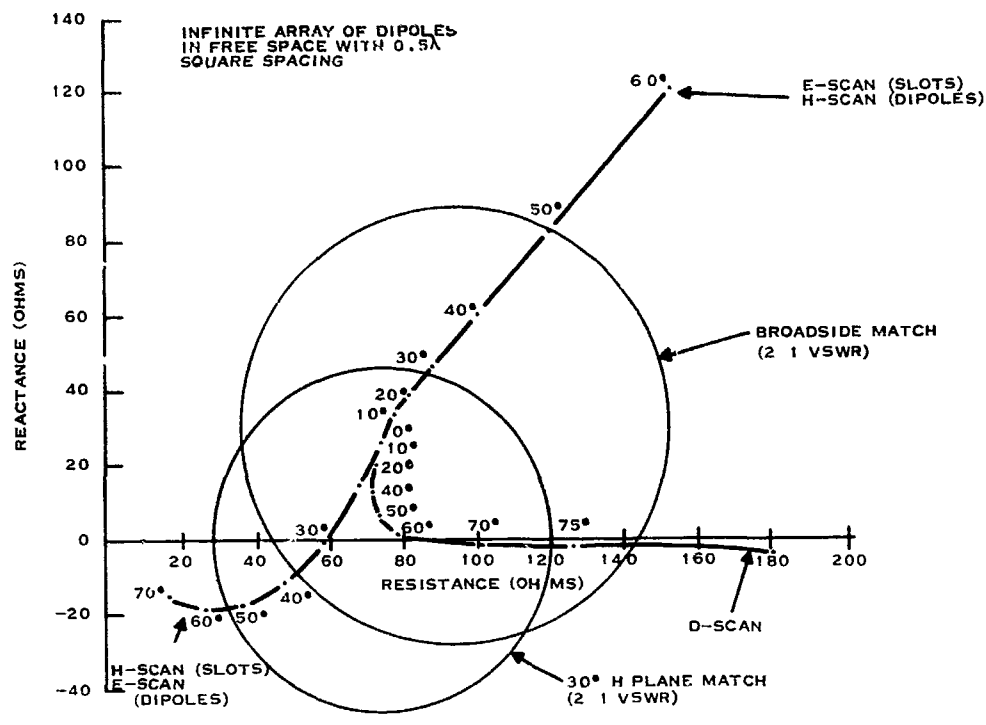


Figure 3-30. Example of H-Plane Optimum Impedance Matching Slots

Table III-6. Maximum Scan Angles of Waveguide Array

VSWR	2:1	3:1	
H-plane	35 degrees	50 degrees	Broadside Match
E-plane	50 degrees	60 degrees	
Diagonal plane	75 degrees	80 degrees	
H-plane	30 degrees	45 degrees	Isolated Match
E-plane	55 degrees	60 degrees	
Diagonal plane	75 degrees	--	
H-plane	52 degrees	65 degrees	30 degree H-plane Match
E-plane	27 degrees	48 degrees	
Diagonal plane	74 degrees	80 degrees	

Considerable improvement is achieved for the H-plane scan at the expense of an addition to the broadside mismatch and some degradation of the E-plane match. This means that the antenna characteristics will be improved for the

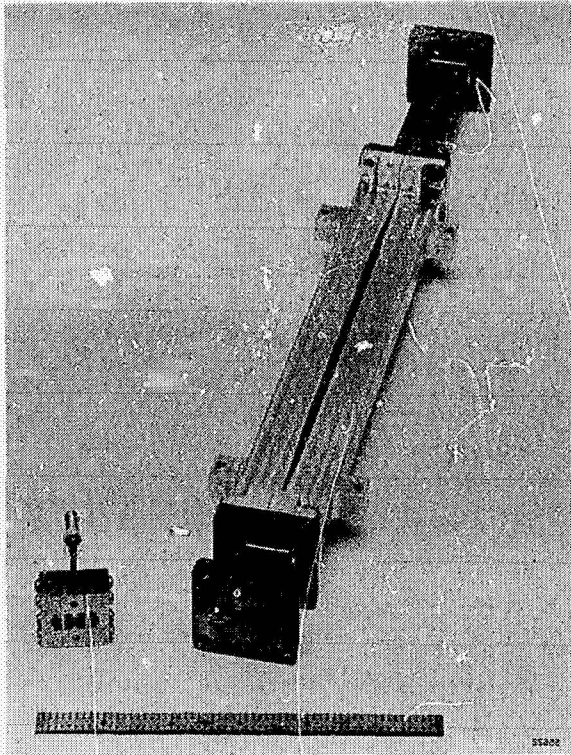


Figure 3-31. 30-Degree H-Plane Simulator

case of scanning over a wide angle in azimuth and over a more limited elevation angle. For other system requirements (conical scan, E-plane scan, etc.), this matching technique would result in a different matching impedance. However, the same trade-off analysis can be used to decrease the VSWR in the desired scan plane at the expense of a slightly increased VSWR in the other scan plane.

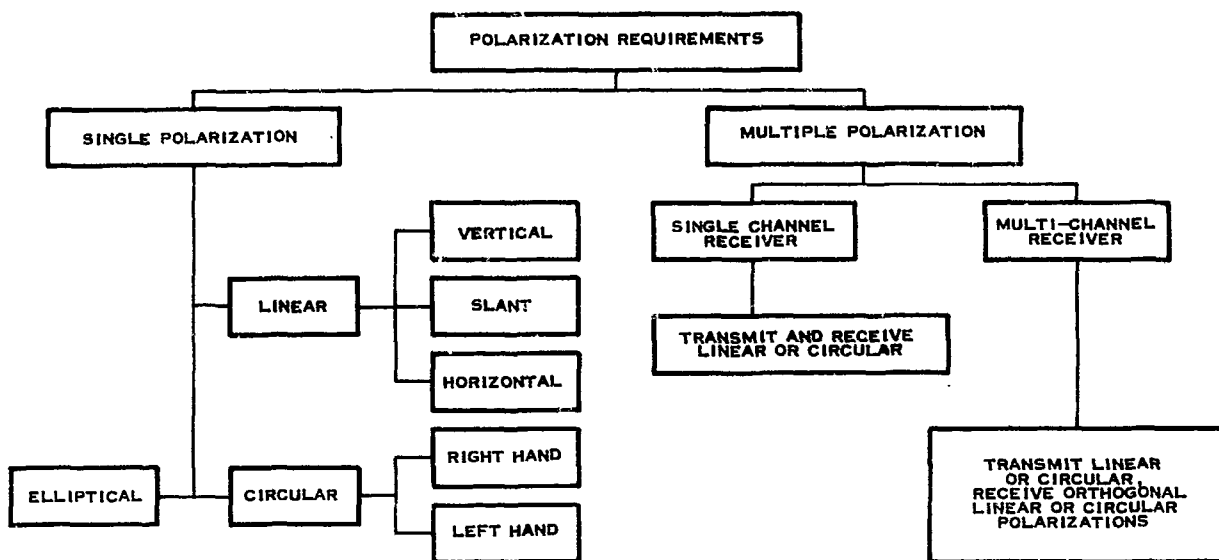
The implementation of an optimum match is generally a complex problem because an equivalent circuit of the element and matching structure must be determined. However, for the 30-degree H-plane scan match it is possible to build an infinite array simulator which can be used to experimentally match the elements. The simulator contains one "whole" and two "half" elements; however, the two half elements cannot propagate and only the whole element acts as an active element. For that reason, the array

simulator can be used quite readily to implement a 30-degree H-plane match. The simulator section shown in Figure 3-31 was used at Texas Instruments to experimentally match a phased-array element to the optimum matching impedance for a similar array.

3.1.4.3 Polarization Requirements and Implementation

The purpose of a radar usually defines the polarization of the antenna. The most basic single polarization antenna is linearly polarized, followed by circular polarization. These antenna polarization features are illustrated in Figure 3-32. Multiple polarization finds application for radars such as the proposed CAT radar.

Multipolarizations are generated with a combination of antenna radiating elements and other components such as hybrids, diode or ferrite phasors, and ferrite circulators. When orthogonal polarizations are desired, two-antenna-element ports can be used to distinguish the two polarizations. Multipolarized antennas can have single or multichannel receivers and are capable of transmitting and receiving more than a single polarization. For



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Figure 3-32. Polarization Requirements

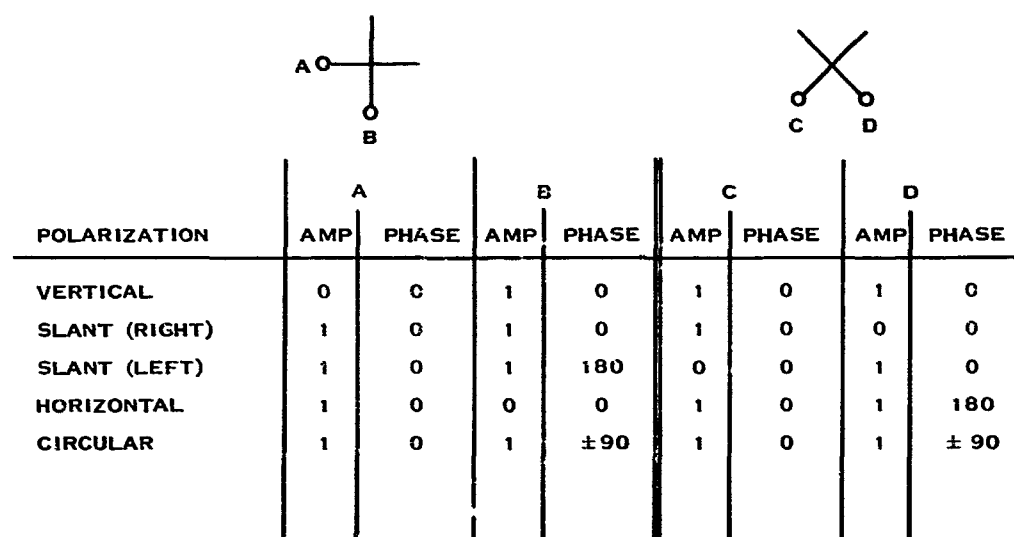
instance, the target can be illuminated with circular polarization, with both orthogonal linear polarizations detected in the receive mode for target cross-section polarization analysis. If a single receiver is used, it can be switched between linear polarizations during the receive mode. If dual receivers are used, then both polarizations can be detected simultaneously.

Figure 3-33 illustrates two types of crossed linearly polarized probes which may be used to radiate various polarizations. The table in Figure 3-33 indicates the required phase and amplitude necessary for a particular polarization condition. The probes shown represent various types of linearly polarized antenna elements, such as dipoles or open-end waveguides.

3.1.4.3.1 Reflectarray Polarization Diversity

Space-fed phased-array polarization diversity may be achieved in either the radiating module or the feed structure. The exact design of the space-fed array components will involve tradeoffs between the feedhorn configuration and the module. If the feed structure is used for polarization implementation, the modules must be polarization insensitive. If module switching is used for polarization diversity, only a single polarization is required at the feedhorn.

Assuming the use of polarization-insensitive elements, Figure 3-34 illustrates some of the possible feed polarization circuits which indicate both a transmit and receive polarization. Using combinations of quarter-



63810

Figure 3-33. Probe Orientation Versus Transmitted Polarization

wave plates, orthomode transducers, hybrids, circulators, switches, and phase shifters, a number of polarization schemes are available. Advantages of feed polarization schemes include a minimum amount of additional polarization logic and easier implementation. Disadvantages center around the requirement of switching higher power rf devices, resulting in rf power handling problems and inherent lower reliability.

Figure 3-35 gives an example of polarization diversity achieved in the module using just a vertically polarized feed structure. Figure 3-35a illustrates the characteristics of a 90-degree hybrid (or branch-line) coupler. Rf power entering one port experiences a 3-dB power split and the relative phase is advanced 90 degrees in one leg. If both right-hand legs are shorted, all the power entering the lower left-hand leg will be routed out the upper left-hand leg. Figure 3-35b shows this effect of shorted legs and illustrates the module operation for a vertically polarized feedhorn. The array element essentially transforms the vertical polarization to horizontal polarization when the phase shifters are set to identical phase states. Likewise, on the receive cycle, the returning horizontal polarization will be converted back to vertical polarization for the feedhorn.

If the relative phase between phase shifters in the two legs is different by 180 degrees, then the operation of the hybrid network is just the opposite. As shown in Figure 3-35c, all the power entering one leg is returned

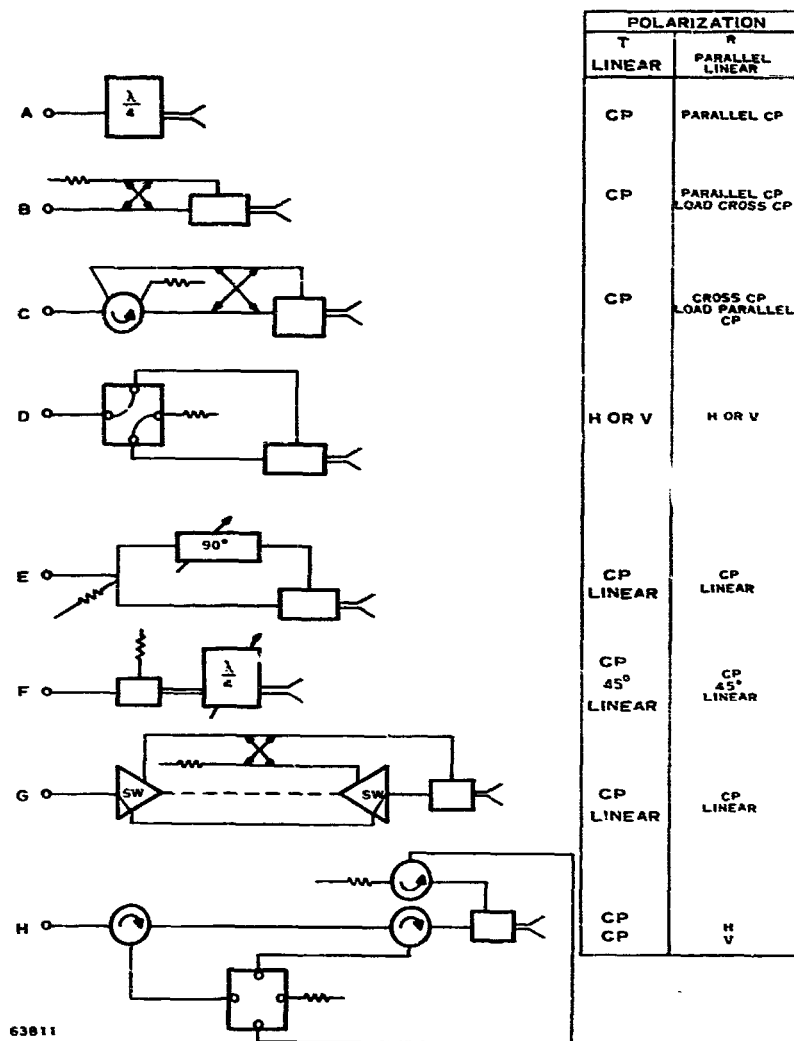


Figure 3-34. Feed Polarization Circuits

to the same leg and can be used to transmit the same-sense polarization as illustrated.

Circular polarization can be generated using the same network. Circular polarization can be considered to consist of two equal linearly polarized vectors, orthogonal in both space and time. Figure 3-35d illustrates the use of the hybrid/phase-shift pair to provide a 3-dB power split. When there is a relative phase difference of 90 degrees between the two hybrid legs, there will be an equal power split on the returning signal. The orthogonal waveguide probes shown exciting the TE_{01} and TE_{10} modes provide the proper space orientation of the transmitted signal, and the extra $\lambda/4$ (90 degrees) line length in the horizontal polarization leg provides the required 90-degree time difference necessary for circular polarization. Of course, the device is reciprocal and the returning circular polarization is converted back to the vertical polarization transmitted by the feed structure. The placement of the extra $\lambda/4$ line length in the module does not affect the vertical or horizontal

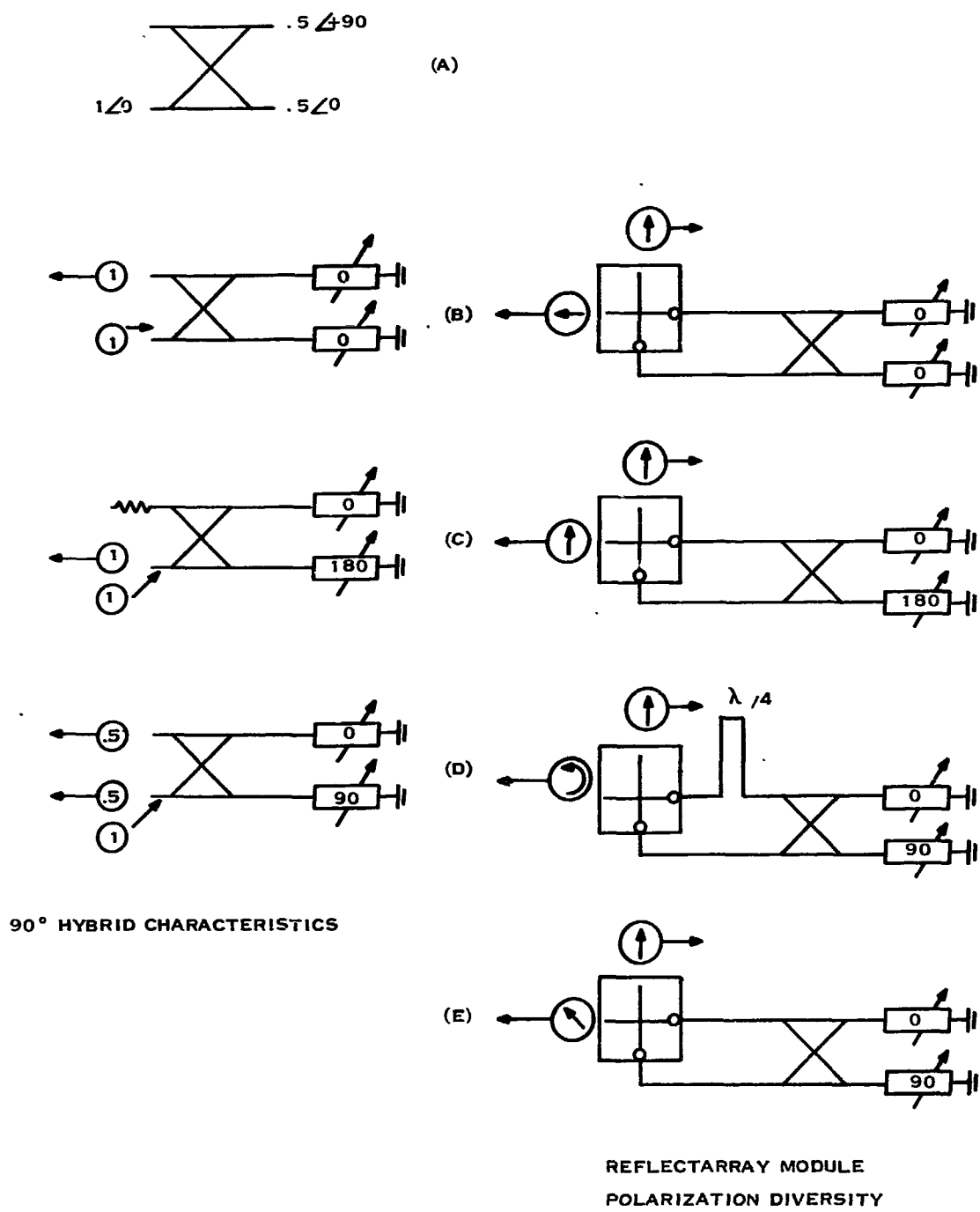


Figure 3-35. Reflectarray Module Polarization Diversity

polarization modes as discussed above. Figure 3-35e points out that the transmitted polarization will be 45-degree-slant linear for the case of 90-degrees phase difference (between phase shifters) without the extra $\lambda/4$ line length inserted. Obviously, various orientations of linear polarization, as well as elliptical polarizations, are available for other values of phase shifter difference phases.

3.1.4.3.2 Rear-Fed Element Polarization Diversity

A rear-fed element can be either a transmissive space-fed array element or a corporate-fed array element. Compared to the reflect-array element just discussed, the rear-fed element offers more methods to achieve polarization diversity, since additional processing may be used.

Several polarization schemes are given in Table III-7 with a the processing required for each. The first configuration shown, (A) is for a case of transmitting and receiving the same linear polarization. This scheme uses the simplest antenna element and processing. The next case (B) is the transmission and reception of the same sense of circular polarization. This polarization scheme is used to detect complex targets which give an even number of reflections or to suppress returns from symmetrical targets such as rain clutter suppression for terrestrial systems. The next configuration (C) will do the opposite of (B), that is, it will enhance symmetrical targets and reject even number bounce targets. The reason for this enhancement or rejection is that the sense of circular polarization is changed for each reflection. Thus, a single reflection return will be in the wrong sense of circular polarization to excite the antenna.

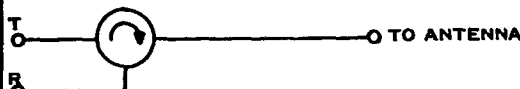
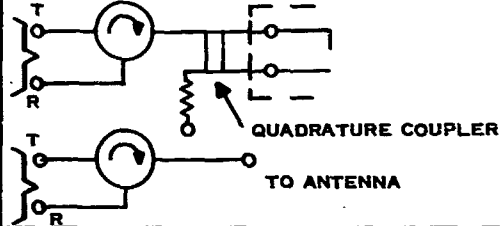
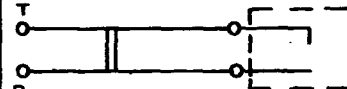
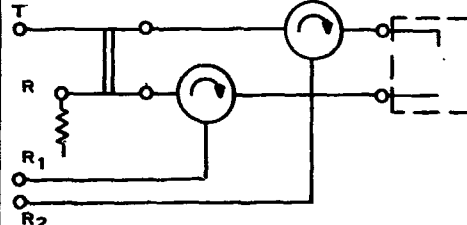
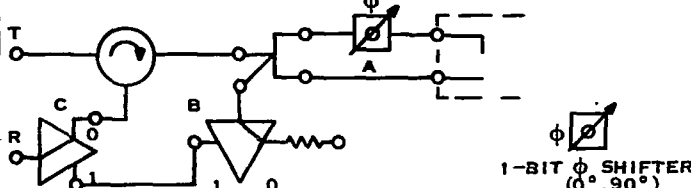
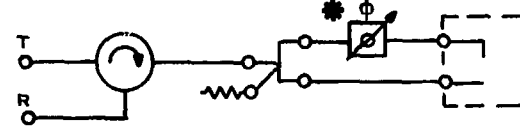
The polarization configuration (D) is useful for polarization classification of targets, but this feature is accomplished at the expense of more lossy elements in the processing, plus two receive channels. The configuration (E) is included to show the increase in complexity of the processing when several polarizations are required.

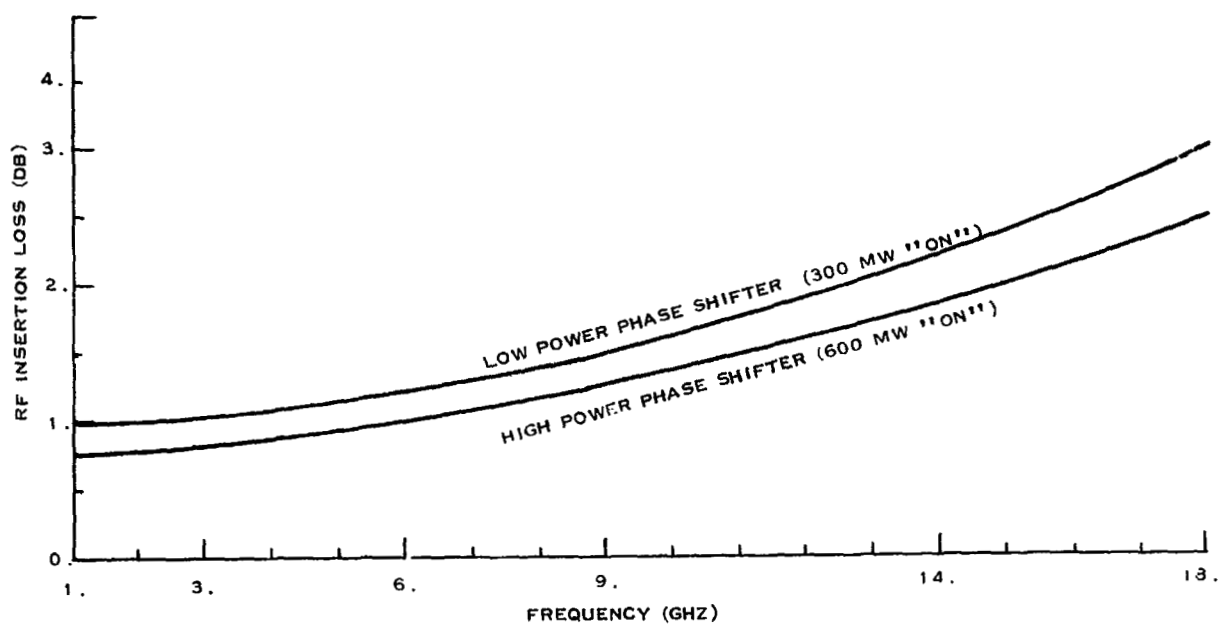
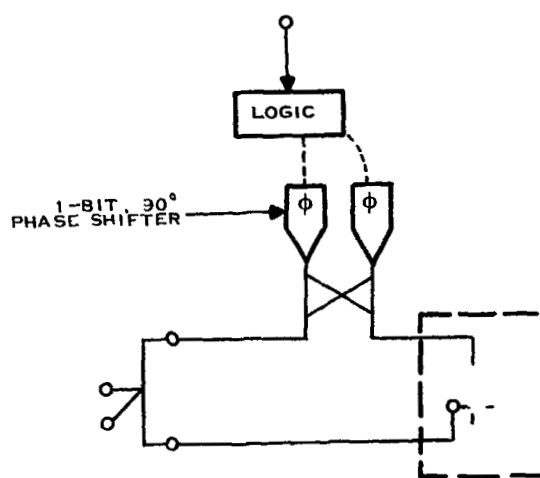
Configuration (F) is similar to (B) and (E). A quadrature coupler plus a 1-bit (90-degree) phase shifter is used to generate either circular or linear polarization.

A major disadvantage of polarization diversity is that extra system losses are introduced, as well as requirements for additional dc power in most cases. In a space radar, where available power is at a premium, polarization diversity may be considered a disadvantage except in the particular case where mission requirements define the use of variable polarization.

Figure 3-36 gives an example of using a one-bit, 90-degree, hybrid-type diode phase shifter to achieve polarization diversity. If higher dc powers are used to drive this type of phase shifter, lower rf insertion loss

Table III-7. Possible Polarization Configurations

POSSIBLE POLARIZATION CONFIGURATIONS																													
	TRANSMIT	RECEIVE	ANTENNA	PROCESSING																									
A	LINEAR	LINEAR	1. DIPOLE 2. SLOT 3. WAVEGUIDE																										
B	CIRCULAR	CIRCULAR	1. CROSSED DIPOLE 2. WAVEGUIDE 3. SPIRAL 4. HELIX																										
C	CIRCULAR	OPPOSITE SENSE CIRCULAR	1. CROSSED DIPOLE 2. WAVEGUIDE																										
D	CIRCULAR	ORTHOGONAL LINEARS	1. CROSSED DIPOLE 2. WAVEGUIDE																										
E	1. LINEAR 2. CIRCULAR	1. LINEAR 2., 3., BOTH SENSES OF CIRCULAR	1. CROSSED DIPOLE 2. WAVEGUIDE	 <table><tr><th colspan="2">POLARIZATION</th><th colspan="3">LOGIC</th></tr><tr><th>TRANSMIT</th><th>RECEIVE</th><th>A</th><th>B</th><th>C</th></tr><tr><td>1. LINEAR</td><td>LINEAR</td><td>0°</td><td>0</td><td>0</td></tr><tr><td>2. CIRCULAR</td><td>CIRCULAR</td><td>90°</td><td>0</td><td>0</td></tr><tr><td>3. CIRCULAR</td><td>ORTHOGONAL CIRCULAR</td><td>90°</td><td>1</td><td>1</td></tr></table>	POLARIZATION		LOGIC			TRANSMIT	RECEIVE	A	B	C	1. LINEAR	LINEAR	0°	0	0	2. CIRCULAR	CIRCULAR	90°	0	0	3. CIRCULAR	ORTHOGONAL CIRCULAR	90°	1	1
POLARIZATION		LOGIC																											
TRANSMIT	RECEIVE	A	B	C																									
1. LINEAR	LINEAR	0°	0	0																									
2. CIRCULAR	CIRCULAR	90°	0	0																									
3. CIRCULAR	ORTHOGONAL CIRCULAR	90°	1	1																									
F	1. LINEAR 2. CIRCULAR	1. LINEAR 2. CIRCULAR	1. CROSSED DIPOLE 2. WAVEGUIDE	 <table><tr><th colspan="2">POLARIZATION</th><th>* LOGIC</th></tr><tr><td>1. LINEAR</td><td></td><td>0</td></tr><tr><td>2. CIRCULAR</td><td></td><td>90°</td></tr></table>	POLARIZATION		* LOGIC	1. LINEAR		0	2. CIRCULAR		90°																
POLARIZATION		* LOGIC																											
1. LINEAR		0																											
2. CIRCULAR		90°																											



63813

Figure 3-36. Added Insertion Loss for Polarizer

will result. Represented are dc drive powers of 300 milliwatts and 600 milliwatts "ON" and 60 milliwatts for the "OFF" condition. Rf insertion loss varies from 0.75-1.0 dB at 1.5 GHz to 2.5-3.0 dB at 18 GHz. This example illustrates the reason for considering a linearly polarized radar, unless mission requirements specify the option of polarization diversity.

3.1.4.4 Array Gain Versus Number of Elements

The array gain (G) may be related to the number of elements (N) by the following equation:

$$G(\theta, \phi) = \Pi N \cdot \eta,$$

where η is the array efficiency.

This equation may be derived from Equation (3-5) assuming a broadside matched element at broadside scan where

$$A = N \cdot a \text{ and } a = \frac{\lambda^2}{4},$$

that is, the area allotted to each element is $\frac{\lambda^2}{4}$ or an equivalent element spacing of $\lambda/2$ on a square grid. The array efficiency accounts for ohmic losses in the elements, power in sidelobes due to less than perfect collimation, and taper efficiency. The g in versus number of elements for $\eta = 0.80$ in a corporate-fed array is plotted in Figure 3-37. The gain versus number of array elements for a space-fed array is also shown in Figure 3-37 as the curve for $\eta = 0.55$. The lower efficiency for the space-fed array is attributed to reduced aperture efficiency. This is caused by the array being illuminated by a horn which includes spillover losses.

3.1.4.5 Gain Variation with Scan

One of the limitations of planar phased arrays is the variation of gain with scan. The gain variation can be attributed to mismatch loss and change in the apparent antenna aperture shape. The directivity of a planar array with an aperture area A scanned to the angle θ from broadside is

$$U(\theta) = \frac{4\pi A}{\lambda^2} \cos \theta \cdot \eta, \quad (3-4)$$

where η is the taper efficiency on the array. A spherical coordinate system is used with the array in the x-y plane.

The array gain for independently fed elements with no ohmic losses and including mismatch losses is

$$G(\theta, \phi) = \frac{4\pi A \eta}{\lambda^2} \cos \theta \left[1 - |\Gamma(\theta, \phi)|^2 \right], \quad (3-5)$$

where $\Gamma(\theta, \phi)$ is the voltage reflection coefficient which is a function of θ and ϕ , that is, scan angle and scan plane. For a planar array with N elements each having an area, a , the element gain function is given by

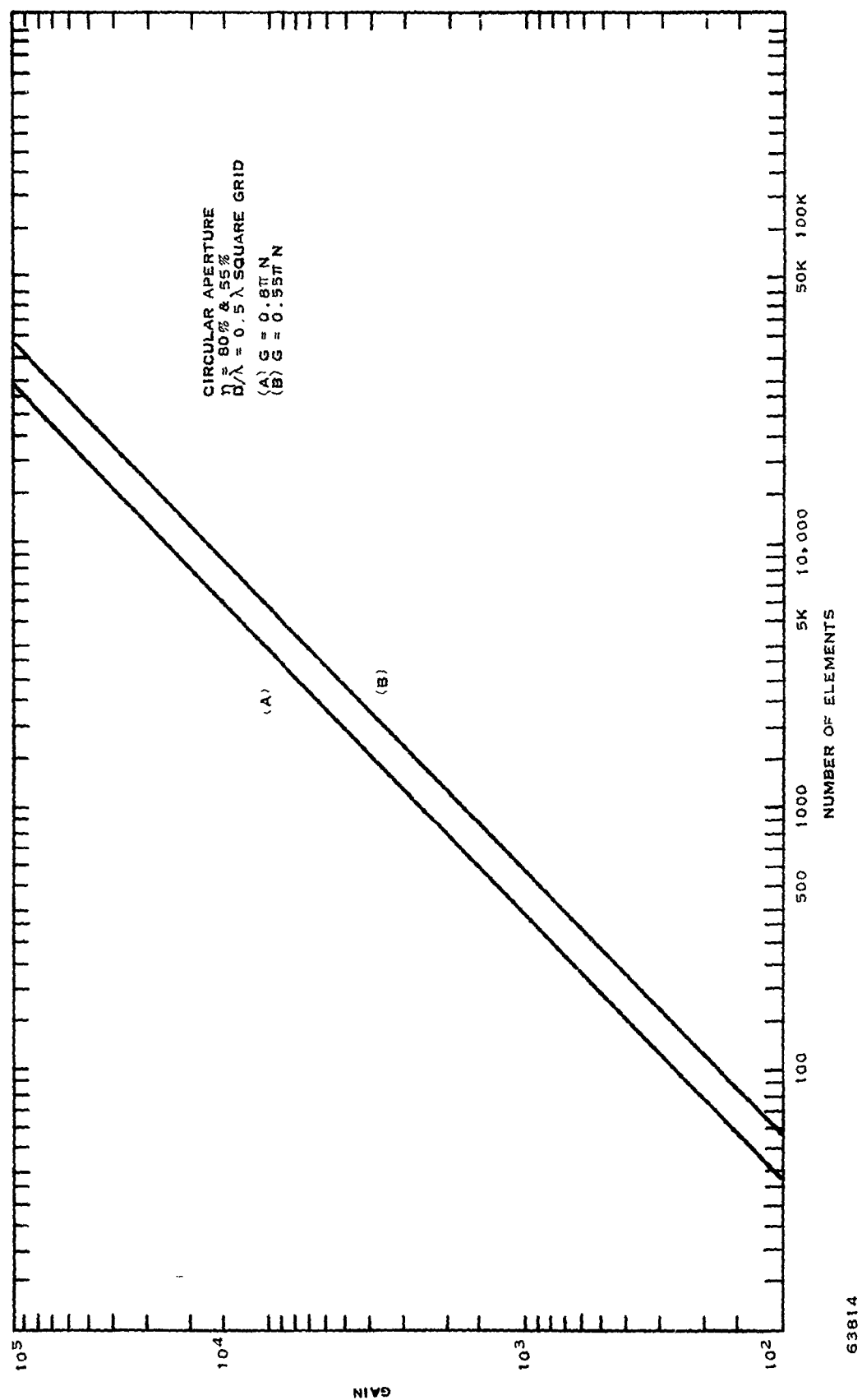


Figure 3-37. Circular Aperture

$$g(\theta, \phi) = \frac{4\pi a}{\lambda^2} \cos \theta [1 - |\Gamma(\theta, \phi)|^2] \quad (3-6)$$

and the array gain is

$$G(\theta, \phi) = g(\theta, \phi) N \cdot \eta. \quad (3-7)$$

The only place that the contributions due to a particular type of element come into play is in $\Gamma(\theta, \phi)$.²⁰ Thus different elements will modify the array gain variation differently through mutual coupling. This analysis assumes identical elements with identical impedance variations.

The gain variation for two match conditions for an array of 0.5λ spaced slots is given in Figure 3-38. The first case (A) is for a broadside scan match and the other (B) is for a 30-degree H-plane scan match.

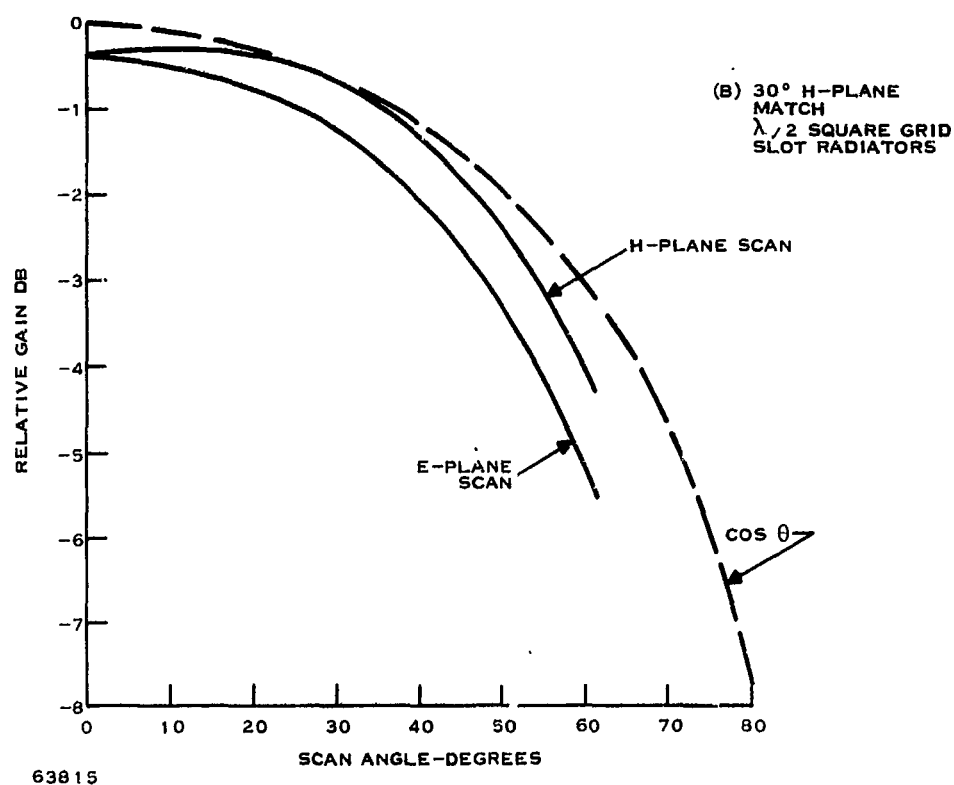
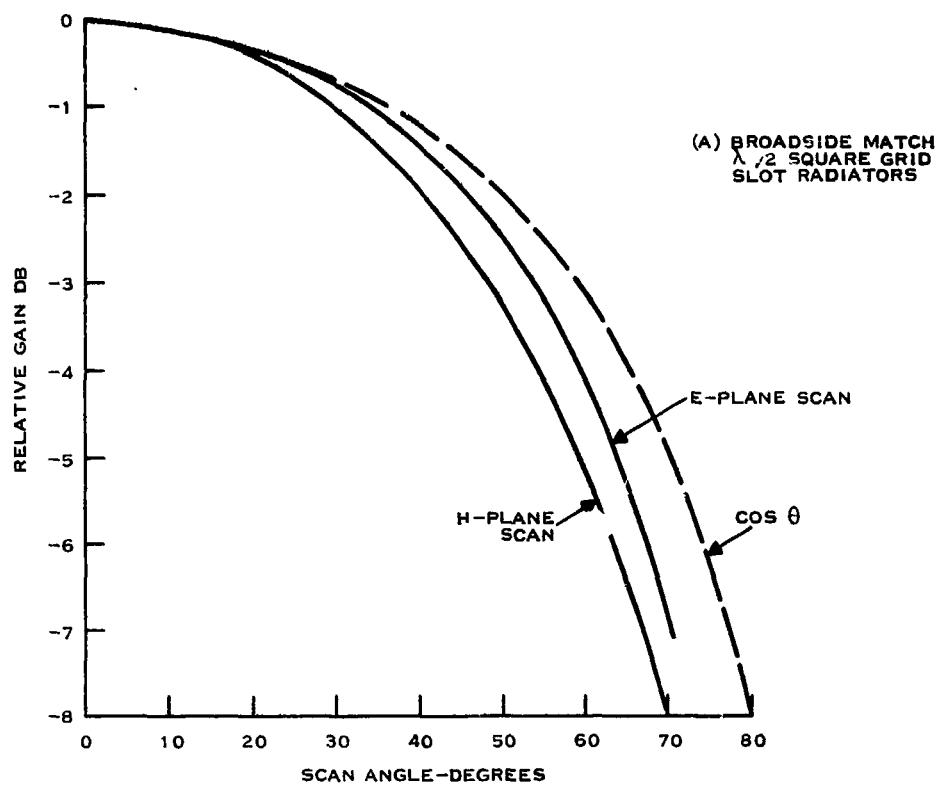
3.1.4.6 Beamwidth Variation with Scan

The beamwidth of a planar phased array will also vary as the array is scanned. This again is caused by a variation of apparent aperture size. The beamwidth will vary as $\frac{1}{\cos \theta}$ or $\sec \theta$, where θ is scan angle from broadside. The variation is shown graphically in Figure 3-39a.²¹ Figure 3-39b shows both the beamwidth and gain variation with scan as actually measured in a phased array.²²

Figure 3-40 illustrates the results from a linear array of dipoles scanned in the H-plane. This array, tested at Texas Instruments, used 4-bit digital diode phase shifters. Formation of a grating lobe at scan angles greater than 45 degrees caused a sudden loss of gain as illustrated in the figure. Gain variation was less than 1 dB for a ± 45 -degree scan angle.

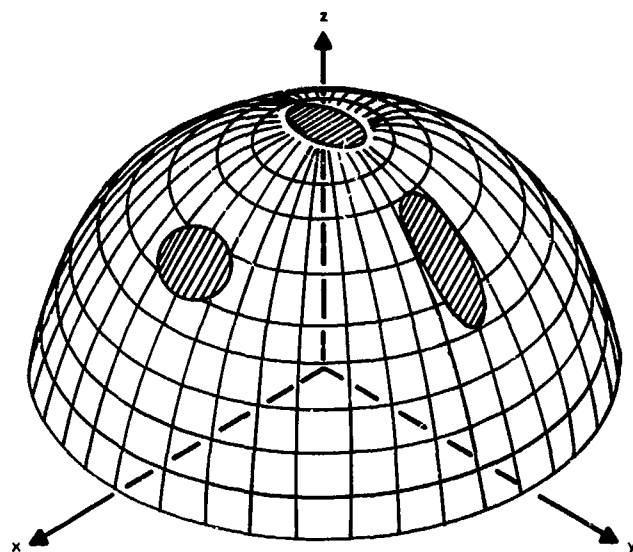
3.1.4.7 Beam-Pointing Error

A beam-pointing error is associated with planar phased arrays which is a function of scan angle but does not include the pointing error introduced by phase quantization. The pointing error is a distortion of the radiated pattern caused by the array element pattern. The radiated pattern is a product of the array pattern and the active element pattern; therefore, the slope of the element pattern will introduce a beam-pointing error as illustrated in Figure 3-41.²³ The pointing error is predictable and can be compensated if it is undesirably high. If the element pattern is circularly symmetric, the error is a function of scan angle (θ). But, if as is usually the case, the element pattern is also a function of ϕ , the correction factor becomes much more complex. The term element pattern is the radiation pattern of the array element in the mutual coupling environment of the phased array being considered. This element pattern is in general different from the element radiation pattern of an isolated element, that is, not in the array environment.

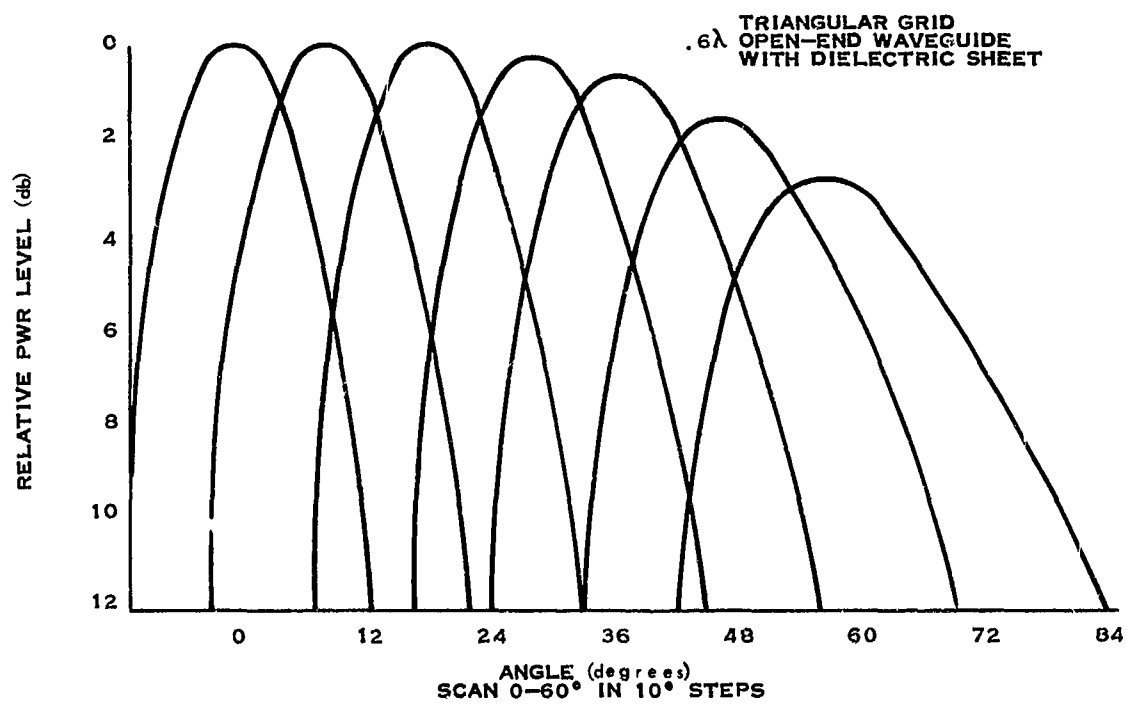


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Figure 3-38. Gain Variation with Scan



(a) BEAMWIDTH VARIATION WITH SCAN



(b) GAIN AND BEAMWIDTH VARIATION WITH SCAN

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Figure 3-39. Gain and Beamwidth Variation with Scan

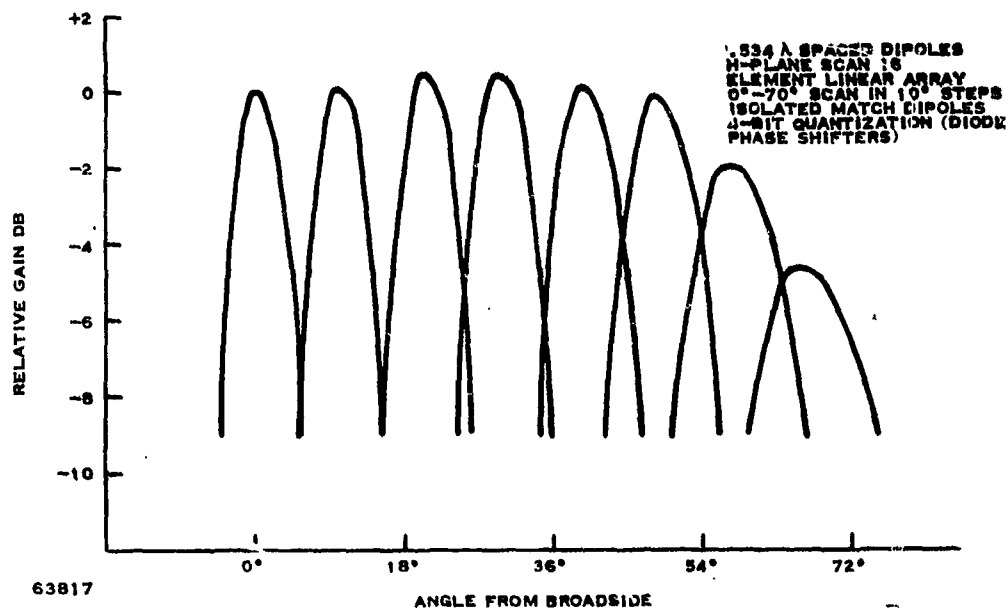


Figure 3-40. Linear Array of Dipoles Scanned in the H Plane

3.1.5 Plasma Effects

There are two plasma effects associated with the space radar, one dealing with the target vehicle and the other with the radar. The target vehicle effect includes the change of the target radar cross section due to a plasma sheath around the vehicle and an ion trail. The second effect is the buildup of a plasma sheath about the radar, thus modifying the radar performance.

3.1.5.1 Target Radar Cross-Section Change

A radar backscatter cross section approaching 100 m^2 has been calculated for satellites with a 1-meter radius and a 3.2-MHz radar frequency.^{24, 25} The plasma frequency for the above calculation was 1.6 MHz and the velocity vector of the target was parallel to the local geomagnetic field which gives the maximum cross section. The vehicle velocity is assumed much greater than the ion thermal velocity and no particles are being introduced due to the vehicle propulsion system.

The vehicle moving through the ionosphere will sweep out the ions in its path, thus leaving a trail that is mildly depleted of electrons. The electron depletion will give a gradient behind the vehicle which will give a radar return at low frequencies. This trail differs from the plasma encountered

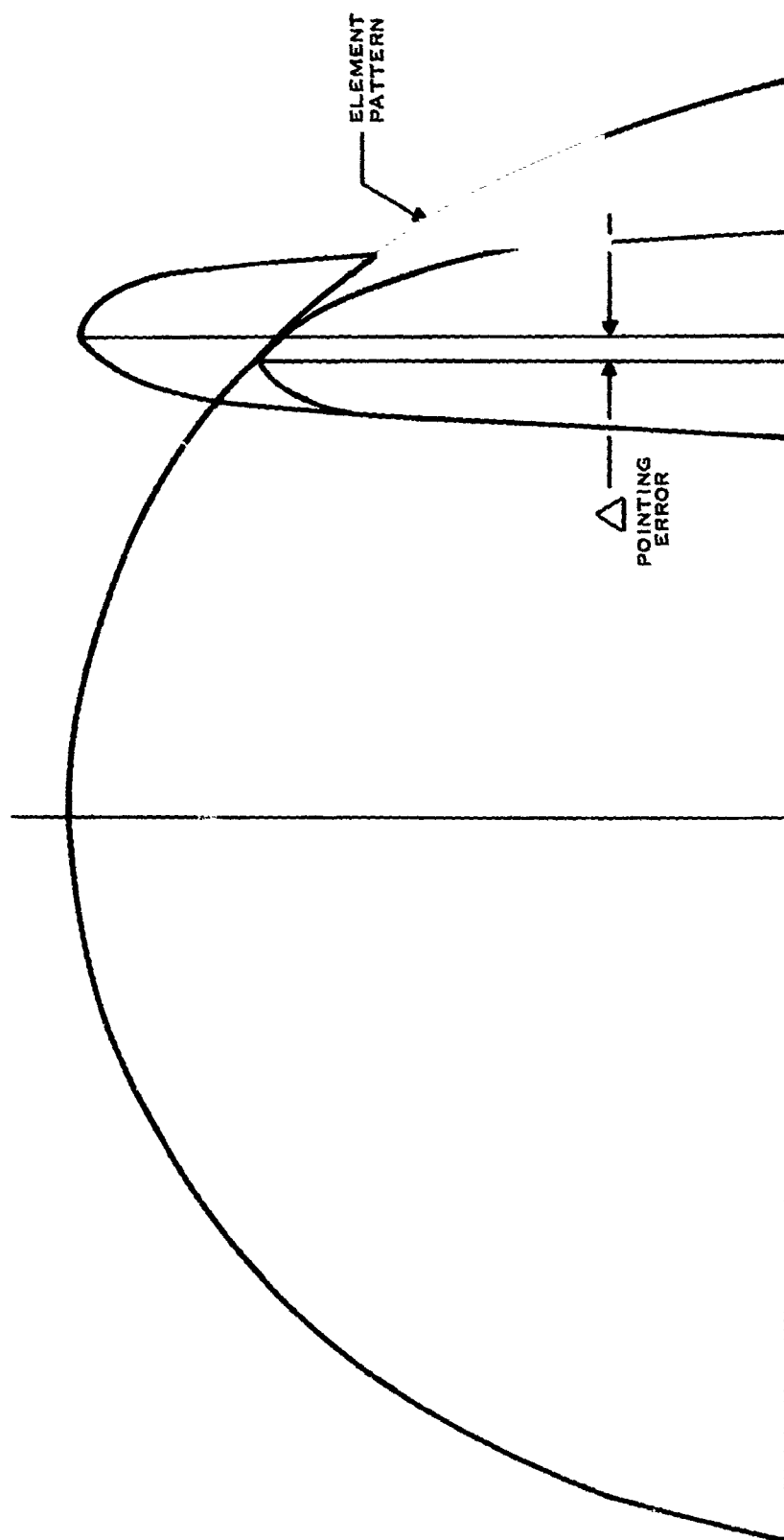


Figure 3-41. Beam-Pointing Error

on reentry, caused by a boundary layer of high electron density around the vehicle resulting from the high reentry temperatures.

This cross-section enhancement is a function of radar frequency. The calculations made in the referenced literature were for frequencies in the H-F range which are considerably below the lowest CAT radar frequency of 1.5 GHz. The backscatter varies at higher frequencies by the reciprocal of frequency to the fifth power as shown by Shea.²⁶ Therefore, the radar cross-section enhancement by the ion trail is negligible at 1.5 GHz and above.

3.1.5.2 Plasma Sheath Effects on Radar Performance

The possible problems to be encountered by a space radar can be divided into two groups: (1) transmission through the sheath and (2) reflection from the sheath. The first group (1) will cause a reduction in transmitted and received power and a distortion of the radiated field. The second group (2) will effectively change the array environment. That is, the mutual coupling will be changed and unwanted power will be coupled back in to the radar, possibly causing decreased array efficiency and radiated beam distortion.

As was discussed for the backscatter problem since the frequency will be above 1.5 GHz, there should be no problem with high attenuation or reflections of large amounts of power by the ion sheath about the radar. A problem may exist with the near field effects of the plasma changing the mutual coupling encountered in the array. The first order effect would be to change the active impedance of the array elements. This can be considered as additional power being reflected back into the array, thus giving a loss in transmitted power. With proper use of circulators and TR switches behind the antenna this should not be a serious problem, being no different, from the module standpoint, than the problem associated with impedance variation with scan.

To more fully evaluate the radar performance in the plasma environment, more information is required on the electron density and other facts on the expected plasma sheath. The area of plasma effects on phased-array space radar performance is recommended for additional study.

3.2 Array Components Analysis

3.2.1 General Array Description

Until recently, small electronically steerable antenna arrays generally used the concept of illuminating an array of phase shifters with a high-power transmitter source and steering the antenna beam by varying the phase-shifter elements. Other techniques used were time-delay scanning or frequency scanning.²⁷

With the development of the concept of using redundant active elements in the array which performed transmitter, receiver and phase-shifter functions (as used in the Texas Instruments MERA radar), the arrays were able to take on quite different configurations. A simplified block diagram of both space-fed and active-element arrays is shown in Figure 3-42. The space-fed phase-scanned antenna array can be either front-fed (reflective) or rear-fed (transmissive). The antenna and feed considerations of this array have been discussed in Subsection 3.1. The active-element array uses a number of low-power, all solid-state, rf modules in parallel and is fed by a low-power exciter. An analysis of the components that make up each of these arrays is given in the remainder of this section with the purpose in mind of reviewing the state-of-the-art and making tradeoff designs that can be used to synthesize a number of radar systems for the CAT radar mission.

3.2.2 Active-Array Components Analysis

3.2.2.1 Introduction

In the preceding section two basic types of electronically scanned phased arrays were described. One, the space-fed array, utilized passive phase-shift elements and a microwave tube transmitter. The other, a corporate-fed array, is an all-solid-state system made up of a number of individual active transmitter/receiver elements (rf modules) and is similar to the MERA radar presently under development on Contract AF 33(615)-2525.

Results to date on MERA have been highly promising. More recent programs, both at Texas Instruments^{28, 29, 30, 31, 32, 33} and other laboratories, ^{34, 35, 36, 37, 38, 39} have further extended the capabilities of microwave integrated circuits. The application of these devices in future phased-array radars and other advanced systems is assured. In this section, the capabilities of microwave integrated circuits are reviewed, and the requirements of active-array elements from 1.5 to 18 GHz for the CAT radar are developed.

The performance numbers are considered neither optimistic nor conservative, but are about what can be done if a system development program were to begin in mid-1968. Since the state of the art is advancing so rapidly, the analysis could be changed at any time. To some extent, near-future capabilities are thus covered. Gunn, LSA, and avalanche transit-time devices are among the existing new prospects. Because of the time-scale assumed, the stringent CAT radar requirements (>200-nm range) and the state of development of these new devices, the transmitter amplifier or amplifier/frequency multiplier chain is considered to be the most efficient and reliable solid-state transmitter power source at this time.

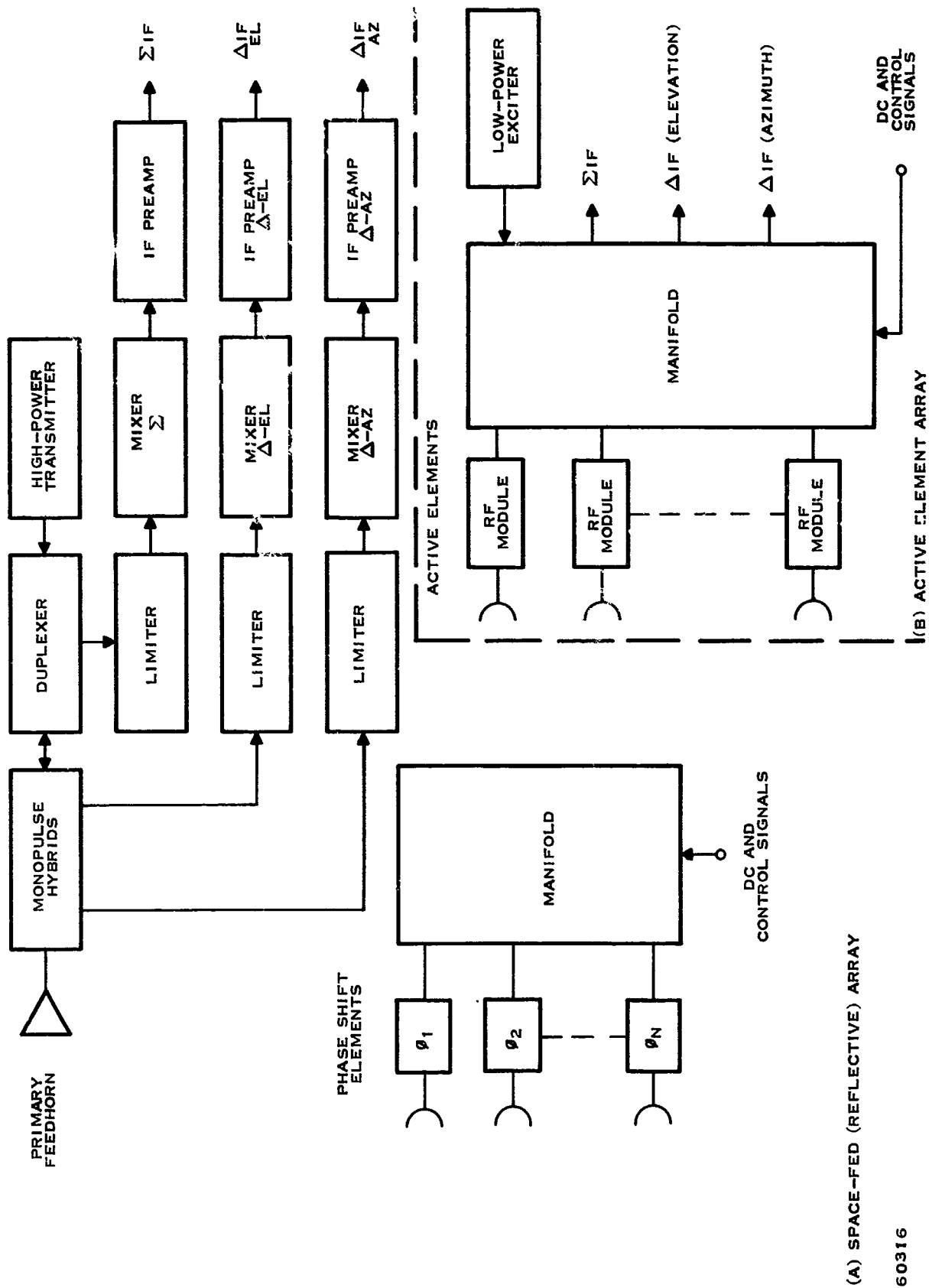


Figure 3-42. Simplified Block Diagram of Space-Fed and Active-Element Arrays

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Recent developments in microwave integrated circuits have led to the application of these devices in advanced systems. Of particular significance to this, and future multifunction radar programs, is the MERA radar presently under development on Contract No. AF 33(615)-2525. The major technological impact of MERA has been the development of microwave integrated circuits required in active phased-array radars.

In addition, work at Texas Instruments has begun on second-generation microwave integrated circuits aimed at upgrading the performance capabilities of the MERA devices. Of considerable interest is a program to develop a 2-watt integrated-circuit X-band transmitter element which includes a transistor power amplifier and a varactor diode multiplier. Parallel effort on this same program is directed toward achieving 10 watts, at X-band, from an integrated injection-locked avalanche transit-time silicon-diode oscillator. In the receiver area, an ultra-low-noise mixer/IF preamplifier has been developed, utilizing advanced circuit techniques and new low-noise transistors. A significant improvement in overall system noise figure is projected for future systems.

Because of the significance of MERA technology in the CAT radar, this program is reviewed briefly in the following paragraphs.

(a) MERA System Description

The MERA radar is an all-solid-state, X-band electronically scanned phased-array radar with approximately 600 active transmitter/receiver elements. As shown in Figure 3-43, the system consists of six units, or "blackboxes," in addition to an array structure and a rack of commercial power supplies. The array structure includes RF, DC, cooling-air distribution manifolds, and 604 plug-in RF modules. The system provides the functions of terrain following, terrain avoidance, ground mapping, and air-to-ground ranging.

Each RF module in the array transmits and receives energy at 9.0 GHz. Transmitter excitation and local oscillator power are uniformly distributed to the modules from the signal generator by the RF distribution manifold. The signal generator output is at S-band, and the necessary frequency multiplication (to 9.0 GHz) on transmit is accomplished in the RF modules. The modules also multiply the frequency of the local oscillator input from 2.125 GHz to 8.500 GHz and down-convert the 9.0 GHz received signal to a 500-MHz IF.

The MERA system utilizes pulse compression techniques to achieve adequate sensitivity and good range resolution with all-solid-state devices. The transmitter excitation signals developed in the signal generator



Figure 3-43. MERA Radar System Block Diagram

permit chirped operation; or operation in an interlaced mode, combining chirped transmission with nonchirped transmission. This permits both close-in coverage and far-range coverage with good range resolution in terrain-following, terrain-avoidance, ground-mapping, and air-to-ground ranging modes.

The 500-MHz if. outputs from the 604 array elements are combined in the rf distribution manifold in accordance with a specified weighted distribution in order to achieve good sidelobe characteristics. The manifold also contains a hybrid comparator network, used to form the vector sum and difference of the combined 500-MHz outputs from the four array quartersections, as required for an elevation monopulse system. The sum and difference signals are fed to the receiver/processor where additional amplification at 500 MHz is provided. A second down-conversion from 500 MHz to 60 MHz is accomplished in the receiver sum-and-difference channels to accommodate the pulse compression filters. The local oscillator signal required for this mixing action is developed in the signal generator.

The receiver/processor also contains monopulse processing circuitry to develop MRI video necessary for the generation of terrain-following commands and for presentation on an E2 display. This unit also produces the amplitude-detected sum video and bipolar difference video required in the range computer. Selected video, consisting of MRI video, sum video, difference video, or a range strobe (depending upon the operating mode), is supplied to the indicator.

The indicator provides presentation on a 5-inch direct-view storage tube (DVST) of either an E2 display (TF) or a sector PPI display (TA, GMP, GMS). In AGR operation, the display consists of a vertical sweep, intensity modulated by a range strobe pulse. The basic indicator, a standard AN/APQ-110 unit, is modified to be compatible with the MERA range, scan, prf, and mode requirements. All sweeps are linear with the exception of the E2 sweep for terrain following.

The terrain-following computer (part of the TF/range computer unit) receives MRI video from the receiver/processor and aircraft flight data (angle of attack, airspeed, pitch) from other aircraft subsystems.

These inputs are used with elevation scan angle information from the synchronizer/scan computer to generate the climb-dive commands necessary for TF operation. The command outputs are supplied to the autopilot for automatic terrain following and to a command indicator (such as ARU-11A) for manual terrain following.

In addition, template video is supplied to the indicator for use as a reference command for the E2 presentation. During flight over areas of low radar reflectivity, where radar returns are below threshold, climb-dive commands are generated from inputs provided by the radar altimeter.

The air-to-ground ranging computer is also located in the TF/range/computer unit. The tracking circuitry requires inputs of sum video and bipolar difference video. Outputs are a video strobe providing a time representation of range for display on the indicator, a dc range analog voltage for use in a weapons release computer, and a lock-on voltage to provide indication that the tracking circuits are engaged.

Primary system timing and gating functions are generated in the synchronizer/scan computer. The basic timing reference is a 10.24-MHz master clock. In addition, this unit develops the scan voltages required by the terrain-following computer and the indicator, providing stabilized outputs as determined by inputs from the various aircraft subsystems and the particular operating mode. The synchronizer/scan computer also generates the phase-shift information required for scanning the beam. The proper phase shift for a particular beam position is determined for each element in the array (by row and column) and routed to the individual array elements in the form of discrete trains of clock pulses. Each rf module contains an OR circuit which accepts the row and column inputs. The output of the OR circuit is fed to a four-stage ripple-through counter which, in turn, controls four digital phase-shift elements. The phase-shift elements are up-dated by the row and column inputs on a pulse-to-pulse basis.

The remaining two units are the control and a rack of dc power supplies. The control provides the switching logic necessary to operate the system in its various modes. It also includes certain control functions (such as, set clearance selection) peculiar to a particular mode. The dc power supplies are commercial units capable of supplying the required power to all other units of the MERA system with the exception of the indicator. The indicator contains its own dc power supplies.

(b) MERA Rf Module Operation and Performance

As discussed previously, the rf module is a complete transmitter-receiver unit, developing approximately 0.6-watt peak power at X-band and providing output signals at a 500-MHz if. Since the array includes 604 of these elements, the peak radiated power for the MERA radar is 360 watts. (A pulse compression ratio of 100:1 results in an effective peak power of 36 kW.)

Figure 3-44 shows the two sides of the module. The entire package, including antenna and covers, is $3.0 \times 0.9 \times 0.3$ inches and weighs approximately 0.08 pound.

The module consists of six individual circuits or circuit groups as shown in Figure 3-45, connected together by special interconnection techniques. The circuits are built on ceramic substrates using thin-film passive components and active devices (transistors and diodes) developed specifically for integration into microstrip transmission lines and circuits.

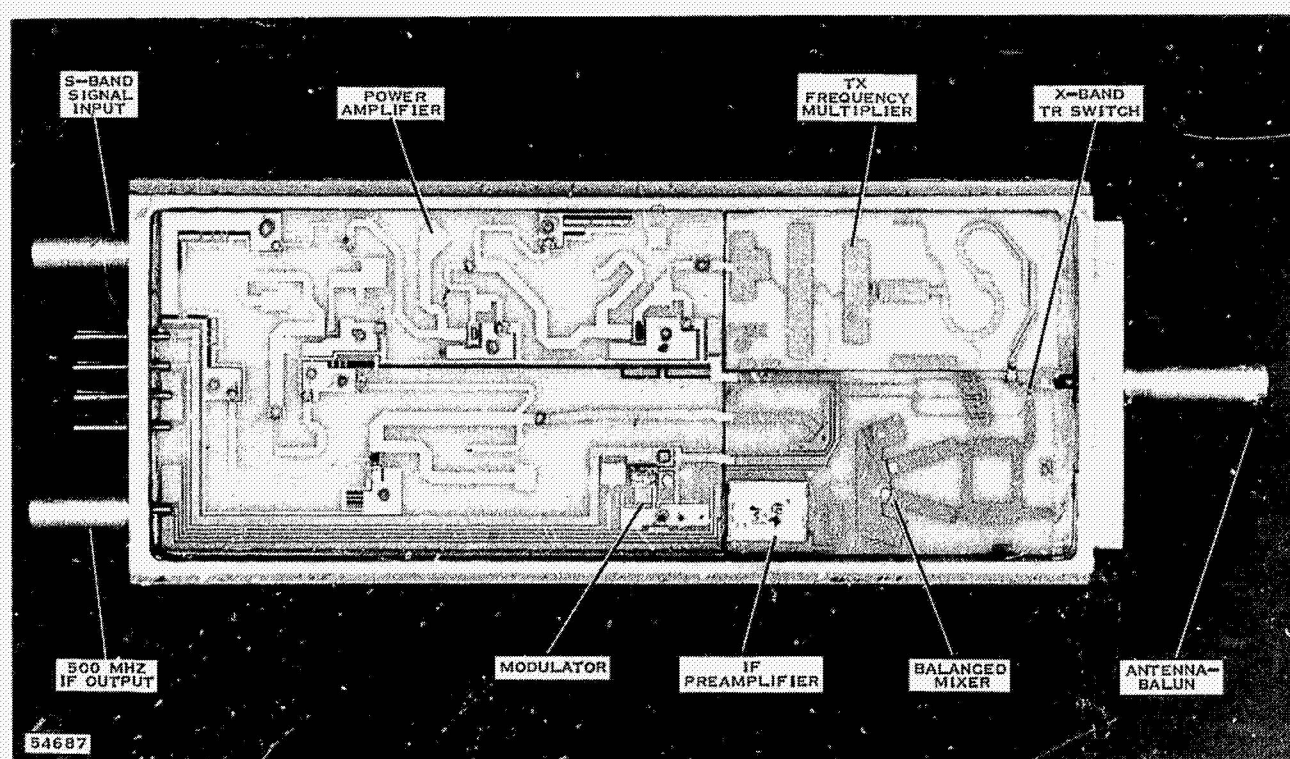
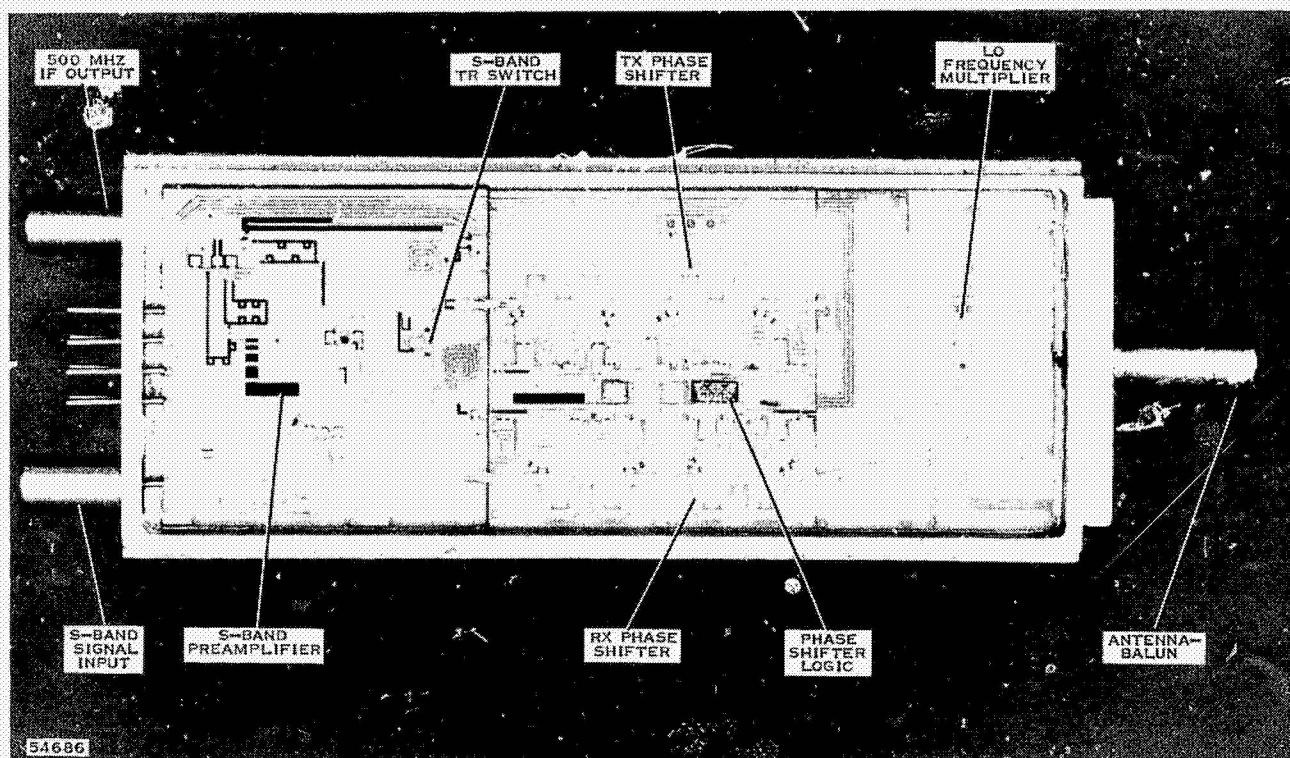


Figure 3-44. Two Sides of the MERA Rf Module

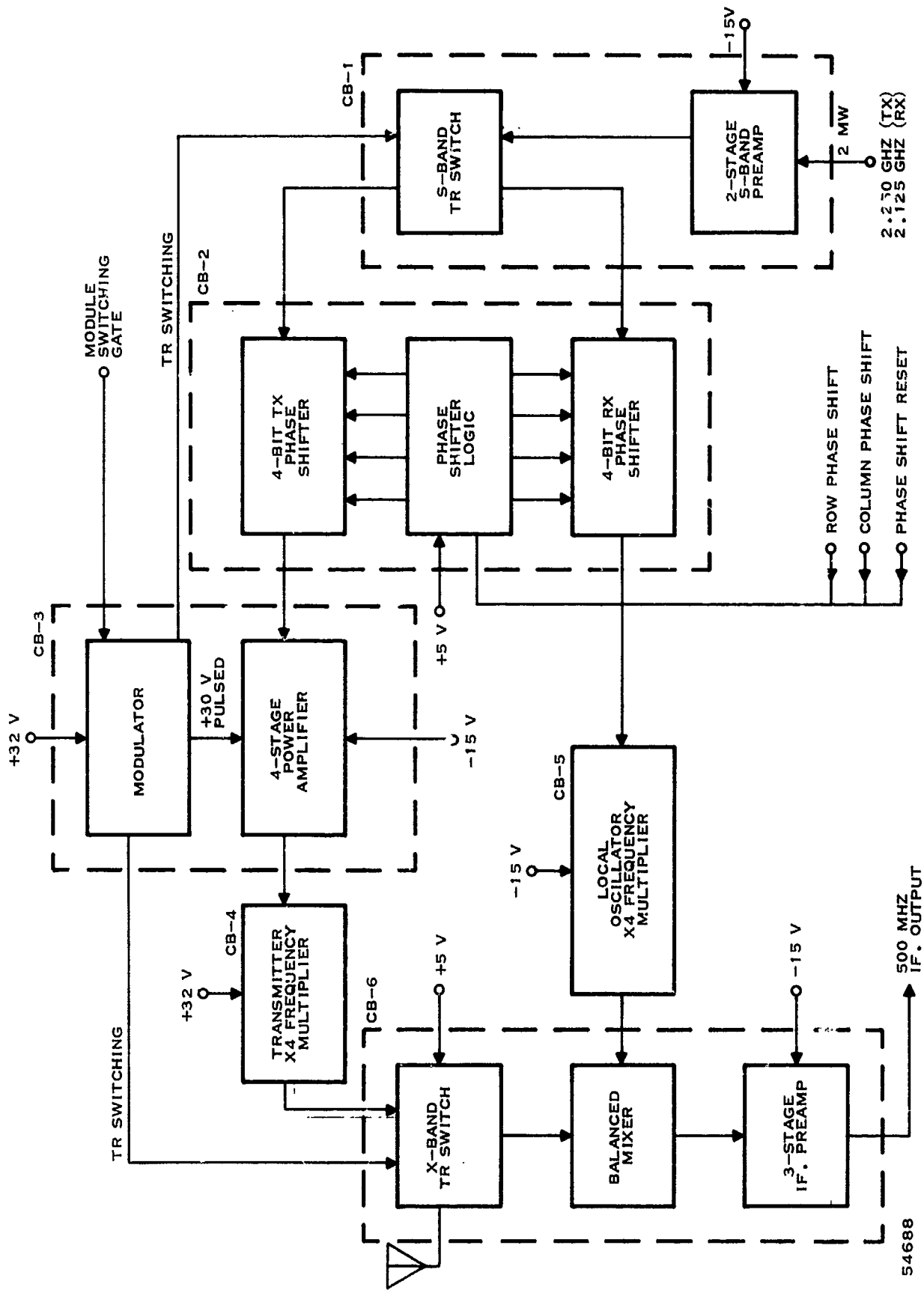


Figure 3-45. Rf Module Block Diagram

The passive components include capacitors, resistors, inductors, and transmission line structures. These devices are formed from metal and dielectric films deposited on the ceramic substrates and processed to form discrete or distributed circuit elements. Capacitors are fabricated using tantalum pentoxide or silicon oxide as the dielectric material. Resistors are formed from tantalum or from nickel-chromium films. Metalized lines consist of plated gold conductors.

As shown in Figure 3-44, only two coaxial connections to the modules are required. One is an S-band input, providing either 2 mW at 2250 MHz during transmit (transmitter excitation) or 2 mW at 2125 MHz during receive (local oscillator excitation). The other coaxial connection routes the 500-MHz if. signal to the combining networks in the rf distribution manifold.

Switching between TX and RX is accomplished in the signal generator prior to the manifold, so only one S-band corporate feed is required. During transmit time, the S-band input is swept linearly from 2250.75 MHz to 2249.25 MHz in a 19.5- μ sec period for chirped operation. For nonchirp operation, the input on transmit is a 0.24- μ sec pulse centered at 2250.00 MHz. During the receive cycle, the S-band input is a 2125.00 MHz CW signal in either operating mode.

The S-band preamplifier is a two-stage, broadband transistor amplifier providing approximately 11-dB gain for either S-band input. The S-band TR switch connects the output of this amplifier to either the TX phase shifter or the RX phase shifter, depending upon the condition of the TR switching signal. Both phase shifters are of the loaded-line type, providing excellent match, high accuracy, and low loss over large bandwidths.

The phase shifters include four digital phase-shift elements which are controlled by the outputs from a four-stage ripple-through counter. The most significant bit represents 45 degrees of phase shift at S-band. The counter is up-dated by the row and column inputs on a pulse-to-pulse basis, with the counter being cleared by the reset pulse just prior to receiving new phase shift inputs. The RX phase shifter uses the complementary outputs of the four-stage counter to compensate for the phase reversal through the mixer.

The pulsed power amplifier, consisting of four-transistor stages, provides a minimum gain of 23 dB at 2250 MHz. The output of this amplifier (a minimum of 2 watts) is fed to a times four varactor diode harmonic multiplier. This circuit includes a composite varactor (diode array) to accommodate the high input-power level and the necessary filtering to suppress unwanted harmonics. Output of the multiplier is 9000 MHz at a power level of approximately 0.7 watt.

The X-band TR switch consists of two surface-oriented switching diodes and associated decoupling chokes and bypass capacitors. This circuit provides greater than 20-dB isolation between transmitter and receiver sections of the module. Insertion loss through either channel is approximately 0.7 dB.

In the received section, the S-band oscillator signal is converted to 8500 MHz by a times four varactor multiplier. This is a low-power device similar to the high-power quadrupler in the transmit channel except that only a single varactor is required.

The received 9000-MHz signal is down-converted to a 500-MHz if. signal in a broadband, balanced mixer which utilizes silicon Schottky barrier diodes. The if. amplifier is a three-stage, video-type circuit providing 20-dB gain. Overall module noise figure is approximately 12 dB.

The modulator develops the 30-volt pulse required by the S-band pulsed amplifier and also provides TR switching signals at the proper levels for operation of the X-band TP switch and S-band TR switch. This circuit is essentially a transistor switch consisting of two monolithic gates and two chip transistors. The modulator is controlled by a low-level "module gate" signal.

The MERA radiating element is a half-wave dipole antenna. The antenna is fed by a split-tube balun mounted in the support structure external to the body of the rf module.

3.2.2.2 Microwave Power Transistors

The MERA program requirement for generating 2 watts of pulse power at 2.25 GHz has resulted in the development of a family of transistors, all utilizing approximately the same material and processing. These devices are used in the MERA four-stage pulsed amplifier and variations of these transistors are also used in the MERA S-band power preamplifier and 500-MHz if. amplifier.

The MERA requirements imposed the specifications in Table III-8 on the power amplifier output transistor.

The technique applied in designing this transistor consisted of scaling appropriate parameters of an existing Texas Instruments device of known characteristics to define a device with the desired characteristics. The starting point was a Texas Instruments transistor with an L-49 geometry (an in-house code number) shown in Figure 3-46. A series of experiments was performed, scaling the L-49 seven-finger transistor to successively a 15-stripe, 25-stripe, 35-stripe, and finally a 65-stripe device, all using the basic L-49 dimensions and spacing. Based on data obtained on these devices

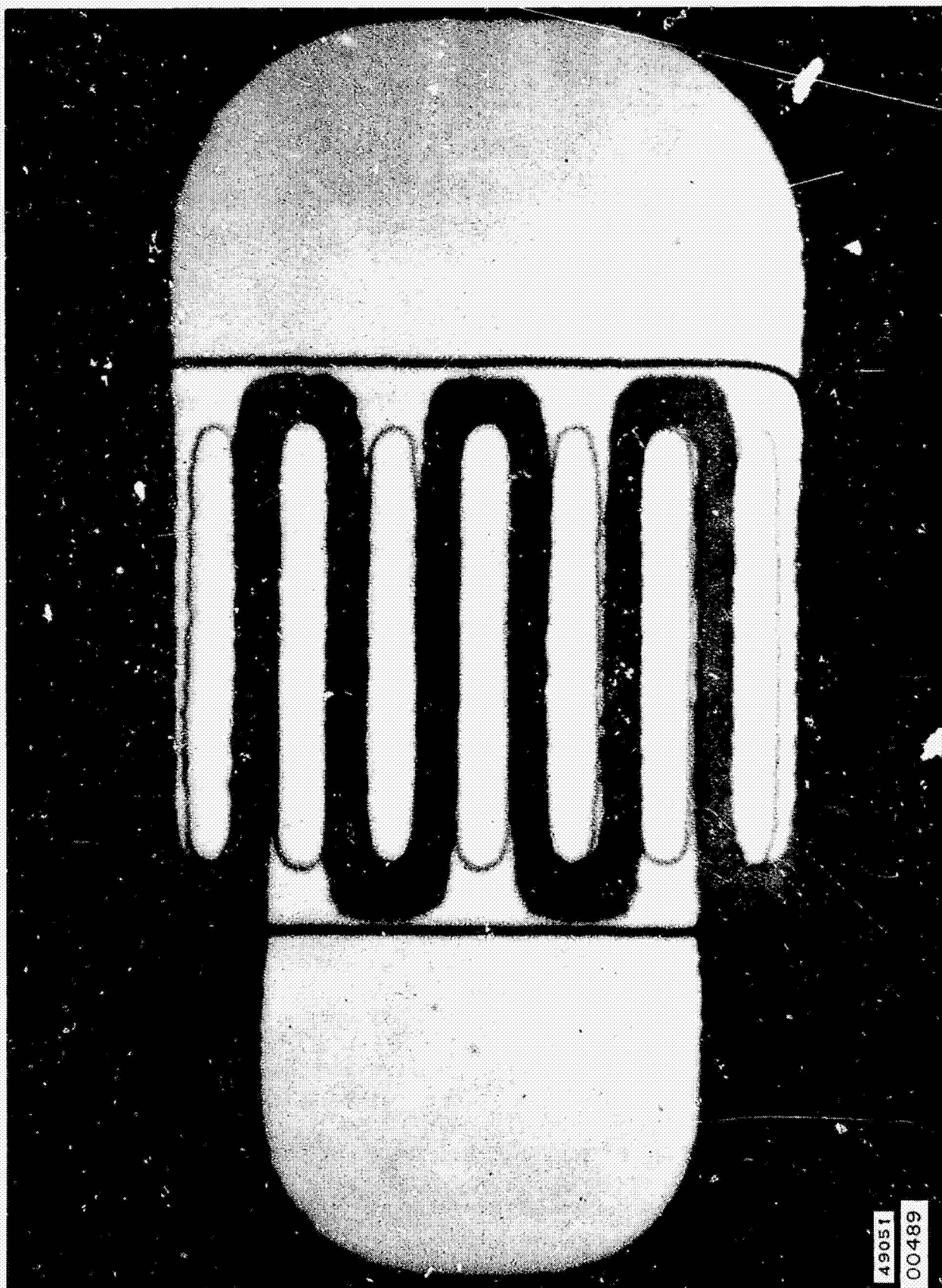


Figure 3-46. L-49 Transistor Geometry

Table III-8. Power Amplifier Output Transistor Parameters

Parameter	Specification
Frequency of operation	2.25 GHz
Power output	2 watts peak
Power gain	3 dB
Pulsewidth	20 μ sec
Duty cycle	10 percent
Operating mode	Class C signal biased

and the L-49, modifications to the stripe width, spacing, and diffusion schedules were made to produce the series of transistors listed in Table III-9, which makes up the four stages of the MERA power amplifier.

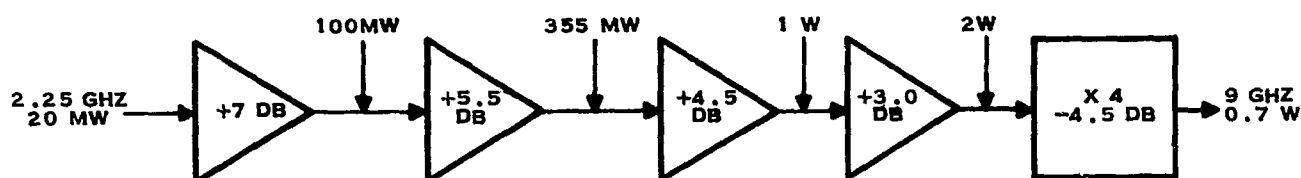
Table III-9. MERA Power Amplifier Stages

Transistor	Frequency	Power Out	Power Gain
L-158A	2.25 GHz	2 watts	3 dB
L-158B	2.25 GHz	1 watt	4 dB
L-158C	2.25 GHz	350 mW	8 dB
L-146(D)	2.25 GHz	100 mW	8 dB

In addition, the L-146 geometry forms the basis for the S-band preamplifier transistors (L-146B) and the 500-MHz if. preamplifier transistors (L-146A).

The end product of that scaling effort was the L-158A, which is the transistor used in the final output stage of the MERA power amplifier chain shown in Figure 3-47.

Figure 3-48 is a photomicrograph of the L-158A device geometry. This transistor, when operated with 30-volts bias and signal drive, gives 2.0- to 2.3-watts peak power output at 2.25 GHz with a 3.5-dB power gain. A continuing effort is underway to extend the capability of this type of transistor for use in high-power rf modules. Early this year, a 201 stripe transistor, called the L-195, was developed which produced 6 watts of peak output power at 2.25 GHz with a 7-dB power gain and 25-percent collector efficiency. A photomicrograph of the L-195 device geometry is shown in Figure 3-49. Figure 3-50 shows gain, efficiency and output power of this device versus



63819

Figure 3-47. MERA Module Power Amplifier-Multiplier Chain

frequency. Recent improvements in the L-195 have resulted in an improvement in efficiency, shown by the projected efficiency curve, indicating it has not been tested across the entire frequency range shown.

Still another advancement in power transistor capability was made recently when design and packaging changes in the L-195 yielded 10-watts peak output power with 35-percent collector efficiency at 2.0 GHz. This was accomplished by paralleling two 201-stripe devices on the same chip. The performance of this high-power amplifier is shown in Figure 3-51. The results in the 2- to 2.25-GHz range are measured and those lower in frequency are projected.

In devices of this type, the maximum frequency of oscillation primarily determines both the power gain and the collector efficiency, assuming negligible losses in input and output matching networks. Improvements in shallow diffusion technology have previously resulted in significant increases in the cutoff frequency (and thus frequency of oscillation) obtainable in large devices of this type, and still further improvements are anticipated. The area in which the greatest improvement in transistor performance may be achieved is device geometry refinement. Photolithographic technology has advanced significantly in just the past year, making it now possible to fabricate a device such as the L-195 with much smaller stripe widths and spacings than those presently used. With the improvements possible, both in device geometry and diffusion technology, the transistors should improve to about 50 percent in collector efficiency at 2.25 GHz.

Transistor performance at 1.5 and 2.25 GHz will be used as the basis for developing high rf power in the active rf modules for the CAT radar.

Several other methods of developing power were considered; including phase-locked avalanche, Gunn, and LSA oscillators. Each of these,

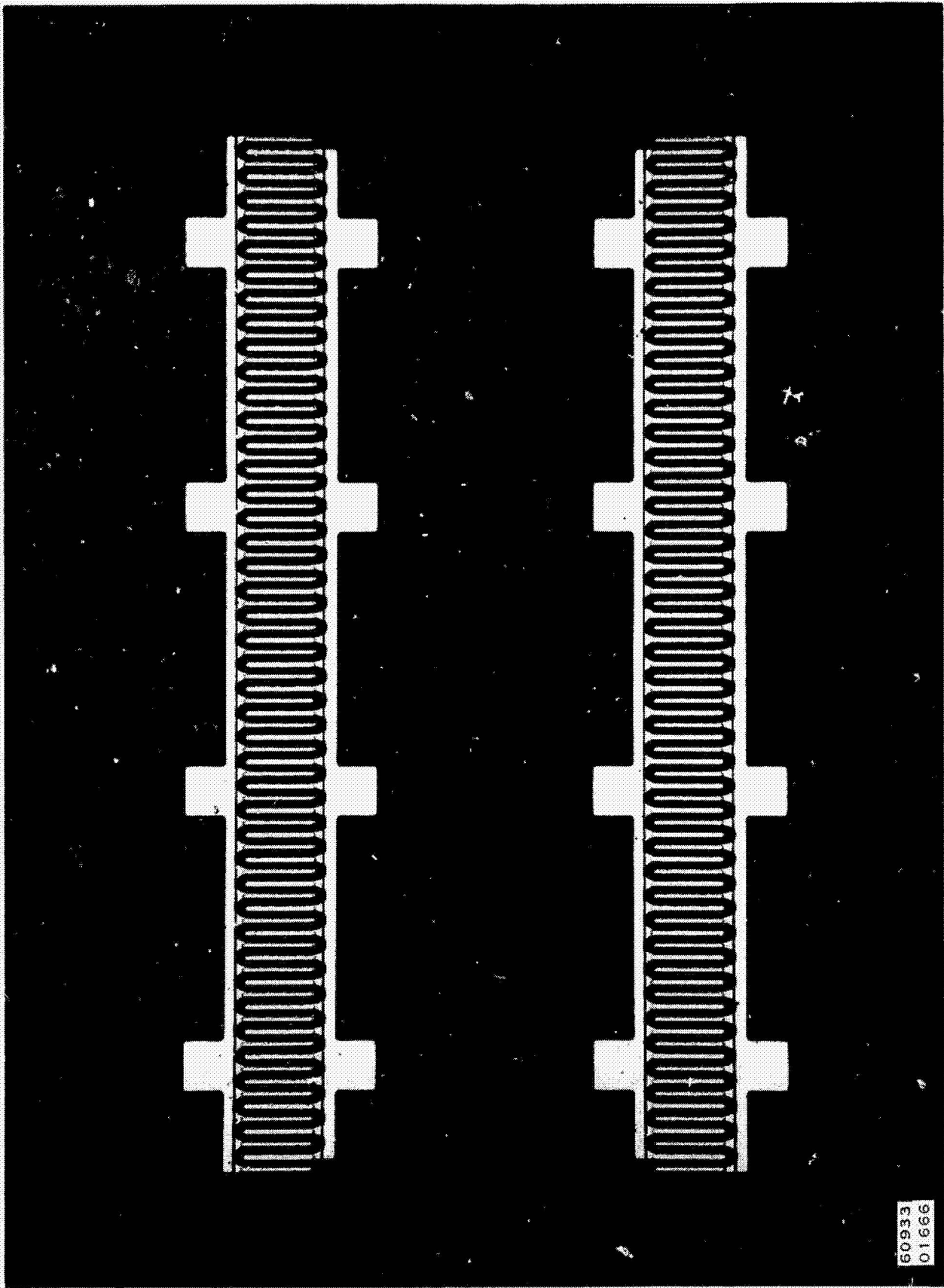


Figure 3-48. L-158A Geometry

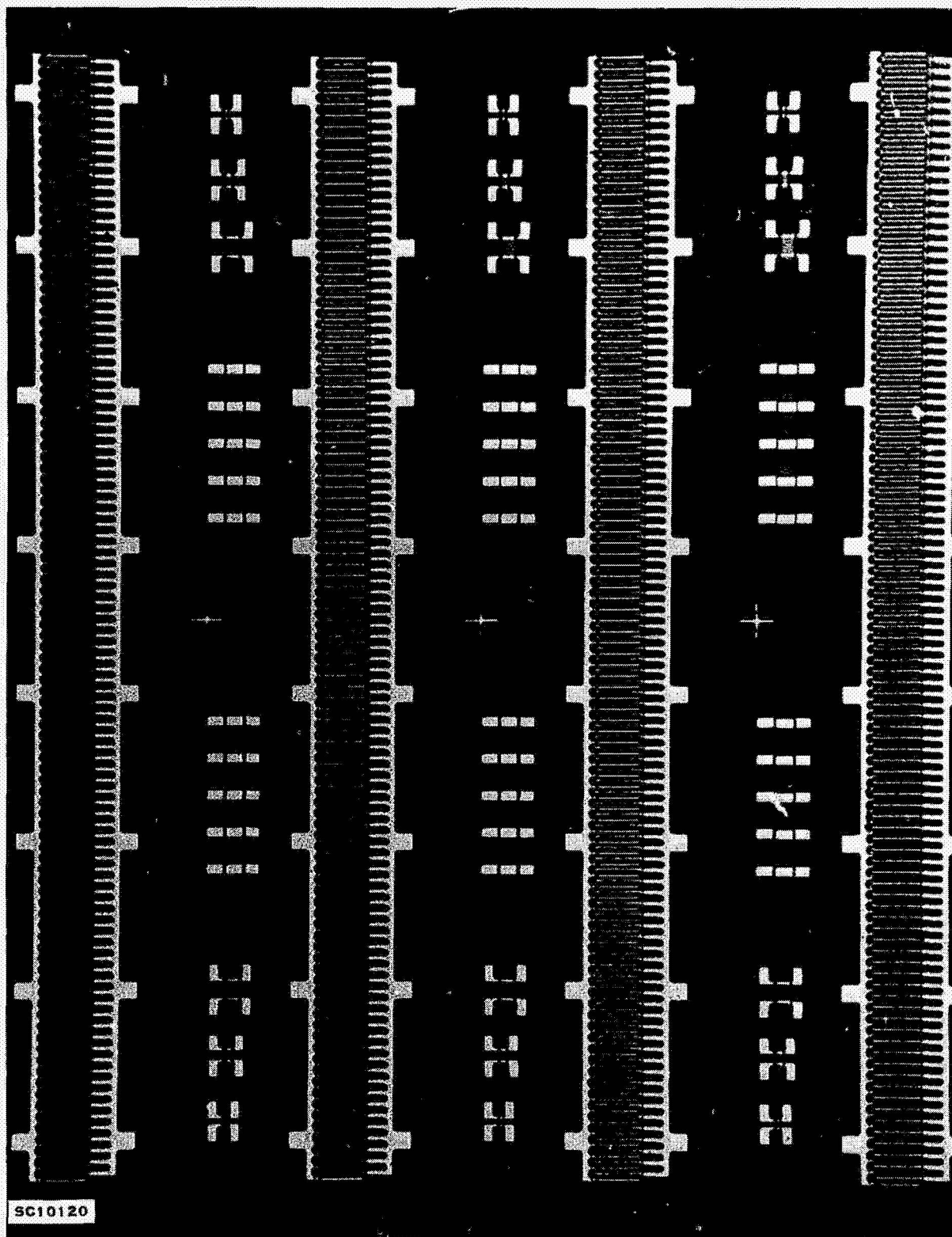
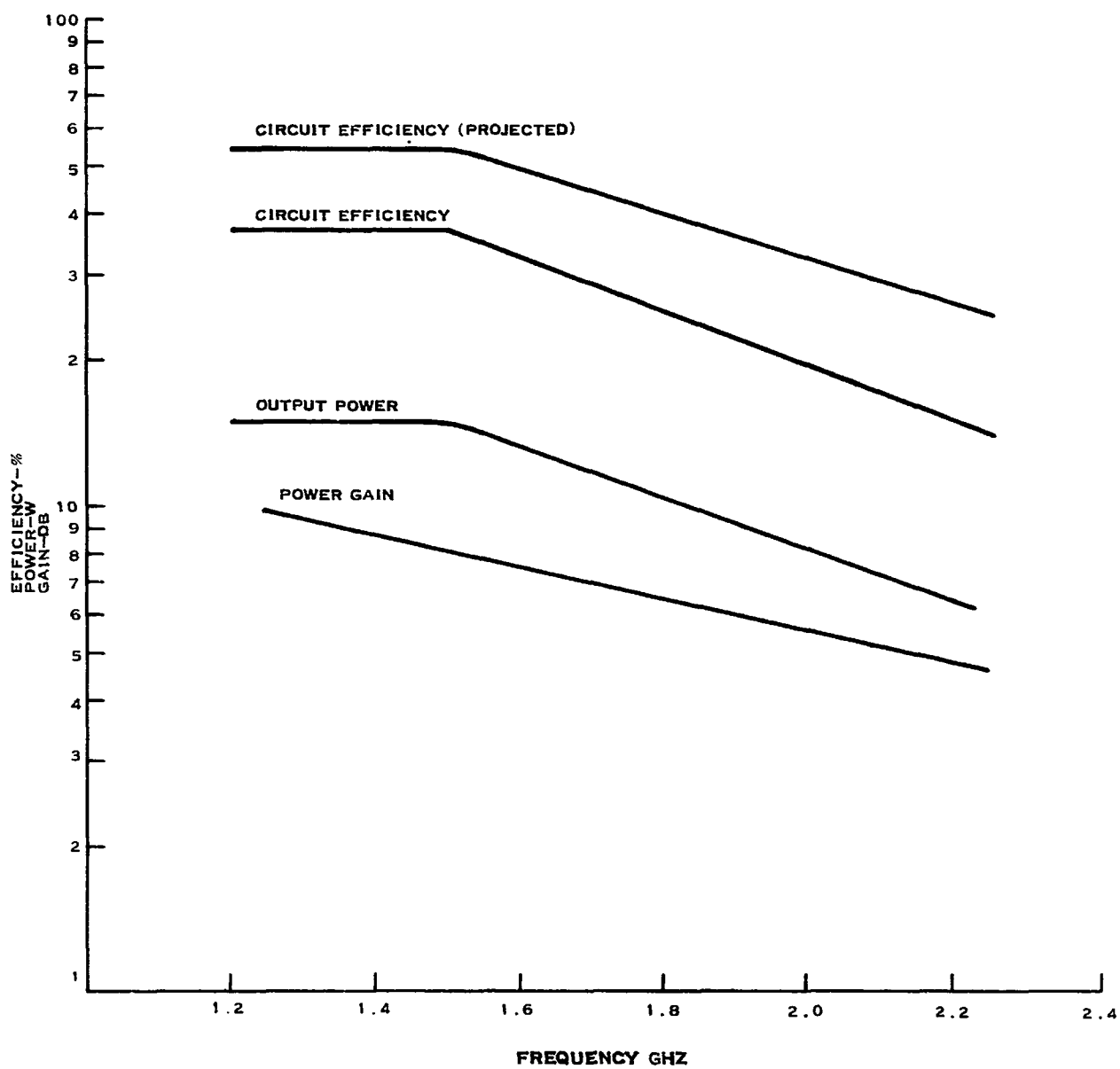
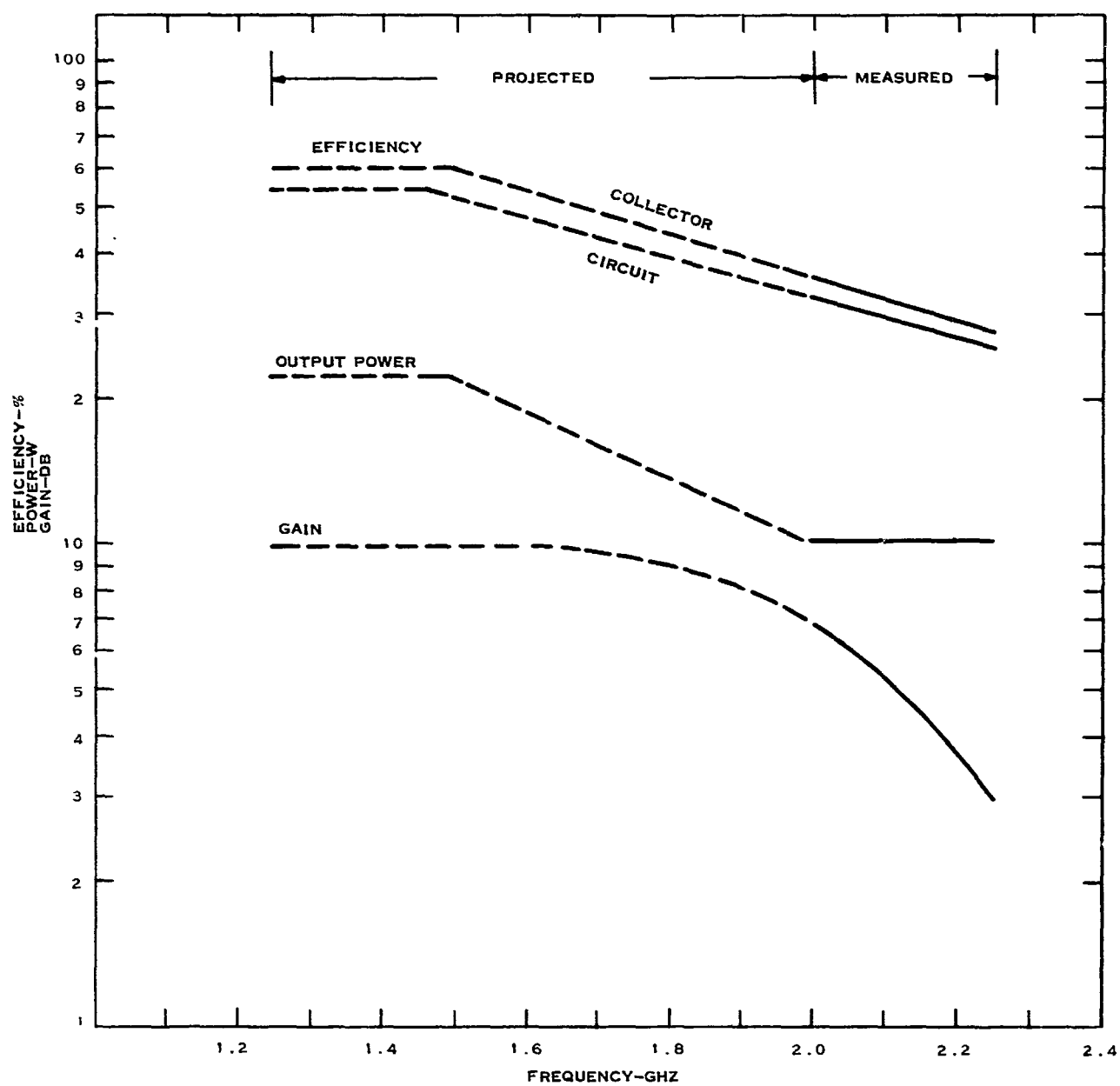


Figure 3-49. L-195 Transistors



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Figure 3-50. L-195 Pulsed Amplifier Performance



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Figure 3-51. Pulsed L-195 Amplifier with 2 Chips

conceivably, could provide the power output required. However, a major consideration in selecting an approach is the efficiency of rf power generation. The transistor or transistor-frequency multiplier chain is direct and has been proven in MERA modules and at still higher power outputs in laboratory devices. It offers an even greater opportunity for achieving high-efficiency operation by further improvement of an existing family of multiplier diodes and power transistors. For these reasons, better efficiency and a more mature technology, this is considered the best approach for the CAT radar active-array requirement. Further developments could influence this decision at a later date, and due consideration should be given to competing techniques prior to actual system design.

3.2.2.3 Microwave Diodes

A review of active rf modules reveals three general functions requiring microwave diodes; that is, switching, frequency multiplication, and mixing. Switching is accomplished on MERA by using beam-lead surface-oriented diodes (Figure 3-52) while frequency-multiplication circuits employ planar beam-lead varactors (Figure 3-53). Single chip, axial-type, Schottky barrier diodes (Figure 3-54) are used for mixing. All of these devices have been developed for integration into microstrip circuitry with a minimum of parasitics.

Surface-oriented diodes employ junctions diffused parallel to the surface of high-resistivity silicon ($>300 \Omega\text{-cm}$), as shown in Figure 3-55. However, the electrically active regions of the junction are the mutually adjacent edges. This permits the reduction of device capacitance to an extremely low value (on the order of 0.01 pF) while maintaining a large contact area, thus providing low resistance and good contact adhesion.

An early problem with surface-oriented diodes was the conductivity modulated forward resistance. This problem was solved by maximizing diffused layer doping levels while maintaining the proper layer-depth to layer-separation ratio. Capacitance of the surface-oriented diode is controlled primarily by the length of opposing P+ and N+ regions and secondarily by diffused layer depth. Fringing capacitance accounts for 50 to 70 percent of the total in a typical device. There is no contribution by MOS (Metal-Oxide Semiconductor) capacity under the beam contacts because of the high resistivity of the silicon. This combination of low capacitance, low resistance, and ease of beam-lead construction makes the surface-oriented diode an ideal device for performing switching functions in microstrip transmission circuitry. Stable, reproducible devices have been developed for the MERA antenna TR switch, S-band duplexer, and S-band phase shifters.

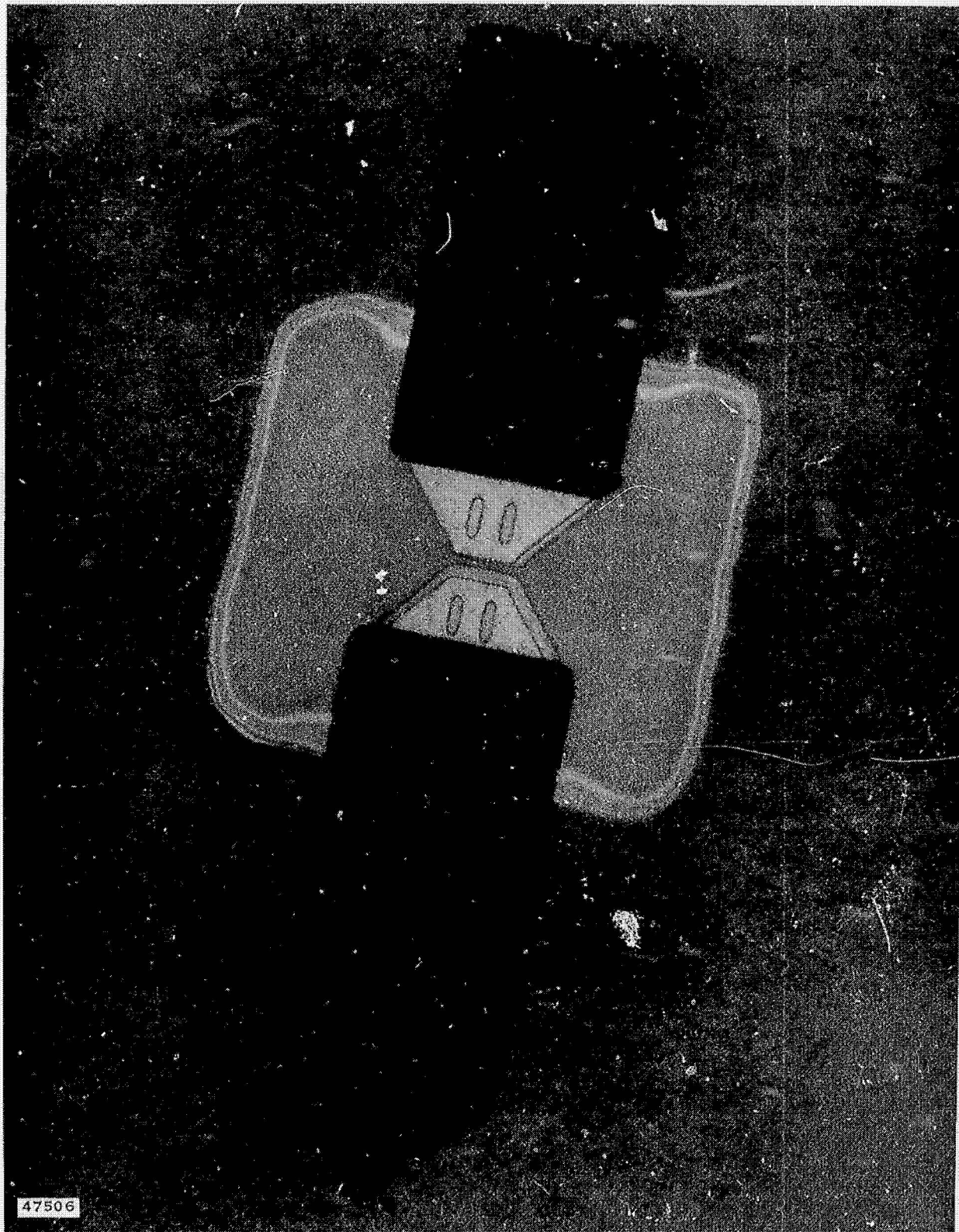


Figure 3-52. Beam-Lead Surface-Oriented Switching Diode

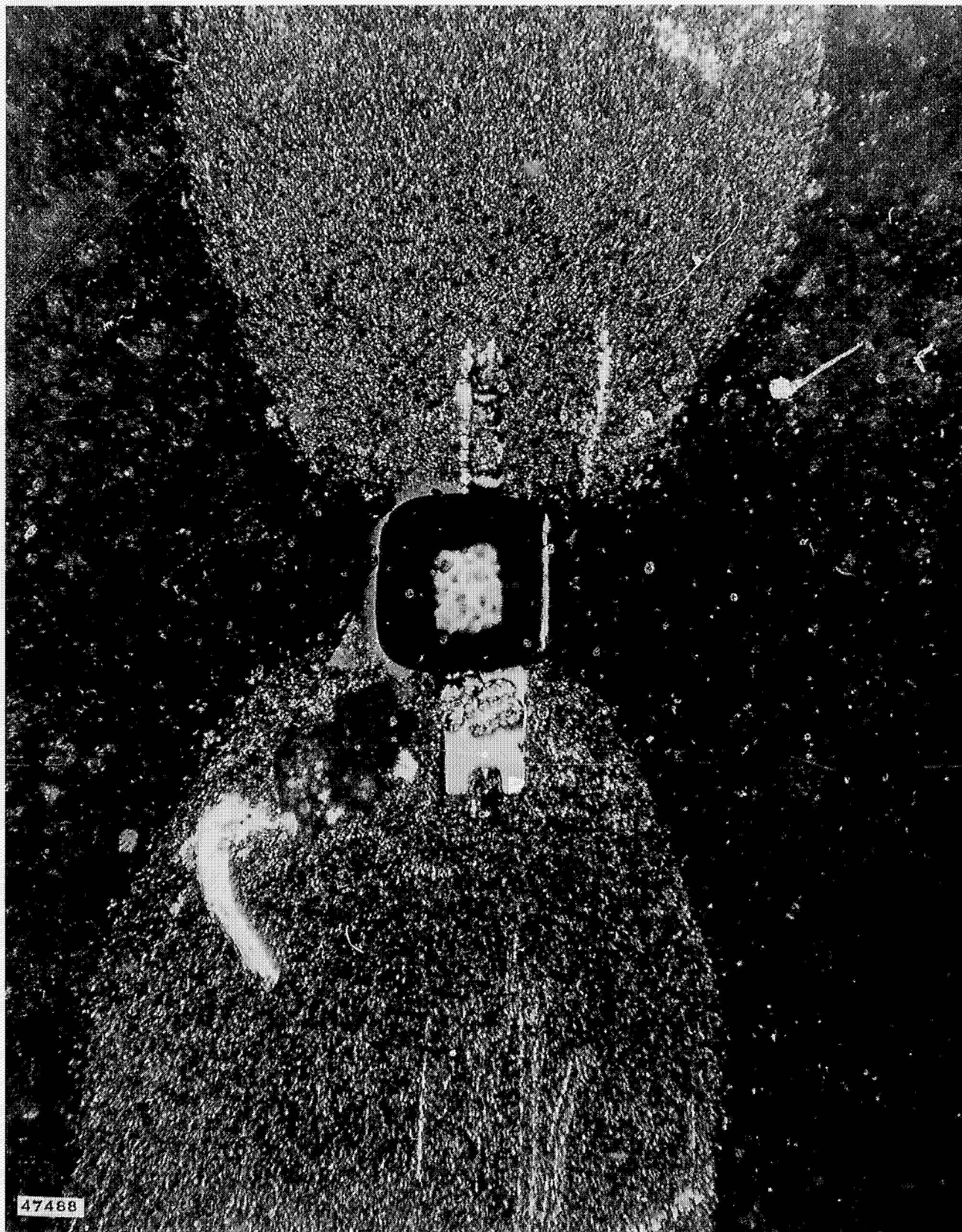


Figure 3-53. Beam-Lead Planar Varactor Diode

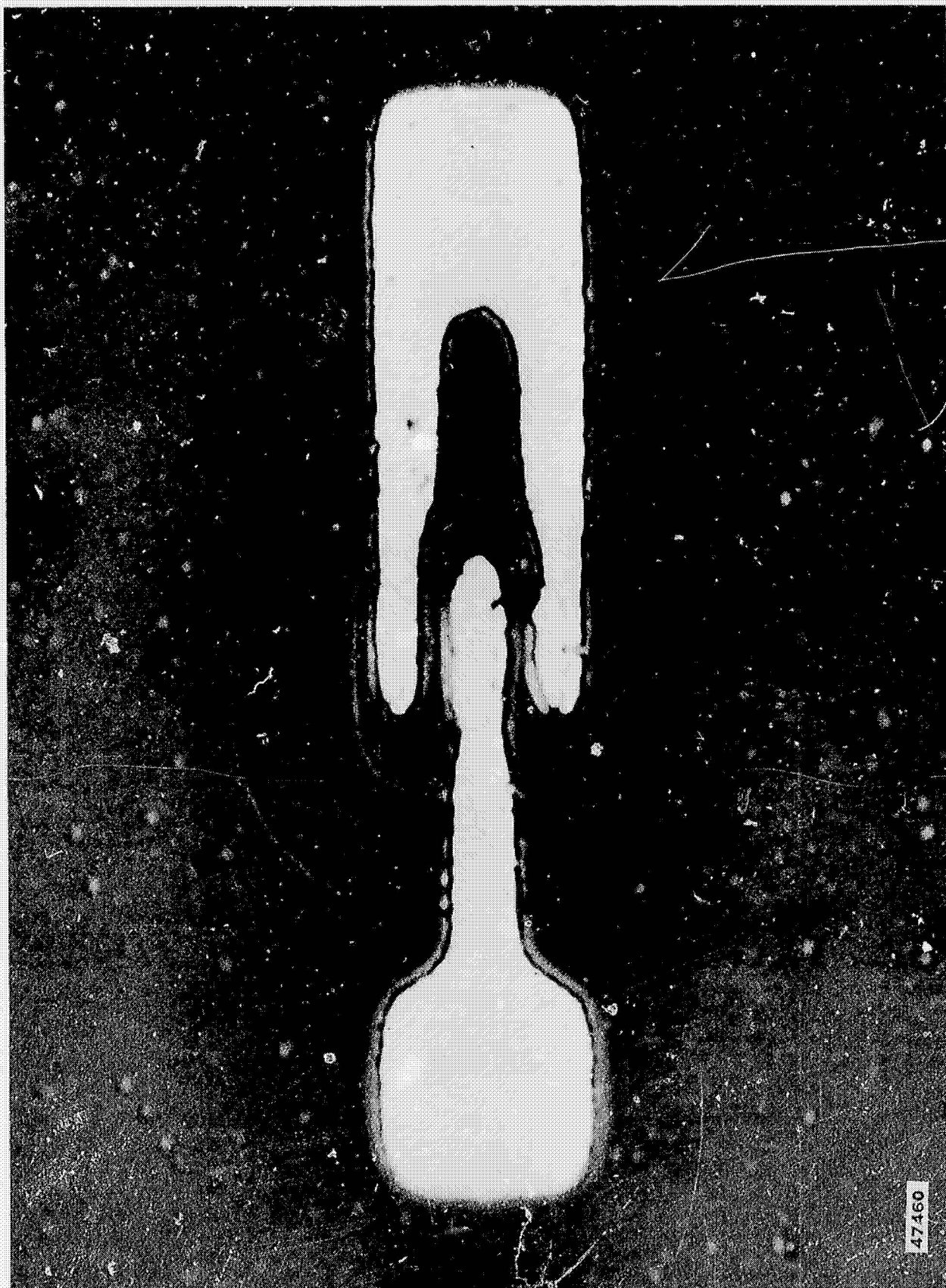
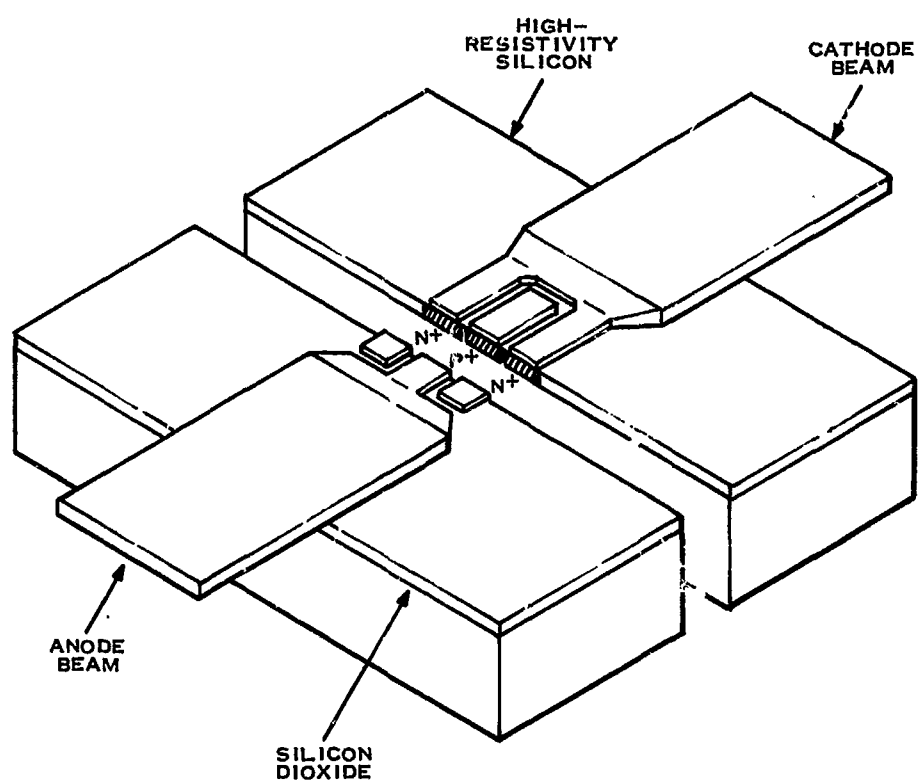


Figure 3-54. Schottky Barrier Diode



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Figure 3-55. Beam-Lead Surface-Oriented Diode Structure

The planar beam-lead varactor is shown in Figure 3-56. This device is constructed from epitaxial material. Both the epitaxial layer and the diffused regions are quite thin in order to minimize series resistance. Two types of these devices have been developed for the MERA program: (1) a conventional "varactor" type, where capacitance variation in the reverse bias region is the mechanism employed in harmonic generation, and (2) a more recent "step-elastance" type which utilizes charge storage nonlinearities to generate harmonics. Table III-10 illustrates important characteristics of two of these devices presently being produced.

Table III-10. Varactor Diode and Step Elastance Diode Characteristics

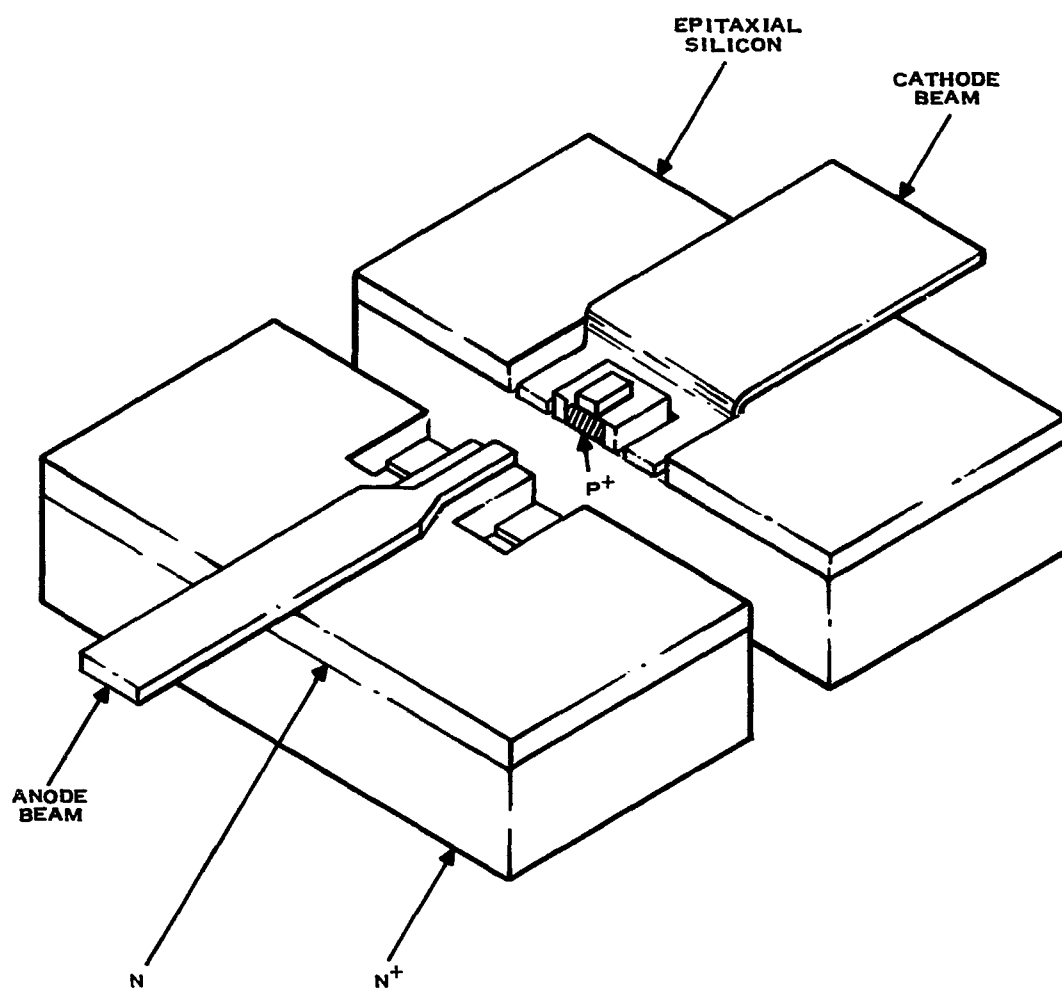
Type	V_B	C_j (at -6V)	f_c
Varactor diode	25V	0.2 pF	450 GHz
Step-elastance diode	60V	0.3 pF	250 GHz

The structure of both the varactor and step-elastance diodes permits a stored charge to be completely contained and easily controlled to achieve optimum frequency-multiplier performance. Junction capacitance is controlled by junction area rather than periphery. Fringing capacitance accounts for about 20 percent of the total in typical devices. MOS capacitance under the anode beam lead limits this device in low capacitance switching applications. Series resistance, while low, is higher than that obtained with surface-oriented diodes. However, the surface-oriented diode is not usable as a frequency multiplier because the stored charge is not conveniently controlled.

In order to handle high rf power with good efficiency, a special varactor has been developed. It consists of several individual varactors, series and parallel connected, on a single silicon ship. A drawing of this varactor, called a centipede varactor, is shown in Figure 3-57. This enables one to use many low-voltage, higher cutoff frequency devices, rather than a few series-connected beam-lead varactors. The use of this varactor should make possible the achievement of a 5.0-dB typical conversion loss with 10 to 20-watt pulse input power for a quadrupler and 3.0-dB loss for a doubler.

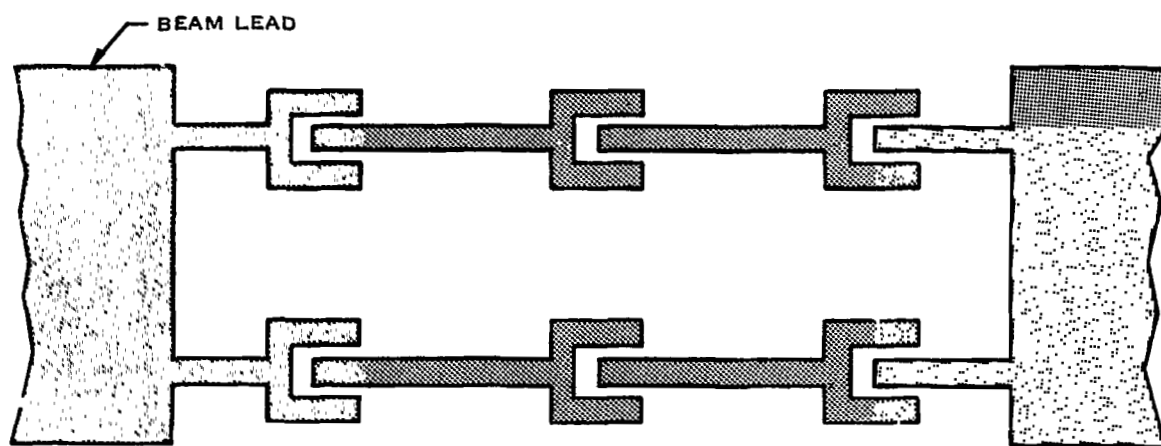
3.2.2.4 Transmitter Chains

In order to make a parametric radar system study across the frequency range of 2 to 20 GHz using active rf modules, five modules were chosen in this band and one at a slightly lower frequency where transistor efficiency is much improved. The output frequencies are 1.5, 2.25, 3.0, 6.0, 9.0, and 18.0 GHz.



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Figure 3-56. Planar Beam-Lead Varactor Structure



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Figure 3-57. Monolithic Structure for Series-Parallel Diode Combinations

As noted previously, current transistors fall off rapidly in gain, power and efficiency above 2.25 GHz. For this reason, the transmitter chains at 1.5 and 2.25 GHz are formed with transistor amplifiers only. The higher frequency modules use these basic chains to drive varactor frequency multipliers. Emphasis has been placed on achieving maximum output power at maximum efficiency. This will produce minimum system heat load, size, and weight. Later tradeoffs could include reducing the output power per module in order to have more rf modules and thus increase antenna gain; and also to reduce module thermal load to a level acceptable to a passive heat removal system. Size and weight penalties would likely be in order to achieve the same system performance.

Various power amplifier-frequency multiplier chains are shown in Figure 3-58. Figure 3-58a describes the present MERA module transmitter chain. A pulsed 20-mW input signal at 2.25 GHz is amplified to 2-watts peak power with 11-percent efficiency and then frequency multiplied to 9.0 GHz providing 0.7-watt peak power with 3.9-percent overall efficiency. Peak input power to the chain is about 12 watts, but this should not be used to calculate heat load because the first stage is operated in Class A, the other three are operated Class C and there is a dc feed requirement.

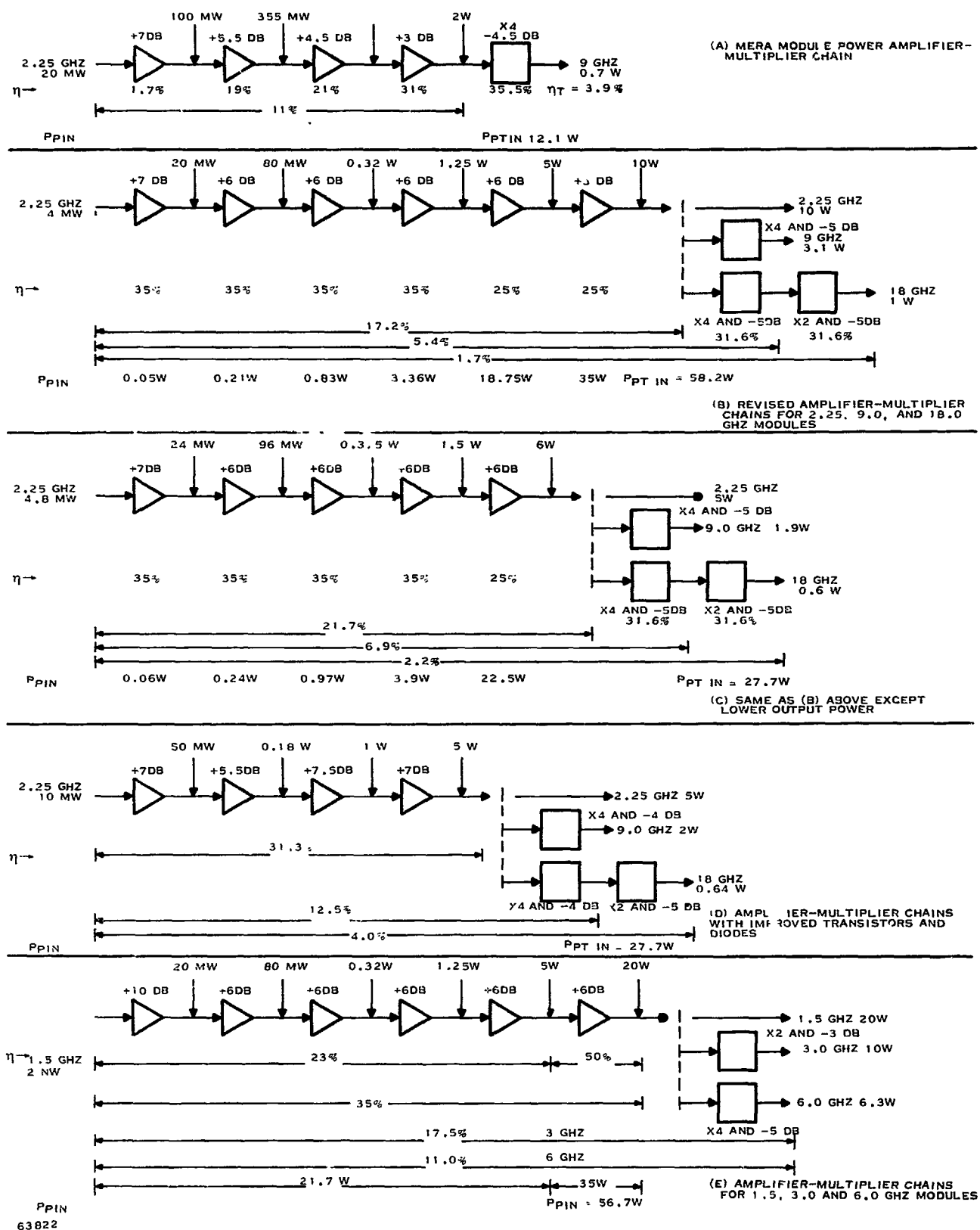


Figure 3-58. Amplifier-Multiplier Chains for Rf Modules Showing Gain, Power and Efficiency

Figure 3-58b describes revised amplifier-multiplier chains using the improved L-195 transistors. These chains show how the transmitter power is generated in the 2.25-, 9.0-, and 18.0-GHz modules. The 9.0-GHz chain generates 3.1 watts at 5.4-percent efficiency, and the 18-GHz chain generates 1 watt at 1.7-percent efficiency. The peak input power, P_{pin} , per stage is shown and the total peak input power, P_{ptin} , is 58.2 watts. The average dissipated power, or heat load, in a chain is simply $P_H = (58.2 \text{ watts} - \text{Peak rf power}) (\text{duty cycle})$, since all stages are operated Class C.

The X4 frequency multiplier has 5-dB conversion loss due to its higher power handling requirement. There is some difficulty involved in building the X2 multiplier shown due to the high-input power and the high frequency (K_u -band) at which it operates, for this reason the conversion loss is assumed to be 5 dB. Circuit elements will be small at 18 GHz and more difficult to design and fabricate. To handle the 3 watts of input power, the varactor requirements call for very low impedance. A multielement centipede varactor should be capable of handling the load.

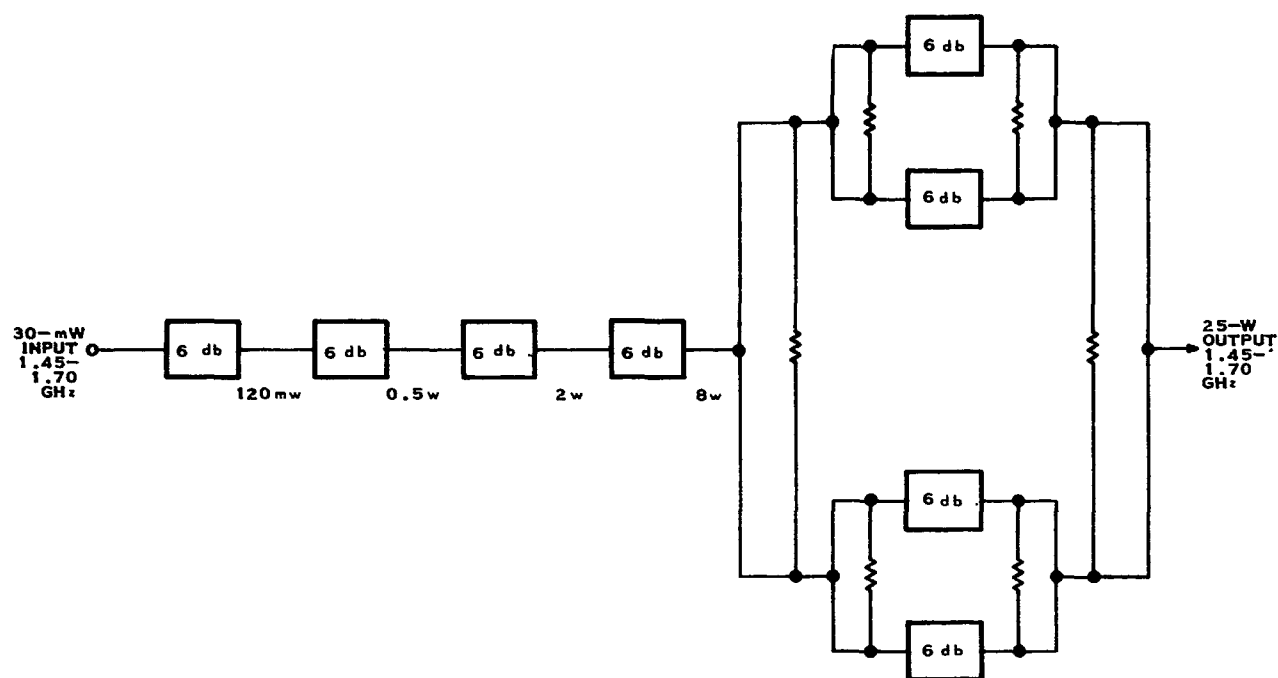
Figure 3-58c is essentially the same as Figure 3-58b except a lower power output is accepted in order to obtain better efficiency and to reduce the heat load problem. The single output transistor should have better reliability than the paralleled devices used in the previous chain.

If further device development were undertaken we could expect to see efficiencies as shown in Figure 3-58d. These devices are considered to be within the capability of today's technology but have not been developed. This places them outside the constraints of the CAT radar study; nevertheless, they may be available by the time actual hardware development begins.

Figure 3-58e shows the amplifier-multiplier chains used in the 1.5-, 3.0-, and 6.0-GHz modules. The latest version of the L-195, whose performance was described in Figure 3-51, is used in the final amplifier. The overall amplifier efficiency is very good, 35 percent, for generating 20 watts of peak power.

This circuit generates more power with better efficiency (10 watts at 17.5 percent) at 3.0 GHz than the straight amplifier approach. As microwave transistors improve in performance, the multiplier can be eliminated. A peak power of 6.3 watts is generated at 6.0 GHz with 11 percent efficiency.

The total input peak power is 56.7 watts which will present heating problems at high duty cycle. The last amplifier stage may be eliminated for some systems configurations.



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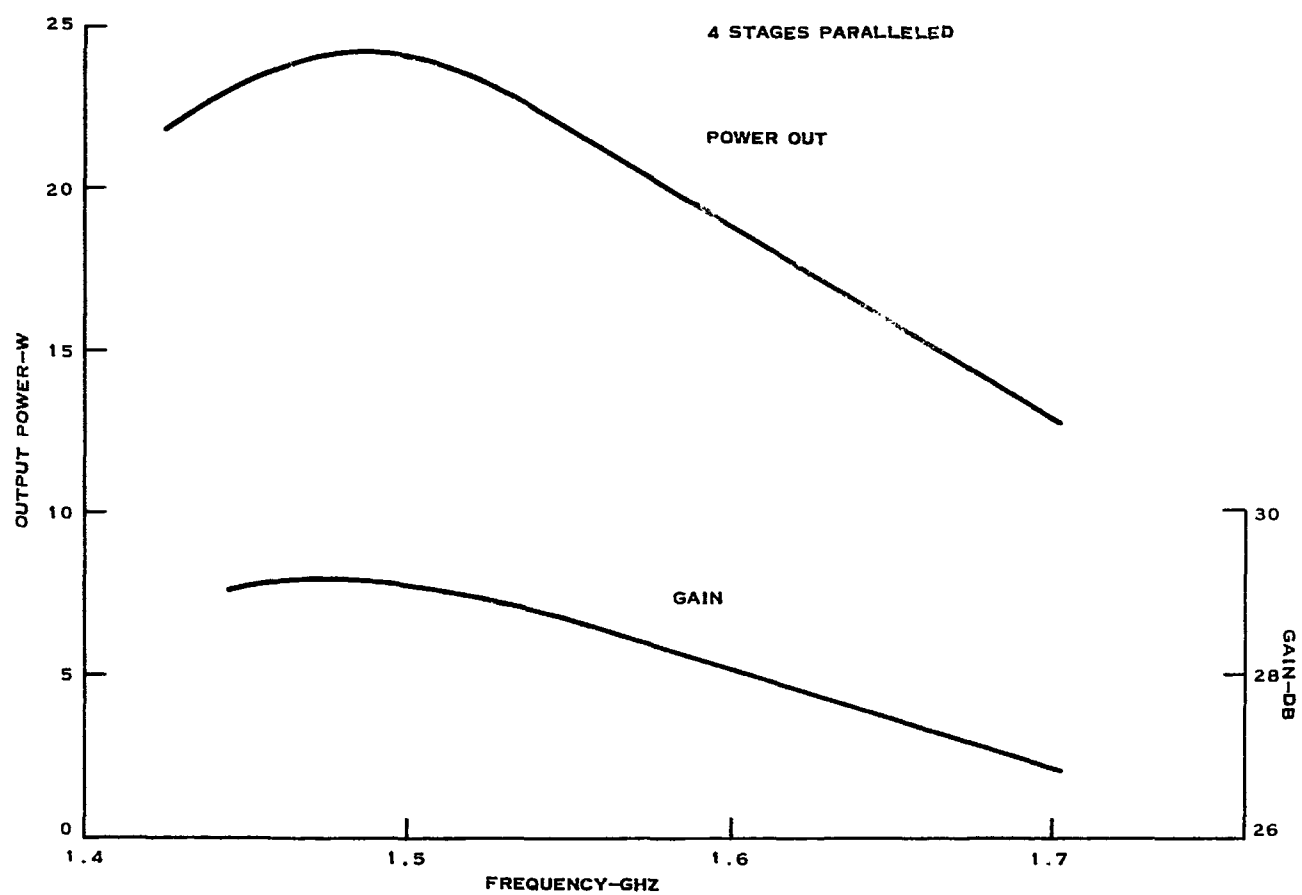
Figure 3-59. Simplified Block Diagram of 25-Watt L-Band Amplifier

Many other amplifier-multiplier chains are possible, but these should provide suitable data for our parametric tradeoffs. Lower powers can be used and higher powers can be generated by utilizing parallel amplifiers or parallel amplifier-multiplier combinations. For example, Texas Instruments recently generated 25 watts of peak power at L-band by the power amplifier circuit of Figure 3-59. Performance curves for this amplifier are shown in Figure 3-60.

The L-band amplifiers have been operated at a duty cycle up to 10 percent with pulselengths up to 350 μ sec. It is also expected that duty cycles up to 30 percent are feasible if the pulselength is decreased.

3.2.2.5 Phase-Shift Networks

Two types of phase shifters were considered for use in the active rf modules, digital-diode, and latching ferrite types. The thin-film, or thin-plate, ferrite phase shifters would seem to require much less dc power, but would be larger in size. The technology needed on the ferrite phase shifter, for use in active rf modules, is not as far advanced as that of the diode phase shifters and it is not feasible to consider them for these rf modules. As the technology advances in the next year, they will be very strong contenders due to their lower power requirements and perhaps better reliability.



63824

Figure 3-60. L-Band Transistor Amplifier Pulsed Power and Gain Versus Frequency

A brief description of the diode phase shifter is covered here and additional discussion is made in Subsection 3.2.3.5.

Either three- or four-bit phase shifters were assumed to be usable in the CAT radar because the sidelobe problems should not be very severe with a clutterless background. A decision on which to use was made mainly on their dc power requirements. A quick review of some of the phase shift networks presently being used indicated that they all require too much average power for the CAT radar; for instance, the MERA phase shift requires about 300 mW and an S-band circuit for ground radar application about 1 watt. These circuits use standard logic and generally drive the diodes quite hard into forward conduction to reduce the insertion loss.

Low power logic is available for the SN5400 and SN7493 circuits which requires 0.1 of the power that the standard circuits require. The low power logic is slower, about 3-MHz maximum, than the standard circuits which can operate in the 10- to 20-MHz range. Initial study of system requirements indicate the maximum range will limit the prf to about 400 Hz to avoid range ambiguities. This low prf would then allow the use of low-power logic. If a module had a single phase shifter for both transmit and receive, using a mixer in the receiver, the phase shifter then would need to shift in less time than that required for the transmitted pulse to reach the closest target and return in order to keep the antenna beam pointed in the same direction.

The low-power logic can run up to 3 MHz, so a period of about 330 nsec can be used for a single pulse. The pulse would be 165 nsec long. At the maximum load, 16 pulses would be needed for a four-bit phase shifter or eight pulses for a three-bit. This corresponds to a total count period of 5.32 μ sec for the four-bit phase shifter and 2.66 μ sec for the three-bit. Since the row and column input signals to the phase shift network logic must be added in series, the total time required to send in the information is 10.64 μ sec for the four-bit network and 5.32 μ sec for the three-bit.

The approximate round trip time for 1-nm range is 12 μ sec. Thus, the minimum range of 5 nm for the CAT radar is quite compatible with the use of low-power logic.

If the mission were extended to permit rendezvous or inspection down to 500-foot range, the total time allowed for switching would need to be less than 1 μ sec in this mode of operation. Use of the standard logic would even be marginal since its maximum specified speed would be exceeded, or at least marginal. In this case, it would be best to use two phase-shift networks and still retain the low-power logic.

Planar diodes are generally driven at 5 mA at 5 volts and SOD diodes are driven at 10 mA at 5 volts for minimum insertion loss. The curve of loss versus drive current begins to flatten out at about 3 mA; hence, 4 mA was selected as the bias point for the planar diode and 7 mA for the SOD diodes. The SOD diodes are generally used for small phase-shift bits, 22.5 degrees or less.

Of the digital diode phase shifters, the loaded-line type is preferred over the line-length switch in the frequency ranges of interest; because, it has lower loss variation with phase-shift position, and is generally (for small phase shifts) smaller and requires fewer diodes. There are exceptions, of course, and each requirement must be looked at separately.

In the case of the three or four-bit phase shifter (180 degrees -90 degrees -45 degrees -22.5 degrees) that we would probably use in the 1.5- and 2.25-GHz modules, a combination of loaded line and line length switches would be optimum. Such a switch is shown in Figure 3-61. The loaded-line phase shifter requires a quarterwave transmission line with a shunt diode at each end. This can be noted in the 22.5-degree and 45-degree bits. These two could be combined on a single quarterwave line as done in the MERA module to reduce loss and size. The loaded-line phase shifter is limited to 45 degrees per quarterwave section, so a 90-degree or 180-degree bit requires two or four quarterwave lines respectively. Two parallel diodes in the same phase bit may be combined as seen in the 90-degree bit. The 180-degree bit is a line-length switch because it requires one less diode than the loaded line and is approximately one-half wavelength shorter, providing a reduction in loss and size.

The square loops seen in Figure 3-61 symbolize the inductor and capacitor used to decouple the dc bias circuit from the rf circuit and provide the rf short to ground. The logic network consists of a dual input NAND gate to accommodate row and column phase signals, a four-stage ripple-through counter, and power gates for controlling the diodes. A reset pulse enters the counter to clear it before being updated.

A summary of phase-shift network dc power requirements for candidate diode phase shifters for the active rf modules is given in Table III-11. Phase-shift networks, using low-power logic, require about one-half the power that the standard logic requires; hence, they will be used exclusively.

The networks selected for use are the following: the 1.5-, 2.25- and 3.0-GHz modules use a three-bit phase shifter requiring 250 mW and the 6.0-, 9.0-, and 18.0-GHz modules use a four-bit phase shifter requiring 125 mW.

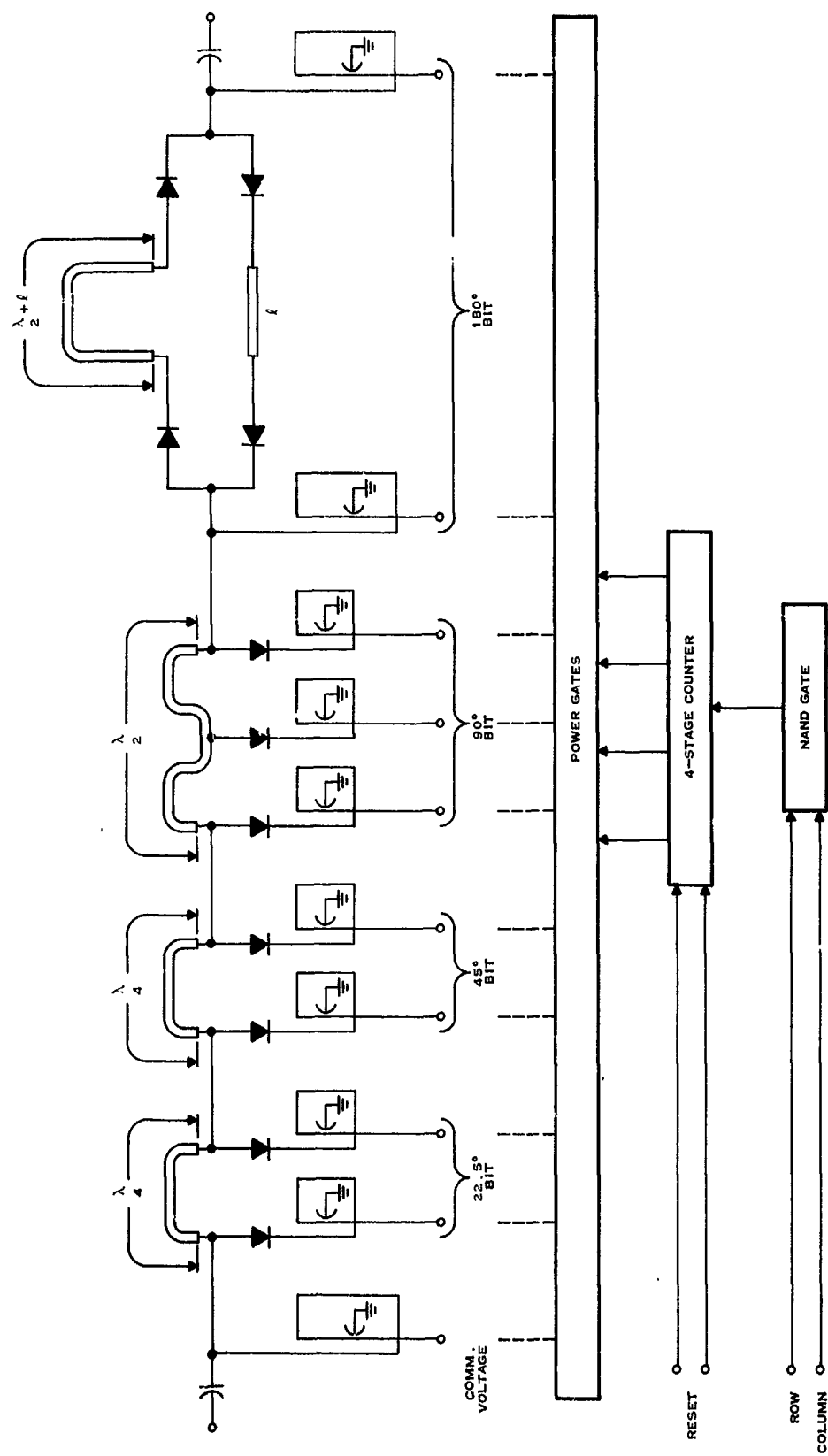


Figure 3-61. Combination Loaded-Line/Line-Length Switch

Table III-11. Power Requirements for Diode Phase-Shift Networks

Phase Shifts	Input Power (mW)			
	Four-bit		Three-bit	
	1 Ps ¹	2 Ps	1 Ps	2 Ps
180° -90° -45° -22.5°				
Standard Logic	605	610	535	540
Low-Power Logic	318	288	248	216
90° -45° -22.5° -11.25°				
Standard Logic	535	500	465	430
Low-Power Logic	294	248	224	228
45° -22.5° -11.25° -5.62°				
Standard Logic	---	300	---	248
Low-Power Logic	---	125	---	120

- NOTES: (1) 1 Ps means logic controls only one phase-shift network
 2 Ps means logic controls two phase-shift networks
 (2) Phase shifters operate in the 1.5- to 3.0-GHz range

3.2.2.6 Receiver Circuits

This section briefly reviews active module receiver front-end circuits. The receiver front-end circuits considered in the CAT radar active array study were

Low-noise microwave transistor amplifier
 Tunnel-diode amplifier
 Parametric amplifier
 Balanced mixer with VHF if. preamplifier
 Balanced mixer with UHF if. preamplifier
 Image-terminated mixer.

Each of these is covered in the following discussion. The low-noise silicon or germanium transistors, that are available today, outperform the conventional mixer up to mid S-band. Gain and noise figure for a typical device, the Texas Instruments L-186, is shown in Figure 3-62. This device was used in several of the rf modules. A newer device currently under

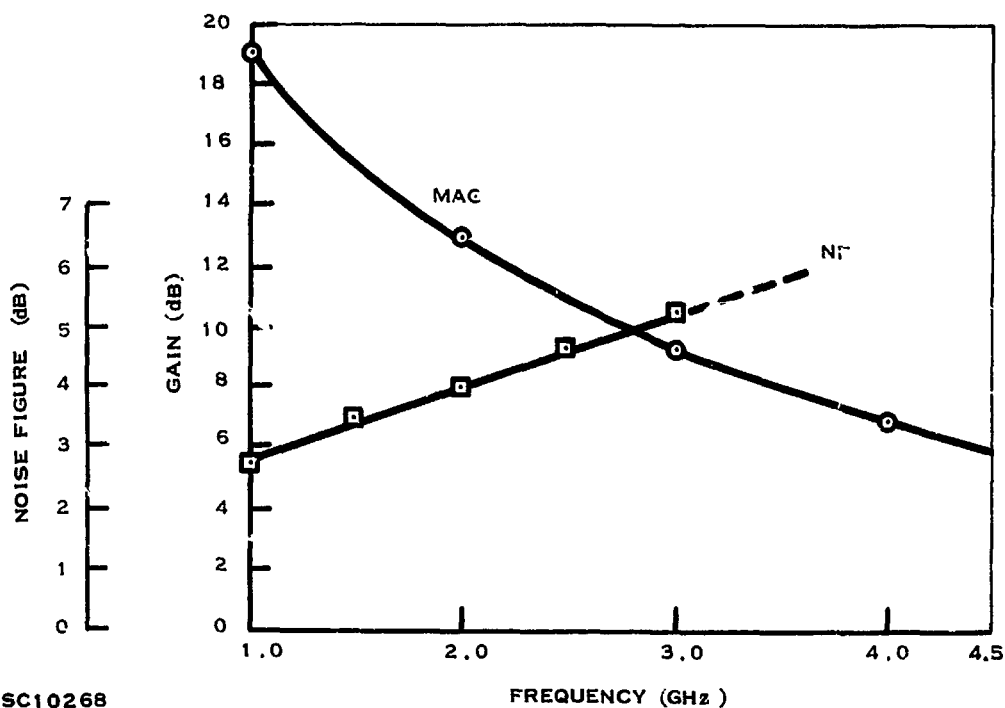
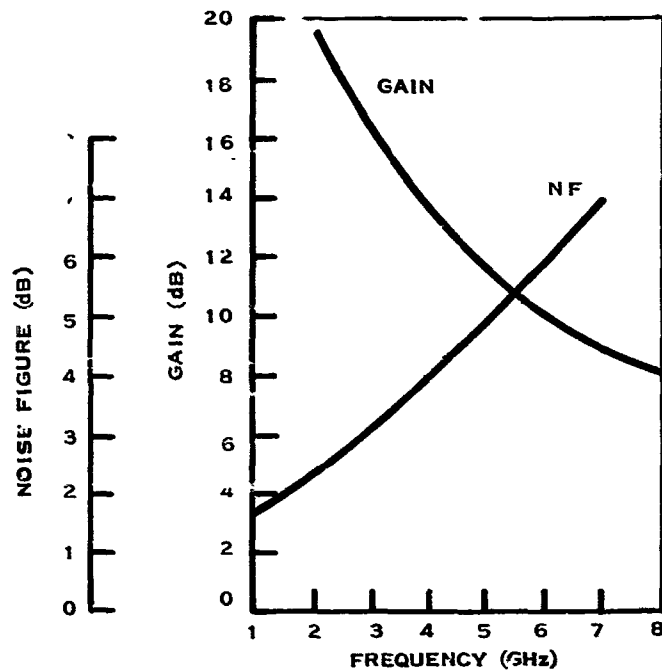


Figure 3-62. Maximum Available, Unneutralized, Common-Emitter Gain and Optimum Noise Figure

development for USAEC is the Texas Instruments L-196; its performance is shown in Figure 3-63. The L-196 data and information is given as a basis for prediction that a 2-dB further decrease in receiver noise figure at L- and S-band may be possible in the near future.

Figure 3-64 is a graphic illustration of noise-figure comparisons between the various circuits mentioned; it shows that a transistor amplifier is exceeded only by the parametric amplifier up to about 3 GHz.

An integrated-circuit tunnel-diode amplifier has been reported⁴⁰ which had a noise figure of 6.4 dB at 9.0 GHz. This is about 1 dB worse than that obtainable in conventional circuitry. The tunnel-diode amplifier (TDA) is a single-port negative resistance amplifier which is much simpler than the parametric amplifiers. The TDA is relatively small (0.5 × 1.0 inch) and generally requires 10 to 30 mW of bias power. It has less dynamic range than any of the circuits considered but this may not be a significant problem on the CAT radar.



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Figure 3-63. Predicted Noise Figure and Gain of L-196 Transistor

The nondegenerate parametric amplifier will provide the lowest receiver noise figure. Three disadvantages are: (1) it is a negative resistance device with multiple tuned circuits and would be considered the least reliable approach; (2) the parametric amplifier front end would be the largest and thus heavier than any approach; (3) it would require a pump source which most likely would be a Gunn or avalanche diode CW oscillator at X- to K_a-bands, depending on frequency of operation. Pump power of 10 to 50 mW is generally required which means 0.2 to 1.0 watt of input power per module would be required for a five-percent efficient oscillator.

The balanced mixer curves are essentially 2 dB at 450 MHz because Schottky Barrier diodes are used. The only difference is, one uses a lower frequency (10 to 100 MHz) if. amplifier and the other uses the MERA 500-MHz if. amplifier. A low-noise hybrid integrated circuit (0.5 × 0.5 inch) has been developed at Texas Instruments which has less than 1.5-dB noise figure (see Figure 3-65). The MERA if. amplifiers typically have a 3.5-dB noise figure. There are advantages to having a high if., though, and when a new transistor currently under development (1.5 to 2.0 dB at 450 MHz) is completed this would probably be the more attractive approach.

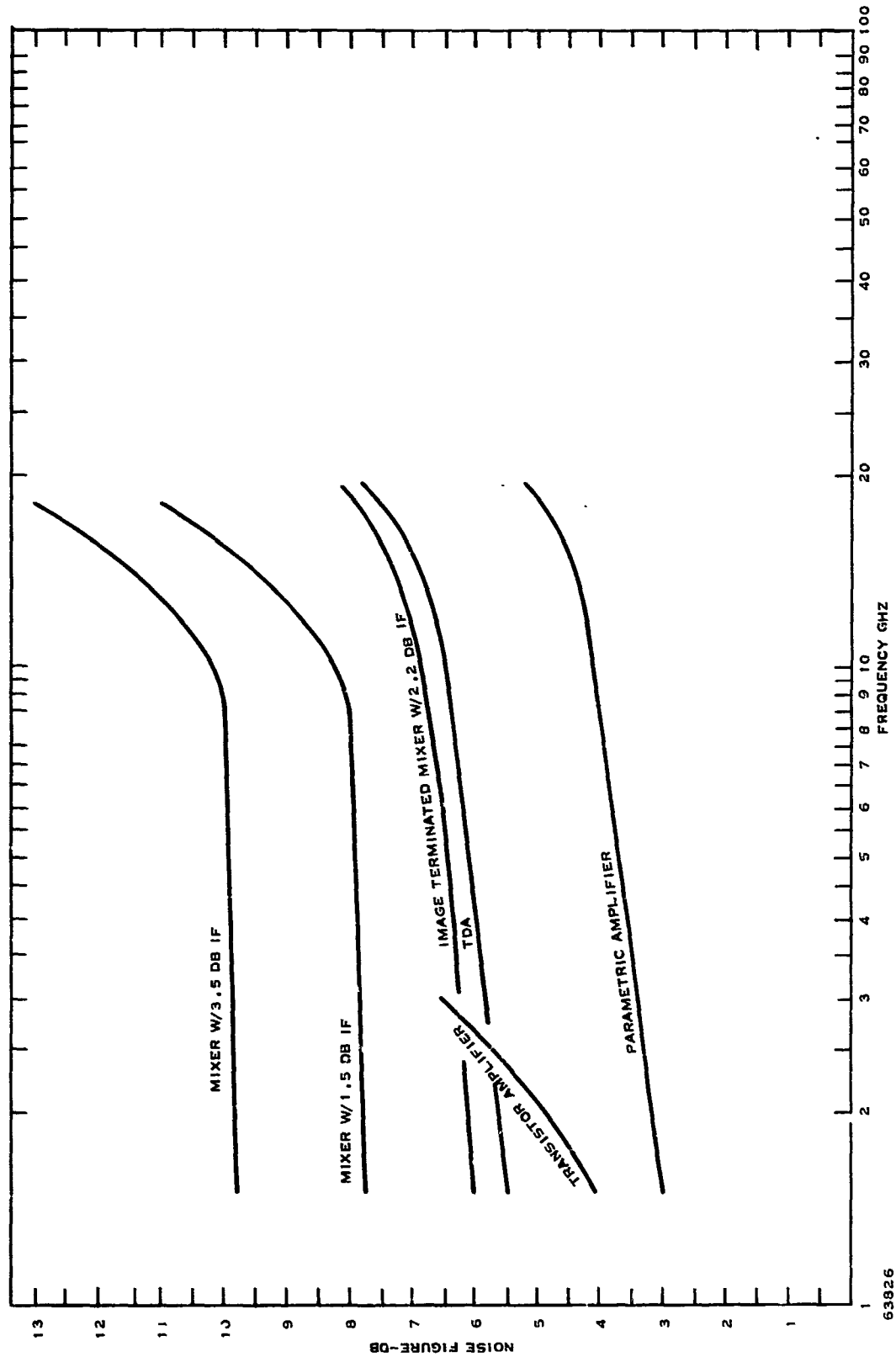


Figure 3-64. Receiver Performance for Rf Modules

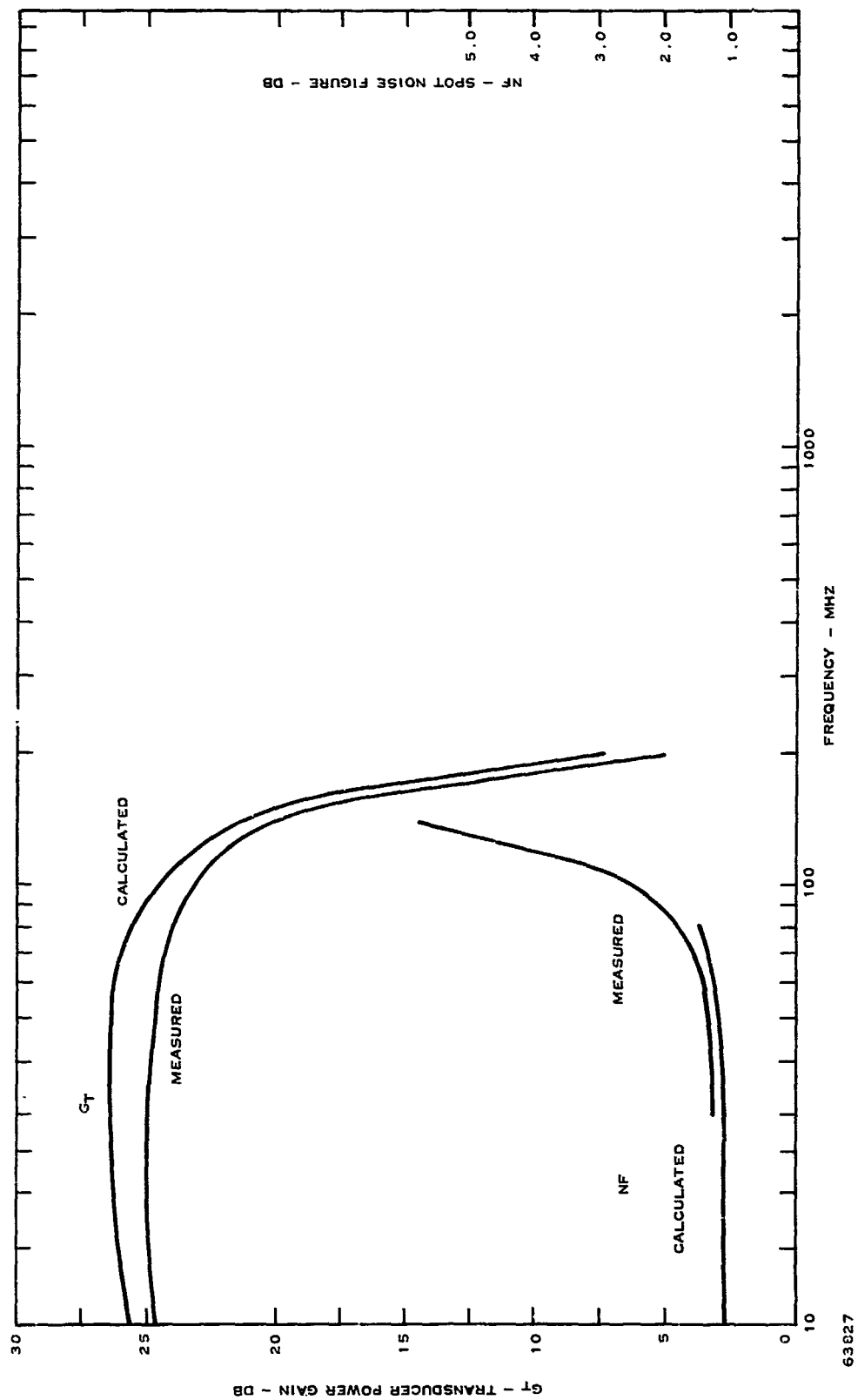


Figure 3-65. 10 to 110 MHz Video Amplifier

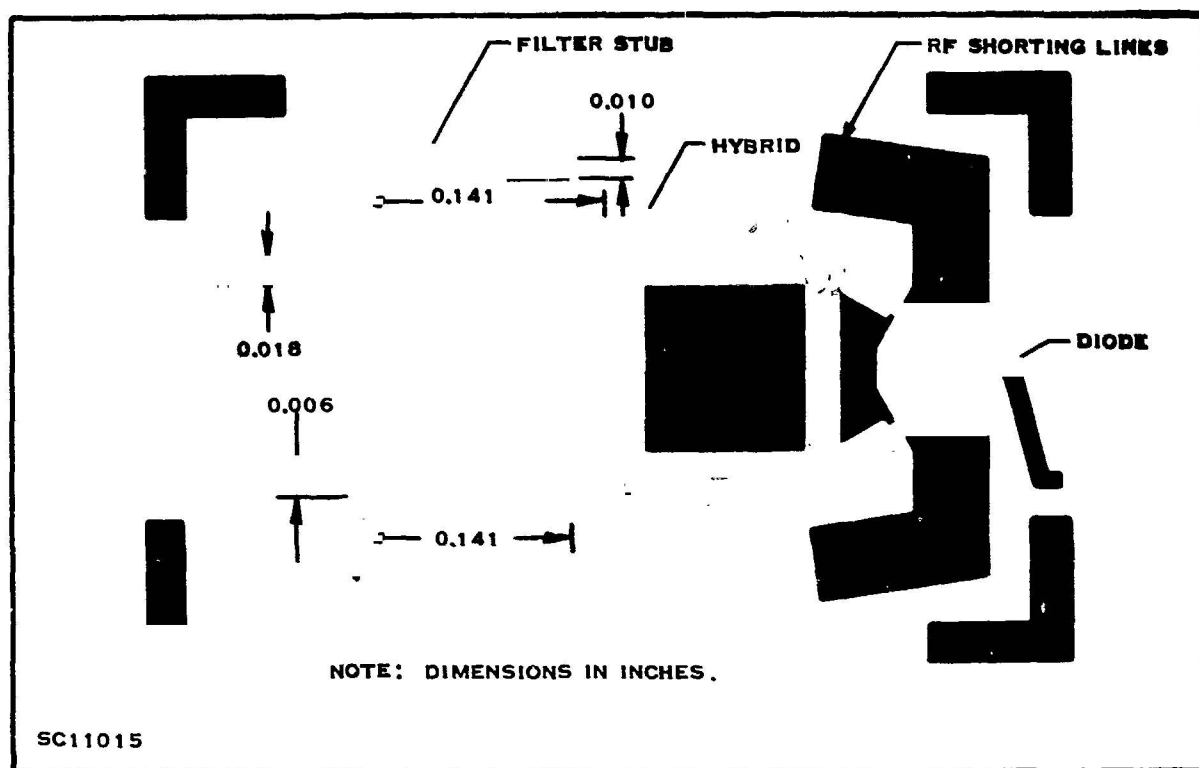
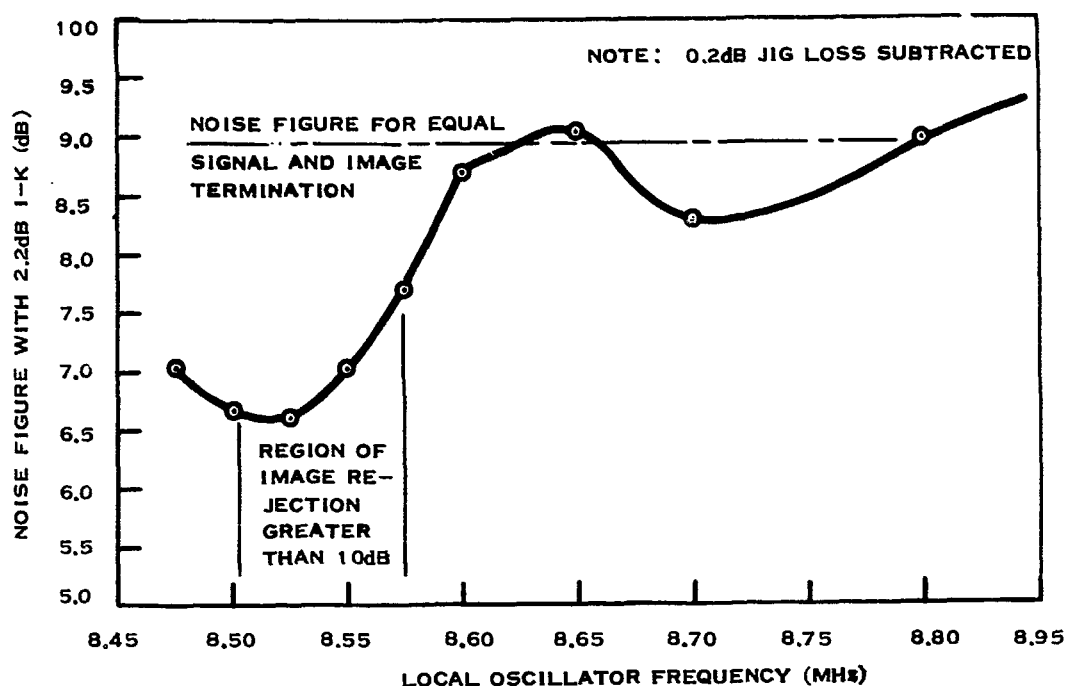


Figure 3-66. Image Terminated Mixer

For some applications, another solution is to replace the conventional balanced mixer with one in which the image is reactively terminated in an open or short circuit. The conventional mixer presents the same impedance to the image and signal terminals, resulting in one-half the energy supplied by the signal source being dissipated at the image frequency. An otherwise lossless converter must then have an inherent conversion loss of 3 dB. If, on the other hand, a purely reactive load is placed across the image terminals, the power dissipated is zero, and on this basis a lossless converter can be hypothesized. The low series resistance of Schottky barrier diodes⁴¹ makes possible the achievement of significantly lower noise figures with the image so terminated.

Texas Instruments has developed [on Contract AF 33(615)-67-C-1639] a low-noise mixer/preamplifier compatible with installation in the present MERA rf module. The image-terminated mixer⁴² utilizes low-loss GaAs Schottky diodes, and improved low-noise transistors are used in the preamplifier. The entire mixer/ifa amplifier occupies an area of only 0.38 square inch and has a noise figure of 6.6 dB. Figures 3-66 and 3-67 show the mixer and its performance. This mixer has a disadvantage; the if. must be at least twice the signal beamwidth and must be sufficiently high that an adequate filter can be constructed.



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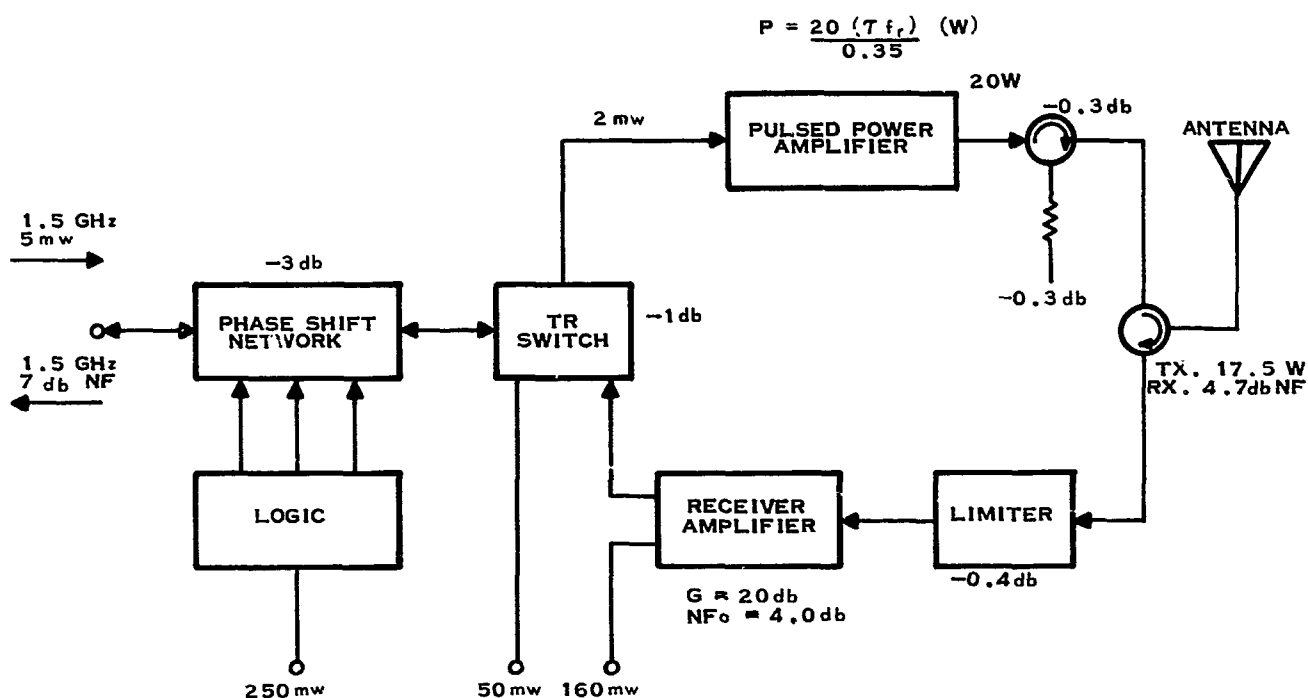
Figure 3-67. Image Terminated Mixer If. Noise Figure with a 2.2-dB If. Thin-Film Amplifier

3.2.2.7 Rf Modules Synthesis

The rf module block diagrams for the active arrays are shown in Figures 3-68 through 3-73. These circuits will be described briefly next, then summary and comparison data will be shown for use in the system trade-off analysis.

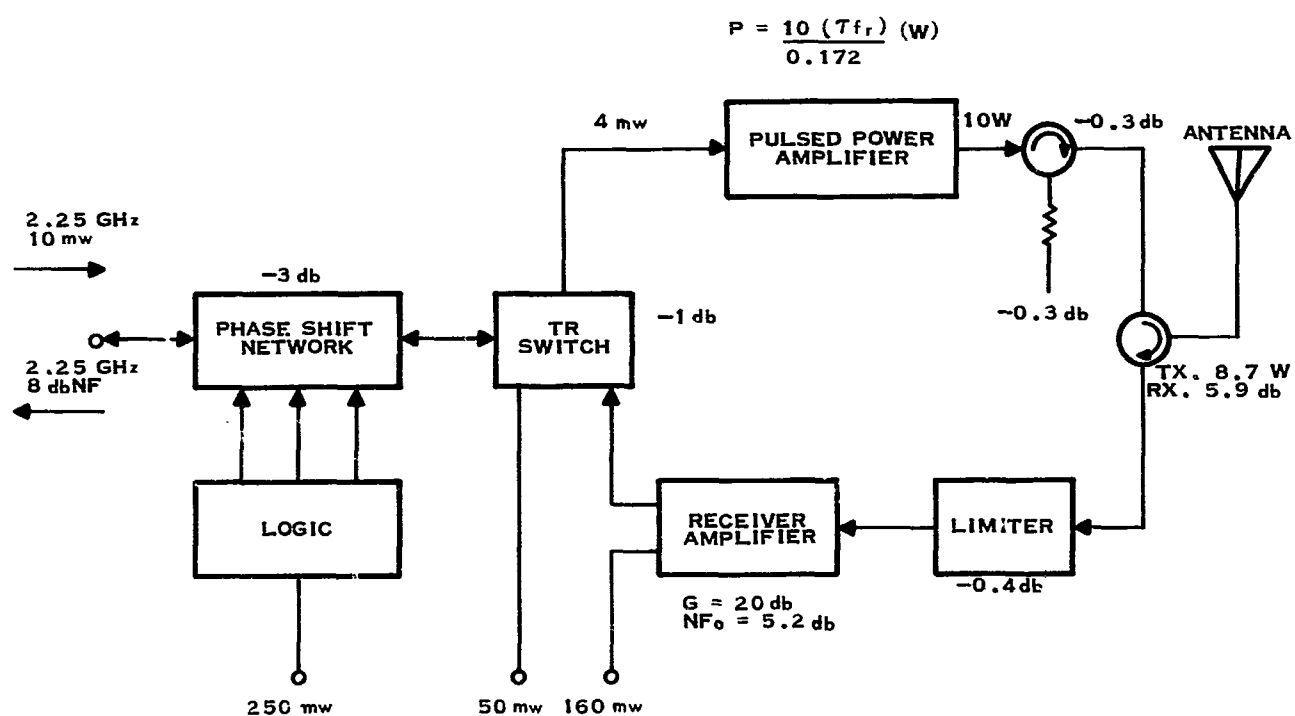
3.2.2.7.1 1.5-GHz Rf Modules

The 1.5-GHz module seen in Figure 3-68 has a single rf port for the input 1.5-GHz pulsed signal and the 1.5-GHz received signal. The module has no mixer in the receiver circuit; so, the antenna beam points in the same direction on receive as well as on transmit. This single frequency module allows the use of a simpler manifold which is smaller and lighter than a multifrequency manifold. A disadvantage, however, is that the same amplitude distribution must be used for the receive circuit as for the transmit circuit. This would be uniform distribution for maximum efficiency and so identical rf modules could be used in the array.



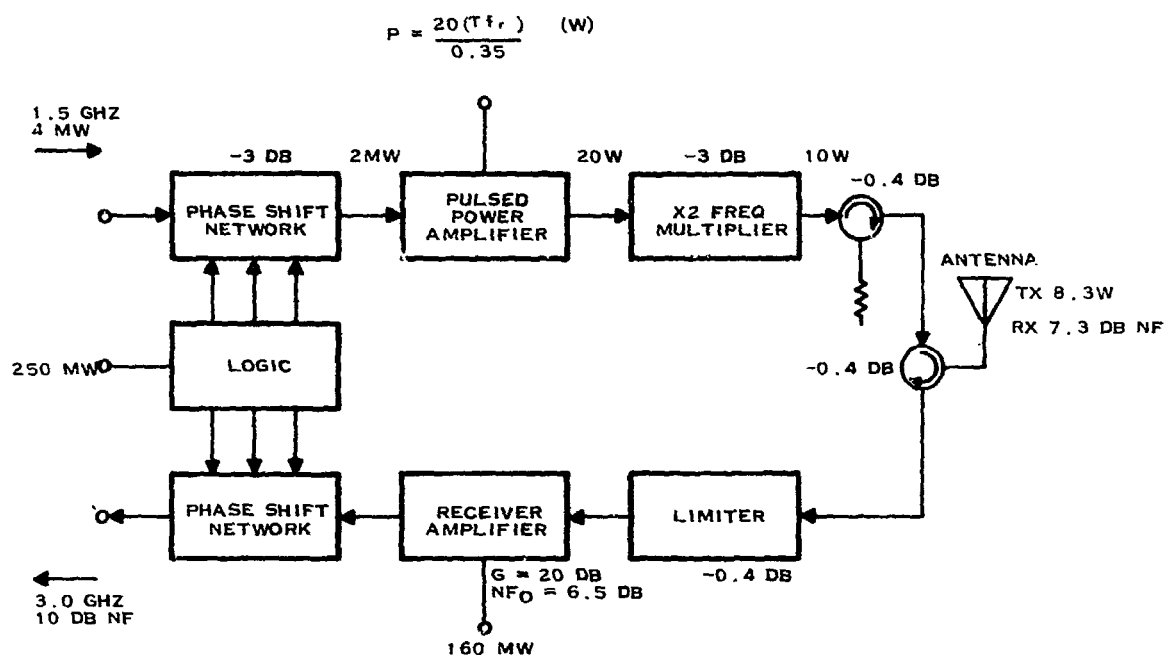
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Figure 3-68. 1.5-GHz Rf Module



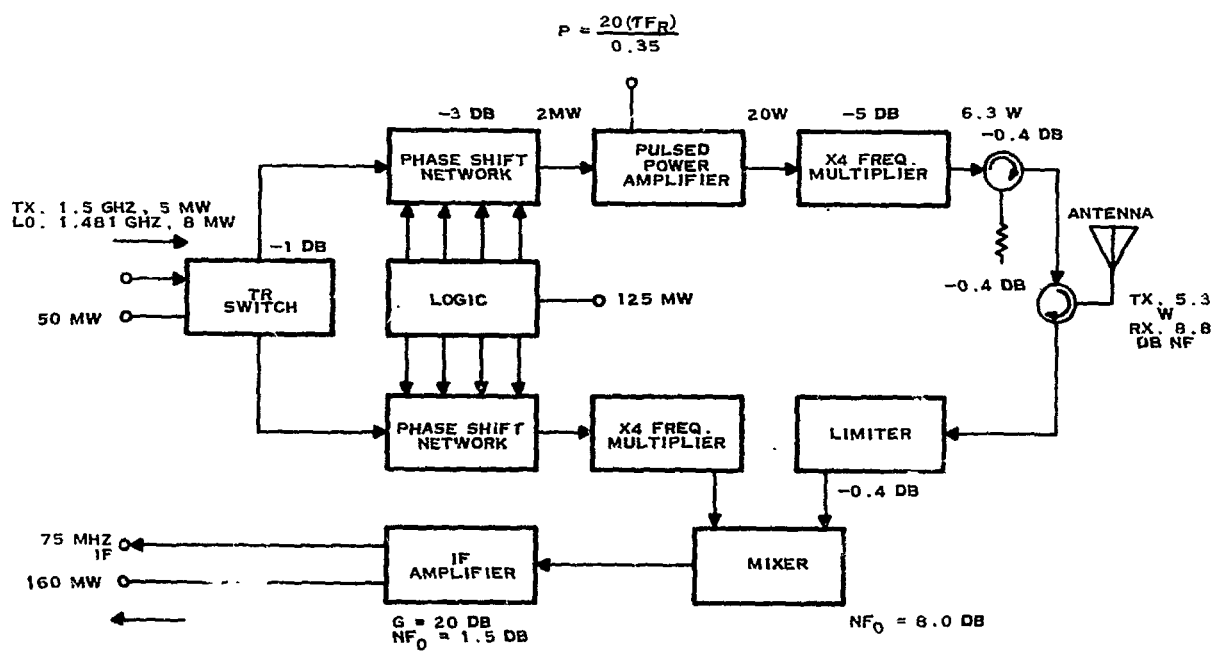
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Figure 3-69. 2.25-GHz Rf Module



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Figure 3-70. 3.0-GHz Rf Module



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Figure 3-71. 6.0-GHz Rf Module

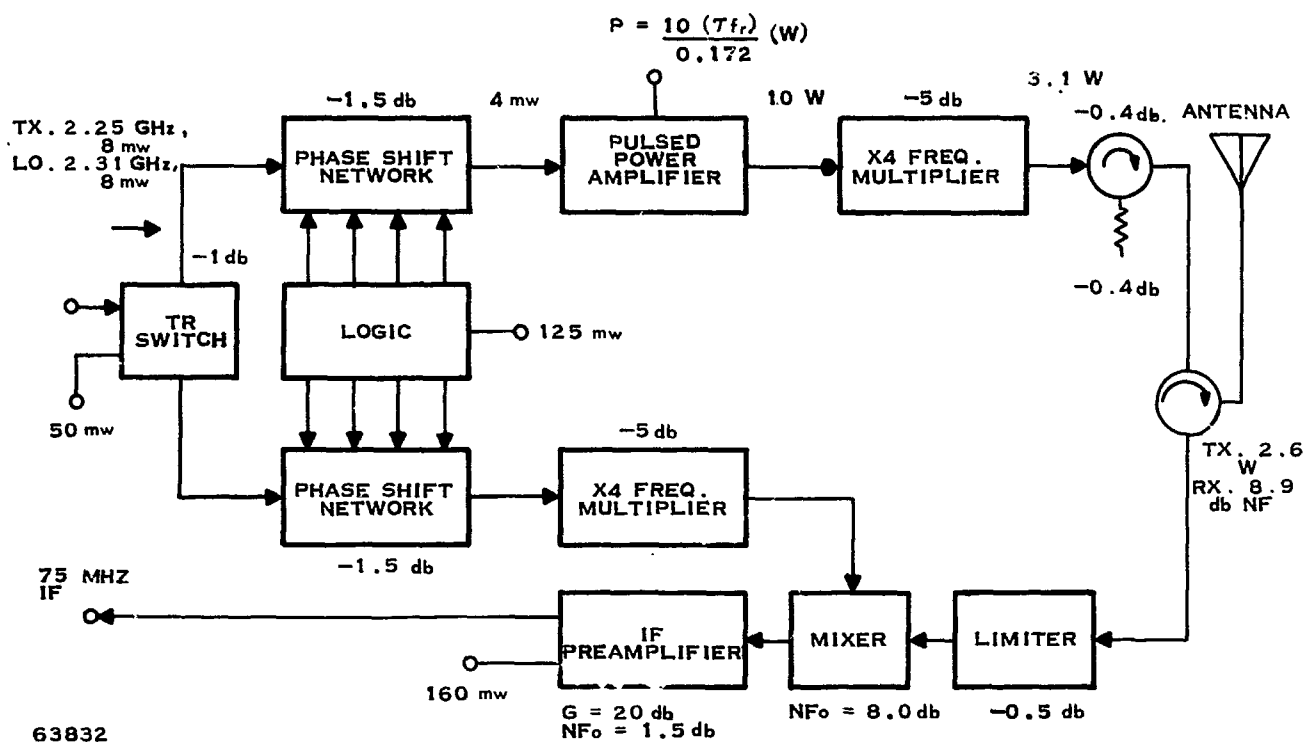


Figure 3-72. 9.0-GHz Rf Module

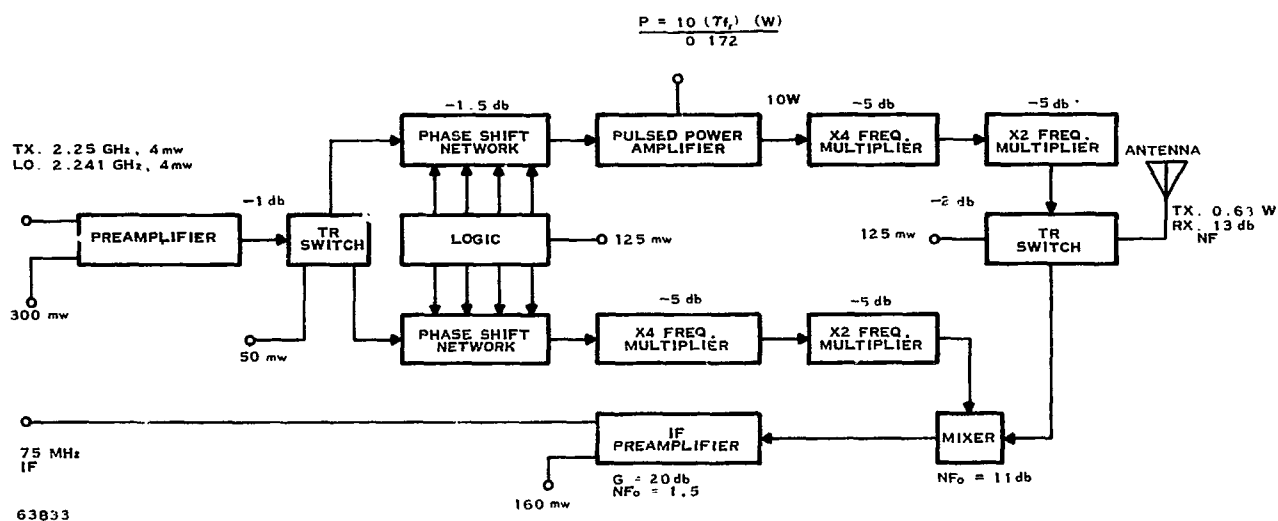


Figure 3-73. 18.0-GHz Rf Module

The phase-shift network, as described previously, has a three-bit phase shifter and requires 250-mW input power. The TR switch is a conventional two-diode switch requiring 50-mW input power. The pulsed power amplifier is the 20-watt output amplifier described in Figure 3-58e.

An isolator is used on the output of the power amplifier to prevent the varying impedance load of the scanning array from affecting its operation. (Circulators, for use in integrated circuits, have been built by Texas Instruments and several other manufacturers.) A circulator is used as the antenna duplexer. It does not require dc power and it will pass any power reflected off the limiter onto the isolator termination. Disadvantages of these ferrite devices as compared with the diode type duplexer are more size and weight.

A diode limiter is placed before the receiver amplifier to reduce receiver saturation effects and prevent possible burnout should a large signal return from the antenna. The limiter is passive in that it requires no dc power.

The receiver amplifier has 20-dB gain and a noise figure, NF_o , of 4.0 dB, which includes the following stage effects. Typical L-186 transistors at 1.5 GHz have a 3.6-dB noise figure. A two-stage amplifier such as this requires about 160 mW dc power.

The peak rf power (P_{pm}) out of the rf module is 17.5 watts. The input noise figure is 4.7 dB. The pulsed power amplifier is operated Class C so the dc input power is a function of the duty cycle (τf_r). The phase-shift network, TR switch, and receiver amplifier, require dc input power continuously. The total power into the module is then

$$P_{in} = 0.46 + \frac{20 (\tau f_r)}{0.35} \text{ watts.} \quad (3-8)$$

3.2.2.7.2 2.25-GHz Rf Module

The 2.25-GHz module block diagram is shown in Figure 3-69. It is essentially the same as the 1.5-GHz module. Peak transmitted power is 8.7 watts using the power-amplifier chain of Figure 3-58b. The input noise figure is 5.9 dB. Power into the module will be

$$P_{pm} = 0.46 + \frac{10 (\tau f_r)}{0.172} \text{ watts.} \quad (3-9)$$

3.2.2.7.3 3.0-GHz Rf Module

The 3.0-GHz module is shown in Figure 3-70. It is similar in many ways to the 1.5-GHz module. One difference is the use of a lower frequency power amplifier and a frequency doubler to achieve good efficiency with relatively high output power.

Another difference is the use of 2 three-bit phase shifters, one operating at 1.5 GHz and the other at 3.0 GHz. Low-noise transistor amplifiers are competitive with mixer/if. amplifier combinations at 3.0 GHz, therefore an rf amplifier is used to establish the system noise figure. This approach eliminates the need for a mixer and TR switch, and a separate LO signal. A 3.0-GHz rf manifold circuit would probably be smaller and less costly than a 75- or 500-MHz if. circuit also.

Peak output power from the module is 8.3 watts and the noise figure is 7.3 dB. The input power is

$$P_{in} = 0.41 + \frac{20 (\tau f_r)}{0.35} \text{ watts.} \quad (3-10)$$

3.2.2.7.4 6.0-GHz Rf Module

The 6.0-GHz module is shown in Figure 3-71. The input rf pulse passes through a diode TR switch to a four-bit phase-shift network. It is then amplified and multiplied to 6.3-watts peak at 6.0 GHz. The two circulators reduce this to 5.3 watts at the antenna. A limiter is again provided to protect the receiver.

The mixer /if. amplifier combination provides an 8.0-dB receiver noise figure. The if. should be as high as practical if image frequency filtering is desired. In a phased array with mixers in the rf modules, LO noise can add coherently in the if. manifold and degrade the system noise figure. For this reason, the LO signal is usually well filtered at the signal and image frequencies. For low if.'s then, a high-Q LO filter can be difficult to design. The problem is further complicated if LO tuning is required. In this case, a tunable yttrium-iron-garnet (YIG) filter would be used.

If a relatively narrow bandwidth system is satisfactory, the use of an image terminated mixer will provide a slight improvement in noise figure. Also, if system performance requires the use of the more complex TDA receiver circuit, then a 1.5-dB improvement in noise figure can be obtained.

The input noise figure is 8.8 dB and the input power is

$$P_{in} = 0.355 + \frac{20 (\tau f_r)}{0.35} \text{ watts.} \quad (3-11)$$

3.2.2.7.5 9.0-GHz Rf Module

The 9.0-GHz module is shown in Figure 3-72 and is similar to the 6.0-GHz module, except the power gain occurs at 2.25 GHz instead of 1.5 GHz.

Peak transmitted power is 2.6 watts and the noise figure is 8.9 dB. Module input power is

$$P_{in} = 0.355 + \frac{10 (\tau f_r)}{0.172} \text{ watts.} \quad (3-12)$$

3.2.2.7.6 18.0-GHz Rf Module

Figure 3-73 is a block diagram of the 18.0-GHz module. It is basically an extension of the 9.0-GHz module with X2 frequency multipliers in the transmit and LO circuits. The additional 5-dB loss in the LO circuit now requires the use of an S-band preamplifier to increase the LO drive. This preamplifier requires 300 mW of power which reduces module efficiency and increases the thermal load.

There are several design and packaging problems in this module, such as the X2 multiplier and small chassis size. The multiplier problem was discussed in a previous section. The connector problem could be approached by grouping 4 to 8 modules into a single large module with internal connections made on the ceramic microstrip.

The duplexer shown is a two-diode switch. Circulators suitable for microwave integrated circuits at this frequency have not been developed and are not included here.

The peak transmitted power is 0.63 watt and the noise figure is 13 dB.

The module input power is

$$P_{in} = 0.76 + \frac{10 (\tau f_r)}{0.172} \text{ watts.} \quad (3-13)$$

3.2.2.8 Rf Module Parametric Results

Now that the active rf modules have been described, we can determine various parameters pertaining to performance capability and make comparisons between the different rf modules. These results can then be used in the radar system synthesis to determine radar capabilities and make comparisons between different radars.

Parameters of interest are:

- Peak power per module
- Module noise figure
- Module gain
- Module input power
- Module heat load
- Module efficiency
- Number of modules per unit of prime power
- Array peak and average power
- Duty cycle
- Size
- Weight.

Average prime power to the antenna array is a basic restriction on total number of modules allowable. The number of modules then determines antenna gain and the peak, or average, power radiated. Duty cycle can be varied to make tradeoffs in number of modules, efficiency, and heat load.

Table III-12 provides a summary of the rf module parametric study results. The following discussion will elaborate on these results. A number of comparisons are made on each of the six rf modules when operated over a duty cycle range of 0.001 to 0.100.

3.2.2.8.1 Module Input Power

Module input power is a function of the continuous power, P_c , required in circuits such as the receiver amplifier, phase-shift network, and switching circuits, transmitter output power, P_o , and efficiency, η . Module input power is

$$P_{in} = P_c + \frac{P_o (\tau_{fr})}{\eta}, \quad (3-14)$$

where

P_c = Continuous power (receiver amplifier, phase shift network, TR switch, etc.)

P_o = Peak rf power out of final amplifier

τ_{fr} = Duty cycle of pulsed power-amplifier chain

η = Efficiency of power-amplifier chain.

The rf input power to a module is neglected here because it is much less than 1 percent of the total power required.

Referring to Table III-12, the average power into the 1.5-GHz module, for instance, varies from 0.517 watt for a duty cycle of 0.001, to 6.17 watts for a duty cycle of 0.10. The 0.0 duty cycle column refers to the continuous power, P_c , required in each rf module. A graphic presentation of the module input power versus frequency for fixed duty cycle is shown in Figure 3-74. Input power can be seen to be relatively independent of frequency at the higher duty cycles.

3.2.2.8.2 Module Peak Power Output

Module peak power output, P_{pm} , is also shown graphically in Figure 3-74. Peak output power falls off roughly with f^3 .

3.2.2.8.3 Noise Figure

Noise figure, NF, includes second stage contributions and can be regarded as the system noise figure.

3.2.2.8.4 Average Power Transmitted

Average power transmitted per module is the product of the module peak power times the operating duty cycle.

3.2.2.8.5 Heat Dissipation

Heat dissipation per module is the average power into the module, minus the average power transmitted. A large portion of the heat dissipated at the lower duty cycles is due to the continuous power circuits. Possible means of reducing this heat dissipation might be use of low-power latching ferrite phase shifters and micro-power receiver amplifiers, or gating all circuits off during portions of the receive time.

3.2.2.8.6 Number of Modules per 1 Kilowatt of Prime Power

Number of modules per 1 kilowatt of prime power is determined by dividing the module input power into 1 kilowatt. The 1 kilowatt of prime power in this case, refers only to that supplied the module array. It does not include any contribution due to power conditioning or to the other subsystems of the radar such as the receiver, signal generator, scan computer, etc. Figure 3-75 is a graphic presentation of these results. For fixed duty cycle, the number of modules is relatively constant with frequency. This says that antenna gain is relatively constant with frequency, since antenna gain is proportional to the number of modules (gain is approximately πN times antenna efficiency).

Table III-12. Rf Module Parameter Results

Frequency (GHz)	Module Input Power (W) Duty Cycle		P _{pm} (W)	NF (dB)	Average Power Tx/Module (W) Duty Cycle		Heat Dissipation/Module (W) Duty Cycle	
	0.0	0.001	0.005	0.01	0.05	0.10	(W)	
1.5	0.46	0.517	0.746	1.031	3.32	6.17	17.5	4.7
2.25	0.46	0.517	0.751	1.042	3.37	6.27	8.7	5.9
3.0	0.41	0.467	0.696	0.981	3.27	6.12	8.3	7.3
6.0	0.335	0.392	0.621	0.906	3.195	6.045	5.3	8.8
9.0	0.335	0.392	0.626	0.917	3.245	6.145	2.6	8.9
18.0	0.760	0.817	1.051	1.342	3.67	6.57	0.63	13.0
							0.0031	0.0063
							0.0315	0.0820
							1.052	1.340
							3.643	6.51

Frequency (GHz)	No. Modules per 1 kW of Prime Power Duty Cycle		Peak Power Tx per 1 kW of Prime Power (kW) Duty Cycle		Avg. Power Tx per 1 kW of Prime Power (W) Duty Cycle	
	0.001	0.005	0.01	0.05	0.10	0.001
1.5	1937	1340	970	301	162	33.82
2.25	1937	1330	960	296	160	16.82
3.0	2140	1437	1020	306	163	17.77
6.0	2550	1610	1103	313	166	13.51
9.0	2550	1600	1091	308	162	6.62
18.0	1225	952	743	272	152	0.77
						0.60
						0.47
						0.17
						0.10

Frequency (GHz)	Avg. Power to Tx Ckts. (W) Duty Cycle		Tx Efficiency (%) Duty Cycle		Module Efficiency (%) Duty Cycle		Module Tx Gain db
	0.001	0.005	0.01	0.05	0.10	0.001	0.05
1.5	0.357	0.586	0.871	3.16	6.01	4.7	14.9
2.25	0.357	0.591	0.882	3.21	6.11	2.5	7.4
3.0	0.307	0.536	0.821	3.11	5.96	2.7	7.7
6.0	0.232	0.461	0.746	3.03	5.88	2.3	5.8
9.0	0.232	0.466	0.757	3.08	5.98	1.1	2.7
18.0	0.657	0.891	1.182	3.51	6.41	0.1	0.4
						1.0	0.6
						1.0	1.0
						0.08	0.3
						0.47	0.86
						0.96	0.22

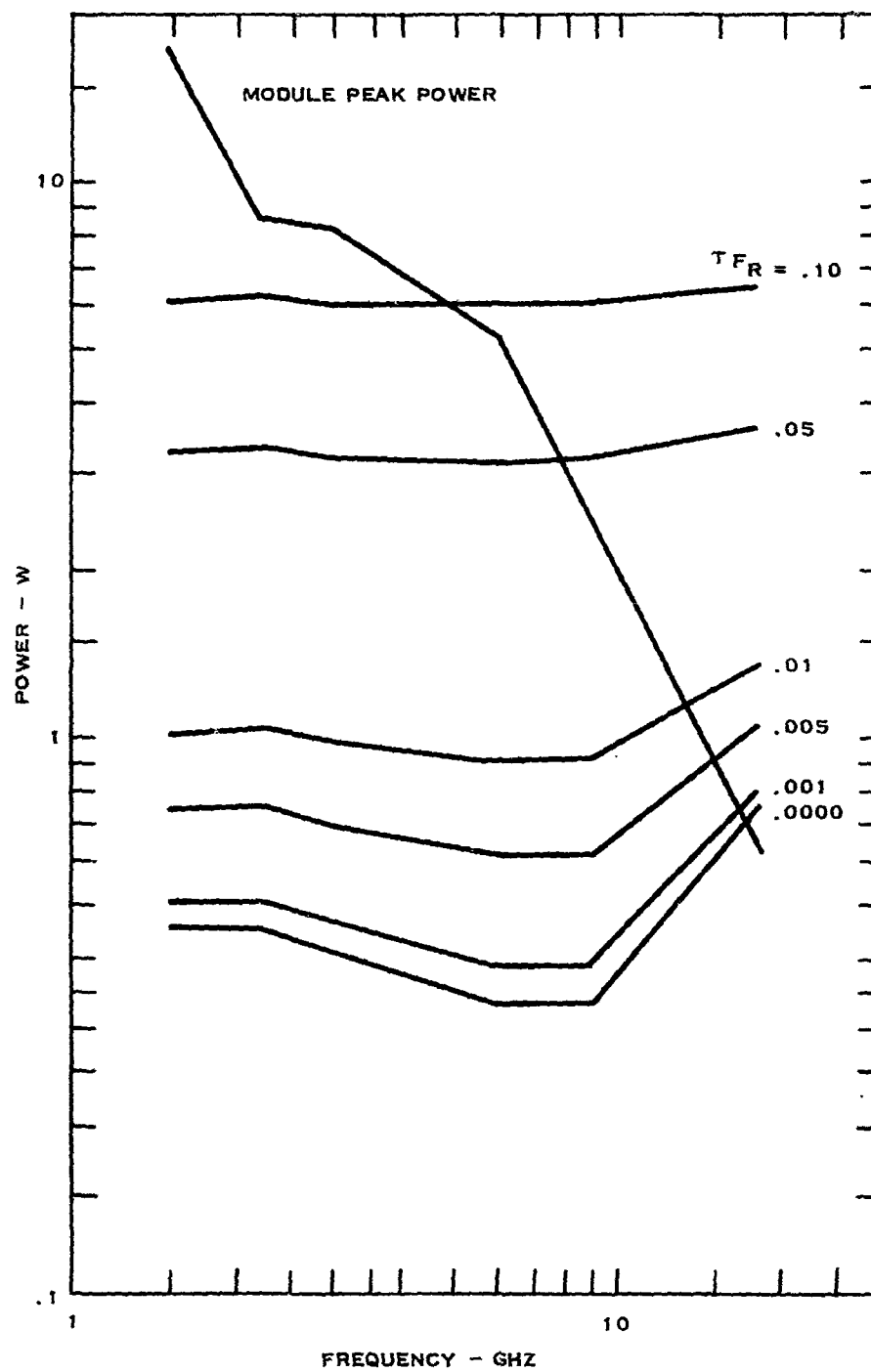
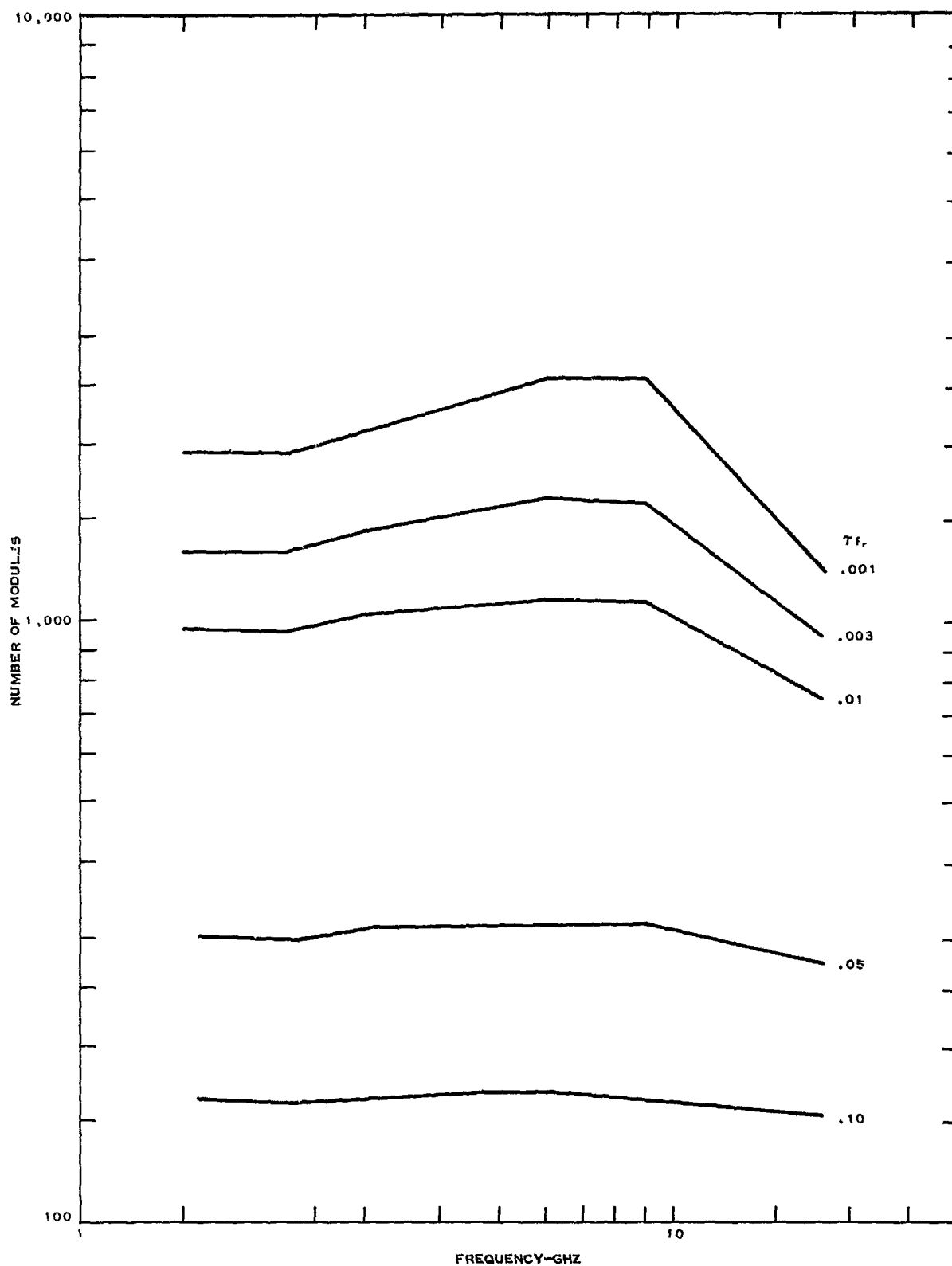


Figure 3-74. Module Input Power Versus Frequency for Fixed Duty Cycle and Module Peak Output Power Versus Frequency



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Figure 3-75. Number of Modules Per 1 kW of Prime Power Versus Frequency for Fixed Duty Cycle

3.2.2.8.7 Peak Power Transmitted per 1 Kilowatt of Prime Power

Peak power transmitted per 1 kilowatt of prime power is peak power per module times the number of modules allowed by 1 kilowatt of prime power. Figure 3-76 is a plot of these results. If additional loss occurs in the output circuits of a module, such as the addition of a polarization diversity circuit, this peak power will be scaled down directly by the power loss ratio. The highest peak power is achieved at the lowest frequency and duty cycle. It is relatively constant in the 2 to 6 GHz range.

3.2.2.8.8 Average Power Transmitted per 1 Kilowatt of Prime Power

Average power transmitted per 1 kilowatt prime power is the product of array peak power and duty cycle. Figure 3-77 is a graphical presentation of these results. The average power numbers are useful in comparing the active rf module array with tube transmitter-space fed array.

3.2.2.8.9 Average Power to Transmitter Circuits

Average power to transmitter circuits includes dc power going to all circuits performing functions required for transmitting; e. g., phase shifter, TR switch, and power amplifier. For all cases here average power is the module input power, minus the 160 mW required by the receiver amplifier.

3.2.2.8.10 Transmit Efficiency

Transmit efficiency is the average power transmitted, divided by the average power to the transmitter circuits. This is also shown in Figure 3-78. Note that efficiency improves with the duty cycle. This improvement is due to the nonactive circuits requiring a smaller proportion of the total required power, thus, efficiency becomes more dependent on the efficiency of the pulsed power amplifier. Module gain is shown and represents overall signal gain from module input to radiated output.

3.2.2.8.11 Module Efficiency

Module efficiency is the average transmitted power, divided by the average input power for the complete module. Figure 3-79 illustrates the drop in efficiency with frequency and duty cycle. As transistor efficiency above 2 GHz improves and better multipliers are developed, the curves will swing up. The lower duty cycle curves will be raised upward if the module continuous power can be reduced.

3.2.2.9 Bulk and Transit Time Device Considerations

Since early 1965, a little over three years ago, the field of solid-state microwave power generation has become very dynamic. Direct

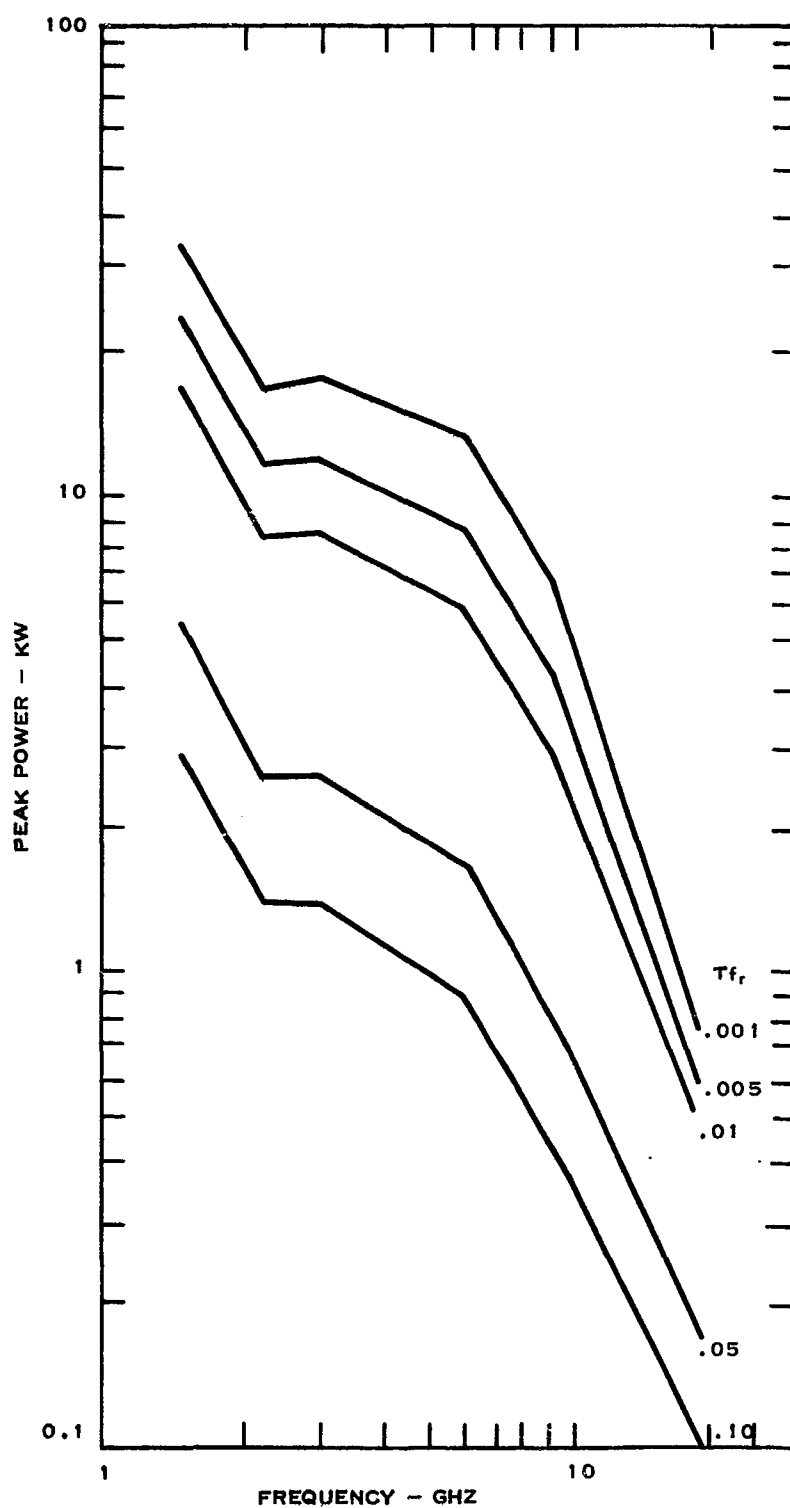


Figure 3-76. Peak Power Transmitted per 1 kW Prime Power Versus Frequency for Fixed Duty Cycle

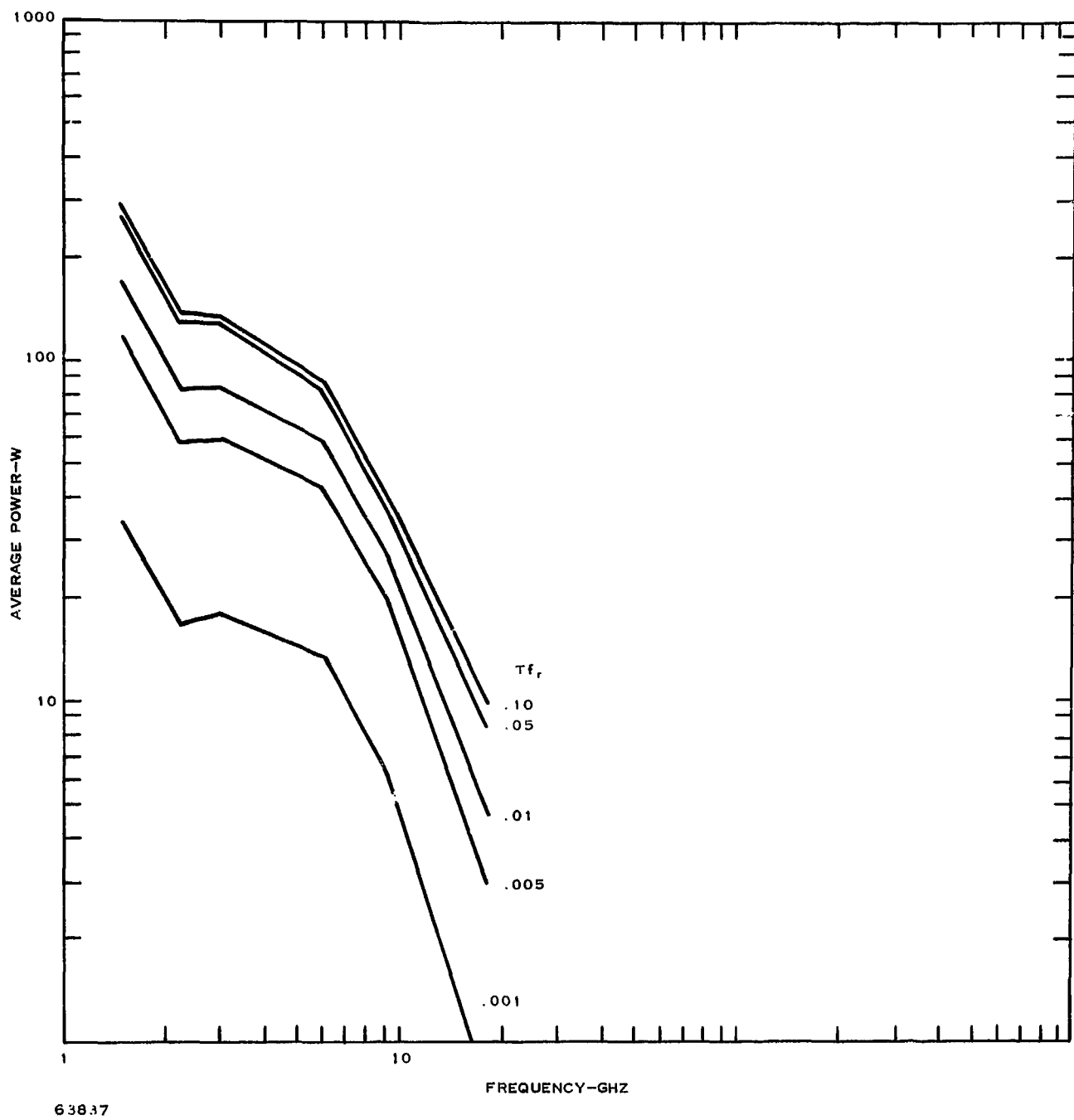
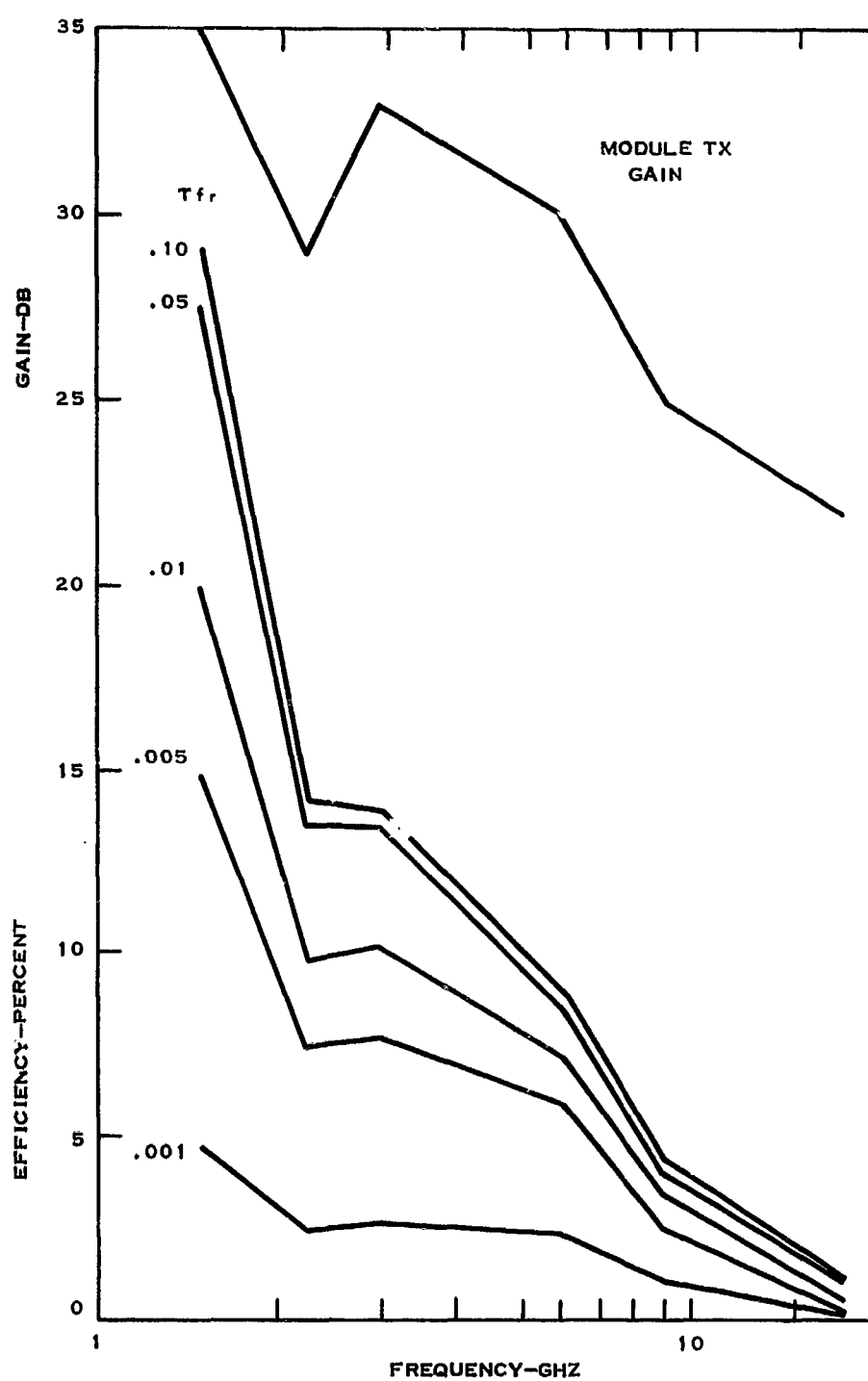
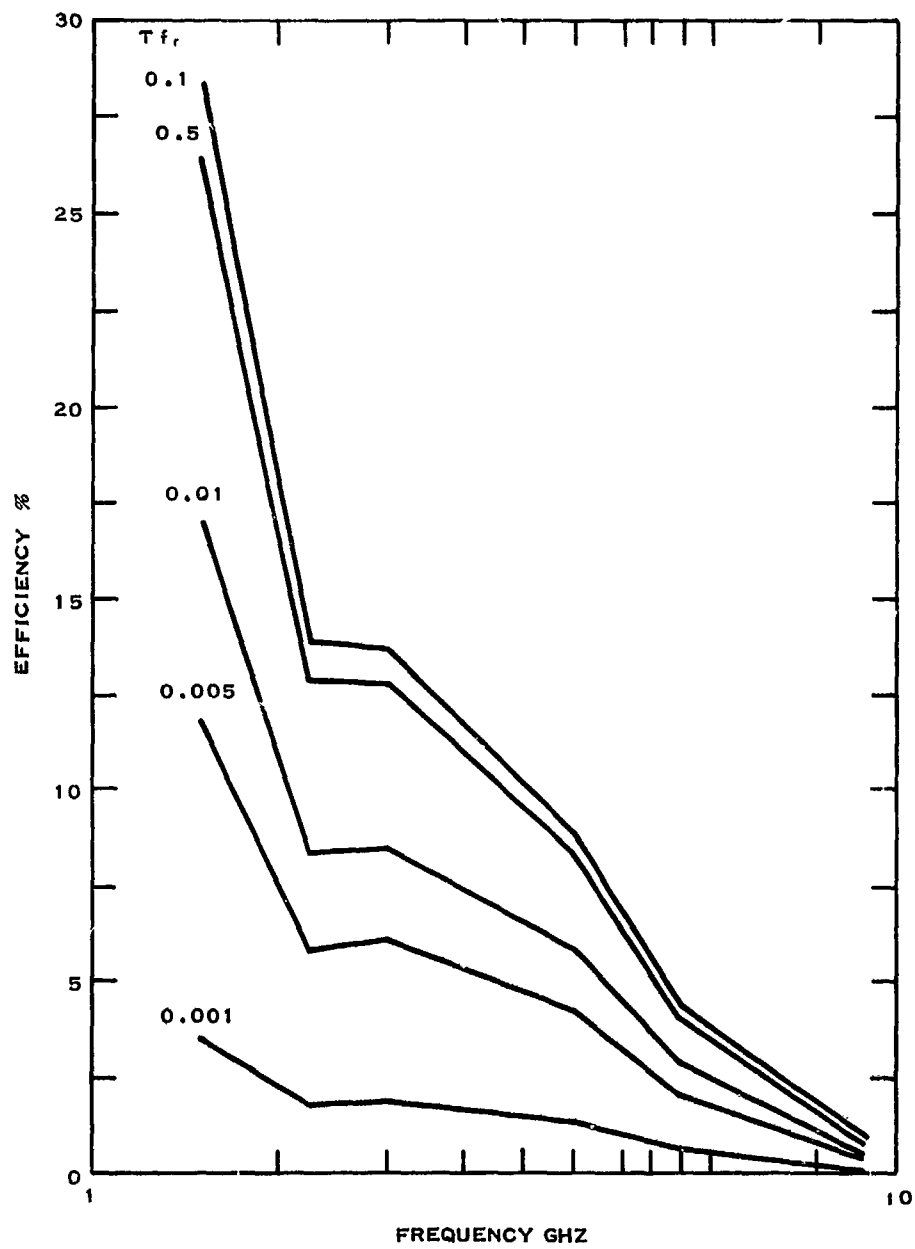


Figure 3-77. Average Power Transmitted per 1 kW Prime Power Versus Frequency for Fixed Duty Cycle



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Figure 3-78. Transmit Efficiency and Gain Versus Frequency for Fixed Duty Cycle



63839

Figure 3-79. Module Efficiency Versus Frequency for Fixed Duty Cycle

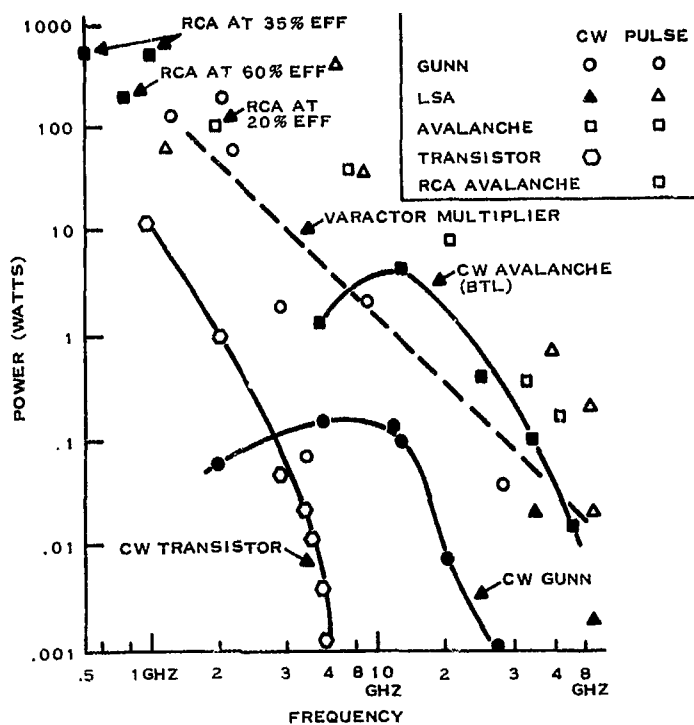


Figure 3-80. Solid-State Device Power to August 1967

3-80 shows the state-of-the-art⁴³ in output power versus frequency for solid-state devices. Although new results are being obtained every day in the laboratories, the output powers in each mode of device operation will generally follow the same $1/f^2$ limitation. The lower frequencies have not received attention due to larger thermal problems associated with the longer thermal path found in the thicker devices. This lack of attention is reflected by the drop in CW power curves for Gunn and avalanche devices.

Efficiency of these devices is reasonably limited to 12 percent for Read and simple p-n structure avalanche devices, 12 percent for the Gunn device, and 18.5 percent for the LSA device. The technology is still in a developmental period. Recent results by RCA on a low-transit-angle mode of avalanche oscillation shows up to 60 percent efficiency in the 1-GHz region. This efficiency has not been demonstrated at the upper microwave frequencies.

The results reported in almost every case are the most recent obtainable, sometimes, just before device destruction. Some of the avalanche and Gunn oscillators, or devices, are becoming reproducible and are being marketed now. Gunn devices will quite readily produce 1 to 20 mW CW and avalanche diodes produce up to 300 mW CW. All generally have 1 to 4 percent

generation of microwave power past the limits of transistor and tunnel-diode oscillators was made possible by the development of avalanche-diode, Gunn, and LSA devices. These new devices have shown the capability of developing considerable amounts of power in the millimeter wavelength region, and at very respectable efficiencies.

The theory of operation for these oscillators will not be discussed here because it is covered quite extensively in the literature (see references). It should be mentioned, however, that some of the intricacies of each oscillator are not completely understood as yet, but the field of endeavor is rapidly advancing in both theory and application.

Most of the effort has been towards generating higher rf power with better efficiency. Figure

efficiency. The avalanche devices are progressing more rapidly primarily because the silicon manufacturing and process control technology is so much more advanced than that of gallium arsenide. Because of this, the avalanche-diode oscillator would probably be the first choice consideration for uses in near future phased-array radar. The LSA mode of operation has an advantage of high power at high frequencies, but its reliability is of concern because of its fast turn-on requirement. The LSA mode oscillator has not had as much attention as has the Gunn oscillator.

Avalanche-diode oscillators may have problems due to high noise content and susceptibility to phase change with dc bias and rf drive. High-Q circuits would reduce the noise problem but would conflict with frequency tuning capability in a "chirped" radar.

Texas Instruments currently is developing a 9.0-GHz avalanche-diode oscillator, capable of 10-watts peak output at 10 percent duty cycle, which will be consistent with the use in MERA-type radars [contract AF 33(615)-67-C-190] .

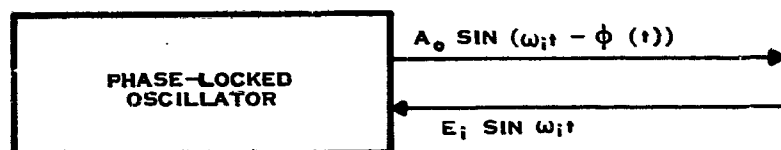
Present indications point to the more probable use of these oscillators at X-band or higher. At lower frequencies, the transistor amplifier/multiplier appears to have the edge.

In a phased-array application, the phase of the oscillator output signal must be controlled.

Modulation of the avalanche oscillator can be accomplished by several means:

- Direct modulation by superimposing baseband modulation on the diode dc bias; the avalanche diode frequency is current tunable
- Direct modulation by baseband driving of a varactor in the oscillator resonant circuit
- Phase modulation by phase-shift networks which switch the rf through different lengths of transmission line
- The oscillator can be locked in phase and frequency to a weak rf signal which is already modulated.

Direct modulation at rf, via the first two categories above, is perhaps the most simple. However, the required baseband controlling signals may differ substantially from diode to diode, circuit to circuit, or varactor to varactor—in short, from oscillator to oscillator.



CA14755

Figure 3-81. Schematic Diagram of Injection Phase Locking

Phase modulation by phase-shift switching networks is perhaps the most developed method and requires no detailed discussion. The fourth method, phase locking of the oscillator by injection of a weak rf signal, will be discussed in some detail, since this is the preferred modulation scheme.

The work of Stover and Shaw^{44, 45, 46} shows that avalanche-diode oscillators can be phase locked, in fact, both the steady-state and transient locking behavior can be described by the Adler equation.²⁷

$$\frac{d\phi(t)}{dt} = (\omega_i - \omega_o) - \Delta_o \sin \phi(t)$$

The phase angle $\phi(t)$, is the phase lag between the input and output signal (see Figure 3-81). The free-running frequency and amplitude of the oscillator are ω_o and A_o ; the output power, P_o , is proportional to A_o^2 . The frequency and amplitude of the injected signal are ω_i and E_i respectively. For a given injected power level P_i ($\propto E_i^2$), the oscillator will lock over a frequency range $2\Delta_o = 2\omega_i \max - \omega_o$. The oscillator diode and circuit may be characterized by Stover's locking figure of merit, η , such that

$$\frac{2\Delta_o}{\omega_o} = \eta \sqrt{P_i/P_o} \quad .$$

This locking figure of merit is best determined experimentally, although it can be related to diode and circuit parameters. The figure of merit has the form of a gain bandwidth product. In order for the above equations to be valid, $P_i \ll P_o$ and $2\Delta_o/\omega_o \ll 1/Q_L$, where Q_L is the loaded circuit Q.

When used as described above, the locked oscillator is equivalent to an FM or PM power amplifier; the power gain is equal to P_o/P_i . Stover⁴⁷ has shown that the modulation sidebands are affected just as though the signal had

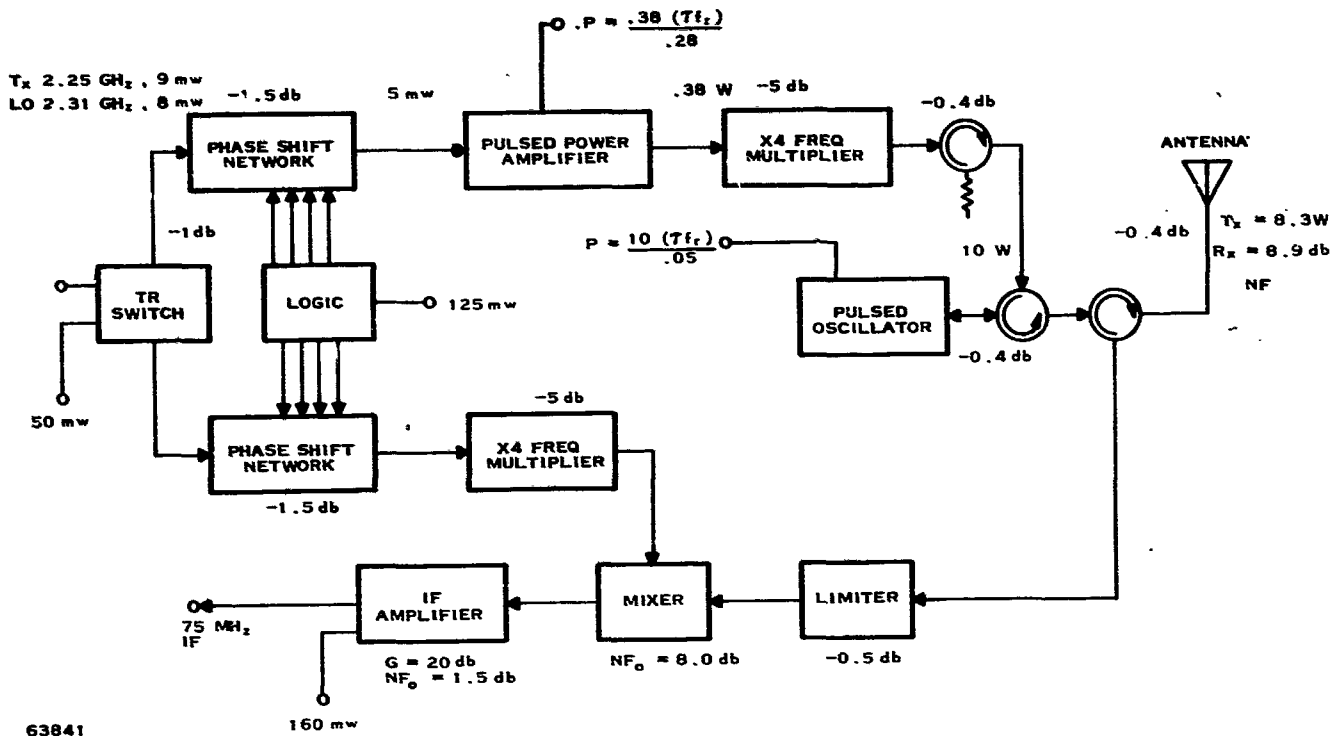


Figure 3-82. 9.0-GHz Rf Module Using Pulsed Oscillator

been passed through a first order Butterworth filter with bandwidth $2\Delta_o$. For a very modest locking figure of merit of $\eta = 0.2$, a 2-percent bandwidth (180 MHz at 9.0 GHz) can be obtained with 20 dB of power gain (that is, $P_i/P_o = -20$ dB; for a 10-watt oscillator, the injected signal is 100 mW).

A block diagram of a phased-array "element" utilizing an avalanche-diode oscillator is shown in Figure 3-82. The module shown is essentially the same as the 9.0-GHz active rf module previously discussed. Here, however, the latter stages of transistor amplification are replaced by the injection-locked avalanche-diode oscillator. The oscillator is essentially a 20-dB gain amplifier. The transistor-amplifier and frequency-multiplier circuits now operate at a lower power level. The oscillator diode junction (5 to 10 milliradians diameter), however, is handling an extremely high current density. This problem can be reduced by utilizing several devices in series or parallel.

Assuming a 5-percent efficiency oscillator is available which can deliver 10-watts pulsed power, then the rf module output is 8.3 watts. The complete transmitter circuit efficiency is 2.2 percent for 0.001 duty cycle and 3.8 percent for 0.01 duty cycle. This compares to 1.1 and 3.5 percent, respectively, on the previously described 9.0-GHz rf module. This shows that

the oscillator efficiency will be approached at 1-percent duty cycle and higher. This approach would be more attractive at K_u -band, where multiplier efficiency produces a much lower overall transmitter efficiency.

In conclusion, it is felt under the CAT radar study constraints of trying to get maximum radar range (200 nm) using today's technology, the avalanche-diode, Gunn and LSA oscillators would not be recommended. Primary considerations were: (1) relative low values of reproducible/reliable output power and efficiency, (2) lack of experimental data showing feasibility of phase-controlled diode oscillators operating in a phased array, (3) junction thermal load, and (4) lag of Gunn and LSA device work due to the immaturity of GaAs technology. The oscillators begin to compare favorably with the transistor amplifier/multiplier approach at X-band and higher.

These conclusions could be outdated at any moment, due to a considerable amount of work being done by Government agencies and industry on the design, development, and use of these devices. It is recommended that just prior to a CAT radar development the state-of-the-art be again reviewed.

3.2.2.10 Rf Manifold

The basic functions of the rf manifold are:

To take the rf signal produced in the signal generator, divide it, and deliver equal-amplitude/equal-phase signals to each rf module in the array

To combine the received signals from each module and deliver them to the receiver. The network may incorporate a monopulse comparator to develop sum-and-difference signals in azimuth and elevation. It may also incorporate a weighted combining network to obtain a particular antenna pattern (lower sidelobes).

For the type of phased array being considered here, a manifold similar to that developed in the MERA program has the best qualifications. That manifold has the following characteristics:

1 rf input 604 outputs 2.125 and 2.25 GHz

604 if.inputs 2 outputs 500 MHz

Approximate Taylor weighting on receive

Two-lobe monopulse

±1-dB amplitude variation

±10-degrees phase variation

Three-dB loss on transmit and receive

Weight, 41 pounds.

Block diagrams of the S-band distribution and if. summation networks are shown in Figures 3-83, 3-84, and 3-85, respectively. The S-band input is split into four parts. Each part is fed to a quarter section of the array. At this point, each signal is split into 39 parts by a 39-way feed section. Each of the 39 signals is then split four ways by a submanifold. This results in 156 equal-amplitude, equal-phase signals per quarter section, but since only 151 are required, five are resistively terminated.

At the if. frequency, groups of four inputs from the rf modules are uniformly summed in the submanifolds. At this point, there are 39 if. signals per quarter section. Each of these signals is amplitude weighted. The 39 weighted signals are summed to a single output in a 39-way combiner. Finally, the four if. signals (one from each quarter section) are combined in a standard monopulse comparator circuit to provide monopulse sum and difference signals.

Physically, the manifold structure consists of several layers of stripline boards, stacked parallel to the plane of the array, as shown in Figures 3-86 and 3-87. Connections consist of right angle miniature stripline-to-coaxial connectors.

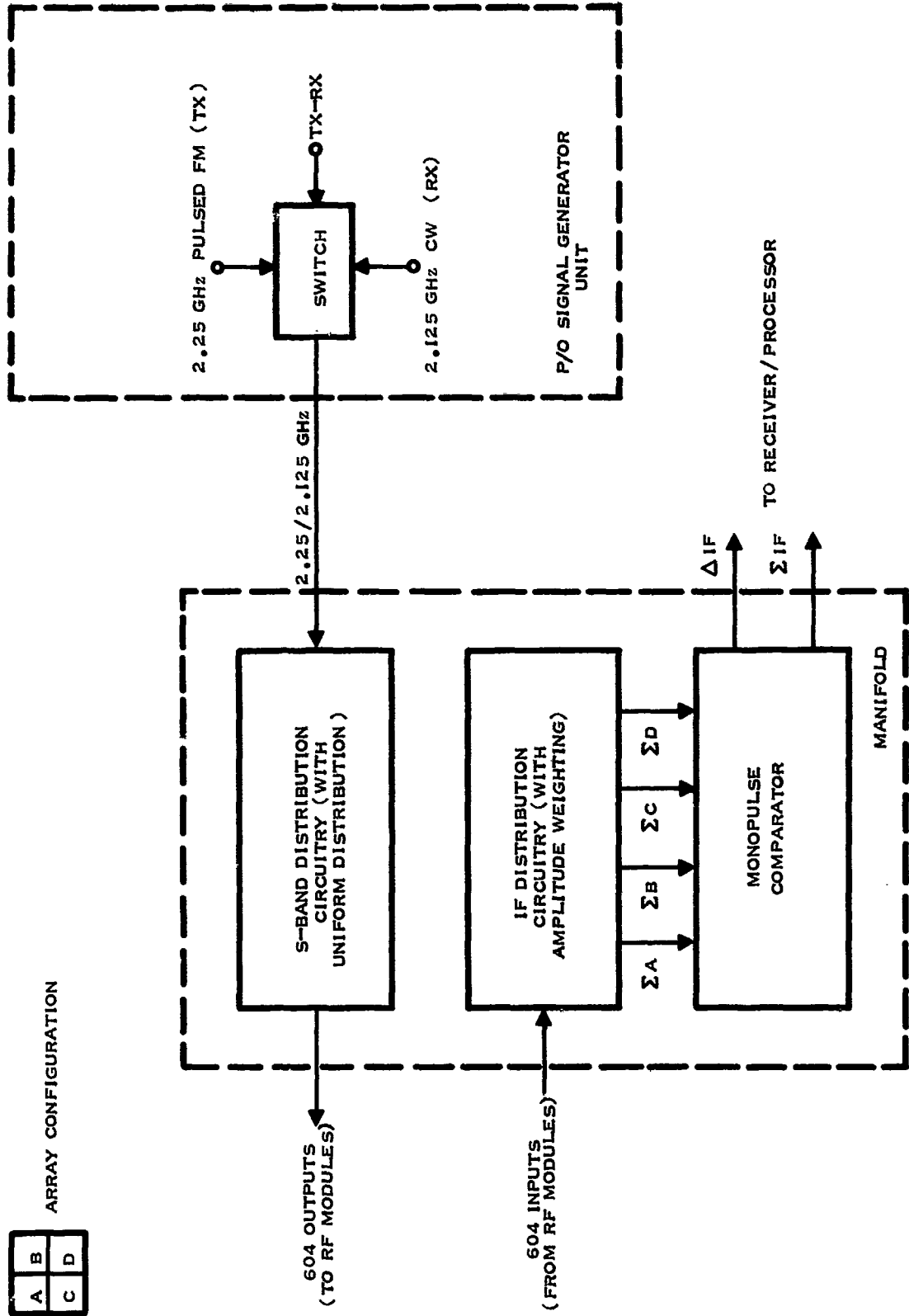
Figures 3-88, 3-89, and 3-90 are photographs of the manifold.

Figures 3-91 and 3-92 show S-band amplitude and phase distribution, respectively, for all 604 outputs. Figure 3-91 shows actual insertion loss from the manifold S-band input to any given output. Since theoretical insertion loss due to distribution is 28 dB, average insertion loss over and above distribution loss is about 3 dB at both 2125 MHz and 2250 MHz. Although total amplitude variation at either frequency is greater than ± 1.0 dB in some places, the distribution does have roughly a Gaussian shape.

The S-band phase distribution in Figure 3-92 is shown in the form of relative phase between all of the 604 outputs rather than in the form of total insertion phase. Again, the tendency is toward a Gaussian shape.

Figure 3-93 shows the 500-MHz phase distribution in the form of relative phase between all of the 604 inputs. Seventy-nine percent of all 604 inputs are within ± 5 degrees of each other, and 99 percent are within ± 10 degrees.

Deviation from the required Taylor amplitude distribution for each of the 604 if. inputs is shown in Figure 3-94. These numbers were obtained by computing the average dissipative insertion loss of each if. quarter section, subtracting this average loss from the total insertion loss measured at each port, and comparing the numbers thus obtained with the Taylor distribution. As is seen, the deviations are generally only a few tenths of a decibel with the exception of the ports around the outside of the array. In each quarter section there is a four-port 48-dB coupler which feeds these outside ports. At this coupling level, the rate of change of coupling with line spacing between coupled lines is about 0.1 dB/0.001 inch; in addition, etching tolerances



SCO 5082A

Figure 3-83. Rf Manifold Block Diagram

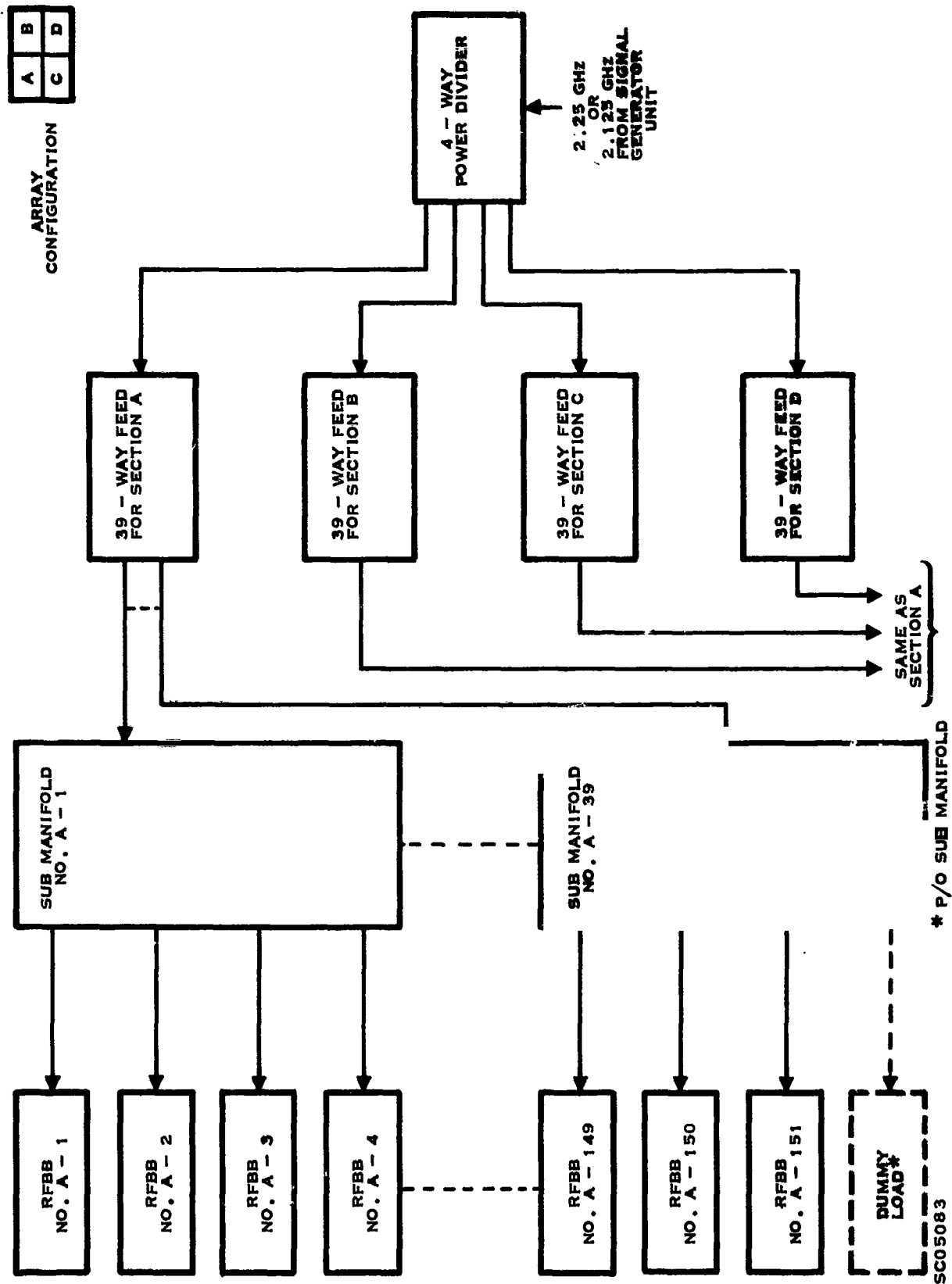


Figure 3-84. S-Band Distribution System

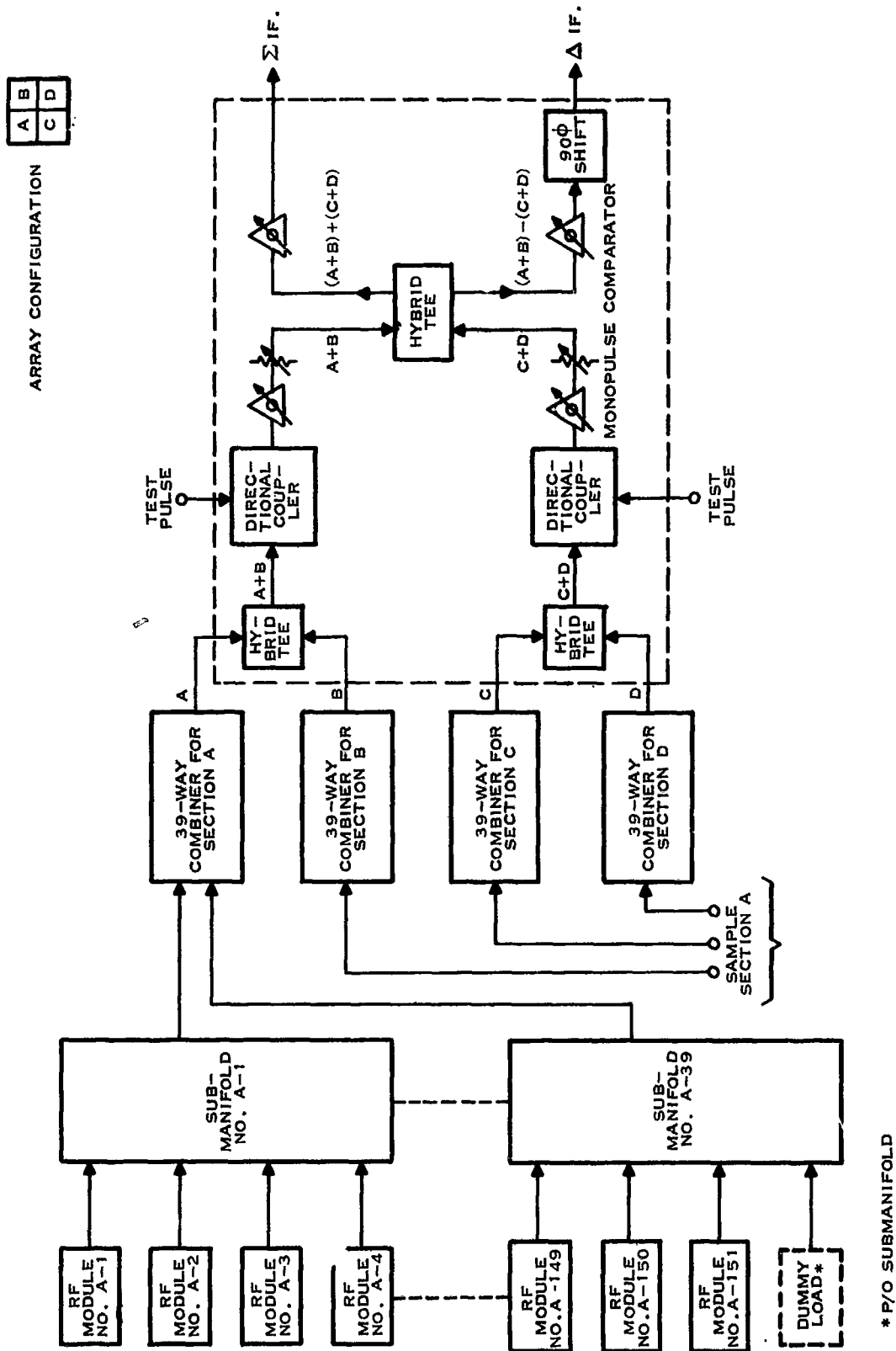
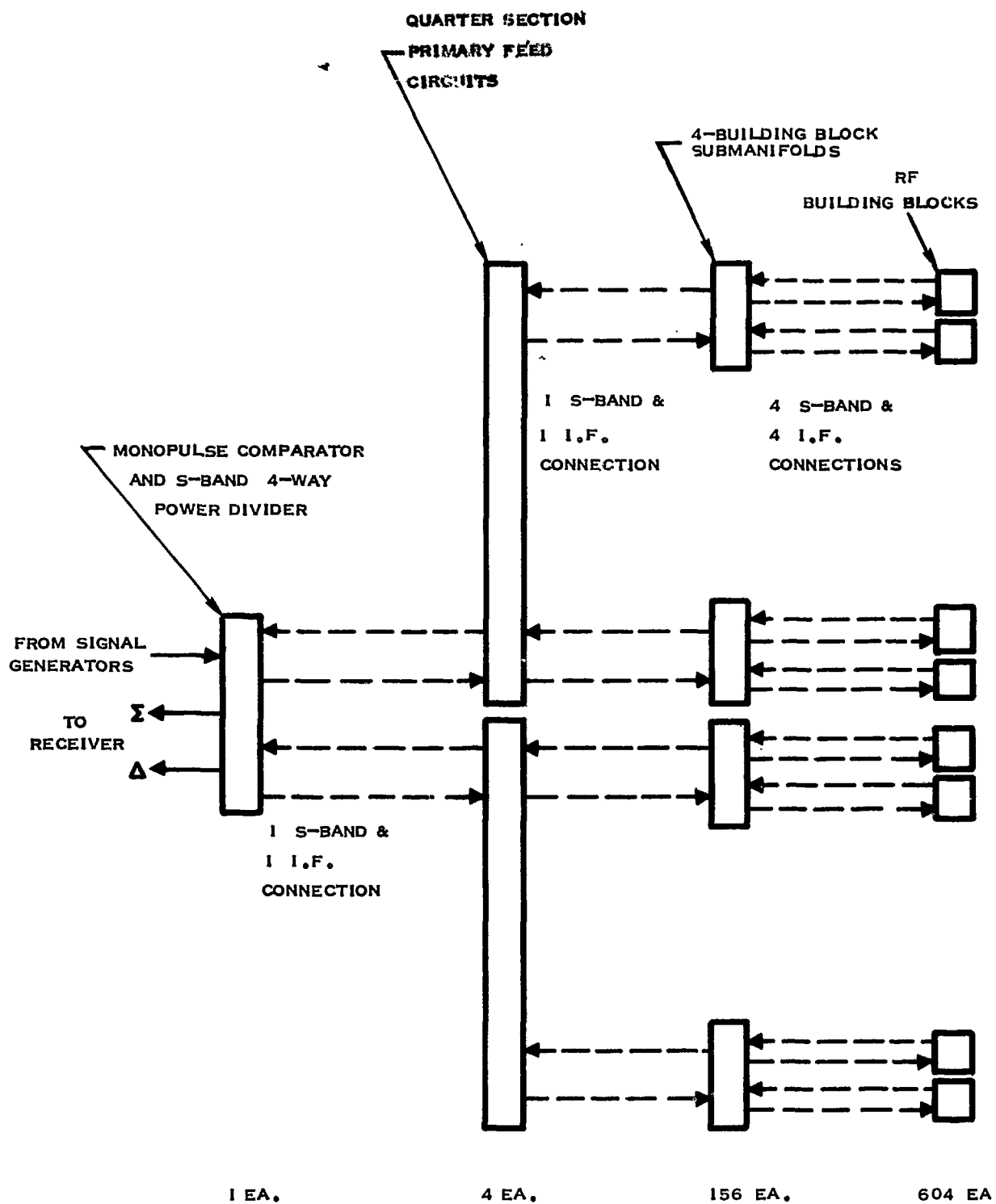


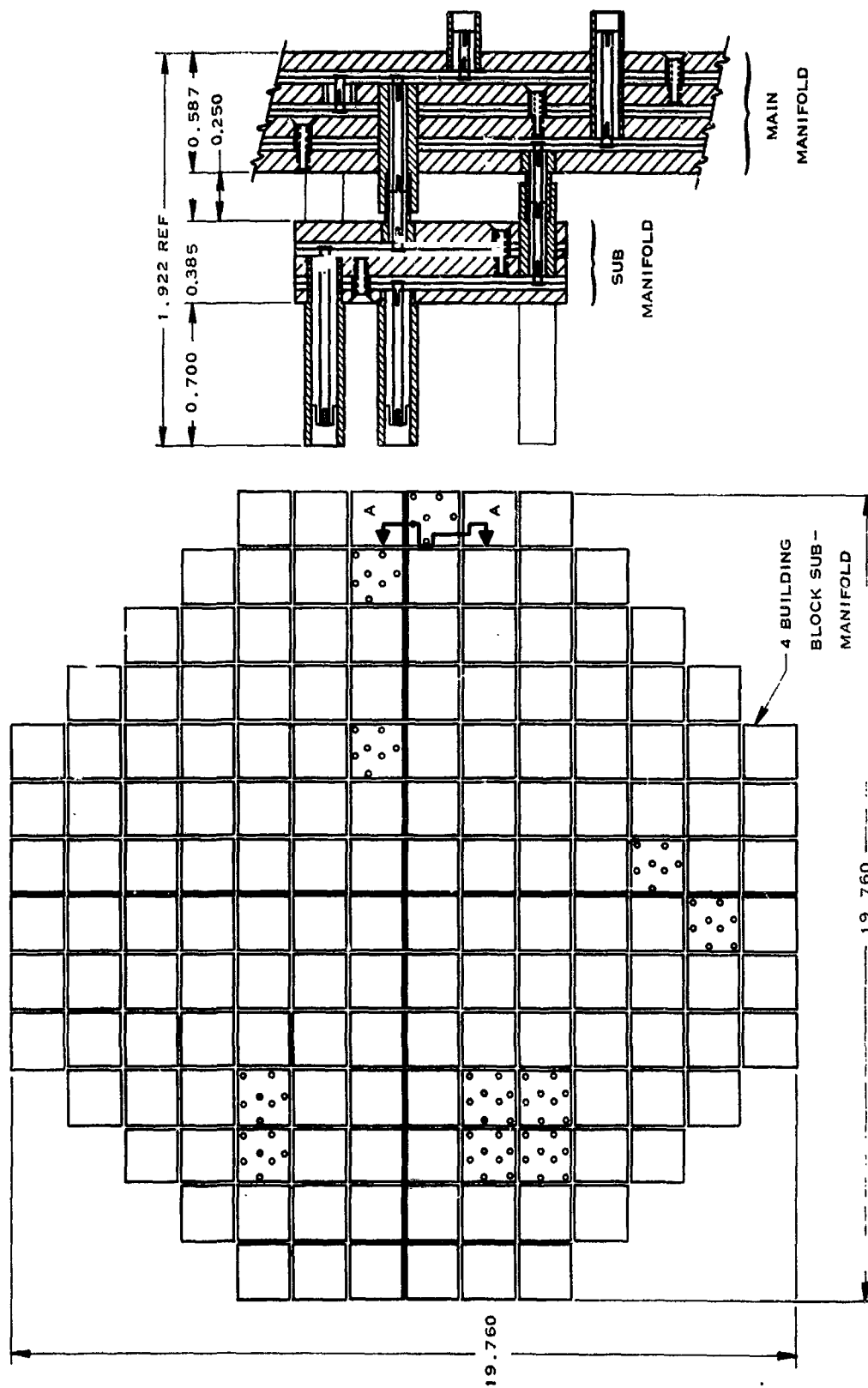
Figure 3-85. If. Distribution and Monopulse Comparator

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SC07433

Figure 3-86. Corporate Feed Structure



SC07432

Figure 3-87. Rf Manifold Configuration

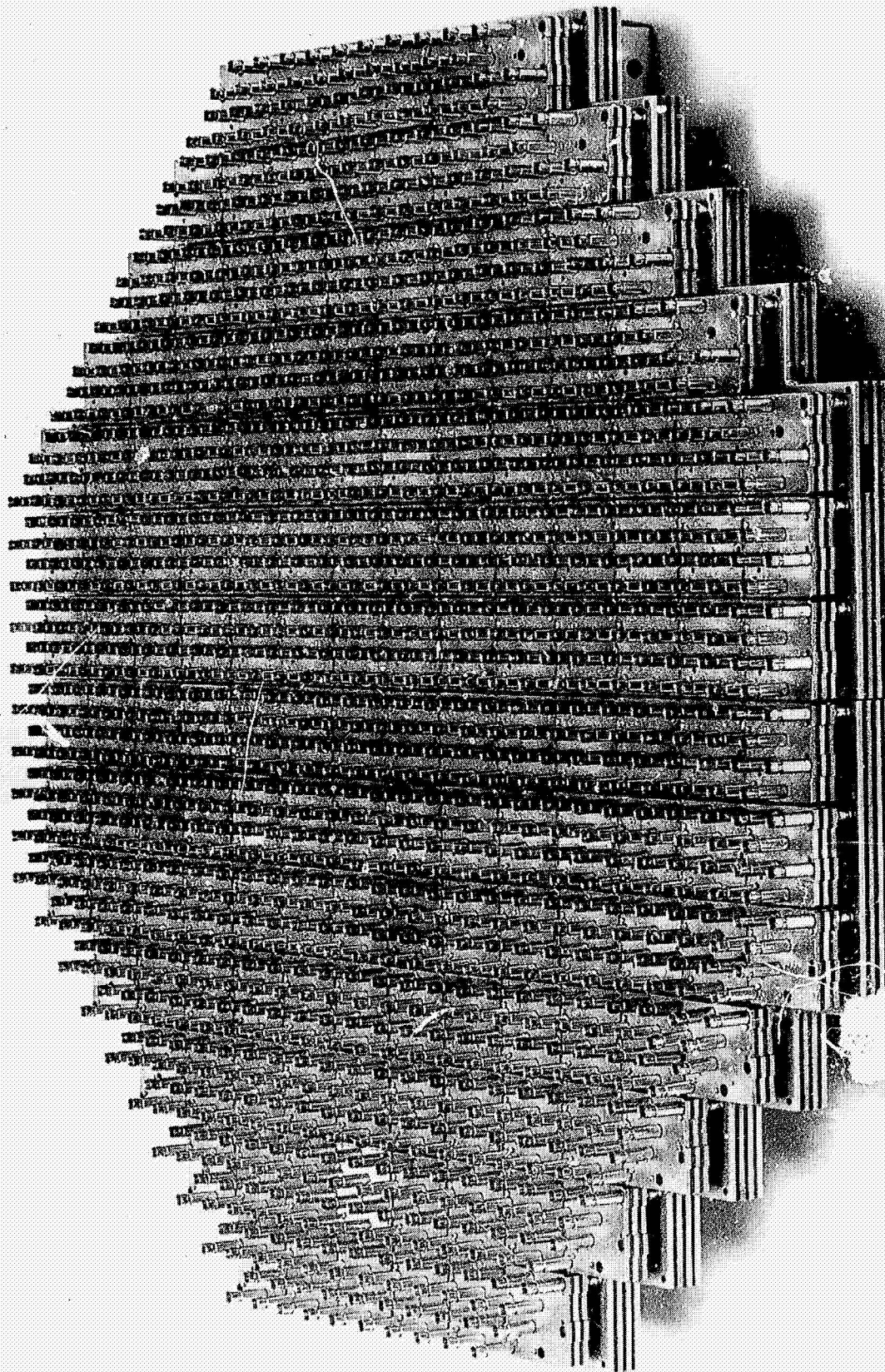


Figure 3-88. Rf Manifold Top View

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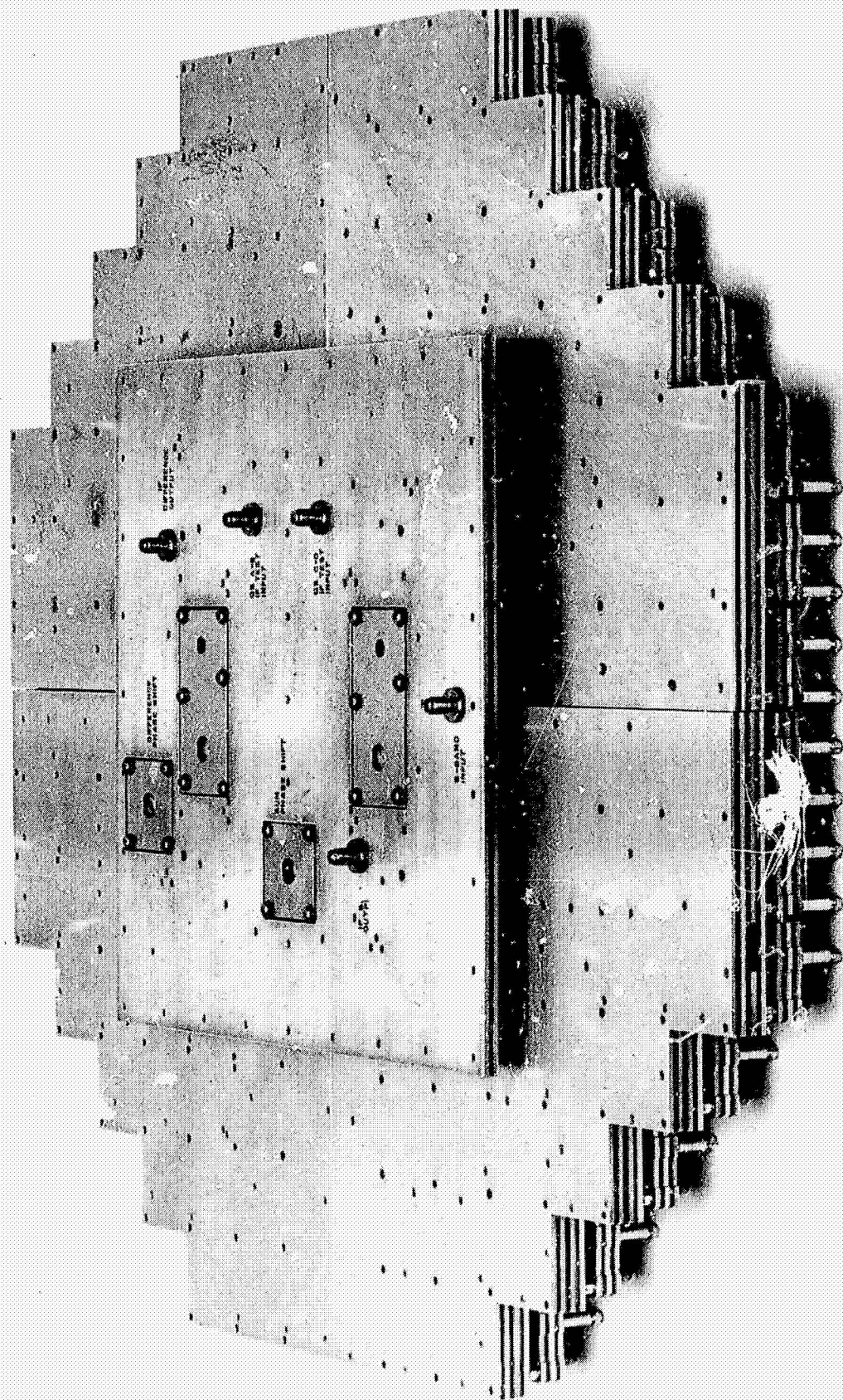


Figure 3-89. Rf Manifold Bottom View

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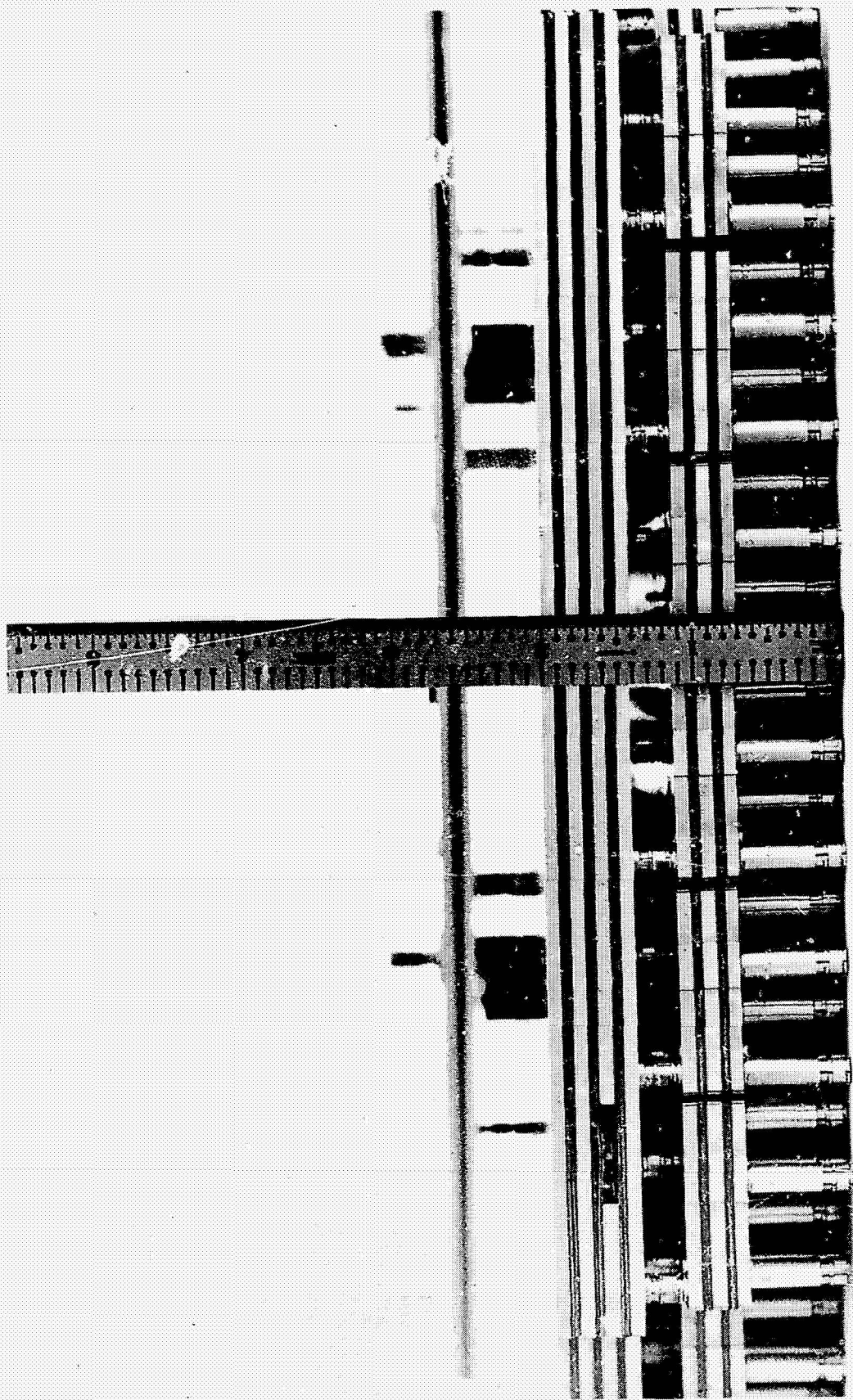
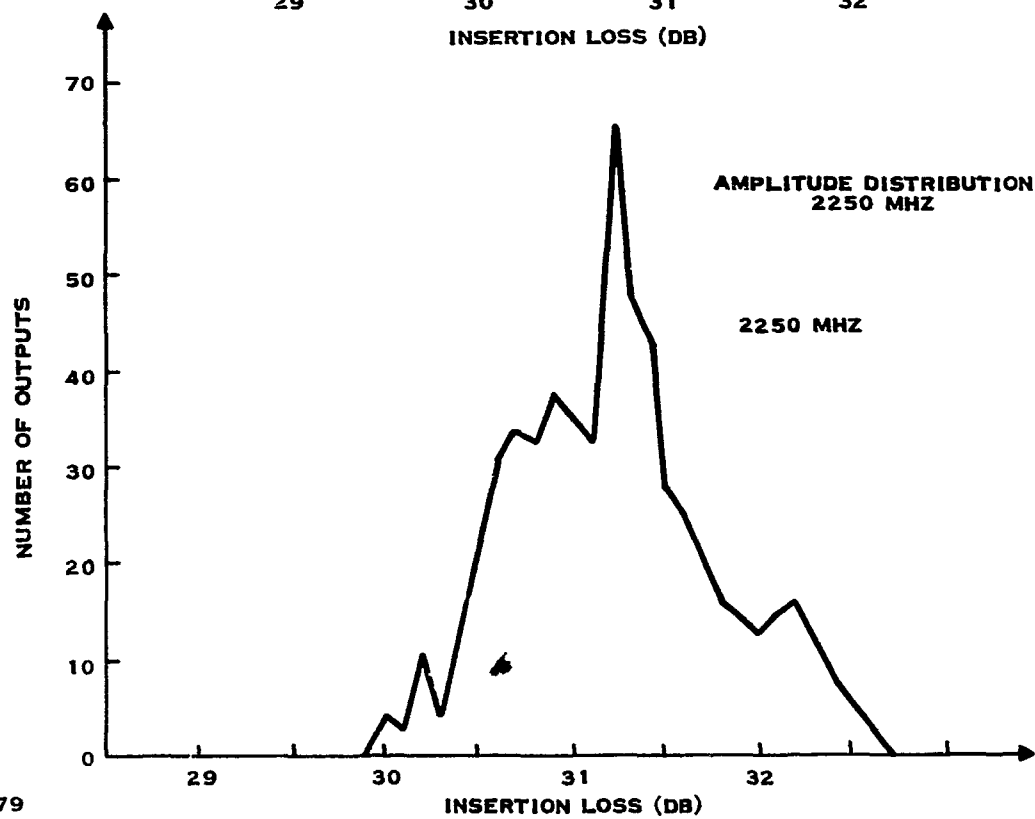
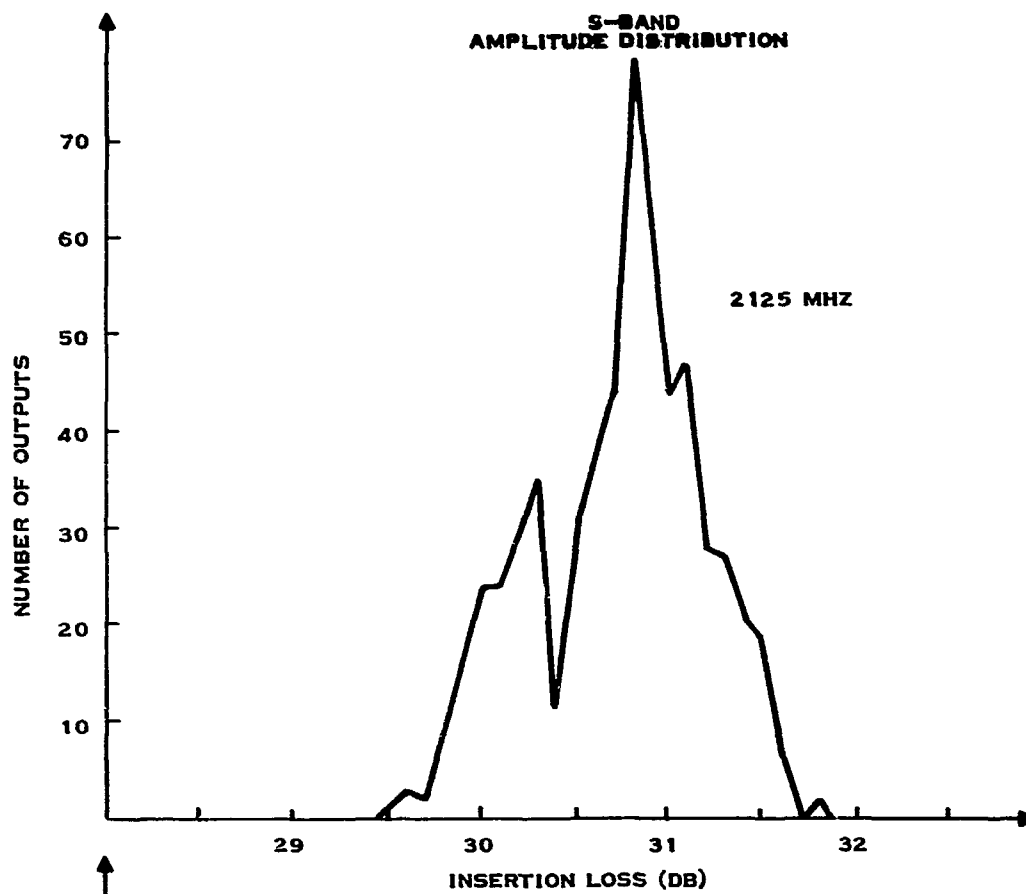


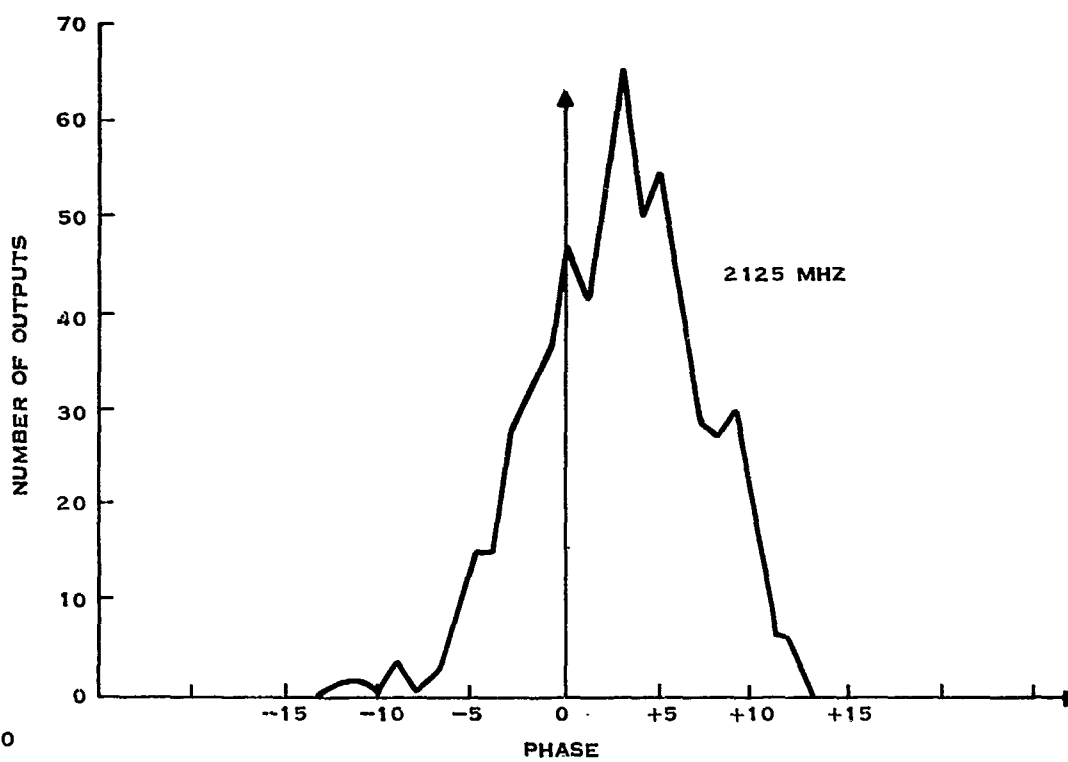
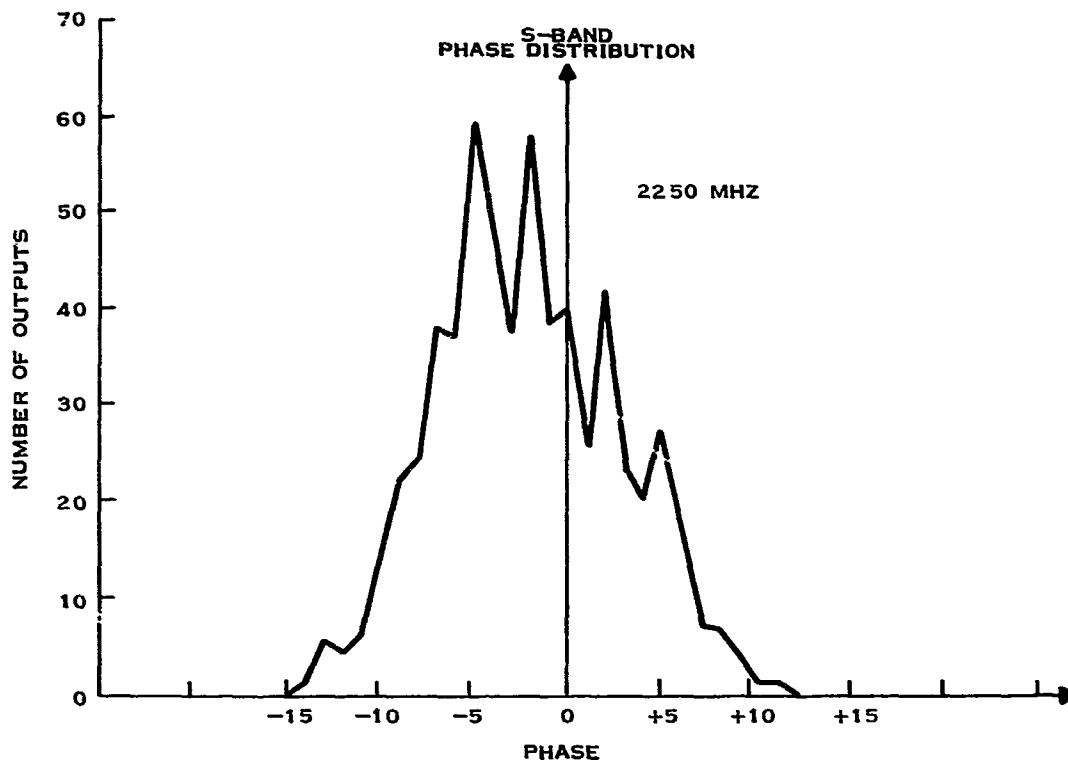
Figure 3-90. Rf Manifold Side View

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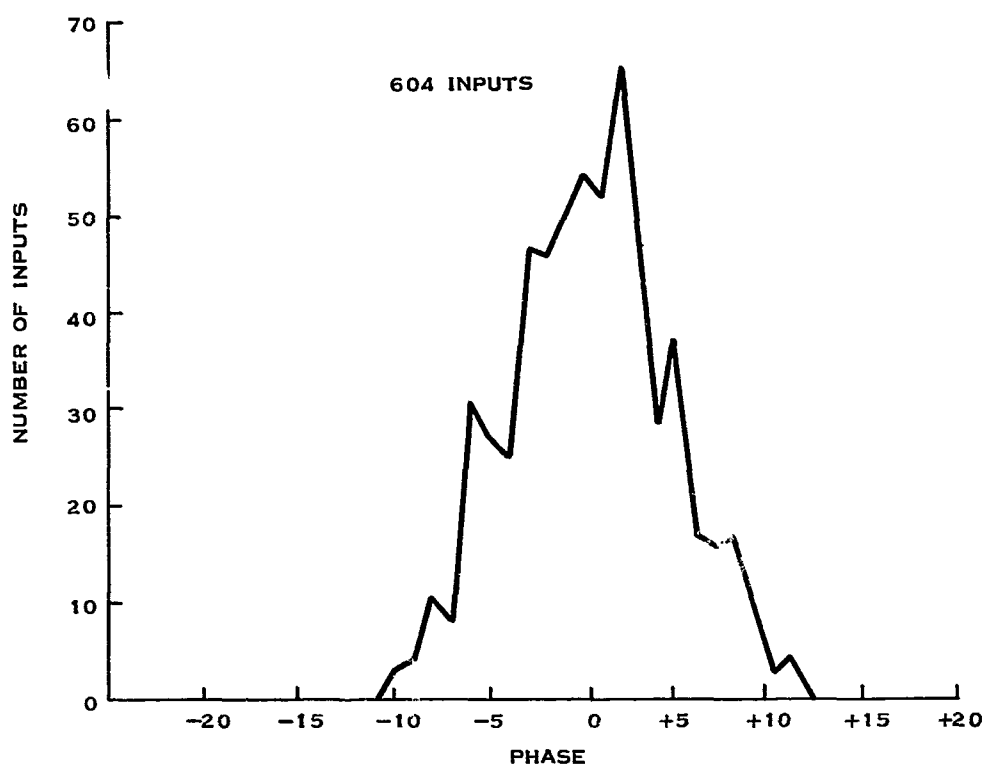
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Figure 3-91. Manifold S-Band Amplitude Distribution



54180

Figure 3-92. Manifold S-Band Phase Distribution (1 Degree Increments)



56677

Figure 3-93. Manifold 500-MHz Phase Distribution (1 Degree Increments)

usually result in line spacings which are too wide. This, in itself, does not entirely account for the rather large deviations, but is probably a large contributing factor.

The deviations in Figure 3-94 were measured by operating the 500-MHz quarter sections as power dividers rather than as power summers. In the normal summing mode of operation, the insertion losses for each quarter section (feeding equal amplitude, equal phase signals in each input) are as follows: QS-A -3.44 dB; QS-B -3.94 dB; QS-C -3.74 dB; QS-D -4.34 dB.

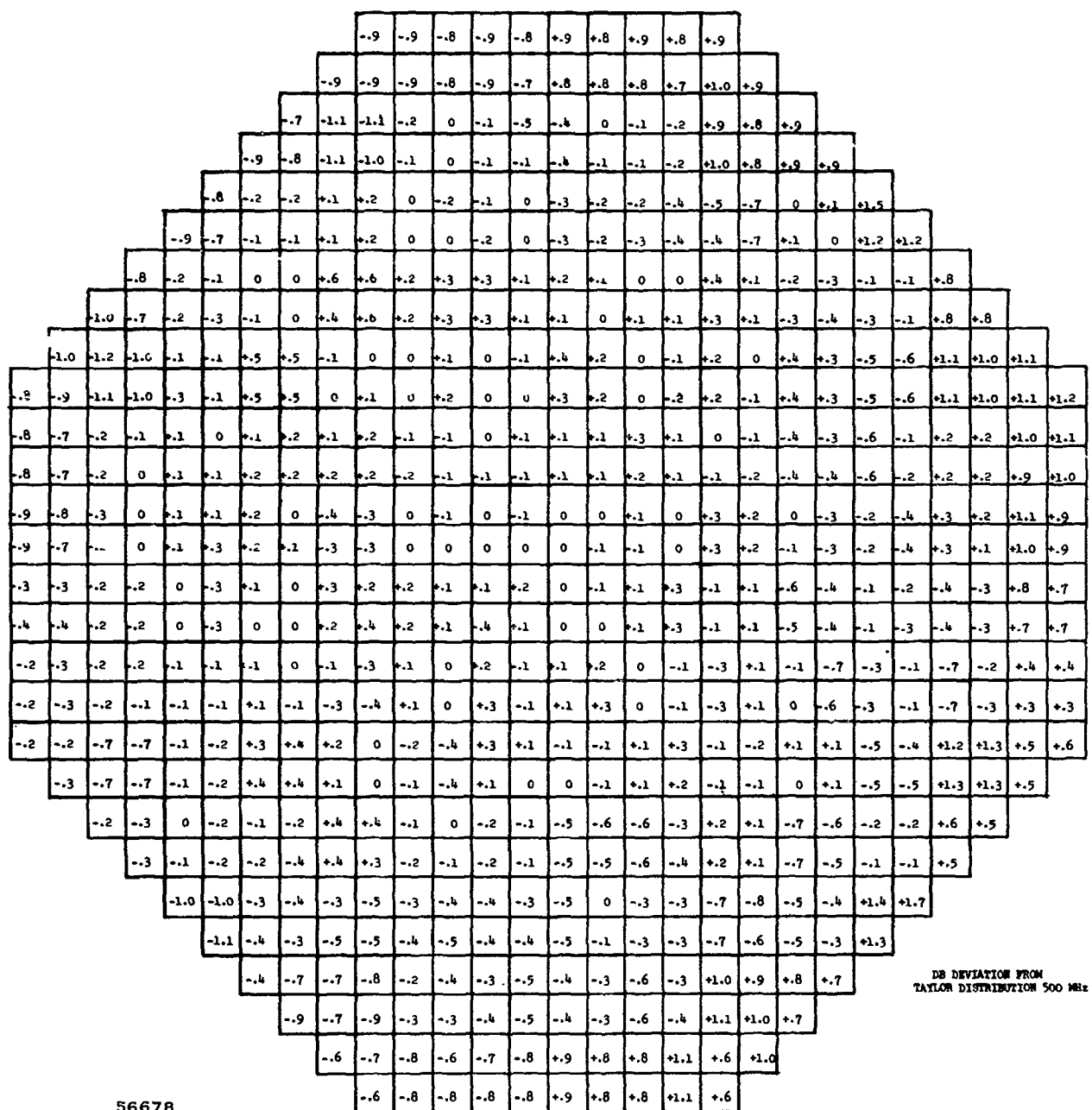


Figure 3-94. Manifold Deviation from Required 500-MHz Amplitude Distribution

Relative phases of the four summed signals are within 4 degrees. In each case, about 2 dB of the measured insertion loss is due to intentional loss in the isolation resistors in order to achieve the required if. amplitude distribution across the array. Total if. summing insertion loss to the sum channel output is about 5 dB.

Other characteristics of the complete manifold are the following:

	<u>500 MHz</u>	<u>2125 MHz</u>	<u>2250 MHz</u>
Input VSWR	Less than 1.2	1.2	1.3
Output VSWR	Less than 1.1	Less than 1.5	Less than 1.5
Isolation	Greater than 15 dB	Greater than 15 dB	Greater than 15 dB.

The monopulse comparator circuit produced a null depth of greater than 50 dB when energized by two test inputs of roughly equal amplitude and phase. When using input signals having a 3-dB difference in amplitude and roughly equal phase, it was still possible to obtain a null depth of 30 dB.

The distribution networks are composed of a large number of hybrid power dividers forming a corporate feed; that is, a parallel-type feed which has equal path lengths to all output parts. This feed is inherently broadband.

The hybrids provide isolation between the individual rf modules. Thus, if a failure occurs in a module, the resulting impedance mismatch has little effect on the phase and amplitude of the signals elsewhere in the array. This is an important feature of this approach. Module failures are isolated failures in a redundant network with slow degradation, not catastrophic, occurring in the array.

The manifold can be lightened considerably for the CAT radar by bonding the stripline board together and eliminating the metal backing plates. Also, it is quite feasible to replace the rf connectors presently used on the manifold and rf modules with a type that can be soldered or welded in place. This would virtually eliminate the possibility of a failure due to a loose connection.

For some array configurations only one rf distribution circuit is needed, which then requires a diode switch on the input for duplexing.

If a low if. (75 MHz) is used in the array, the stripline power-divider network would be replaced by a lumped constant, RLC, equivalent circuit or a combination network of stripline and ferrite torroid power dividers.

For small arrays it becomes feasible to consider the use of ceramic substrate to reduce the circuits size.

3.2.2.11 Signal Generator

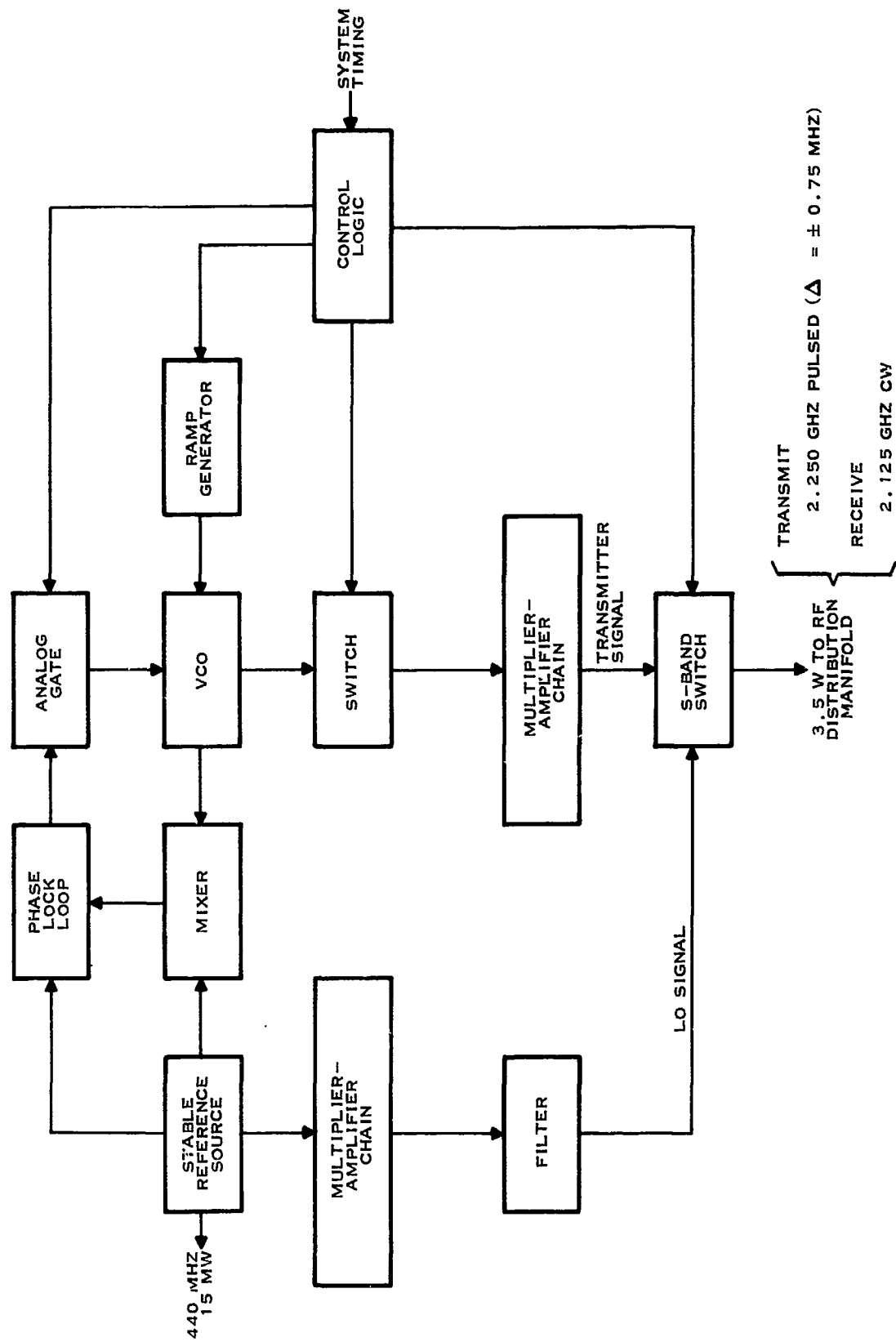
The signal generator unit provides the transmitter and local oscillator excitations required for the rf modules, and a second local oscillator source if a double conversion receiver is used. Depending on the CAT requirements, the all solid-state signal generator is quite capable of generating:

1. An L- or S-band excitation with crystal stability
2. A FM "chirp" signal with good linearity
3. Short and long-pulse, "interlace," excitation for close- and long-range radar operation
4. Frequency diversity signals
5. Pulse and phase coded signals
6. Signals for coherent processing
7. Various combinations.

As an example, Figure 3-95 shows the signal generator developed for the MERA radar.

The signal generator unit contains the circuitry necessary to provide transmitter excitation signals (both chirped and nonchirped) and local oscillator power to the 604 rf modules in the array. A second local oscillator source is included also to accomplish a 500-MHz to 60-MHz down conversion in the receiver sum-and-difference channels.

The output to the rf distribution manifold is an S-band signal at 3.5 watts. During transmit time, the output is swept linearly from 2250.75 MHz to 2249.25 MHz in a 19.53- μ sec period with the system operating in the chirp mode. For nonchirp operation, the output on transmit is a 0.244- μ sec pulse centered at 2250.00 MHz. During the receive cycle, the S-band output is a 2125.00-MHz CW signal in either operating mode. The second local oscillator signal is a 440-MHz, 15-mW output.



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Figure 3-95. Signal Generator Block Diagram

Inputs to the unit include the timing signals necessary to switch the S-band output between transmit and receive and to permit operation in the chirp mode only or in an interlaced mode combining chirped and nonchirped operation.

The 2.250-MHz swept-frequency (chirped) signal is derived from a voltage-controlled oscillator (VCO) operating nominally at 93.75 MHz. During radar receive time, the frequency of the VCO is held constant by an automatic-phase-control (APC) loop with reference signals supplied from a crystal-controlled oscillator. During the 19.53- μ sec transmit time, the input to the VCO is switched from the APC loop to the output of a ramp generator. The ramp sweeps the VCO frequency essentially linearly over a 62.5-kHz range. The output of the VCO is switched into a multiplier-amplifier chain during transmit time. A switch is used at this point to provide additional isolation during receive time. For nonchirp operation, the ramp generator is disabled and the VCO remains fixed at 93.75 MHz.

During radar receive time, the output to the rf distribution manifold is the 2125-MHz local oscillator signal. This signal is also derived from a stable crystal-controlled oscillator. In addition to the multiplier-amplifier chain, a bandpass filter is included in the local oscillator channel with cutoff frequencies at 2000 MHz and 2500 MHz. This filter is required since output at these frequencies during receive time appears as coherent noise within the if. passband.

Any number of configurations is possible; so, further consideration of system requirements and constraints is necessary to finalize a paper design.

3.2.3 Space-Fed Array Components Analysis

3.2.3.1 Introduction

This section discusses aspects of the space-fed array and how they will impact the synthesis of an optimum space-fed array for the CAT radar. The space-fed array is concerned with front-fed reflective phase-shifter elements and rear-fed transmissive elements. The antenna and feed circuit considerations have been discussed in Subsection 3.1 and are covered quite extensively in the referenced literature. This section also considers aspects of transmitter tubes and modulators, and the receiver circuits. Of major concern is the dc power and insertion loss of the phase-shifter elements, reliability of the transmitter tubes, and arcing/breakdown problems in the high-voltage circuits and waveguide.

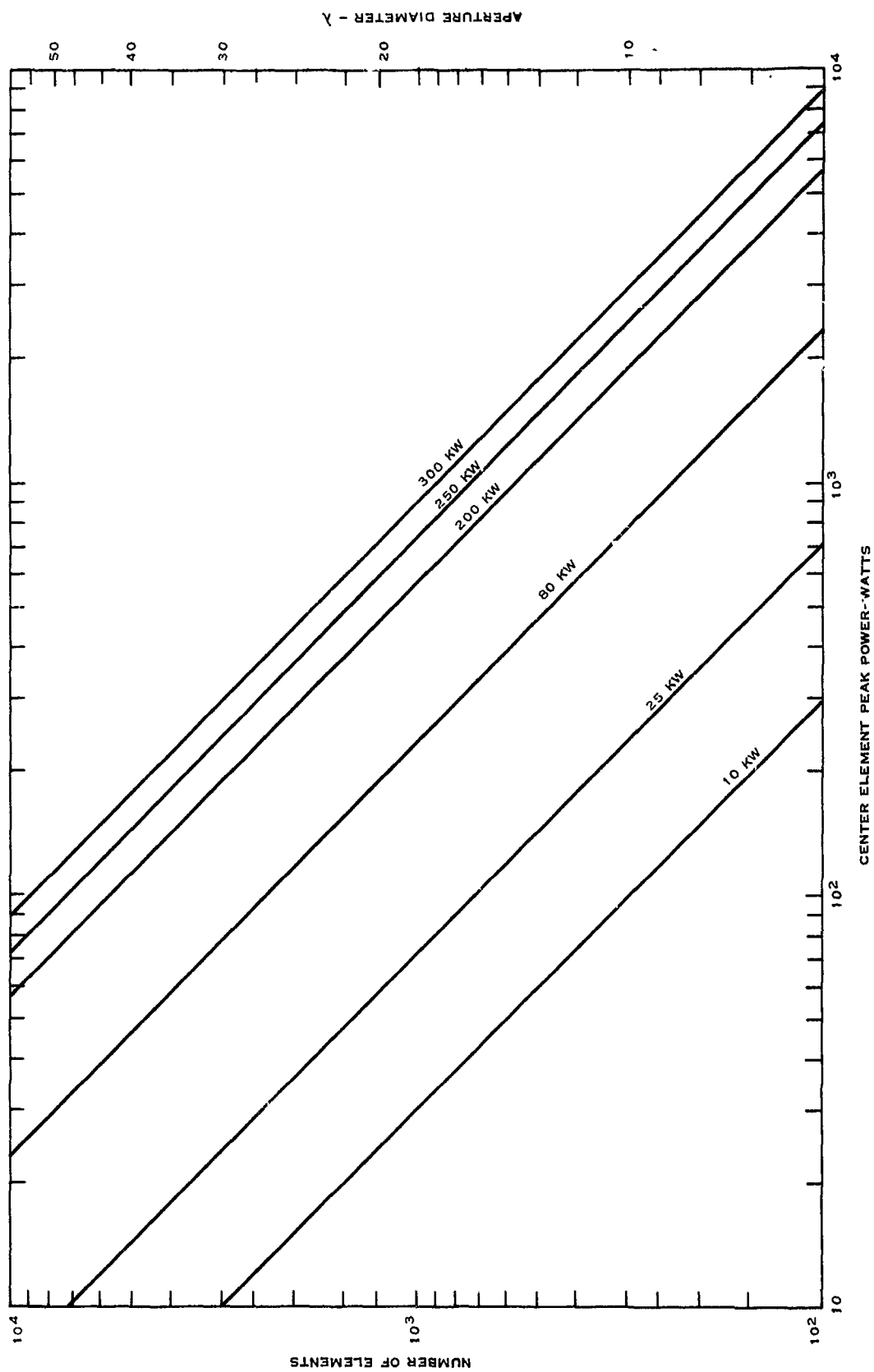
3.2.3.2 Phase-Shifter Consideration

The phase shifter is one of the most critical components in the space-fed electronically scanned array radar. Not only must its performance be compatible with system requirements, but it can and does represent a large percentage of total system cost in a radar of several thousand elements. Its selection then must be carefully made, and tradeoffs with other system considerations are generally required for a particular application.

Phase shifters can be designed to accept analog or digital control information. Digital devices are usually preferred for stability reasons and because their control circuitry is less complex. In addition, digital design is generally considered more efficient in most applications where ferrite devices are used because they have the capability of being "latched" in either of two remnant states. Each bit of the phaser is controlled by current pulses which switch a dc magnetic field between these states. This means that the phase shifter uses drive power only when moving the beam to a new position. No "holding" power is required. Phasers also have the property of being reciprocal or nonreciprocal. In the reciprocal device, the differential phase shift is independent of the direction of the rf field. Thus, a reciprocal phase shifter can be used in either a transmissive or a reflective array. A non-reciprocal phase shifter can be used in a transmissive array providing switching time between transmit and receive is compatible with system requirements. If such were the case, the nonreciprocal phase shifter could offer an overall savings in control power requirements in some applications where data rates were sufficiently low.

One of the first considerations in selecting a phaser for a particular application is to determine the power handling requirements of each element. This is a function not only of transmitter output, but the size of the array, the number, and spacing of elements in the array, and the illumination taper necessary to meet sidelobe requirements. For example, if a circular aperture is illuminated with a symmetrical pattern from the feedhorn, the elements in the center of the array will require the highest power handling capacity. Figure 3-96 illustrates the required center element peak power for various size arrays and primary power levels in a typical space-fed array, designed to provide sidelobes of approximately 25 dB. Consider the case of an array of 1000 elements illuminated with 251-kW peak power. As shown, the phasers in the center portion of the array must be capable of handling approximately 80-watt peak power.

Many other factors must be considered in selecting a phase shifter for a particular application. The most important of these are listed in Table III-13. Several different types of phase shifters were considered for the CAT radar application. These included: (1) waveguide diode phase shifters; (2) diode microwave integrated circuit phase shifters; and (3) ferrite finite phase shifters. Each type is discussed in the following paragraphs.



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Figure 3-96. Center Element Peak Power for Space-Fed Array

Table III-13. Phase Shifter Characteristics

Frequency	
PEAK POWER	RF Bandwidth
AVERAGE POWER	Polarization
Rf Loss	Reciprocal/Nonreciprocal
Loss Deviation with Phase Shift	Transmissive/Reflective
Switching Rate	Phase Accuracy
Switching Time	Phase Uniformity
Switching Energy	Insertion Loss Uniformity
Holding Power	Number of Bits
Maximum Pulsewidth	Phase Range
Temperature Sensitivity	Size
Weight	Cooling
Driver Complexity	Computation Requirements
MTBF	COST

3.2.3.3 Waveguide Diode Phase Shifters

For applications requiring high peak and average power, waveguide diode phase shifters are often used. Recent work in this area⁴⁸ offers considerable promise for the development of compact, low-loss devices utilizing discrete-diode modules in an all-pass phase shift network. Phase shifters of this type are low-loss elements, comparing favorably with discrete ferrite devices. However, they cannot be realized in extremely lightweight structures as generally desired for space-borne applications. In addition, since they are fabricated from discrete devices in waveguide, their ultimate cost for the CAT radar application will probably be significantly greater than integrated circuit phase shifters of either the diode or ferrite types.

3.2.3.4 Diode Microwave Integrated Circuit Phase Shifters

3.2.3.4.1 Introduction

Recent advances in microwave diode and integrated circuit technology make possible small, lightweight phase-shift elements having power handling capabilities compatible with the CAT radar requirements. These

units, while having slightly more insertion loss than comparable discrete ferrite devices, represent significant reductions in system weight and driver circuitry complexity. In addition, they are temperature-insensitive, providing excellent antenna pattern control when operated in the expected environments without requiring compensating devices and circuitry.

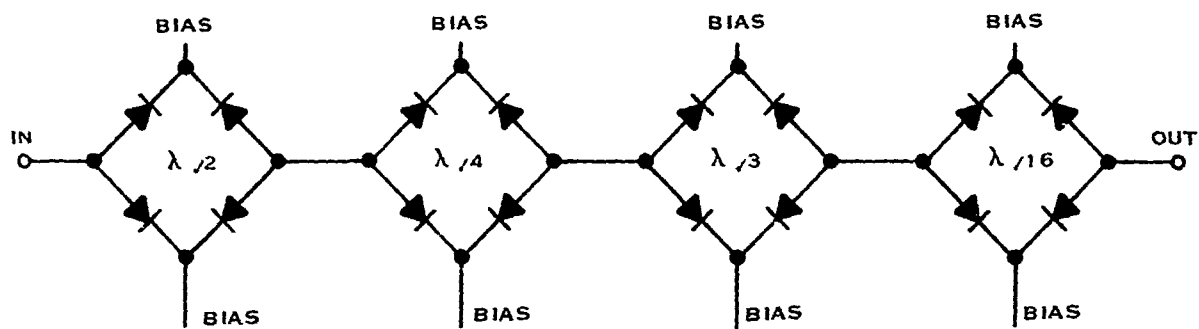
At present, diode integrated circuit phase shifters are being constructed both in monolithic form and in hybrid form on silicon and ceramic substrates. In the monolithic approach, all semiconductor devices are formed into the substrate with the circuit, while hybrid techniques involve attaching devices in chip form to the circuit. The approach selected is dependent upon the requirements of the particular application, with monolithic circuits being utilized wherever performance objectives can be achieved because fewer manual operations are required in the fabrication process.

3.2.3.4.2 Types of Diode Phase Shifters

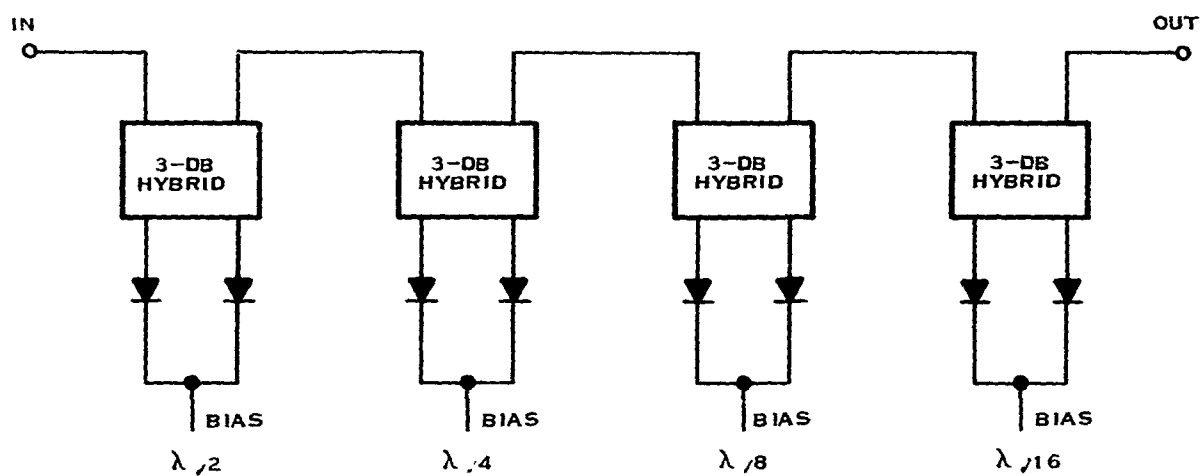
Diode phase shifters may be designed to provide continuous phase variation over a specified range or discrete changes of phase in any number of steps. The continuous or analog phase shifter utilizes the variable capacitance of varactor diodes while the step or digital phase shifter makes use of PIN switching diodes. For this application, the step phase shifter represents the preferred approach. PIN diodes are capable of handling much greater power than varactors and can be implemented in circuits with lower insertion losses. In addition, the variable capacitance of the varactor diode is sensitive to temperature and bias voltage changes, imposing limitations on achievable phase stability.

Most of the circuit forms suitable for discrete diodes are amenable to microwave integrated circuit techniques. Figure 3-97 shows the three basic circuit types which are currently being used for most applications requiring digital diode phase shifters. Figure 3-97a is a four-bit switched-line phase shifter which used PIN diodes to switch the rf signal through alternate transmission lines of different lengths. As shown, each phase bit requires four diodes, all of which must handle the same rf power (less losses in the preceding bits). Thus, all 16 diodes in this circuit must have essentially the same peak-power handling capability. A photograph of an actual L-band, integrated circuit four-bit switched-line phase shifter is shown in Figure 3-98. This is a hybrid circuit, consisting of four chips, each chip includes four diodes, bonded to a thin-film circuit consisting of transmission lines, chokes, and bypass capacitors.

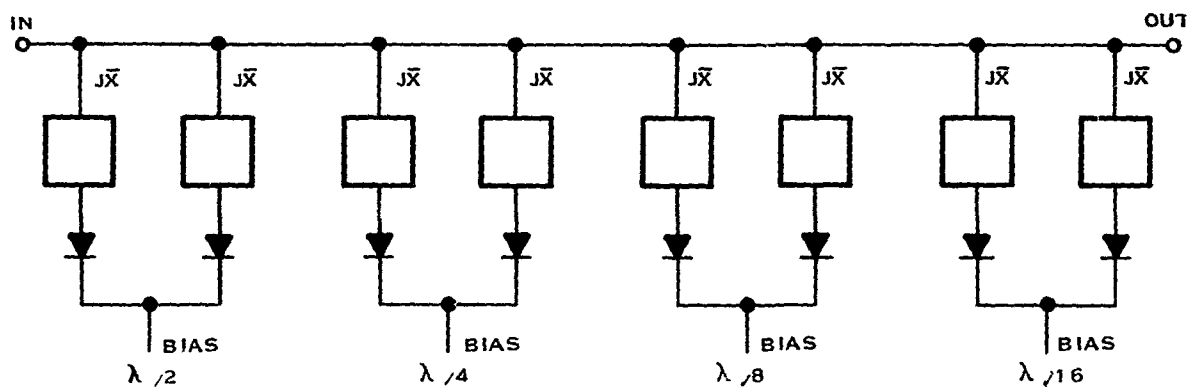
A hybrid-coupled phase shifter is shown in Figure 3-97b. This circuit consists of an array of 3-dB hybrids with balanced phase bits connected to the coupled arms. Each phase-shift bit produces two different phase shifts corresponding to the ON and OFF states of the diodes. As shown, only two diodes are required for each bit. In addition, only one-half of the device input



(A) 4-BIT SWITCHED-LINE PHASE SHIFTER



(B) 4-BIT HYBRID-COUPLED PHASE SHIFTER



(C) 4-BIT LOADED-LINE PHASE SHIFTER

63843

Figure 3-97. Diode Phase-Shifter Types

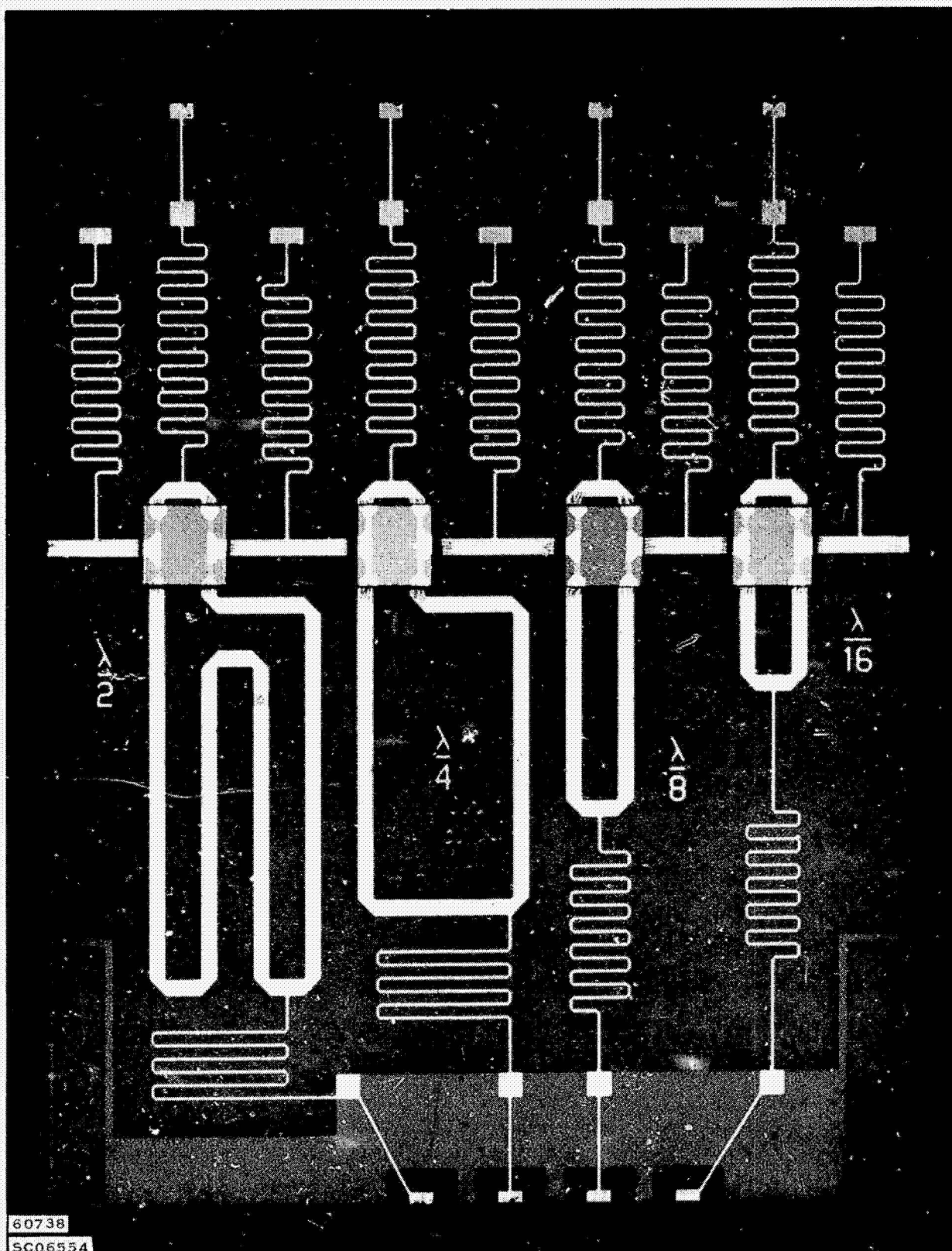


Figure 3-98. L-Band Phase-Shift Network

power is incident on each diode due to the nature of the 3-dB hybrid. The required breakdown voltage and insertion loss are also functions of the bit size. For these reasons (fewer diodes, less severe breakdown voltage requirements, and lower insertion loss) the hybrid-coupled phase shifter is considered superior to the switched-line circuit for most applications.

The third circuit shown in Figure 3-97c is a loaded-line phase shifter. This design uses switched loading susceptances spaced along a transmission line. By using pairs of equal loading susceptances that are small and spaced approximately one-quarter wavelength apart, minimum perturbations may be achieved in either of the two switching states. A modified version of this circuit could permit operation of a four-bit phase shifter with a total of six diodes.

Performance of the loaded-line phase shifter is comparable to that of the hybrid-coupled device. The former is somewhat less lossy and is more compact, making it particularly attractive where size and weight are important considerations.

Figure 3-99 is a photograph of a four-bit digital loaded-line phase shifter for S-band operation. The circuit is processed on 20-mil alumina substrate and has deposited thin-film capacitors, beam-lead discrete capacitors, and beam-lead PIN switching diodes. The entire circuit is 0.30 inch by 0.75 inch.

At higher microwave frequencies, phase-shift networks can be fabricated either in monolithic or hybrid forms. However, at K_u -band for example, the placement of diodes in the circuit becomes critical in terms of allowable phase-shift error. The monolithic circuit is then easier and less costly to manufacture because the active devices and the circuit elements are processed into the network with extreme precision. No separate handling or bonding is involved in the fabrication of the circuit and an entire four-bit phase shifter can be built with a size comparable to that of a conventional digital integrated circuit. Work is currently underway at Texas Instruments [Contract No. F33(615)-C-1817] and at Microwave Associates [Contract No. F33(615)-67-C-1816] to develop a monolithic integrated K_u -band phase-shift network for utilization in airborne phased-array radars.

The program at Texas Instruments will include the development, fabrication, and flight testing of a front-fed illuminated array radar utilizing 624 of these K_u -band elements. The elements are being packaged in groups of four in a single module. Each module, including the driver circuits, four phase shifters, and four radiating elements, weighs approximately 0.03 pound. For a 1000-element array, the total element weight would be approximately 75 pounds. This corresponds to a weight of 35-50 pounds for 1000 elements of the discrete ferrite type. This phase-shifter design offers a low-weight low-cost efficient phaser for many applications.

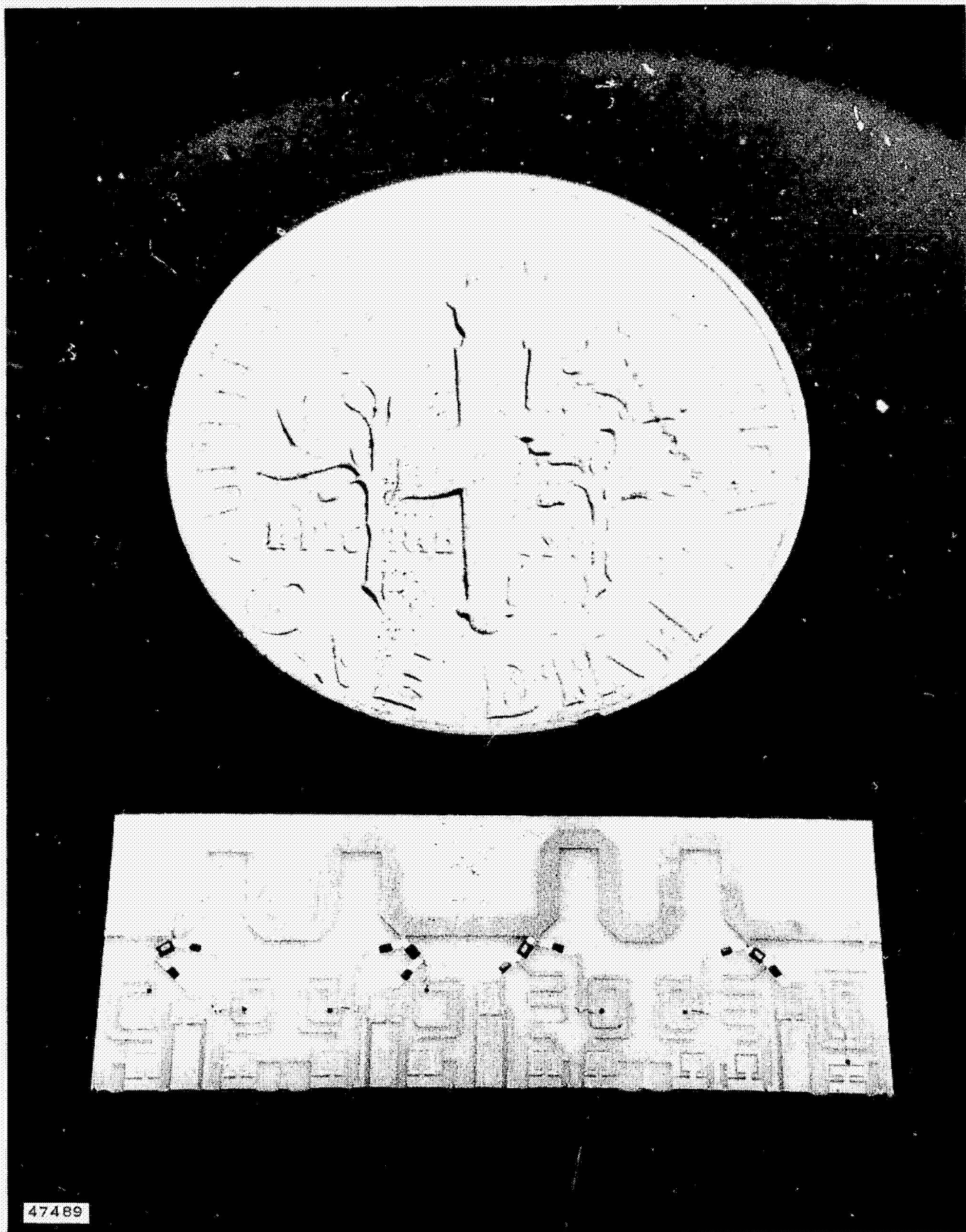
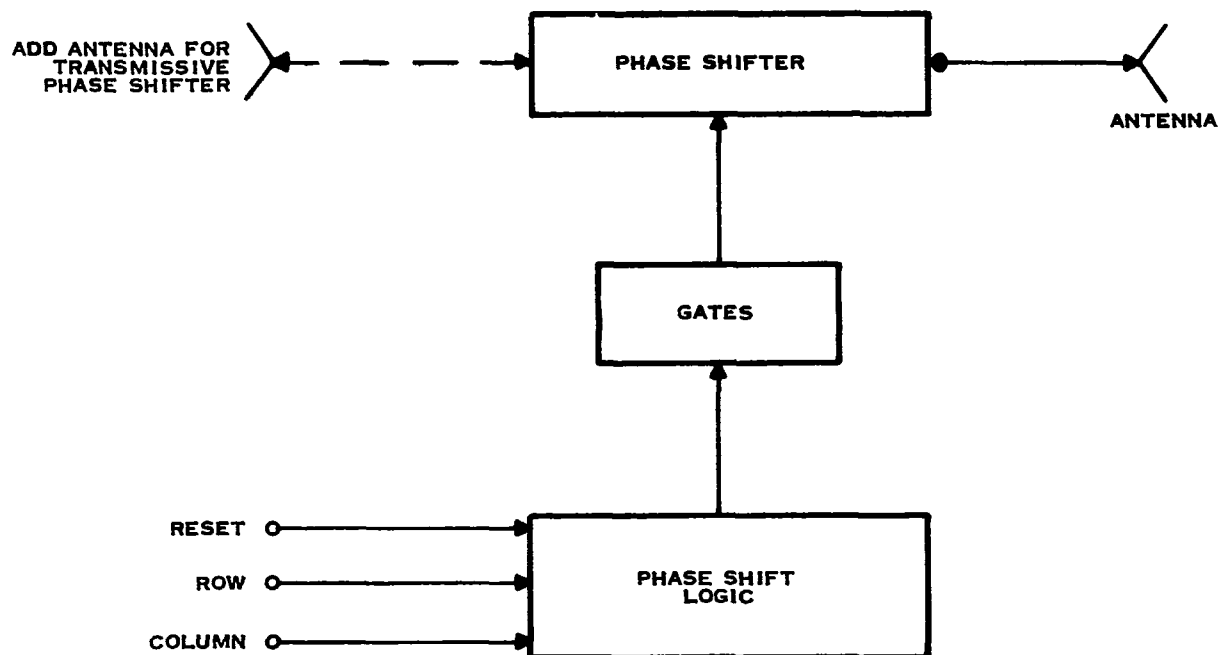


Figure 3-99. Four-Bit S-Band Phase Shifter



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Figure 3-100. Diode Phase-Shifter Module Block Diagram

3.2.3.5 Diode Phase-Shifter Modules

A block diagram of the diode phase-shifter module for use in the CAT radar space-fed array is shown in Figure 3-100. Only one antenna is required for the reflective-array module. Two are needed for the transmissive-array module (one on each face of the module). Reset and steering commands from the scan computer are sent to each module through a multilayer dc manifold. The commands, digital binary signals, are processed through a ripple-through counter and a group of gates to set the appropriate bias current on each diode.

Since the power for biasing the diodes is relatively high in this application, a three-bit phase shifter will be specified in order to use a minimum number of diodes. The reflective phase shifter uses 90 to 45 to 22.5-degree bits with a short circuit on one end. The round-trip phase shift then is the desired 180 to 90 to 45 degrees. The transmissive phase shifter will require about twice the number of diodes to achieve the same phase shift on one pass.

Table III-14 summarizes the module capabilities from 1.5 to 18 GHz. The peak rf power input (P_{pin}) improves as we go lower in frequency due to the higher capacitance and quality diodes available. Insertion loss improves somewhat at lower frequencies while dc input power is independent of frequency for the same circuit. Since the logic circuitry is on all the time,

it establishes the minimum power requirement. The average dc input power is taken as half the sum of the minimum and maximum input powers. This is valid assuming the array is composed of many elements and half the phase shift sections are energized and half are not.

Table III-14. Modules for Space-Fed Arrays with Diode Microstrip Phase Shifters

Frequency (GHz)	P _{pin} (W)	Insertion Loss (dB)		Dc Input Power (W)	
		Reflective	Transmissive	Reflective (P _{mr})	Transmissive (P _{mt})
1.5	1000	1.5	1.5	0.6	1.2
3.0	600	1.5	1.5	0.6	1.2
6.0	300	1.6	1.6	0.6	1.2
9.0	100	1.8	1.8	0.6	1.2
18.0	100	2.0	2.0	0.6	1.2

3 -Bit
Single Polarization

3.2.3.6 Ferrite Phase Shifters

3.2.3.6.1 General

Ferrite phasers can use either ferrite switches for real-time delays or can use ferrite transmission lines in which the propagation time is altered. The use of circulators, used as switching elements, is an example of a switched line device. In other devices, the propagation constant of the transmission line is varied by changing the permeability of the ferrite medium. This variable transmission line technique is generally preferable because it is less dense and has electrical advantages; it can be made either reciprocal or nonreciprocal.

Rectangular guide, circular guide, coaxial, stripline, and microstrip techniques are all applicable to ferrite techniques. Some of the more recent work has been done in microstrip configuration, which lends itself to efficient miniaturization, very comparable to that of the diode phase shifters.

3.2.3.6.2 Ferrite Phaser Properties

Ferrite phasers can be analogue (AFPS) or digital (DFPS). The digital types can be made latching, and require only a high current pulse for a

very short period to reset and set them to a new position. For large phased-array applications, this property is very important. In aerospace use, a savings of several orders of magnitude of drive power can result with a reciprocal latching design as compared with an analog design, or any type of diode phase shifter. This can be a significant savings in primary power.

The nonreciprocal property of ferrite devices is the one factor which makes them highly suitable for circulators, isolators, switches, non-reciprocal phasers, and polarization switches. The nonreciprocal phase shifter differential phase shift is dependent upon the direction of rf energy flow or the direction of the applied magnetic field. This type phaser must be switched between transmit and receive in order to maintain proper phase relationships. This imposes a minimum range limitation and the drive power is a function of the PRF. Some advantage is obtained from the nonreciprocal properties in being able to maintain low sidelobes as the impedance changes with scan angle. The degree of this advantage must be analyzed in more detail. A high PRF system may require more drive energy for a nonreciprocal design than is feasible to supply. A reciprocal phase shifter requires drive power only during the time at which the antenna pointing position is changed. This advantage plus no minimum range limitation for a particular system design, may outweigh the other advantages the nonreciprocal designs have.

3.2.3.6.3 Ferrite Phaser Characteristics

Many study programs have been carried on in several organizations, most of which have been Government funded. Table III-15 lists parameters for most of the ferrite phase shifters applicable to airborne or spaceborne systems. Most phase effort has been done in rectangular waveguide. Phasers 1, 2, 3, and 4 are examples of this design technique. The most successful designs of this type are phasers 2 and 4, which are nonreciprocal designs. So far, a reciprocal design, having low insertion loss and good temperature and frequency versus differential phase-shift characteristics, is not known to have been designed. Programs are still in progress to obtain better characteristics in these respects. Some rectangular nonreciprocal designs have achieved full waveguide peak power capability.

About the most efficient, reciprocal, and compact designs reported in standard guide are phasers 5 and 6, which are done in circular guide using the Faraday effect. Phaser 5 is peak-power limited and analog-controlled, using considerable drive power. Details as to how latching is achieved in circular guide are not available for the phaser 6 design, as is its temperature performance.

Little developmental work has taken place in TEM lines (coaxial). Some work has been reported on stripline (TEM), as is indicated by phaser number 7. This phaser size has no advantage over the others, nor are its performance characteristics very good.

FOLDOUT FRAME 1

Table II

CHARACTERISTICS	1	2	3	4	
Trans. or Refl.	Trans.	Trans.	Refl.	Trans.	Refl.
Directivity	Nonreciprocal	Nonreciprocal	Reciprocal	Nonreciprocal	Reciprocal
Frequency	8.5 - 9.5 GHz	8.0 - 10.0 GHz	15.8 - 17.5 GHz	C-Band	K _u -Band
Bandwidth	1.0 GHz (Bit Change Flat vs Freq)	2.0 GHz (Bit Change Flat vs Freq)	1.7 GHz (0.076%/MHz Bit Change)	500 MHz	±5%
Analog					
A. Range					360°
B. Holding power					0.33 W/
C. Switching power					175 μJou
D. Accuracy					9° rms
E. (Nonlinearity phase-current)					30°/360
Digital					
A. No. Bits	4	4	3	4	
B. Holding Power	None	?	--	--	
C. Switching Power	450 μJoules (Inc. Driver)	?	1000 μJoules (Inc. Driver)	750 μJoules (Inc. Driver)	
D. Accuracy	±5° Setability	±5°	?	4.2°/360°	
Switching Time	3.0 μsec	6.0 μsec	5.0 μsec	6.0 μsec	<50 μsec
Switching Rate (Max)	5 KC Max any bit	?	2 KC Max any bit	10 kHz	?
Insertion Loss	360°/dB	510°/dB	180°/dB	480°/dB	300°/dB
Peak Rf Power	20 kW	2 kW	400 W	1.2 kW	93 W
Avg Rf Power	200 W	20 W	4 W	--	1.3 W
Temperature	Any 10° Range between 20° and 50° C	±15° over -40 to +200°F Range	ΔPhase change linear from 80°F	ΔPhase change 0.2%/°F (70-120°F)	3.2°/F Δφ Fair 155°F
L W H	8.3 × 1.9 × 3.3 (Inc. Driver)	6" of RG/52	2.5 × 0.700 × 0.200 (no driver)	6" large X	0.43 Dia Long
Weight	32 oz	?	64 oz (No driver)	3 oz	0.85 oz 0.75 oz Driv
Type Trans Line	Rect. W/G	Rect. W/G	Rect. W/G	Rect. W/G	Circular
Remarks	Water cooled	Flux Drive - operate on pulse length - has single torroid	Flux Drive - can be Temp & Freq compensated	Made approx. 400 units	Now in Qual test

FOLDOUT FRAME 2

Table III-15. Examples of Ferrite Phaser Designs

4	5	6	7	8
Reciprocal	Refl. Reciprocal K _u -Band ±5%	Trans. Reciprocal K _u -Band ?	Trans. Reciprocal C-Band 5.4 - 6.0 GHz (22% linear variation each bit)	Trans. Nonreciprocal X-Band 10% (Relative insertion Phase +2% of total phaser length)
	360° 0.33 W/360° 175 μJoules 9° rms 30°/360° (Inc. Hyst)			
		4	?	4
		--	--	Logic Power
μJoules (Inc. Driver)		600 μJoules (Inc. Driver)	500 μJoules (Inc. Driver)	25 μJoules
50°		?	?	±5%
sec	<50 μsec	<150 μsec	5 μsec	?
	?	?	?	1 μsec
B	300°/dB	300°/dB	180°/dB	120°/dB
	93 W	3 kW	Medium Power	1 kW
	1.3 W	10 W	--	10 W
change F (70- 155° F	3.2°/F (Ins. phase) Δφ Fairly Constant	?	0 - 120° F (Δφ change 20° - 280°)	?
ge X	0.43 Diam × 2.0" Long	0.4 Diam - 6.0" Long (Incl. Dr.)	6-5/8" × 1" × 3" (Driver Incl.)	1 × 0.35 × 0.08 (No Driver)
	0.85 oz (Incl. Dr.) 0.75 oz (Not Incl. Driver)	0.7 oz Inc. Dr.	18.4 oz	≈0.3 oz
W/G	Circular W/G	Circular W/G	Stripline	Microstrip
approx. 400	Now in production Qual test made	Built about 200 units	Only Feasibility Model Flux-Driven	Feasibility Design Flux-Driven Single Torroid

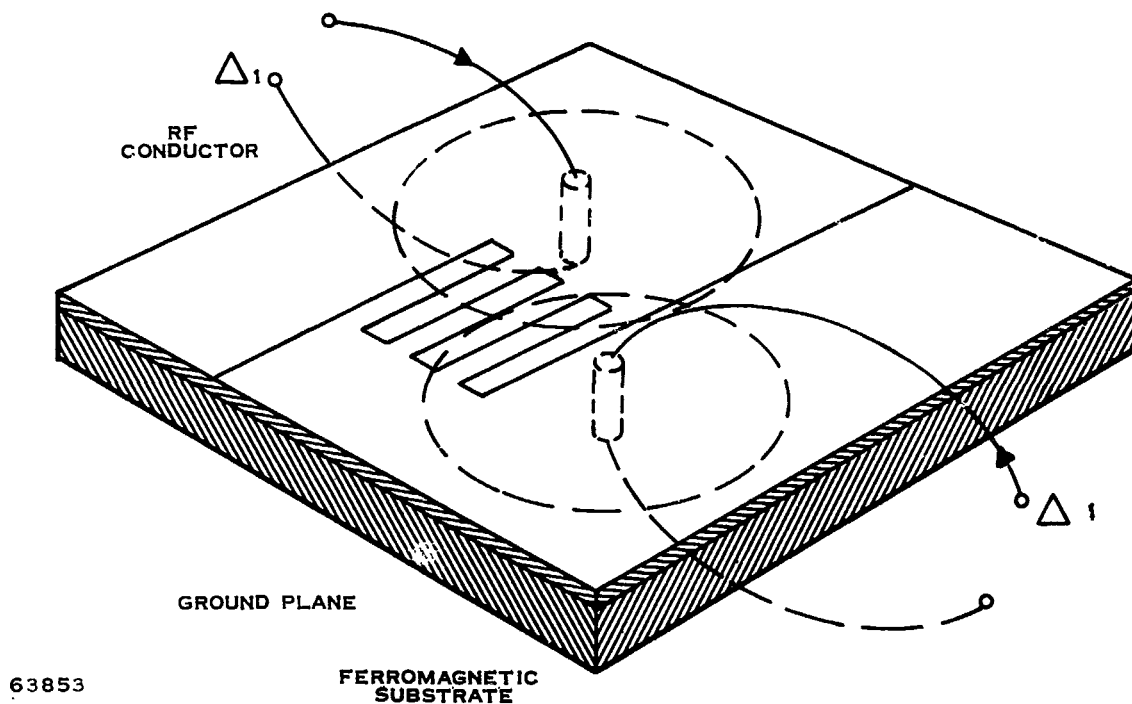


Figure 3-101. Reciprocal Latching Phase Shifter

Phaser number 8 is still in the experimental stage and a promising microstrip phaser, as reported by Syracuse University Research Corporation. It is a nonreciprocal element, designed for an eight-element array.⁴⁹ It demonstrates definite capability in this technique for a low density, highly efficient design for aerospace applications. About a 1-dB penalty is paid in insertion loss over standard line types of phasers. However, drive power savings amounting to as much as several orders of magnitude over all other techniques, including diode phasers, can easily make up for the increased insertion loss in terms of primary power requirements. The insertion loss of diode and ferrite microstrip phasers is about the same. Peak powers of 1 kW or more are believed obtainable with the ferrite microstrip phasers.

3.2.3.6.4 Description of the Integrated Circuit Ferrite Phaser

The integrated circuit ferrite technique uses thin slabs of ferrite about 0.020 inch or less in thickness. On this slab of ferrite is deposited a metallic conductor so that a transmission line is formed. The line is usually formed in a meandering path in order to achieve compactness and other desired characteristics. Figure 3-101 depicts a reciprocal latching phaser. The lines are separated about five thicknesses in order to reduce coupling effects. Two

latching holes, through which a loop is passed to apply the latching pulse in order to create the toroidal fields, are shown. When the fields are parallel to the lines, the permeability the applied rf wave sees is different than when the applied fields are perpendicular to the lines. The length of the lines can be made four or five wavelengths which is enough to provide 360 degrees of differential phase shift. The phase setting of the element can be controlled as follows:

- (1) a saturating pulse resets the phaser to a reference phase shift resulting from the applied current creating field lines perpendicular to the transmission lines,
- (2) and, a set pulse (either magnitude or time controlled) sets the phaser to a minor hysteresis loop in the direction parallel to lines.

This method is called the flux driven technique and only requires one torroid per phase shifter.

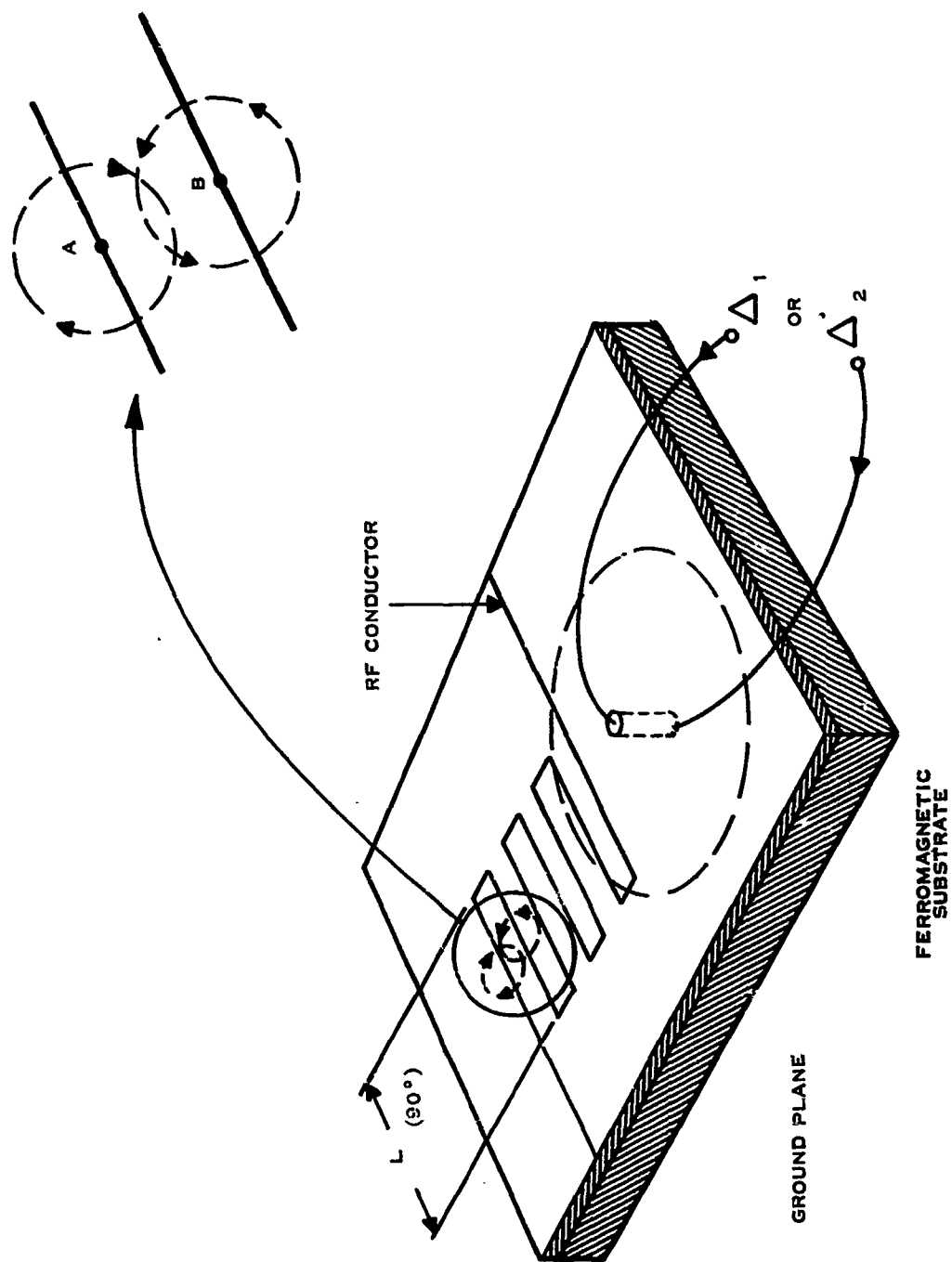
A nonreciprocal phaser can be made as depicted in Figure 3-102. One of the requirements of nonreciprocal phase response is that the rf field must have circular polarized components. This means there must be two orthogonal components 90 degrees out of phase. The length (L) of a line segment is made 90 degrees, so that at points A and B, the fields are 90 degrees out of phase and a degree of orthogonal coupling of the rf field lines concentric to the conductor produces the nonreciprocal effect. Only one latching hole with the applied field parallel to the lines is required. If the phaser is set in a clockwise latched position for transmit (and a positive pulse achieved this), then a negative pulse of equal amplitude, or time must drive it to a new latched position for receive. This will maintain the same $\Delta\phi$ shift for both conditions.

Interpolation of various data indicates that drive energy requirements for a transmissive phase-shift element is approximately as shown in Figure 3-103. Drive energy for reflective elements would be one-half that shown. These are indeed small power requirements when compared to diode-type phase shifters. Figure 3-104 shows approximate variations expected in insertion loss. Elements for a reflective array would have about the same loss. Weight estimates for the ferrite phase shifter alone range from 0.03 to 0.54 ounce, from K_u-band to L-band, respectively. Switching time varies from 1 to 18 μ sec for the phase shifters plus 5 to 6 more for the logic circuitry.

3.2.3.7 Ferrite Phase-Shifter Modules

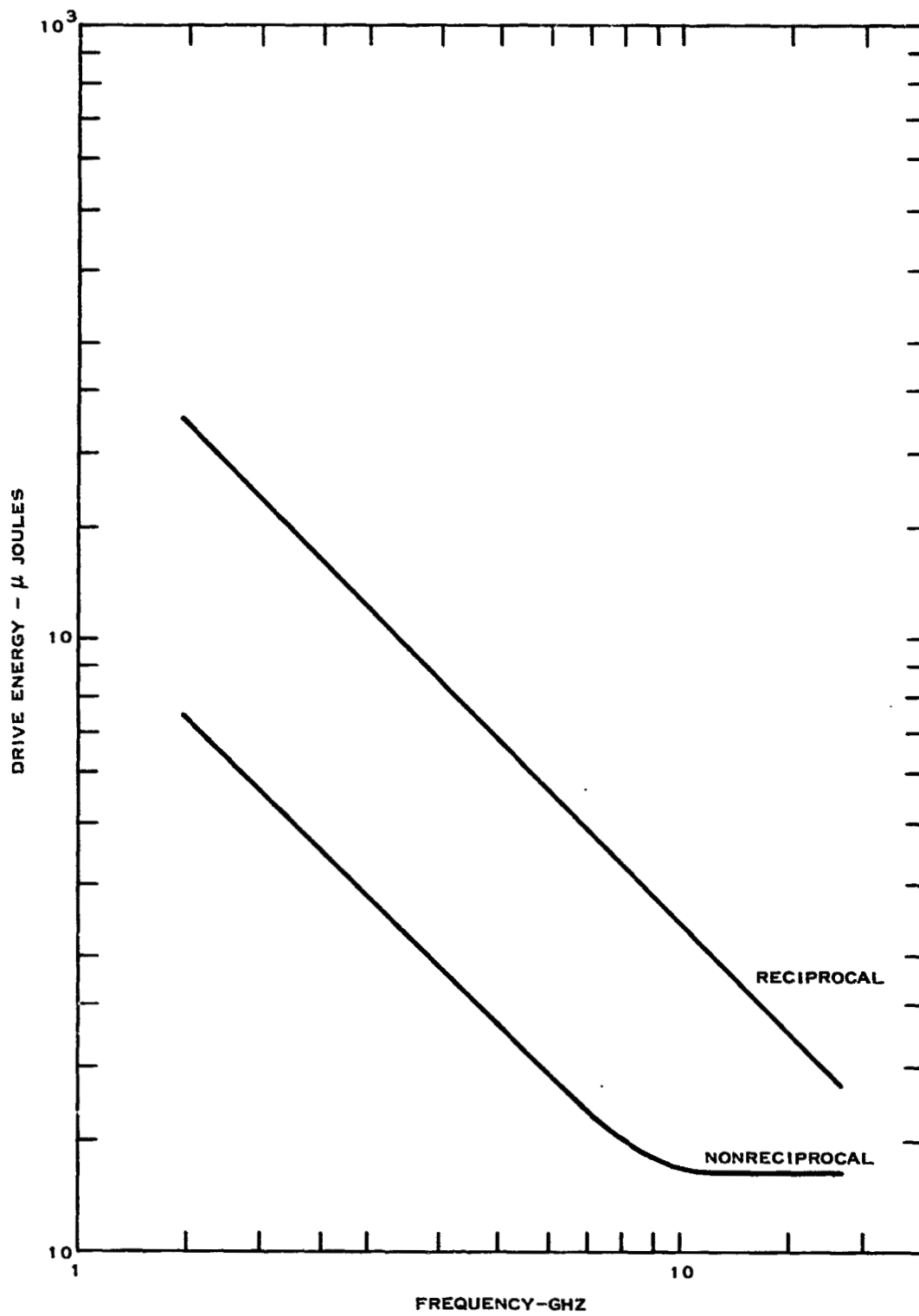
Due to size, weight, and power consumption considerations, none of the ferrite phase shifters currently available are deemed suitable for the CAT radar. With continued development, the microstrip latching ferrite phase shifters are quite promising, with the principal advantages being lower dc power and higher rf power capability.

ORTHOGONAL FIELDS



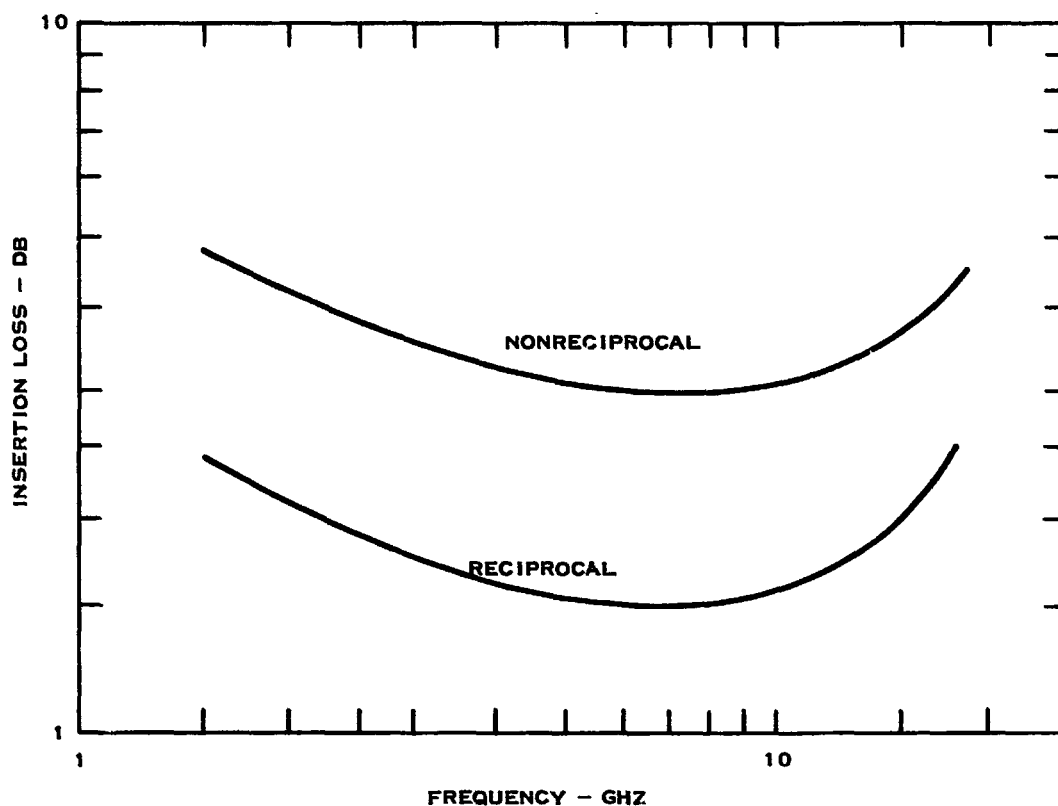
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Figure 3-102. Nonreciprocal Latching Phase Shifter



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Figure 3-103. Drive Energy-Transmissive Ferrite Phaser in Microstrip



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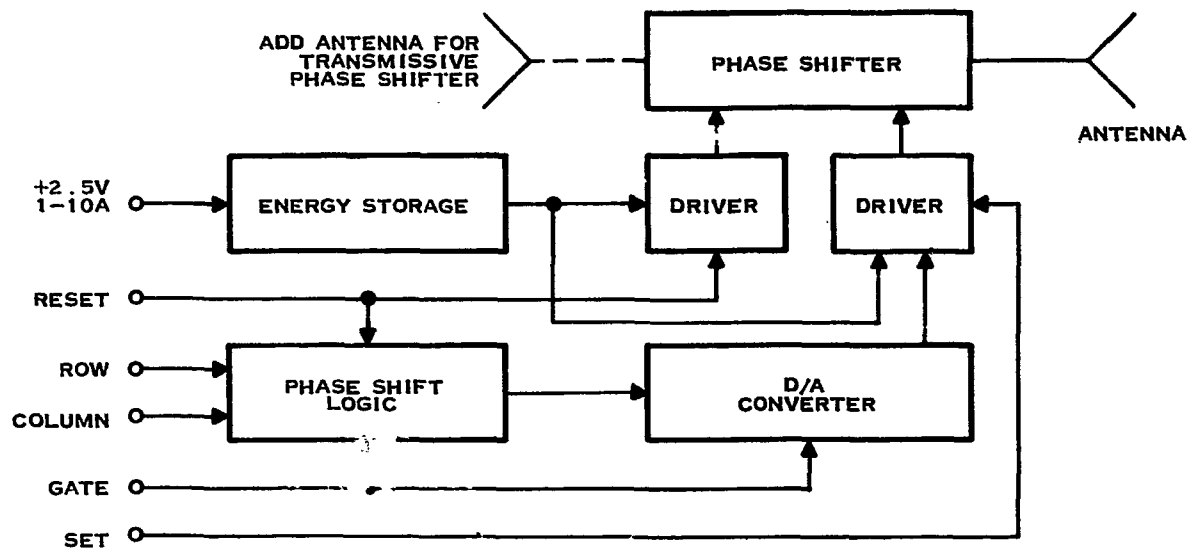
Figure 3-104. Insertion Loss - Transmissive Ferrite Phaser in Microstrip

Module block diagrams for reciprocal and nonreciprocal phase shifters are shown in Figures 3-105 and 3-106. These depict flux-driven phasers at low voltage-high current levels. For large phased arrays, it is much simpler to use a digital scan programmer to steer the array. Thus, digital signals are supplied to the individual phase shift modules, then converted to analog signals to set the drive current level. Various digital control signals are supplied for correct timing. The digital and driver circuits will require more power (≈ 100 mW) than the latching phase shifter. Since the circuits are only needed for updating the phase shifter, they may be gated off during most of the interpulse period.

3.2.3.8 Transmitter Analysis

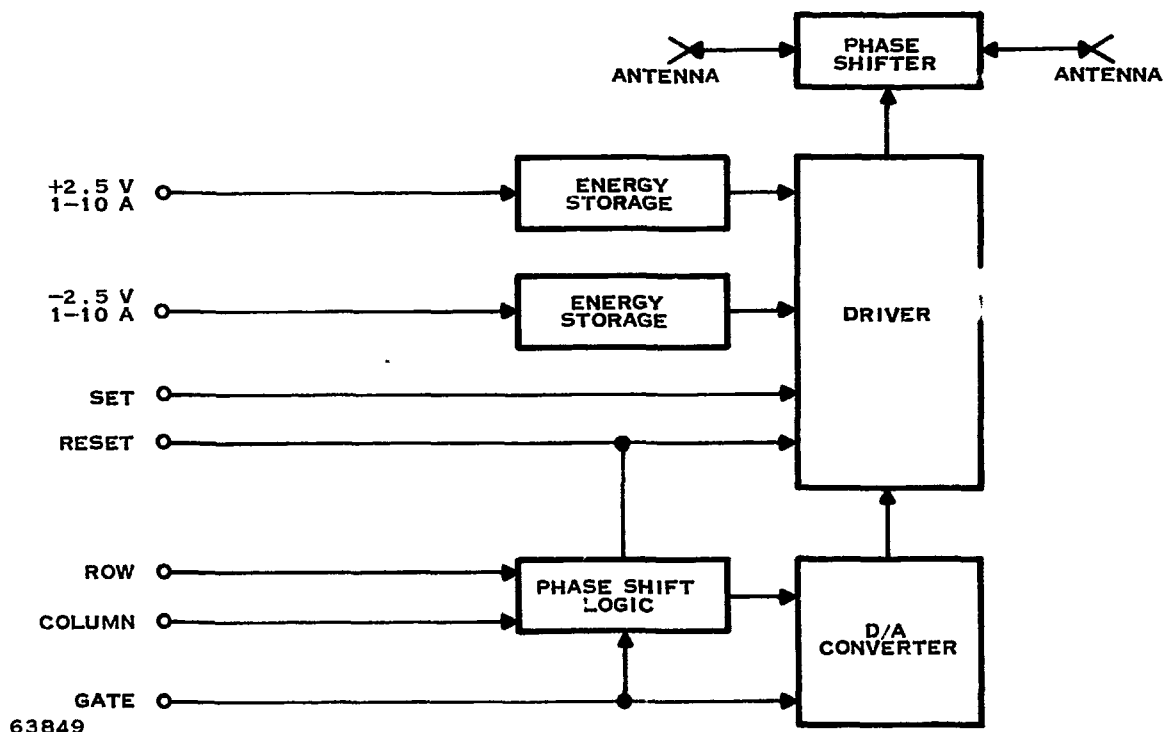
3.2.3.8.1 General

Microwave vacuum tube power sources fall into two general categories, coherent and noncoherent transmitters. The basic difference is that the coherent transmitter is a microwave amplifier and the noncoherent transmitter is a microwave oscillator. Several considerations involved in making a tradeoff analysis are:



63848

Figure 3-105. Reciprocal, Reflective or Transmissive Ferrite Phase-Shifter Module Block Diagram



63849

Figure 3-106. Nonreciprocal, Transmissive Ferrite Phase-Shifter Module Block Diagram

Peak power
Average power
Efficiency
Bandwidth
Reliability
Size
Weight
High voltage
Pulse length
Duty cycle
Gain
Prf.

The noncoherent transmitter is a better choice on the basis of design risk; that is, as a general rule it is simpler to design. For a given power level, the noncoherent transmitter can weigh as much as 30 percent less than an equivalent coherent transmitter. In general, a noncoherent transmitter will cost less. One disadvantage of the noncoherent transmitter is that high peak power is required to get high average power. This causes problems with waveguide arcing, pressurization and switching.

A modulator is generally required to turn the microwave tube on and off in a precise manner. In a broader sense, the modulator supplies the energy which causes an rf source to oscillate or amplify, thus creating a burst or pulse of rf energy.

Pulse modulators depend on the storage of electrical energy in an electrostatic field, and the subsequent discharge of all or part of this energy into the load (transmitter tube). Those that discharge only part of the stored energy are known as "hard tube modulators", those that discharge all of their stored energy are called "line type modulators". A solid-state magnetic modulator, in the line type category, is recommended here because it is smaller, lighter, more reliable, and has better efficiency.

3.2.3.8.2 Microwave Tubes

There are basically five types of transmitter tubes for use in pulsed-radar systems—pulsed traveling-wave-tube amplifiers (PTWT), klystron pulse amplifiers (KPA), crossed-field amplifiers (CFA), magnetrons, and planar triodes. Table III-16 summarizes some of the characteristics of these tubes.

Table III-16. Transmitter Tubes for Space-Fed Array

<u>Tube</u>	<u>Remarks</u>
1. Pulsed Traveling Wave Tube	Power Out— $1-10^6$ W; Efficiency—10-30 percent; high gain (60 dB), wide bandwidth, high voltage, best reliability data, coherent operation, long pulse, CW tubes space qualified
2. Magnetron	Power Out— $0.1-10^6$ W; Efficiency—15-35 percent; noncoherent, high voltage, short pulse, low duty cycle
3. Crossed-Field Amplifier	Power Out— $0.1-10^6$ W; Efficiency—35-55 percent; low gain (15 dB), coherent, high voltage, medium duty cycle, long pulse, wide bandwidth, no heater
4. Klystron Pulse Amplifier	Power Out— $0.1-10^6$ W; Efficiency—20-40 percent; high gain, narrow bandwidth, high average power, medium pulse, heavy
5. Planar Triode	Power Out— <1 kW; Efficiency—up to 65 percent; low frequency (<6 GHz), low voltage, short pulse, very small, space qualified (Gemini), noncoherent

The PTWT's are generally low-power (<25 kW) devices, but some high-power units (>250 kW) are available for special ground-based applications. Efficiency is in the 10- to 30-percent region currently and is expected to increase to 40-50 percent in the next two or three years. Its gain is quite high (up to 60 dB), thus it can convert milliwatts to kilowatts. This eliminates the need for a driver stage. Typical pulselengths are 2 to 20 μ sec. Only about 3 to 5 percent of the energy injected into a PTWT is dissipated in its slow-wave structure (as contrasted with 50 percent for CFA's). This also makes the PTWT's available for medium and high prf duty-cycle applications. The PTWT's wide bandwidth capability, up to 4 GHz, is more than adequate for any pulse compression or frequency diversity configuration the CAT radar could have. Traveling wave tubes have been used extensively in spaceborne

communication systems. These are all low-power CW tubes. Nevertheless, the design techniques, construction and back-up reliability data indicates the PTWT is the most dependable tube available.

Magnetrons are used more extensively in radar systems than any of the other tubes. These are available over a wide range of power levels and have efficiencies about the same as PTWT's. Significant disadvantages include noncoherent operation (pulse compression not possible) and short pulse-low duty cycle limitation (which requires high peak power to get the average power up). Magnetrons are not known to be long life devices.

CFA's inherently have low gain (20-dB maximum, 13-dB nominal), which is a definite disadvantage. Nonlinear beam CFA's are bandwidth limited due to some electron beam feedback. This feedback, however, yields greater efficiency (35 to 55 percent), an outstanding advantage. CFA's have been built with peak-power capabilities in the megawatt region and are typically used only for medium and high-power applications. The duty cycle, and therefore average power, is usually limited to less than 0.5 percent. This limitation is due to the heat dissipating slow wave structure. Because of the high efficiency, CFA's are the likely choice for most high-power, low-prf coherent transmitters. In a few instances, one might find a CFA to use in a medium prf system, but the TWT is more likely to fill this slot.

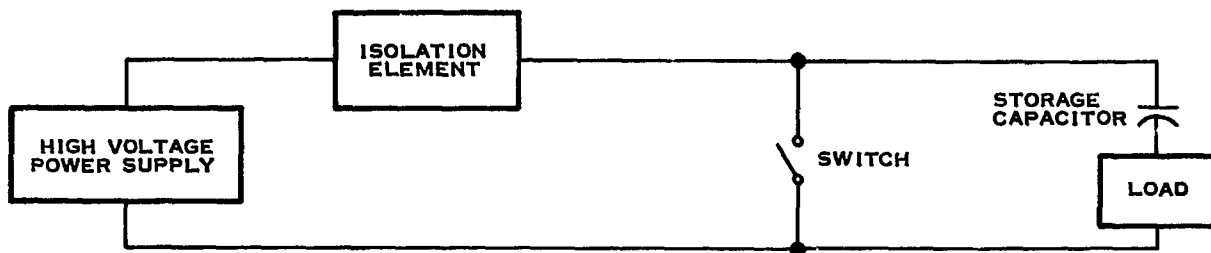
A competitor of the TWT for high-prf applications is a klystron power amplifier (KPA). This is a tuned cavity tube and therefore has a very narrow bandwidth (15 to 25 MHz). The rest of its qualifications are on the plus side; that is, high gain (40 to 60 dB), low voltage requirements (compared to the PTWT), rugged internal heat dissipation structure, good efficiency (≤ 40 percent), and high-average-power capability. Most KPA's are designed for high-power applications and are quite large.

The planar triode, a small tube, has limited power and frequency capability. It was used on the Gemini rendezvous radar, a cooperative radar, and required only 1 kW of peak power. Its high efficiency makes it a good candidate for L- and S-band systems with a limited amount of power.

3.2.3.8.3 Modulators

Modulator considerations⁵⁰ are sometimes ignored in comparing various transmitter systems, so a little background will be given here on modulator techniques.

Hard-tube modulators were common in early radar sets and are used today mostly in high-prf systems. A hard-tube modulator is illustrated in Figure 3-107. The capacitor is charged to a high voltage and then



63850

Figure 3-107. Hard-Tube Modulator

partially discharged into the load by closing the switch momentarily. The switch is typically a triode, tetrode, or beam-tetrode tube. The high-voltage power supply is part of the modulator and adds to its size and weight. Also, a large amount of power is used in generating the high voltage.

The line-type modulator is illustrated in Figure 3-108. In a line-type modulator, the energy storage element is usually a transmission line section with lumped inductances and capacitances. This line section is called a pulse-forming network, or PFN, and its function is to store energy and shape the output pulse. The hydrogen thyatron tube is generally used to trigger the high-voltage pulse.

Figure 3-109 shows a relatively new type of modulator which is the all-solid-state magnetic modulator. In this circuit, the output pulse energy is in a capacitor at low voltage. At a specified time, an SCR discharges the capacitor into the saturable reactor. The voltage is stepped up and applied to the PFN. At some later time, the reactor "saturates" and behaves as a closed switch and discharges the PFN. A pulse transformer steps up the voltage even higher and applies it to the load.

Advantages of solid-state magnetic modulators as compared to a hard-tube or soft-tube modulator are:

1. Reduced size and weight
2. Greater efficiency
3. Higher inherent reliability
4. Longer operating life and higher MTBF
5. Lower maintenance
6. Less pulse-time jitter (than soft-tube modulators)
7. No high-voltage supply required.

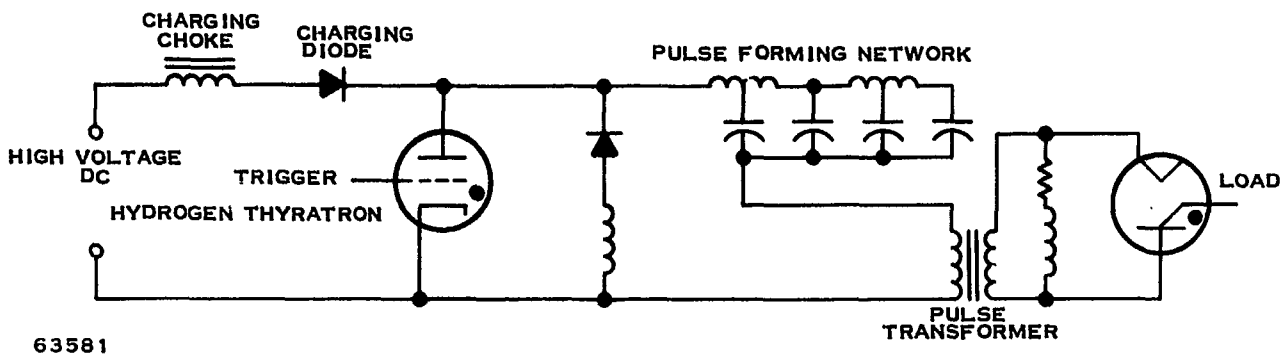


Figure 3-108. Line-Type Modulator Diagram

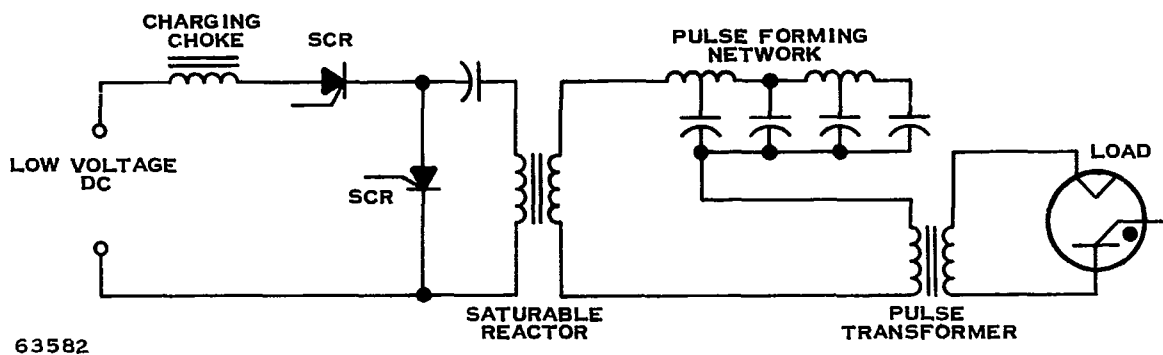


Figure 3-109. Magnetic Modulator Diagram

The disadvantages are:

1. Pulsewidth not continuously variable
2. Lower maximum prf than hard-tube units
3. Cannot withstand as large a load impedance mismatch as hard-tube units
4. Lower output power than soft-tube units.

Some typical parameters of solid-state magnetic modulators are shown in Table III-17.

3.2.3.8.4 Rf Breakdown in Waveguide

One significant problem in the space-fed array is the requirement for a single-channel feed for the transmitter power from the tube to the

radiating horn. The waveguide and components carrying this power are susceptible to breakdown at some rather low power levels if the gas pressure is near the paschan point. Figure 3-110 shows the variation of pulse breakdown power in standard waveguide, versus air pressure and W/G height. The curves show that W/G breakdown power is about three orders of magnitude worse at about the 100, 000-foot level than at sea level. The curves also show that the W/G can handle the shorter pulses better. In the complete absence of air, little problem would exist. Studies of the Gemini rendezvous radar (1 kW peak pulse power) and other applications reveal the likelihood of having some gas density due to out-gassing of materials or spacecraft propulsion gas. Transmission line VSWR's will further degrade the power-handling capability.

Perhaps the best solution to this problem is pressurization of the waveguide circuit (pressurization was used on the Gemini radar). This would present many design problems, especially on long missions, and would have to be carefully considered in a development program on the CAT radar.

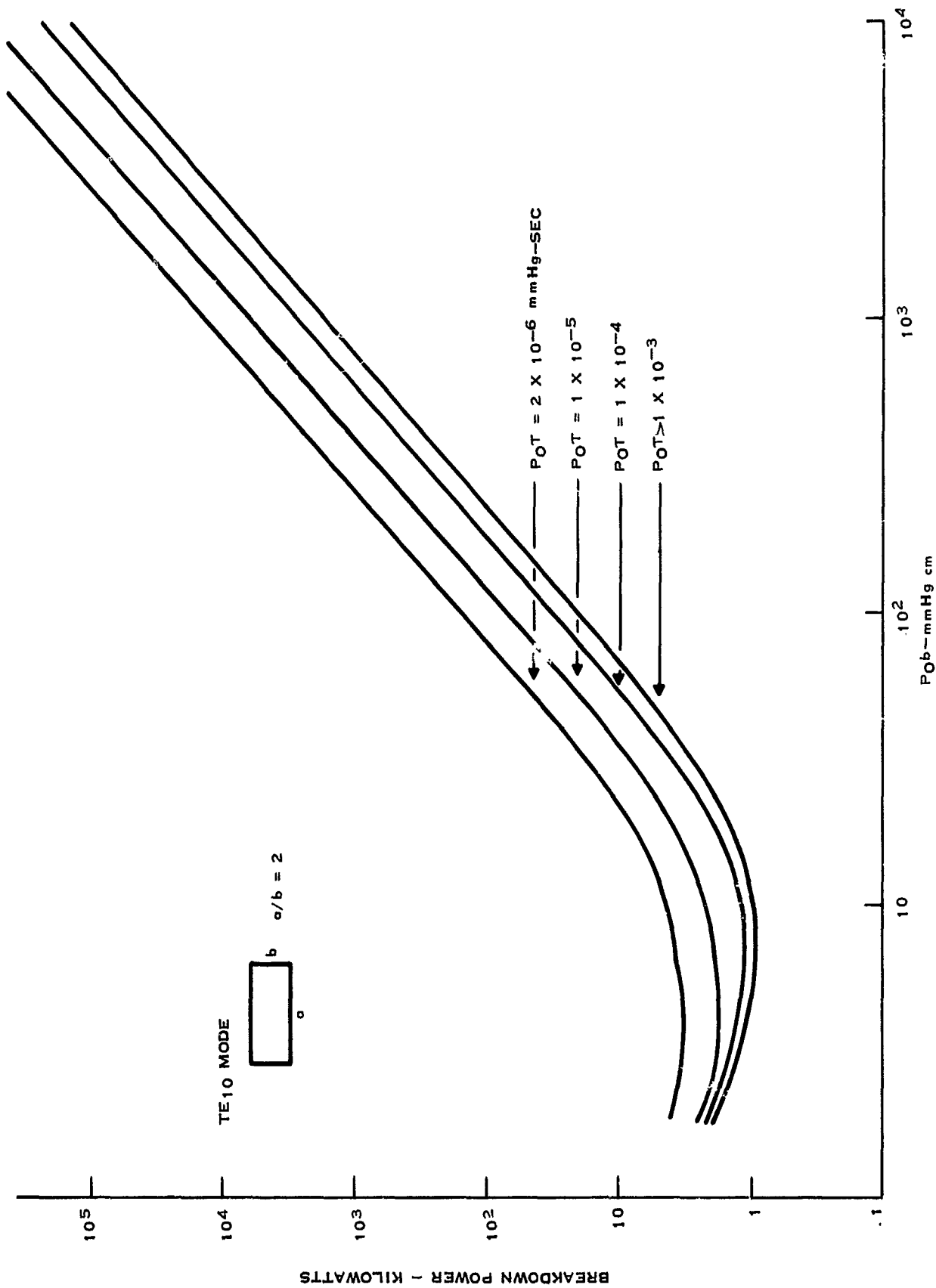
Table III-17. Solid-State Magnetic Modulator

Modulator Output	5 kV at 5A	10 kV at 10A	16 kV at 16A
Peak Power	25 kW	100 kW	256 kW
Pulsewidth	2.5 μ sec	2.5 μ sec	2.5 μ sec
Prf	400 pps	400 pps	400 pps
Weight	4.5 lb	6 lb	12 lb
Efficiency	60%	60%	60%

3.2.3.9 Space-Fed Array Analysis

The front end of the space-fed array is shown in the block diagram in Figure 3-111. The transmitter circuit includes the transmitter tube (or paralleled tubes) with its modulator, some transmission-line circuitry, and the phase-shift modules. As will be shown later, most of the input power, P_T (assumed to be 1 kW), will go to the phase-shift modules for maximum radar range. A single PTWT or paralleled PTWT's are recommended along with an all-solid-state magnetic modulator. Optimistic efficiencies for each are assumed which give an overall efficiency of 20 percent for the average power output, P_A , from the transmitter. This power suffers an insertion loss, L_T , which reduces it to the average transmitted power, P_{TA} , out of the array elements. The transmitter losses for systems from 1.5 to 18 GHz are tabulated in Table III-18. The transmitter overall efficiency is

$$\eta = \frac{P_{TA}}{P_T} \quad (3-15)$$



63853

Figure 3-110. Single-Pulse Breakdown Power for Air-Filled Standard Waveguides

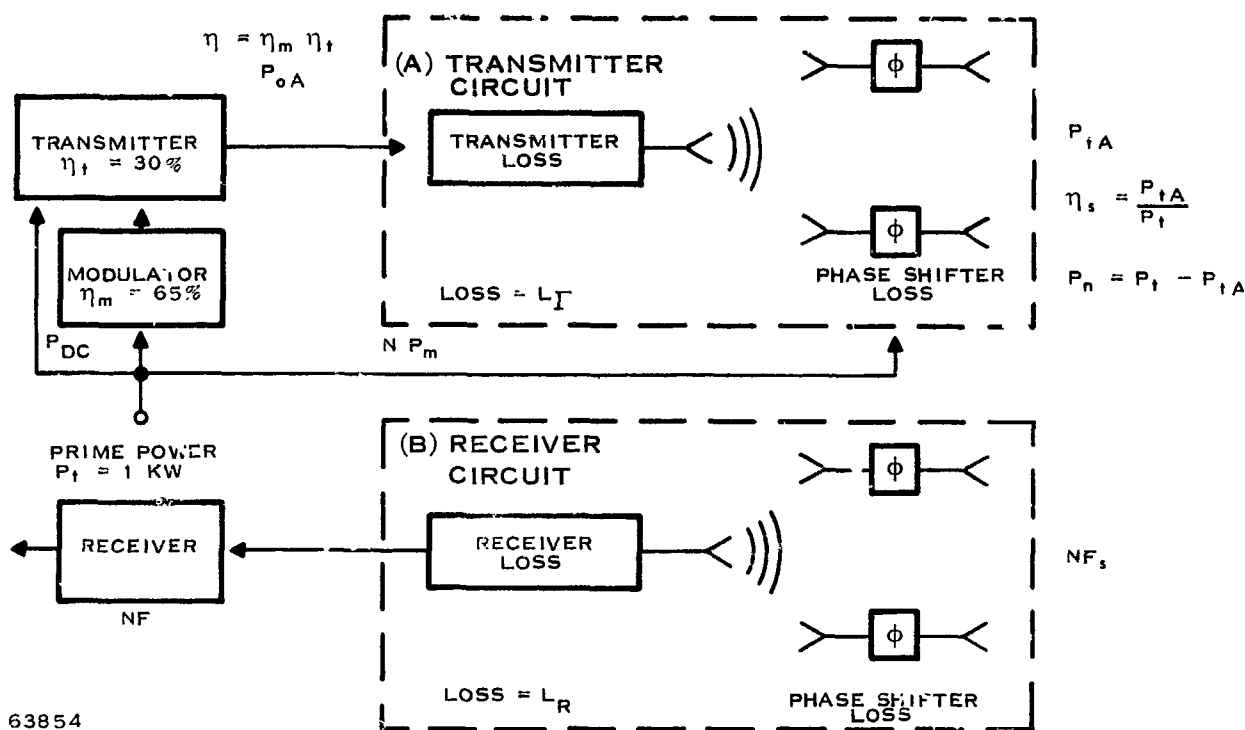


Figure 3-111. Transmitter and Receiver Circuit Description for Space-Fed Array

The system synthesis section will show that this efficiency will vary from 3-10 percent and the heat load, P_H , will vary from 905-970 watts, depending on frequency and certain size/weight restrictions.

The receiver circuit is shown in Figure 3-111b. Again loss is suffered in the phase-shift modules and in transmission line elements. These are summarized in Table III-18.

Reducing the system noise figure as much as possible is highly desirable. A 1-dB improvement in noise figure is equivalent to increasing the transmitter power 1 dB. Unfortunately, increasing the transmitter output power costs a great deal in input power and heat load because of poor efficiency in the transmitter circuits.

The receiver front ends considered were the conventional, balanced-mixer, tunnel-diode amplifier (TDA), parametric amplifier and transistor amplifier. All are solid-state devices and will not be discussed in detail here because of their coverage in the literature (Subsection 3.2.2.6). Receiver noise figures using each of these front ends are summarized in Table III-18. Uncooled parametric amplifiers will produce the lowest noise figures but are the most complex and least reliable of those considered. The higher frequency parametric amplifiers would require pump sources at K_a -band or higher. Use of a TDA or transistor amplifier if possible is

recommended. A tradeoff between the TDA and balanced mixer should consider reliability and dynamic range.

Table III-18. Transmitter-Receiver Contributions for Space-Fed Array

Freq. (GHz)	Receiver NF				Transmitter Losses			
	Mixer (dB)	TDA (dB)	Paramp (dB)	Trans- sis (dB)	Array (dB)	Dupl. (dB)	Line (dB)	Total (dB)
1.5	6.0	4.5	2.5	4.0	1.5	0.3	0.5	2.3
3.0	6.0	4.5	3.0	6.5	1.5	0.3	0.5	2.3
6.0	6.5	5.0	3.5	—	1.6	0.4	0.6	2.6
9.0	6.5	5.5	3.5	—	1.8	0.5	0.7	3.0
18.0	7.5	6.5	4.0	—	2.0	0.6	0.9	3.5

Freq. (GHz)	Receiver Losses (dB)			
	Array	Duplexer/Limiter	Line	Total
1.5	1.5	0.9	0.5	2.9
3.0	1.5	0.9	0.5	2.9
6.0	1.6	1.0	0.6	3.2
9.0	1.8	1.2	0.7	3.7
18.0	2.0	1.4	0.9	4.3

Another front-end element considered, but not listed, was the image-terminated mixer. Recent work by Barber⁵³ and others shows lower noise figures can be obtained in balanced mixers by terminating the image frequency in either a short or open circuit. Barber, for instance, reported getting 4 dB at 6 GHz. Some circuit problems, such as narrow bandwidth and filter losses, need to be worked out. As lower noise UHF transistors become available this mixer will be applicable in narrowband systems.

3.3 Signal Processing Analysis

Signal-processing circuitry for the CAT radar must perform three functions: target detection, target tracking in range, and target tracking in angle. In this section, various subsystems will be discussed which contribute to the performance of these functions. In particular, three types of range-tracking techniques and three types of angle-tracking techniques will be considered, and performance measures for the subsystems will be given. Before proceeding with the discussion, it would be well to view the overall signal-processing system to obtain a clear idea of the interrelationships among the various subsystems. Figure 3-112 shows a simplified block diagram of the portion of the signal-processing system which is invariant with the various tracking techniques. Thus, it serves as the foundation upon which the complete processing system is constructed. The signals from the individual modules, at an intermediate frequency, are combined in the signal-combining circuitry to form one or more composite signals; regardless of the type of tracking technique, one of the composite signals is always the sum of the signals from all N-array modules. The sum signal is amplified and filtered in the if. strip; in the simplest case, the if. strip will consist of bandpass amplifiers, but if pulse-compression techniques are used, it will also include the pulse-compression filter. Envelope detection of the output of the if. strip gives the video (compressed or uncompressed) pulses which are integrated to improve the signal-to-noise ratio. If the level of the integrated signal plus noise exceeds the threshold of the threshold device, a detection is declared. The time of occurrence of the detection decision can be measured to obtain an estimate of target range, or the detection decision can be used to trigger the tracking subsystems. Various range-tracking techniques will be discussed next.

3.3.1 Range-Tracking Subsystems

There are three types of range-tracking techniques which are pertinent to pulsed radars. The first of these, an open-loop measurement, involves the measurement of the time of occurrence of the detection decision that a target is present. The other two techniques, although different in detail, are similar in concept. They involve the closed-loop positioning of a locally generated gate pulse, so that its time of occurrence is the same as that of a pulse returned from a target. Thus, the position of the gate pulse can be used to estimate the range to the target. These techniques are now discussed in more detail, and performance measures derived.

3.3.1.1 Open-Loop Range Measurement

Figure 3-113 shows the block diagram of a system which combines the detection decision with a measurement of target range. Transmission of a pulse sets the flip-flop to the "on" position and triggers the lower clock. The AND gate will then be on, and the upper clock counts in synchronism with the lower clock. If level of the integrated video signal plus noise exceeds the bias level of the flip-flop, the flip-flop is reset to the "off" position, the AND gate

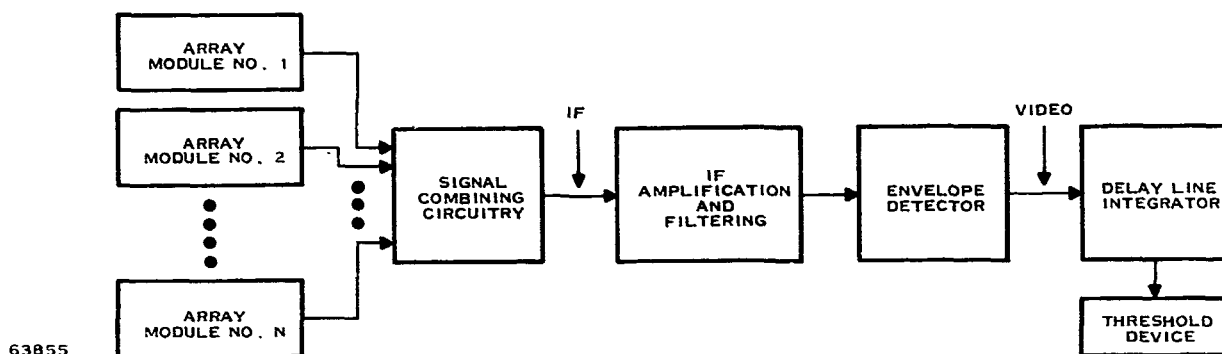


Figure 3-112. Invariant Portion of Signal-Processing System

is turned off, and the upper clock stops counting. The count of the clock is then a measure of the time difference between transmission of a pulse and crossing of the threshold. Assuming that the accuracy of the technique is not limited by the clock rates, then the maximum theoretical accuracy in the measurement of the time of accuracy of the threshold crossing can be shown to be:

$$\sigma_t = \frac{1}{\sqrt{2} B_r \sqrt{\left(\frac{S}{N}\right)}} \quad S/N \gg 1, \quad (3-16)$$

where

σ_t = standard deviation of time measurement (sec)
 = rms thermal noise error in time measurement (sec)

B_r = if. bandwidth of receiver (Hz)

τ = duration of envelope-detected pulse (sec)

S/N = if. signal-to-noise power ratio of receiver for a single pulse.

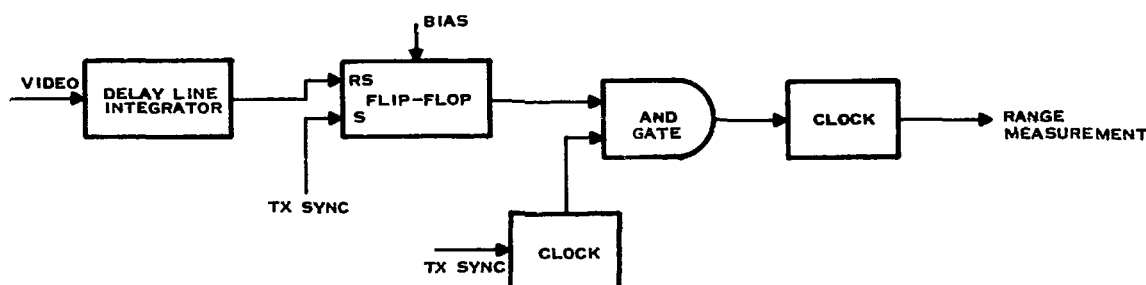


Figure 3-113. Combined Detection and Range Measurement System

Taking into account the improvement of effective signal-to-noise ratio due to pulse integration, Equation (3-16) becomes

$$\sigma_t = \frac{1}{\sqrt{2B_r} \sqrt{nE_i(n)(S/N)}} \quad nE_i(n)S/N \gg 1, \quad (3-17)$$

where

n = number of pulses integrated

$E_i(n)$ = pulse-integration efficiency for video integration

B_r and S/N are defined above.

The measurement just discussed gives the time of occurrence of the threshold crossing; whereas, it is the time of occurrence of the leading edge of the return pulse that is of interest. The difference between the two is directly proportional to threshold setting and inversely proportional to the i.f. bandwidth of the receiver. However, since this error is a constant for a given bandwidth and threshold setting, it can be biased out after the receiver parameters are fixed.

The dynamic response of the open-loop system is limited primarily by the time required to integrate enough pulse to achieve the required S/N . Marcum⁵⁴ shows that the return pulses contribute their full amplitude to the integration when the integration providing the target movement Δr during the integration period satisfies

$$\Delta r \leq \frac{C\tau}{4}. \quad (3-18)$$

But $\Delta r = \dot{R}\Delta t_i$, where $\dot{R}$ is the target velocity and t_i is the integration time. If n is the number of pulses required to be integrated, then $\Delta t_i = n/f_r$, where f_r is the pulse repetition rate. Thus Equation (3-18) can be put in the form

$$\dot{R} \leq \frac{C\tau}{4\Delta t_i} = \frac{C\tau f_r}{4n}. \quad (3-19)$$

The maximum lag error in range measurement due to target velocity is given by

$$\epsilon_{ss}(t) = \frac{2\dot{R}}{C} \Delta t = \frac{1}{2} \tau. \quad (3-20)$$

Measurement of target velocity (or, equivalently, measurement of the velocity of the leading edge of the return pulse) is easily implemented by numerical differentiation of the range measurement. Thus, if measurements R_1 and R_2 are separated in time by T_m seconds, then the target velocity can be estimated by

$$\dot{R} = \frac{R_2 - R_1}{T_m} \quad (3-21)$$

or, in terms of location of the leading edges of the return pulses,

$$\dot{t} = \frac{t_2 - t_1}{T_m} \quad (3-22)$$

If one assumes t_1 and t_2 to be Gaussian independent variables with standard deviation σ_t given by Equation (3-17), then the standard deviation of $\dot{t}$ given in Equation (3-22) is

$$\sigma_{\dot{t}} = \frac{\sigma_t}{\sqrt{2}T_m} \quad (3-23)$$

3.3.1.2 Split-Gate Range Tracker

Figure 3-114 is a block diagram of a split-gate range-tracking system, and Figure 3-115 shows the associated waveforms. Detection of a target by the threshold device causes the video signal to be applied to the range-tracking loop; this applies a dc voltage to the range-loop integrator, causing the range gate to sweep through the desired range interval, searching for the target. The error generator, in effect, generates a voltage which is proportional to the average value of the product of the received pulse and tracking gate pulse. This voltage constitutes an error signal. Under the desired condition of the tracking pulse being centered on the return pulse, the average value of the product is zero. The error signal is integrated, and the resultant voltage is directly proportional to the position of the tracking gate (and thus may serve as an analog measure of range). Moreover, since the integral of the error signal is an analog range signal, the error signal itself can be used as an analog range rate signal. Thus, the system of Figure 3-114 furnishes analog range and range-rate information, and may be easily modified to furnish digital range (by obtaining a digital measure of the range-gate position) and digital range rate (by numerically differentiating the digital range measurement).

Split-gate range-tracking systems are discussed in detail by Barton,⁵⁵ who shows that optimum performance for such a loop is obtained for $B_r \tau = 1.4$, where B_r is the i.f. receiver bandwidth and τ is the duration

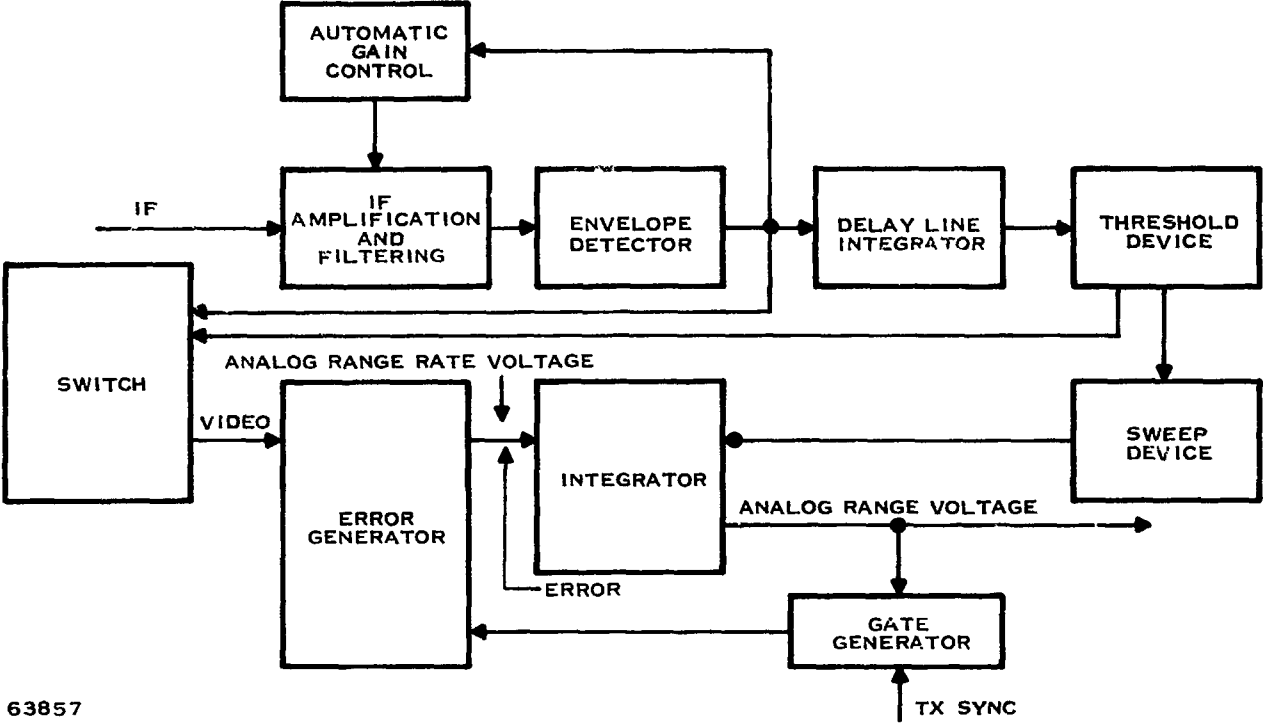


Figure 3-114. Split-Gate Range-Tracking System

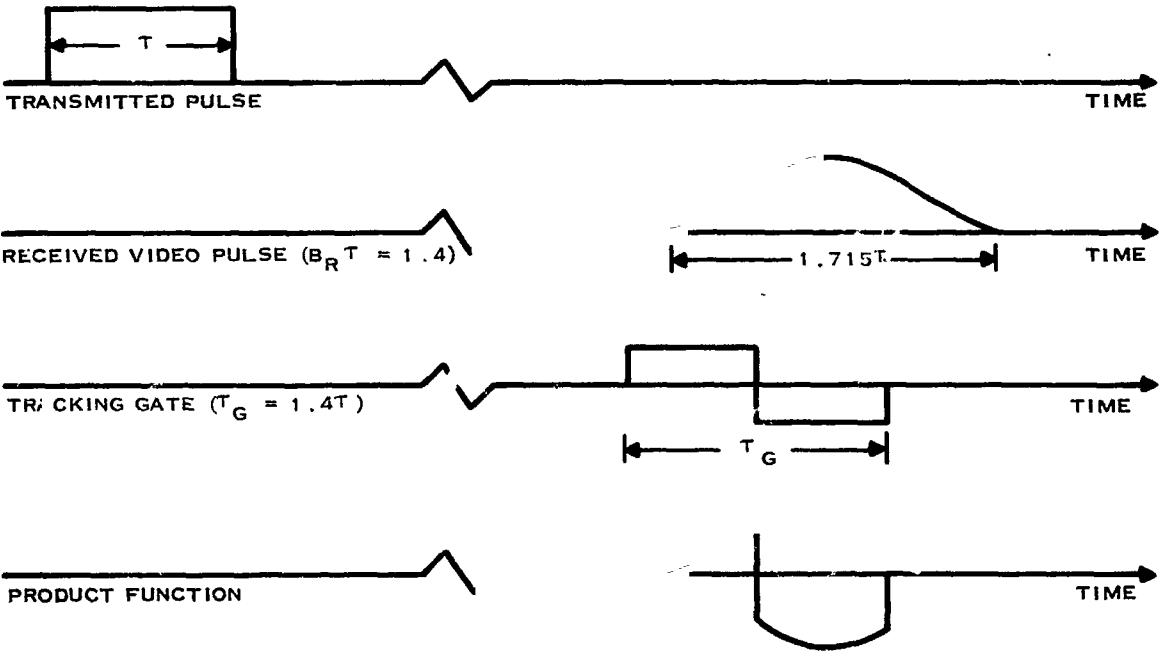


Figure 3-115. Waveforms of Split-Gate Range-Tracking System

of the video (compressed or uncompressed) pulse as $B \rightarrow \infty$, and $\tau_g = 1.4$, where τ_g is the duration of the tracking gate pulse. For these parameters, the thermal noise error is given by

$$\sigma_t = \frac{1}{1.785 B_r \sqrt{\left(\frac{S}{N}\right) \left(\frac{f_r}{B_\ell}\right)}} \quad B_r \tau = 1.4 \quad \tau_g = 1.4 \tau, \quad (3-24)$$

where

B_r = receiver if. bandwidth (Hz)

S/N = if. signal-to-noise power ratio

f_r = pulse repetition rate (sec^{-1})

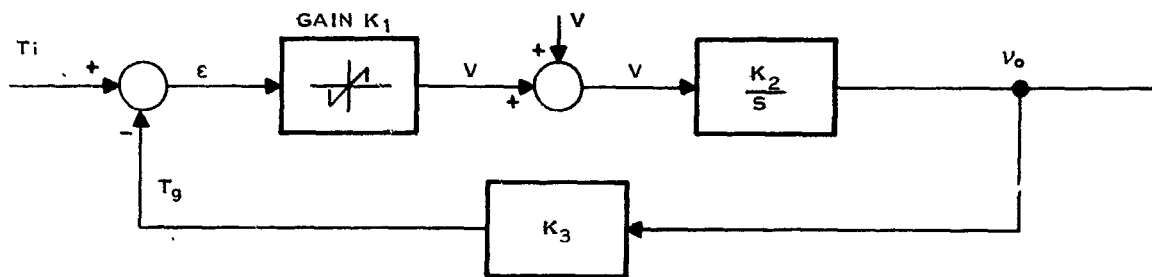
B_ℓ = nominal (high S/N) tracking loop bandwidth (rad/sec)

τ = pulse duration (sec)

τ_g = tracking gate duration (sec).

For low S/N , the tracking loop gain, hence bandwidth, decreases due to the degradation of AGC performance and nonlinear performance of the error detector, and thus tends to improve noise performance; this is counter-balanced, however, by a degradation of S/N in the video detector. Thus, Equation (3-24) is valid for all S/N , but it must be remembered that B_ℓ is not the actual tracking loop bandwidth for low S/N .

In addition to noise response, it is desirable to determine the dynamic response of the tracker. This can be obtained by consideration of Figure 3-116, which illustrates in simplified form the performance of the loop. In this figure, T_i is the location (in time) of the centroid of the return pulse (with respect to the time of the transmitter sync pulse), T_g is the location of the centroid of the gate pulse, V is the dc voltage applied to the integrator to cause the tracking gate to sweep during acquisition, K_1 is the gain (V/sec) of the error generator, which has been partially linearized to remove the effects of input signal amplitudes, K_2 is the integrator gain (V/V), and K_3 is the gain of the gate generator (sec/V). The range gate sweeps in time until it just begins to overlap a return pulse; at this time the error is just within the permissible range for linear operation of the error generator shown in Figure 3-115, namely, $\epsilon = T_j - 1.558\tau$. For this condition, it is easily shown that the response of the system is given (in Laplace transform notation) by



63859

Figure 3-116. Simplified Block Diagram of Split-Gate Range-Tracking System

$$T_g(S) = \frac{K_1 K_2 K_3}{S + K_1 K_2 K_3} T_i(S) + \frac{K_2 K_3}{S + K_1 K_2 K_3} V(S) + \frac{T_g(o+)}{S + K_1 K_2 K_3} \quad (3-25)$$

and

$$\epsilon(S) = \frac{S}{S + K_1 K_2 K_3} T_i(S) - \frac{K_2 K_3}{S + K_1 K_2 K_3} V(S) - \frac{T_g(o+)}{S + K_1 K_2 K_3} \quad (3-26)$$

Calculating the response of the loop to a constant-velocity target, that is, $T_i(t) = T_i + \dot{T}_i t$:

$$T_g(S) = \frac{K_1 K_2 K_3 T_i}{S(S + K_1 K_2 K_3)} + \frac{K_1 K_2 K_3 \dot{T}_i}{S^2(S + K_1 K_2 K_3)} + \frac{K_1 K_2 V}{S(S + K_1 K_2 K_3)} + \frac{T_g(o+)}{S + K_1 K_2 K_3} \quad (3-27)$$

$$\epsilon(S) = \frac{T_i}{S + K_1 K_2 K_3} + \frac{\dot{T}_i}{S(S + K_1 K_2 K_3)} - \frac{K_1 K_2 V}{S(S + K_1 K_2 K_3)} - \frac{T_g(o+)}{S + K_1 K_2 K_3} \quad (3-28)$$

Inverting Equation (3-28) and substituting $T_g(o+) = T_i - 1.558 \tau$,

$$\epsilon(t) = \frac{1}{K_1 K_2 K_3} \left\{ (T_i - K_2 K_3 V) - \left[(\dot{T}_i - K_2 K_3 V) - 1.558 K_1 K_2 K_3 \tau \right] \right\} e^{-K_1 K_2 K_3 t} \quad (3-29)$$

From Equation (3-29), the steady-state error is given by

$$\epsilon_{ss}(t) = \frac{T_i - K_2 K_3 V}{K_1 K_2 K_3} . \quad (3-30)$$

Since the loop will lose lock if $|\epsilon(t)| > 1.558\tau$, we have, for the necessary condition on tracking,

$$\left| \frac{\dot{T}_i - K_2 K_3 V}{K_1 K_2 K_3} \right| < 1.558\tau , \quad (3-31)$$

which can be written

$$-1.558 < \frac{\dot{T}_i - K_2 K_3 V}{K_1 K_2 K_3} < 1.558\tau . \quad (3-32)$$

This equation can be rearranged to give

$$-1.558\tau + K_2 K_3 V T_\ell < \dot{T}_i T_\ell < 1.558\tau + K_2 K_3 V T_\ell , \quad (3-33)$$

where $T_\ell = 1/K_1 K_2 K_3$ is the time constant of the closed-loop system. The time of occurrence of the centroid of the return pulse is related to target range by

$$T_i = 0.858\tau + \frac{2R}{C} , \quad (3-34)$$

where C is the speed of light. Thus $\dot{T}_i = 2\dot{R}/C$. Substituting this equation in Equation (3-33) gives

$$-0.779C\tau + \frac{K_2 K_3 V T_\ell C}{2} < \dot{R} T_\ell < 0.779C\tau + \frac{K_2 K_3 V T_\ell C}{2} . \quad (3-35)$$

Now, in the acquisition mode (while the range gate is sweeping),

$$T_g(t) = K_2 K_3 V t . \quad (3-36)$$

Let ΔR be the range interval to be searched, ΔT_g be the corresponding gate search interval, and ΔT_R be the search time desired for the range loop. Then

$$\Delta T_g = K_2 K_3 V \Delta T_R \quad (3-37)$$

and

$$\Delta R = \frac{C}{2} K_2 K_3 V \Delta T_R . \quad (3-38)$$

Solving for V,

$$V = \frac{2\Delta R}{CK_2 K_3 \Delta T_R} . \quad (3-39)$$

Substituting this result into Equation (3-35),

$$-0.779C\tau + \frac{T_\ell \Delta R}{\Delta T_R} < \dot{R} T_\ell < 0.779C\tau + \frac{T_\ell \Delta R}{\Delta T_R} . \quad (3-40)$$

The number of range segments to be searched is given by $2\Delta R/C\tau$. If n_a is the number of pulses required to be exchanged with each range segment, then the required search time is given by

$$\Delta T_R = \frac{2\Delta R n_a}{C \tau f_r} , \quad (3-41)$$

where f_r is the pulse repetition rate. Substituting this expression into Equation (3-40) gives, after dividing through by T_ℓ

$$-\frac{CT}{2} \left(\frac{1.558}{T_\ell} - \frac{f_r}{n_a} \right) < \dot{R} < \frac{CT}{2} \left(\frac{1.558}{T_\ell} + \frac{f_r}{n_a} \right) . \quad (3-42)$$

Making appropriate substitutions into Equation (3-30) yields, for the steady-state lag error in the position of the centroid of the tracking gate,

$$\epsilon_{ss}(t) = \frac{1}{K_1 K_2 K_3} \left(\frac{2\dot{R}}{C} - \frac{\tau f_r}{n_a} \right) = T_\ell \left(\frac{2\dot{R}}{C} - \frac{\tau f_r}{n_a} \right) , \quad (3-43)$$

where it is to be noted that the product, $K_1 K_2 K_3$, is the closed-loop bandwidth B_ℓ (rad/sec), under the assumed linear operating conditions. As previously noted, the closed-loop bandwidth is reduced at low S/N from its nominal value due to deterioration of AGC action and the nonlinear behavior of the error generator.

In summary of the dynamic behavior of the split-gate range-tracking loop, Equation (3-42) defines the ranges of target velocity which can be tracked, and Equation (3-43) defines the lag error in tracking-gate position due to target velocity.

Two possibilities are available for measurement of target velocity; an analog measurement based on the error signal, and a digital measurement based on numerical differentiation of the digital measurement of the tracking gate position. The digital case is similar to that discussed in connection with the open-loop measurement techniques, and Equation (3-23) again applies: if independent measurements, t_1 and t_2 , are made of the tracking gate centroid, and each has standard deviation, σ_t , given by Equation (3-24), the standard deviation of the estimated $t = (t_2 - t_1)/T_m$ is given by

$$\sigma_t = \frac{\sigma_t}{\sqrt{2}T_m} \quad (3-44)$$

We now derive the standard deviation of the analog range rate measurement. For large if. signal-to-noise power ratios, it can be assumed that the noise power spectral density at the input to the tracking loop is flat and constant over the bandwidth B_r , and that its level is given by A .⁵⁶ The closed-loop power transfer function of the closed-loop system (Figure 3-116) is given by

$$G(\omega) = \frac{K_e^2}{\omega^2 + K_e^2 K_3^2} \quad (3-45)$$

where $K_e = K_1 K_2$. Hence, the noise power of the output is

$$\sigma_o^2 = \frac{1}{2\pi} \int_{-\pi B_r}^{\pi B_r} \frac{AK_e^2}{\omega^2 + K_e^2 K_3^2} d\omega = \frac{AK_e^2}{2\pi} \left[\tan^{-1} \frac{\pi B_r}{K_e K_3} - \tan^{-1} \left(-\frac{\pi B_r}{K_e K_3} \right) \right] \quad (3-46)$$

Now, $K_e K_3$ is proportional to the closed-loop bandwidth, which generally is much less than the if. bandwidth. Equation (3-46) can be approximated by

$$\sigma_o^2 = \frac{AK_e^2}{2} \quad (3-47)$$

The noise power of the signal, T_g , is clearly $K_3^2 \sigma_o^2$; but the noise power of the signal, T_g , is just σ_t given by Equation (3-24), hence,

$$\sigma_t^2 = \frac{AK_e^2 K_3^2}{2} \quad (3-48)$$

so that

$$A = \frac{2\sigma_t^2}{K_e^2 K_3^2} = \frac{2\sigma_t^2}{(K_1 K_2 K_3)^2} \quad (3-49)$$

The power transfer function of the loop between the input and the signal preceding the integrator is

$$H(\omega) = \frac{K_1^2 \omega^2}{\omega^2 + (K_1 K_2 K_3)^2} \quad (3-50)$$

so that the noise power of the signal, v , preceding the integrator is given by

$$\sigma_v^2 = \frac{2\sigma_t^2 K_1^2}{(K_1 K_2 K_3)^2} \frac{1}{2\pi} \int_{\pi B_r}^{\pi B_r} \frac{\omega^2}{\omega^2 + (K_1 K_2 K_3)^2} d\omega \quad (3-51)$$

$$= \frac{2\sigma_t^2}{(K_2 K_3)^2} \frac{1}{2\pi} \left(\omega - K_1 K_2 K_3 \tan^{-1} \frac{\omega}{K_1 K_2 K_3} \right)_{-\pi B_r}^{\pi B_r} \quad (3-52)$$

Again making use of the fact that $B_r \gg K_1 K_2 K_3$, Equation (3-52) reduces to

$$\sigma_v^2 = \frac{2B_r \sigma_t^2}{K_2^2 K_3^2} \quad (3-53)$$

But from Figure 3-116

$$T_g(t) = K_3 v_o(t) \quad (3-54)$$

hence,

$$\dot{T}_g(t) = K_3 \dot{v}_o(t) \quad (3-55)$$

Again from the figure,

$$\dot{v}_o(t) = K_2 v(t) , \quad (3-56)$$

so that

$$\dot{T}_g(t) = K_2 K_3 v(t) . \quad (3-57)$$

Thus, we have finally

$$\sigma_t^2 = K_2 K_3 \sigma_v^2 = \sqrt{2B_r} \sigma_t . \quad (3-58)$$

Position of the centroid of the tracking pulse, whose desired location is coincident with the centroid of the return pulse, has been shown to be subject to two errors: An rms jitter σ_t due to internally generated thermal noise; a bias error ϵ_t due to target velocity. These two errors are defined, respectively, by Equation (3-24) and Equation (3-43). Now, the probability that the loop will lose lock due to the combined effects of these two errors may be determined. Assume that the error signal in the loop is a Gaussian random variable with mean ϵ_{ss} and standard deviation σ_t . It can be described by the probability density function

$$p(\epsilon_t) = \frac{1}{\sqrt{2\pi}\sigma_t} \exp \left[-\frac{(\epsilon_t - \epsilon_{ss})^2}{2\sigma_t^2} \right] . \quad (3-59)$$

As shown in Equation (3-31), the loop will lose lock if the error exceeds 1.558τ , so that the probability of losing lock is

$$P_L = P(\epsilon < -1.558\tau) + P(\epsilon > 1.558\tau)$$

$$= \frac{1}{\sqrt{2\pi}\sigma_t} \int_{-\infty}^{-1.558\tau} \exp \left[-\frac{(\epsilon_t - \epsilon_{ss})^2}{2\sigma_t^2} \right] d\epsilon_t + \frac{1}{\sqrt{2\pi}\sigma_t} \int_{1.558\tau}^{\infty} \exp \left[-\frac{(\epsilon_t - \epsilon_{ss})^2}{2\sigma_t^2} \right] d\epsilon_t . \quad (3-60)$$

With the appropriate change in variables, Equation (3-60) becomes

$$P_L = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{-\frac{1.558 + \epsilon_{ss}}{\sigma_t}} \exp\left[-\frac{y^2}{2}\right] dy + \frac{1}{\sqrt{2\pi}} \int_{\frac{1.558 - \epsilon_{ss}}{\sigma_t}}^{\infty} \exp\left[-\frac{y^2}{2}\right] dy . \quad (3-61)$$

3.3.1.3 Separated Gate Range Tracker

Figure 3-117 is the block diagram of a separated gate range tracker. It is, conceptually, quite similar to the split-gate tracker system, the main differences being in the receiver if. bandwidth, and in the method of error generation. Figure 3-118 shows the waveforms associated with the separated gate tracker.

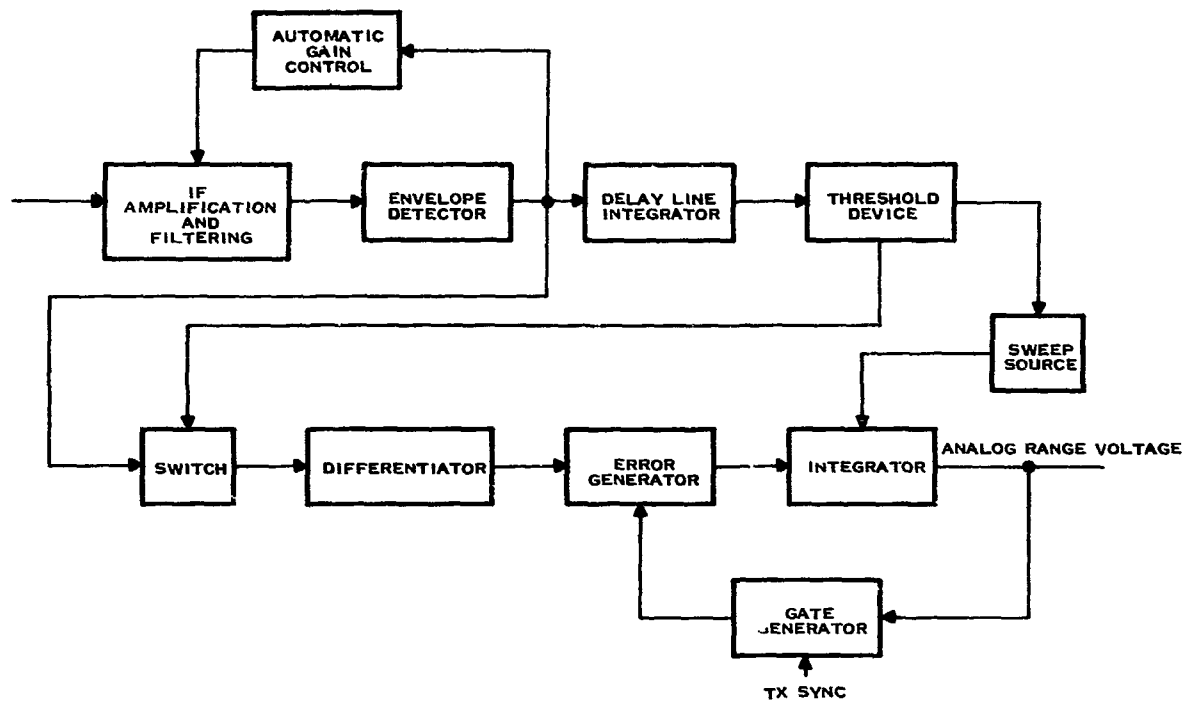
Barton⁵⁷ discusses the separated gate tracker and shows that the thermal-noise error in the position of the centroid of the tracking gate is given by

$$\sigma_t = \frac{1}{1.4 B_r \sqrt{\left(\frac{S}{N}\right) \left(\frac{f_r}{B_l}\right)}} , \quad (3-62)$$

where all symbols are as previously defined. Comparison of this expression with Equation (3-24) shows the split-gate tracker to give superior performance for comparable bandwidths. However, the expression of Equation (3-24) is valid only for $B_r \tau = 1.4$. For a higher $B_r \tau$ product, the performance of the split-gate tracker degrades. Equation (3-62), however, is valid for all $B_r \tau$ products greater than unity. Thus, for a system with a high $B_r \tau$ product, the separated gate system may give superior performance depending on how high the product is.

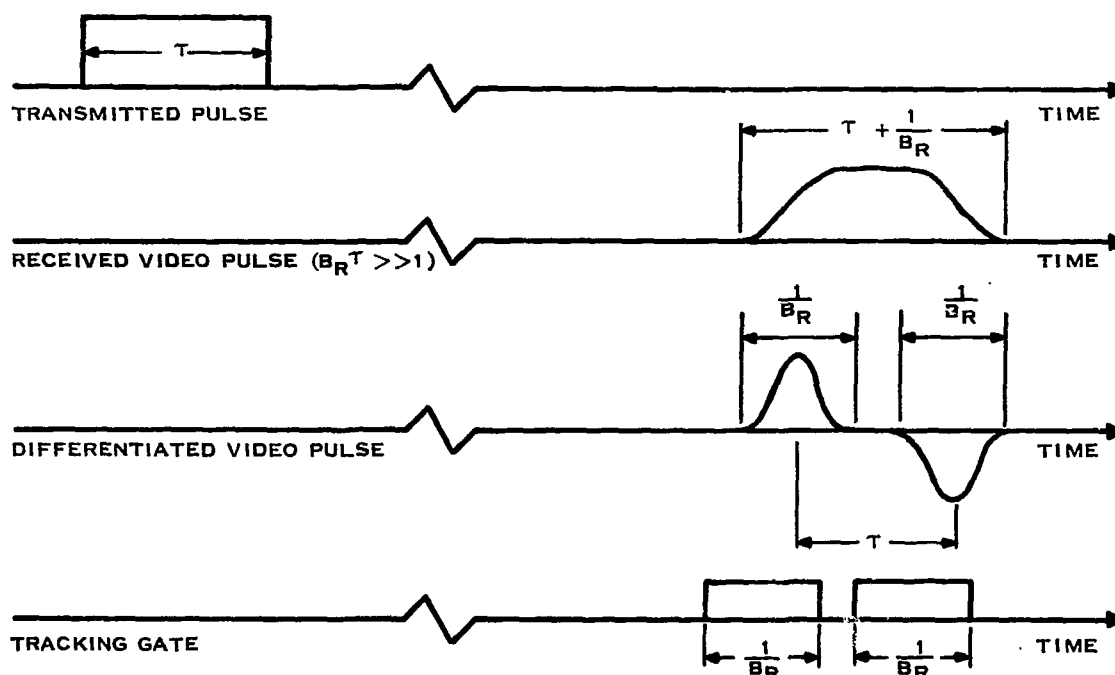
Dynamic response of the separated gate tracker, with suitable linearization of the error generator, proceeds in much the same fashion as for the split-gate tracker, and will not be repeated here. From Figure 3-118 it is clear that the loop will lose lock if the centroid of the tracking gate is in error by an amount greater than $\tau + 1/B$, hence, we have for permissible target velocities

$$-\frac{C\tau}{2} \left(\frac{1 + \frac{1}{B_r \tau}}{T_l} - \frac{f_r}{n_a} \right) < \dot{R} < \frac{CT}{2} \left(\frac{1 + \frac{1}{B_r \tau}}{T_l} + \frac{f_r}{n_a} \right) , \quad (3-63)$$



63860

Figure 3-117. Separated-Gate Range-Tracking System



63861

Figure 3-118. Waveforms of Separated-Gate Range-Tracking System

where all symbols are as defined for the split-gate tracker. As for the split-gate tracker, the lag error due to target velocity is given by

$$\epsilon_{ss}(t) = \frac{1}{T_\ell} \left(\frac{2\dot{R}}{C} - \frac{\tau \dot{f}_r}{n_a} \right) . \quad (3-64)$$

For the assumed linearity of the error generator, the standard deviation of digital and analog range-rate measurements are the same for the separated-gate tracker as for the split-gate tracker.

$$\sigma_t = \frac{\sigma_t}{\sqrt{2T_m}} \quad \text{for the digital measurement} \quad (3-65)$$

and

$$\sigma_t = \sqrt{2B_r} \sigma_t \quad \text{for the analog measurement} . \quad (3-66)$$

Finally, for the separated-gate tracker, the probability of loss of lock is given by

$$P_L = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\frac{\left(\tau + \frac{1}{B_r} + \epsilon_{ss}\right)}{\sigma_t}} \exp\left[-\frac{y^2}{2}\right] dy + \frac{1}{\sqrt{2\pi}} \int_{\frac{\tau + \frac{1}{B_r} - \epsilon_{ss}}{\sigma_t}}^{\infty} \exp\left[-\frac{y^2}{2}\right] dy . \quad (3-67)$$

3.3.2 Angle-Tracking Subsystems

The mission of the CAT radar will be to select a particular target and follow its course in range, angle, and velocity. In this section, angular-tracking techniques and methods of extracting the angular-tracking signals will be considered. The angle trackers discussed here will employ a high-gain pencil beam formed by the array, with two degrees of freedom. The pointing of the beam is controlled by closed-loop angular error feedback when target acquisition has occurred.

The versatility of the phased array makes it possible to mechanize several forms of angular track. Monopulse and conical scan have proved to be the most useful for mechanically scanned antennas, and these will be discussed in detail. The inertialess property of the electronic scan of a phased-array antenna permits variations on these two basic tracking techniques. One technique, designated conopulse, which takes advantage of the better features of both monopulse and conical scan will be discussed.

3.3.2.1 Monopulse

Monopulse, as the name implies, is a tracking technique which derives all the tracking information from a single pulse and generates new and independent error information with each new pulse. Thus, pulse-to-pulse target cross-section amplitude fluctuations (scintillation) do not cause an angle-tracking error. The block diagram of a monopulse radar, for a phased-array antenna, is shown in Figure 3-119 and Figure 3-120.

The distinguishing feature of a monopulse system is the connection of the rf circuits of four phased-array segments such that sum-and-difference signals are obtained simultaneously. The patterns of the array segments overlap and are used for generating the angle-tracking information. The sum signals from separate array segments merge the array segment patterns into a single pattern, illustrated by Figure 3-121 a, while the difference signals produce a null pattern with sharp zero at the center, as illustrated by Figure 3-121 b.

Angle tracking requires simultaneous measurement of two quantities: magnitude and sense of the angle-tracking error. This is accomplished at the phase-sensitive detector where the sum-and-difference signals are phase compared. The relative phase provides the sense, and the slope of the difference pattern provides the magnitude. Thus, the radiation pattern determines the angular error slope; that is, the rate of change of error signal with respect to target angular motion off the tracking axis. This quantity is important in evaluating the noise response of the tracker. The radiation pattern of the sum signal can be approximated by $\sin x/x$; a study of actual antenna⁵⁸ patterns has indicated that the sum pattern is more accurately represented, in the vicinity of the peak, by

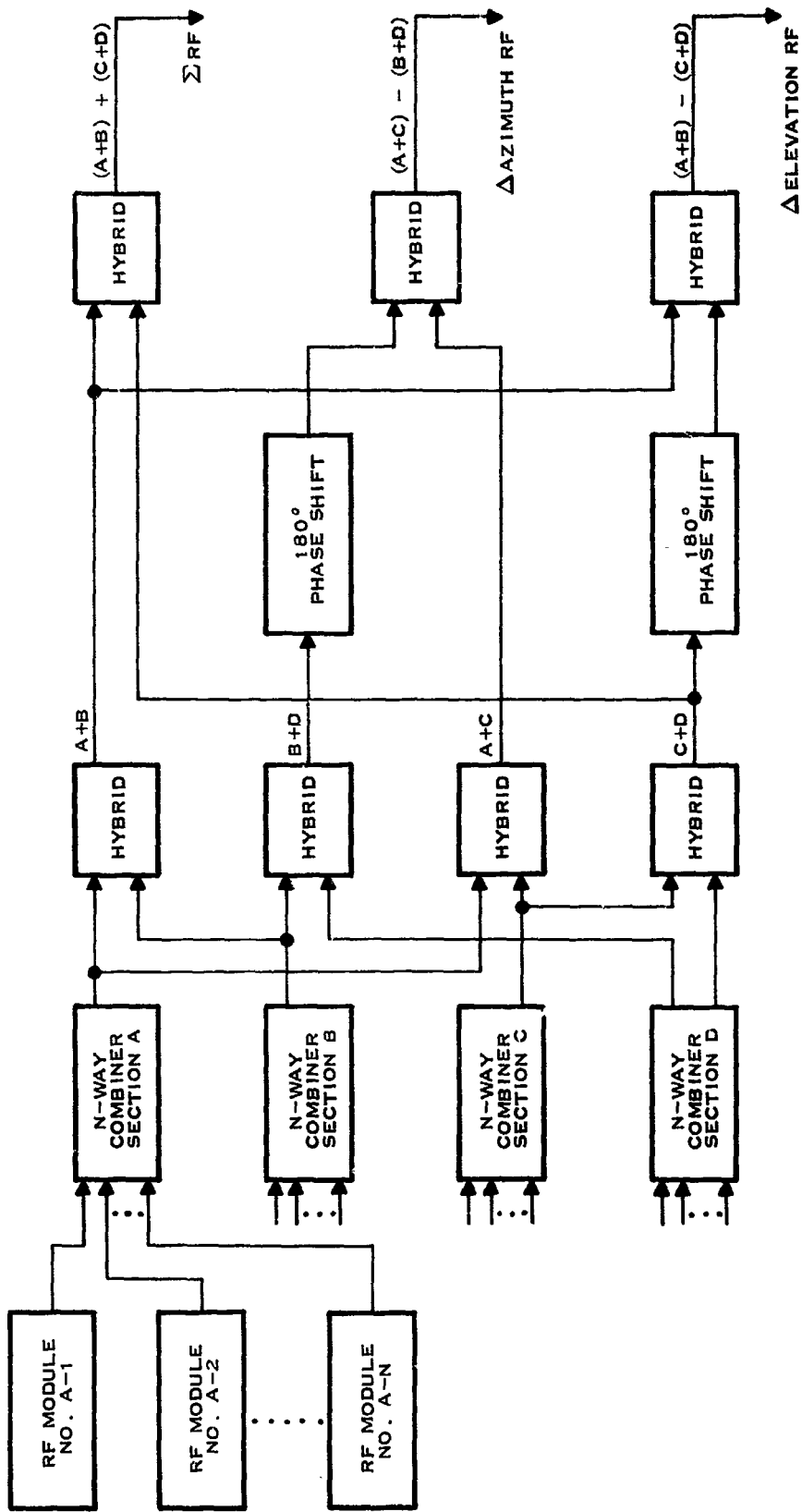
$$E_s = \cos^2 (1.18\xi) \quad (3-68)$$

and the difference pattern by

$$E_o = 0.707 \sin (2.36\xi),$$

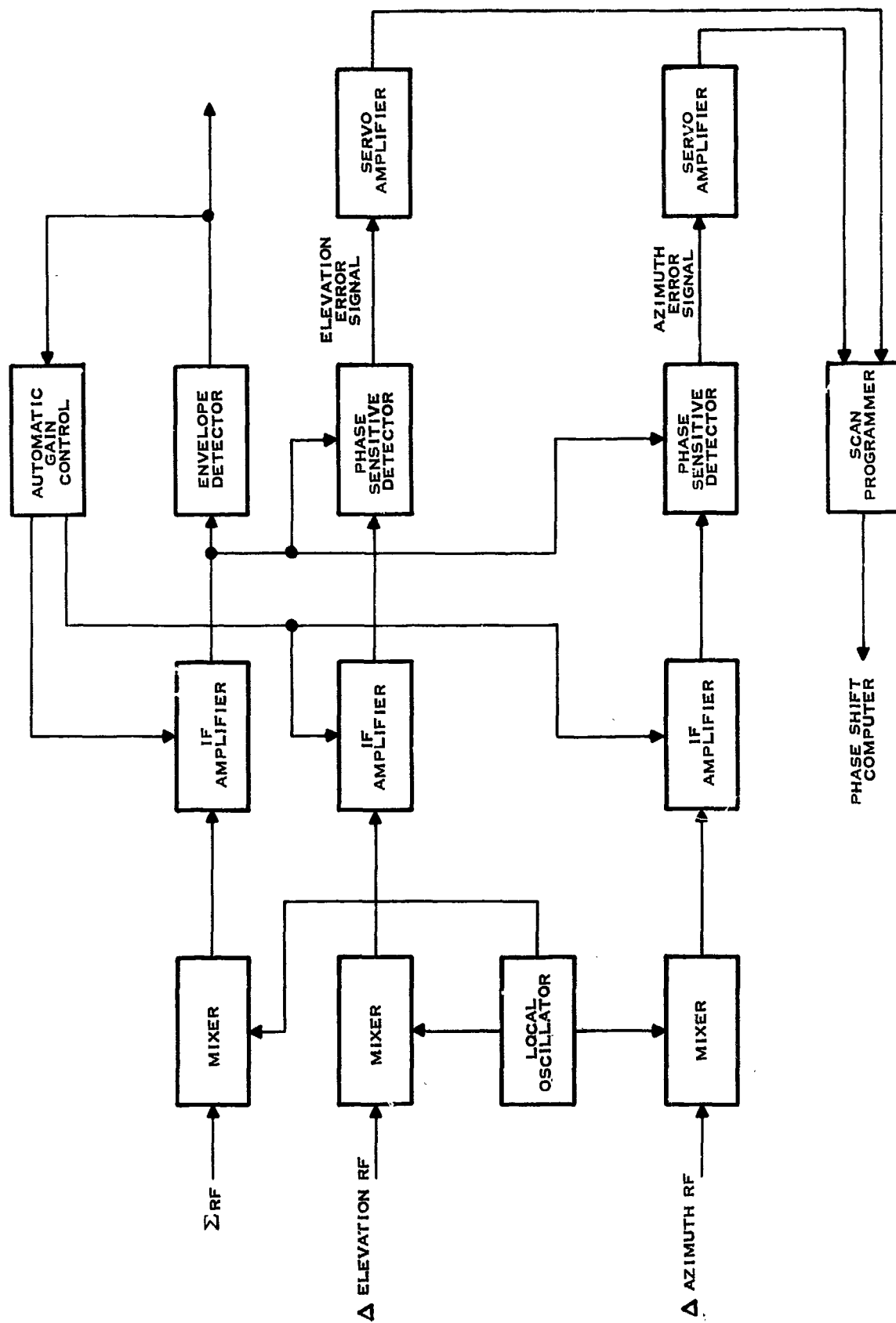
where ξ is the normalized off-beam-center angle in beamwidths, the constants 1.18 and 2.36 have units of radians per beamwidth, and E is the normalized voltage signal. The error slope has been found, experimentally, to be approximately 1.57 for ξ near zero.

The noise response of monopulse angle trackers has been discussed in detail by several authors.^{59, 60, 61, 62, 63, 64} Based on the work of these authors, the thermal noise error of a monopulse angle tracker is given by the following equation.



63862

Figure 3-119. Monopulse RF Receive Circuitry Block Diagram



63863

Figure 3-120. Monopulse Receiving and Processing Circuitry Angle Track Block Diagram

63864

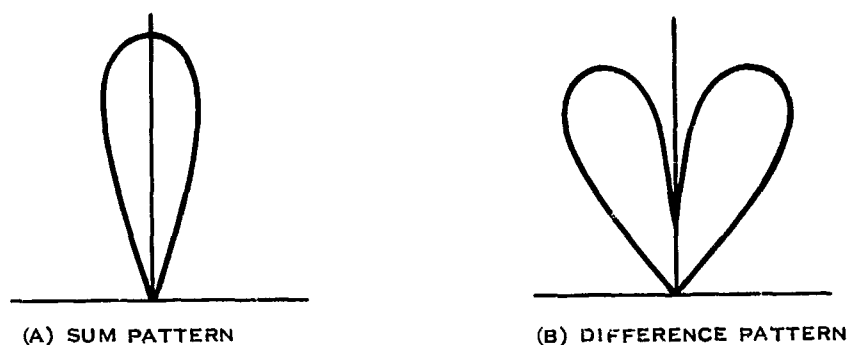


Figure 3-121. Sum-and-Difference Beam Patterns of a Monopulse Tracking System

$$\sigma_a = \frac{\theta}{k_m \sqrt{B_r \tau \left(\frac{S}{N}\right) \left(\frac{f_r}{\beta_\ell}\right)}} , \quad (3-69)$$

where

$$\left. \begin{array}{l} \sigma_a = \text{rms thermal noise error in beam position, per channel} \\ \theta = \text{half-power beamwidth} \end{array} \right\} \text{any consistent units}$$

k_m = error slope

B_r = receiver if. bandwidth (Hz)

τ = duration of transmitted pulse (sec) (or, if pulse compression is used, τ is the duration of the equivalent compressed pulse)

S/N = receiver if. signal-to-noise power ratio

f_r = pulse repetition rate (sec^{-1})

β_ℓ = nominal (high S/N) tracking loop bandwidth (rad/sec).

As was mentioned in connection with the range-tracking loops, the actual tracking loop bandwidth decreases from its nominal value for low S/N . This does not affect the thermal-noise equation because the phase-sensitive detector performance also degrades at low signal-to-noise.

Another source of random error for a monopulse tracker is target glint; an extended target may have multiple reflection points which contribute perturbations to the phase of the return rf signal. This, in turn,

perturbs the phases of the if, sum-and-difference signals from which the tracking signals are derived. An approximate expression for glint error is given by

$$\sigma_g = \frac{1}{4} \frac{L_t}{R} , \quad (3-70)$$

where

$$\left. \begin{array}{l} \sigma_g = \text{rms glint error (rad)} \\ L_t = \text{target span} \\ R = \text{range to target.} \end{array} \right\} \text{any consistent units}$$

Finally, if the monopulse angle-tracking loop is viewed as a first-order servo loop, there will be a constant lag error due to target crossing velocity with respect to the radar. The lag error is given by

$$\epsilon_a = \frac{\omega_t}{\beta_\ell} , \quad (3-71)$$

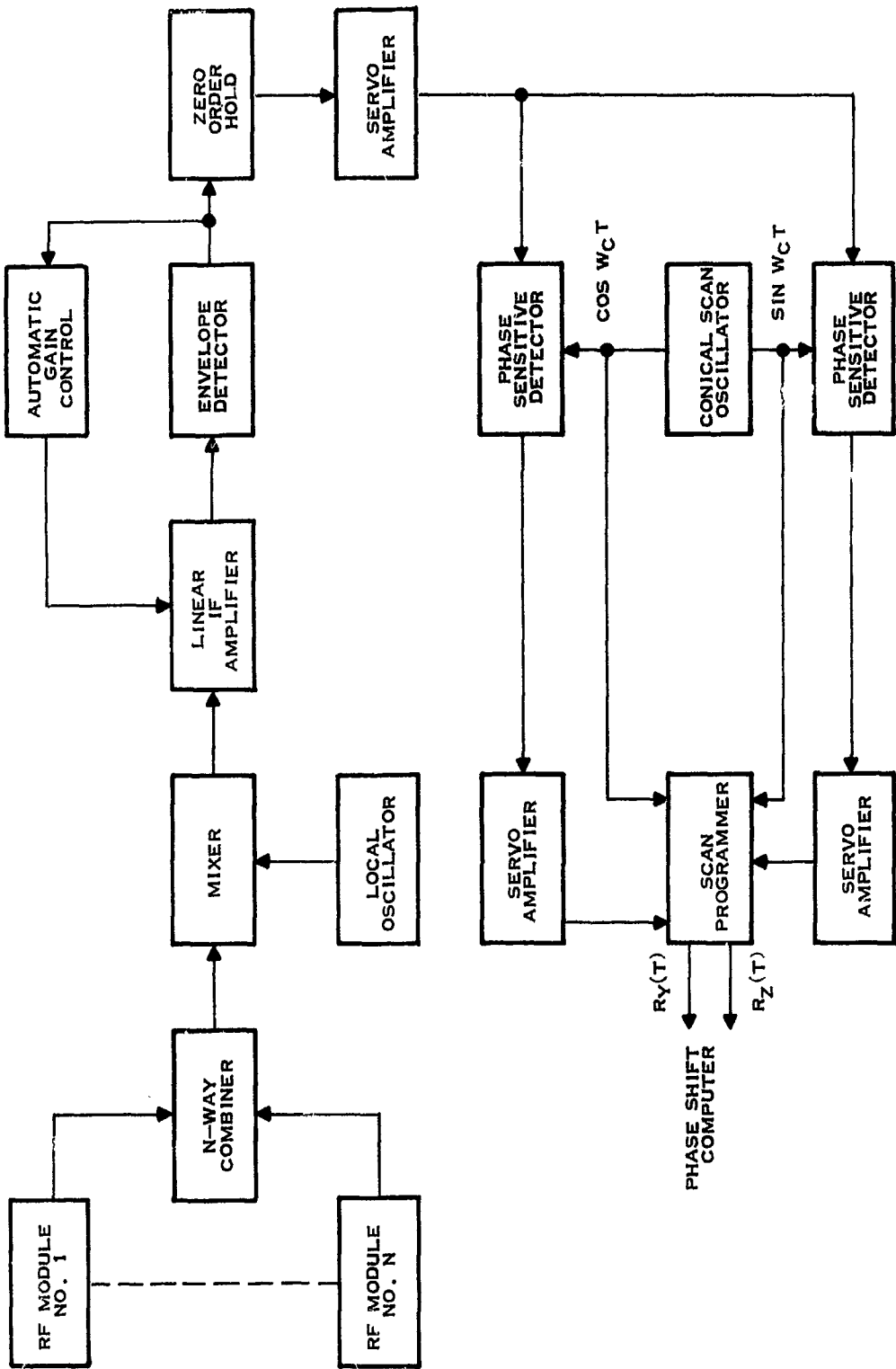
where

$$\begin{array}{l} \epsilon_a = \text{azimuth (elevation) lag error (rad)} \\ \omega_t = \text{azimuth (elevation) target angular velocity (rad/sec)} \\ \beta_\ell = \text{actual azimuth (elevation) tracking loop bandwidth (rad/sec).} \end{array}$$

3.3.2.2 Conical Scan

A typical conical-scan radar angle-tracking block diagram is illustrated by Figure 3-122. The beam is caused to rotate about the tracking axis by application of the conical-scan oscillator to the scan programmer. The relationship of the target, scanning beam, and angular-tracking error signal is shown in Figure 3-123. If the target is not on the tracking axis, the returning pulsetrain will be amplitude modulated. If the time constant of the automatic gain control (AGC) is much greater than the conical-scan period, the AGC average value of the pulsetrain will be held constant, but the modulation will not be suppressed. The pulsetrain is envelope detected, and processed by the zero-order hold and low-pass servoamplifier. Thus, the output of the servoamplifier is given by

$$\begin{aligned} e(t) &= E_o \left[1 + k_c \epsilon \cos (\omega_c t + \phi) \right] \\ &= E_o + E_o k_c \epsilon \cos \phi \cos \omega_c t - E_o k_c \epsilon \sin \phi \sin \omega_c t , \end{aligned} \quad (3-72)$$



63865

Figure 3-122. Conical Scan Radar Angle Track Block Diagram

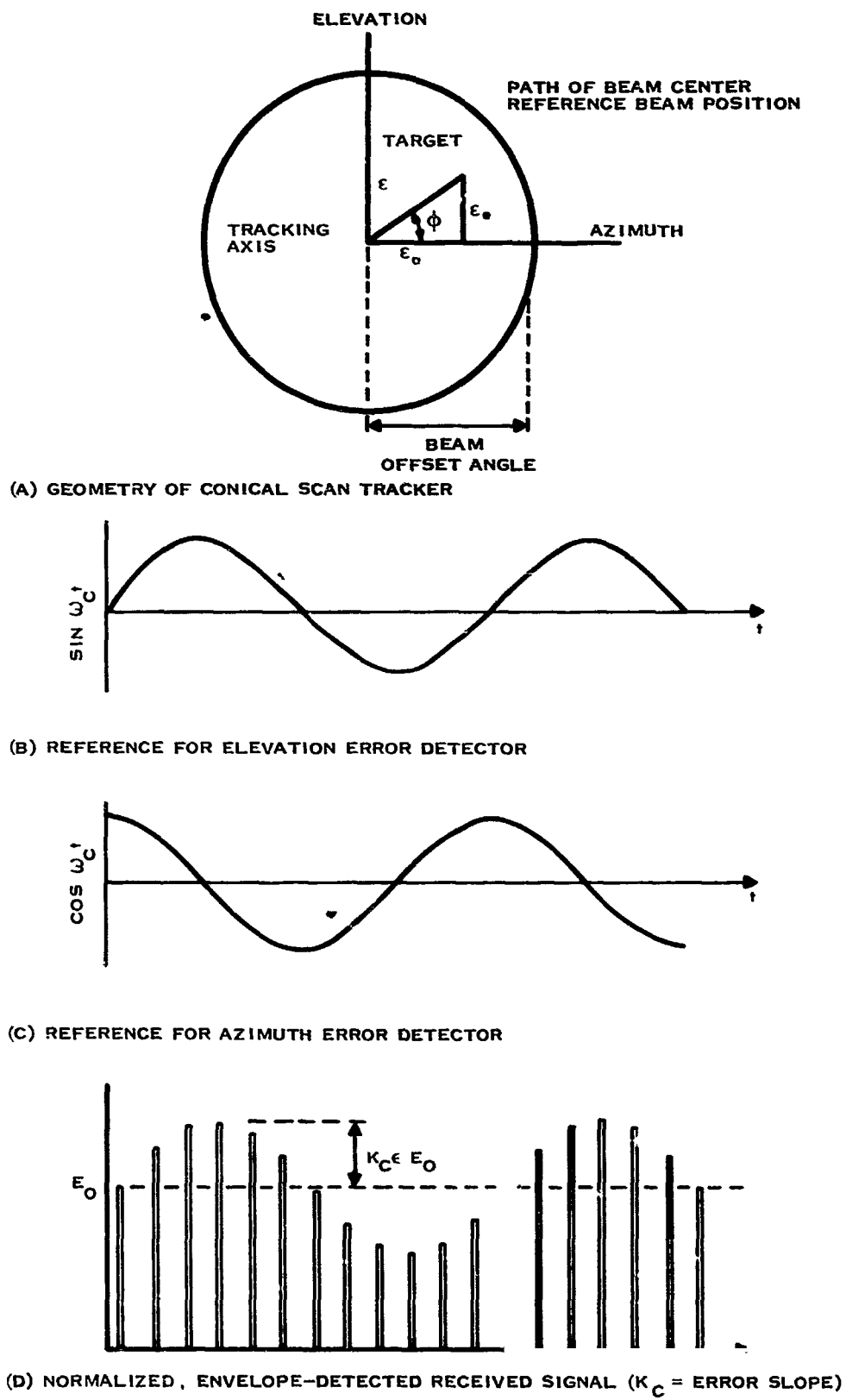


Figure 3-123. Relationships Between Target, Scanning Beam, and Received Signal for Conical-Scan Tracker

so that the output of the phase-sensitive detectors is given by

$$e_a(t) = E_o k_c \epsilon \cos \phi \quad (3-73)$$

and

$$e_e(t) = -E_o k_c \epsilon \sin \phi, \quad (3-74)$$

which, aside from constant scale factors, are the desired error signals.

The angular error slope function can be found by an investigation of the two-way voltage beam pattern as a function of beam offset angle. The two-way voltage beam pattern can be approximated by

$$E(\xi) = \frac{\sin^2 \xi x}{(x\xi)^2}. \quad (3-75)$$

Letting ξ be the angle off beam center in beamwidths, and knowing that $E(\xi) = 0.5$ when $\xi = 0.5$, the value of x is 2.78. The two-way voltage beam pattern is therefore

$$E(\xi) = \frac{\sin^2 (2.78\xi)}{(2.78\xi)^2}. \quad (3-76)$$

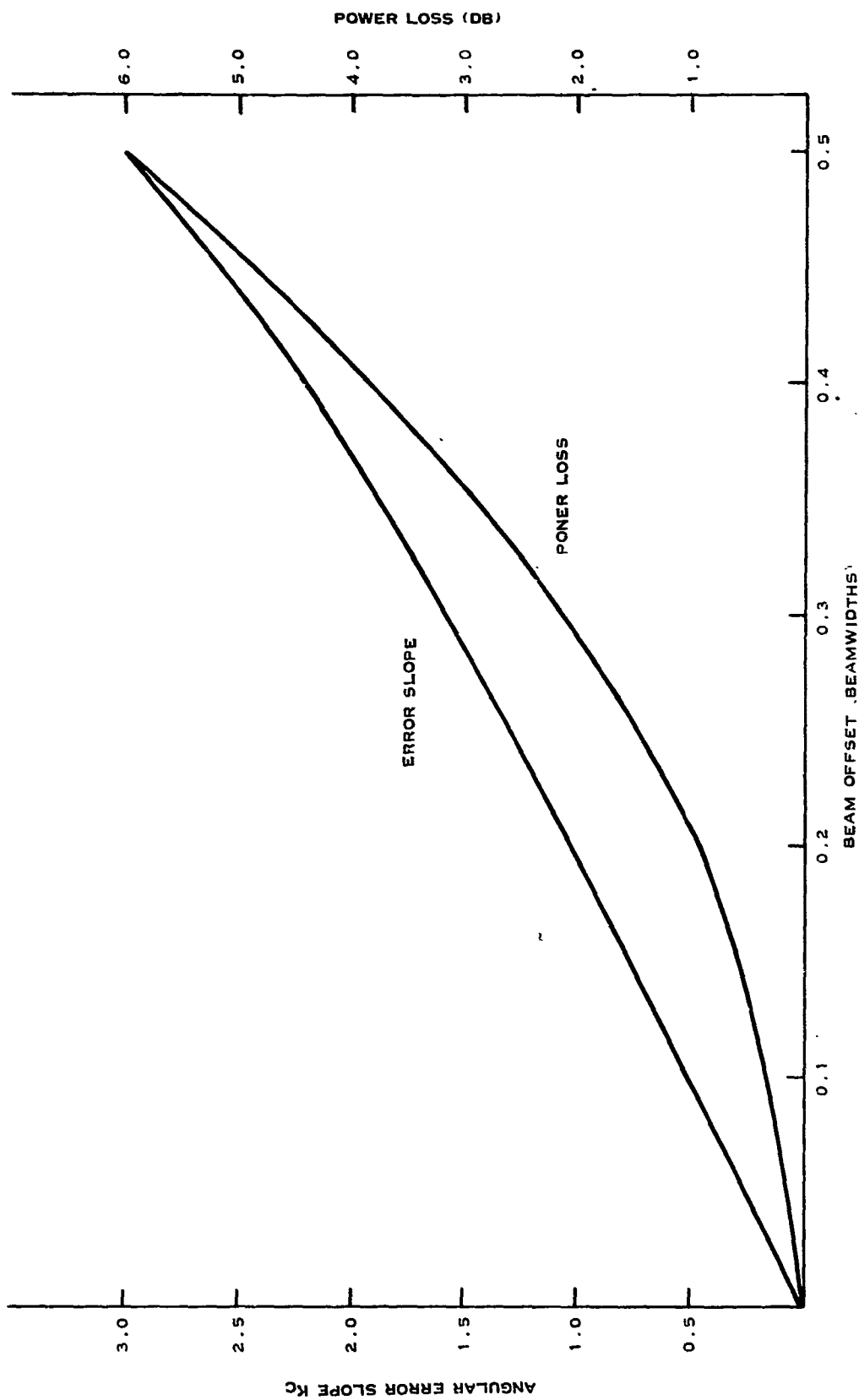
The error slope for an offset angle ξ , is defined by

$$k_c(\xi_1) = \frac{1}{E(\xi_1)} \left[\frac{dE(\xi)}{d\xi} \right]_{\xi=\xi_1}. \quad (3-77)$$

This function is evaluated for the beam pattern of Equation (3-75) and plotted in Figure 3-124. Also plotted in that figure is the loss factor which is due to off-beam-center tracking. That is, a target which is being tracked will not be on the maximum of the beam pattern; hence, the S/N will be degraded by the factor L_x from that which would be obtained for a target on the beam center.

The thermal noise error for a conical-scan angle tracker has been derived by Barton,⁶⁵ and is given by

$$\sigma_a = \frac{1.4 \theta \sqrt[4]{c_d/c_a}}{k_c \sqrt{B_r} \tau \left(\frac{S}{N} \right) \left(\frac{f_r}{\beta_l} \right) \frac{1}{L_c}}, \quad (3-78)$$



63867

Figure 3-124. Angular Error Slope and Power Loss for Conical Scan Tracker

where

$$\begin{aligned}
 \sigma_a &= \text{rms thermal-noise error in beam position, per channel} \\
 \theta &= \text{half-power beamwidth} \\
 k_c &= \text{error slope} \\
 B_r &= \text{receiver if. bandwidth} \\
 \tau &= \text{duration of transmitted pulse (sec) (or, if pulse compression is used, } \tau \text{ is the duration of the equivalent compressed pulse)} \\
 S/N &= \text{if. signal-to-noise power ratio} \\
 f_r &= \text{pulse repetition rate (sec}^{-1}\text{)} \\
 \beta_l &= \text{nominal (high S/N) tracking loop bandwidth (rad/sec)} \\
 L_c &= \text{power loss factor} \\
 c_d &= \text{detector loss factor for low S/N} \cong [2(S/N) + 1] / 2(S/N) \\
 c_a &= \text{AGC loss factor for low S/N} \cong [S/N + 1] / (S/N)
 \end{aligned}
 \left. \begin{array}{l} \sigma_a \\ \theta \end{array} \right\} \text{any consistent unit}$$

Since the error signals for a conical scan system are not derived from rf carrier phase comparisons as in the monopulse system, there is no glint problem. However, scintillation, which refers to pulse-to-pulse amplitude variations of the return due to changing target aspect angle, can cause difficulties by introducing extraneous modulation on the return pulsetrain. Barton⁶⁶ shows that the rms scintillation error is given by

$$\sigma_s = \frac{0.27\theta}{f_c} \sqrt{f_g \beta_l} \quad , \quad (3-79)$$

where

$$\begin{aligned}
 \sigma_s &= \text{rms azimuth (elevation) error due to scintillation} \\
 \theta &= \text{half-power beamwidth} \\
 f_c &= \text{conical-scan frequency (Hz)} \\
 f_g &= \text{azimuth (elevation) scintillation frequency (Hz)} \\
 \beta_l &= \text{actual azimuth (elevation) tracking loop x-band (rad/sec).}
 \end{aligned}
 \left. \begin{array}{l} \sigma_s \\ \theta \end{array} \right\} \text{any consistent unit}$$

The scintillation frequency is given by

$$f_g = \frac{2f_a L_t}{\lambda}, \quad (3-80)$$

where

$$\left. \begin{aligned} f_g &= \text{azimuth (elevation) scintillation frequency (Hz)} \\ f_a &= \text{rate of change of target aspect angle in azimuth (elevation) (Hz)} \\ L_t &= \text{target span in azimuth (elevation)} \\ \lambda &= \text{wavelength} \end{aligned} \right\} \text{any consistent units.}$$

Lag errors due to target velocity for conical scan are given by the same expression, Equation (3-71), as for monopulse.

3.3.2.3 Conopulse

Conceptually, the operation of the conopulse is quite simple. The center of the antenna beam is deflected off the tracking axis by something less than one-half beamwidth at the center of a transmitted pulse. As the pulse is being transmitted, the center of the main beam is made to rotate in a conical fashion about the tracking axis at a frequency called the conopulse frequency, so that several revolutions about the tracking axis occur before the end of the transmitted pulse occurs. At the end of the transmitted pulse, the conopulse radar reverts to the receive mode with the center of the beam returning to the tracking axis and remaining there for the entire receive portion of the cycle. The reflected signal, to be shown in the next section, contains sufficient information so that angle-error signals can be obtained for use in automatically tracking the target. The preceding brief description, together with the more detailed description of the next section, will show that the conopulse angle-tracker technique combines the desirable feature of both monopulse and conical-scan systems. Tracking information is derived from a single pulse, which was a significant advantage of monopulse over conical scan in that scintillation is eliminated. On the other hand, the error signals are not derived from the phase of the rf carrier, so that it also has the advantage which conical scan has over monopulse, elimination of glint. Finally, conopulse requires only a single if. channel, unlike monopulse which requires three if. channels. Finally, the loss factor due to off-beam-center tracking is less than for conical scan, since it depends on the one-way beam pattern rather than the two-way pattern.

3.3.2.3.1 System Description

A system which should perform the functions of a conopulse tracking radar described above is shown in block diagram form in Figure 3-125. The purpose of this section is to describe in detail how this conopulse radar system works.

FOLDOUT FRAME

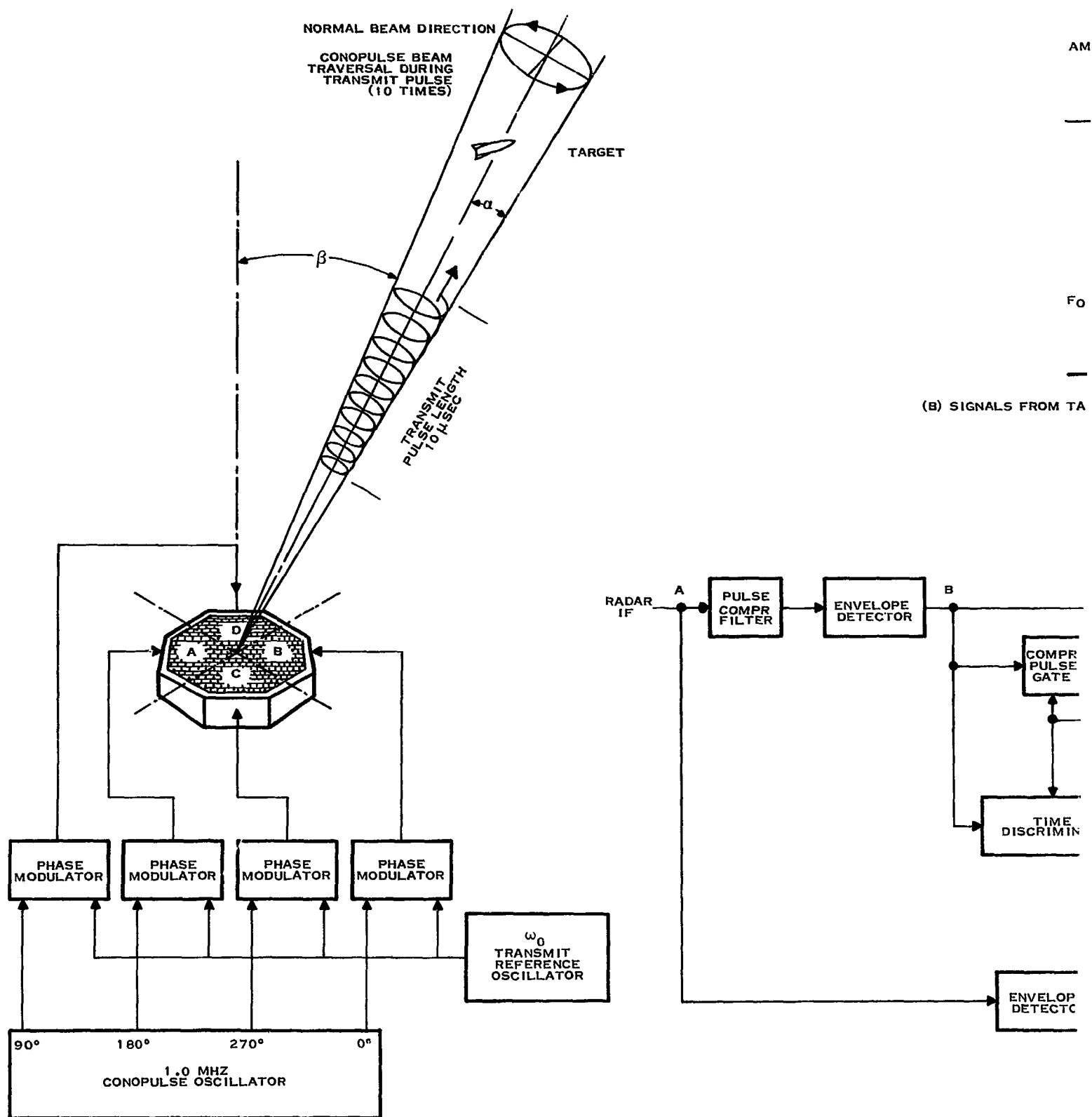
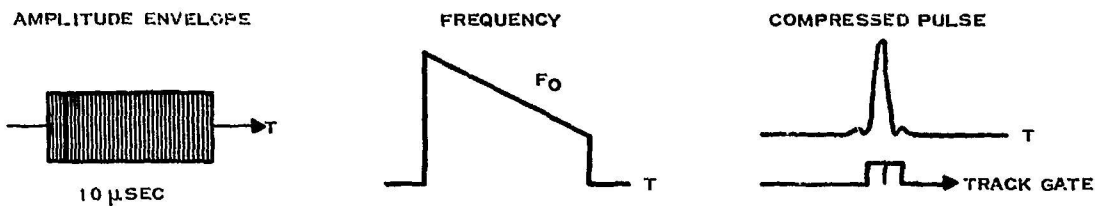
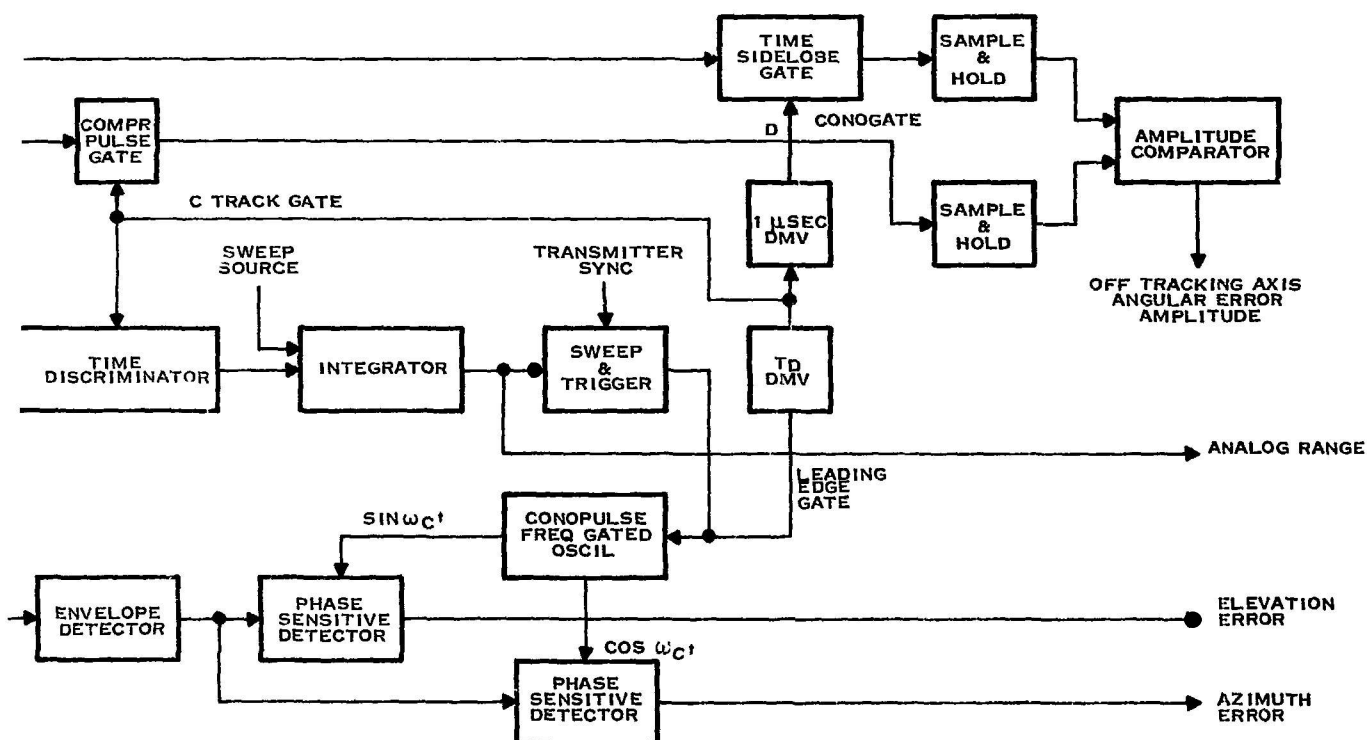
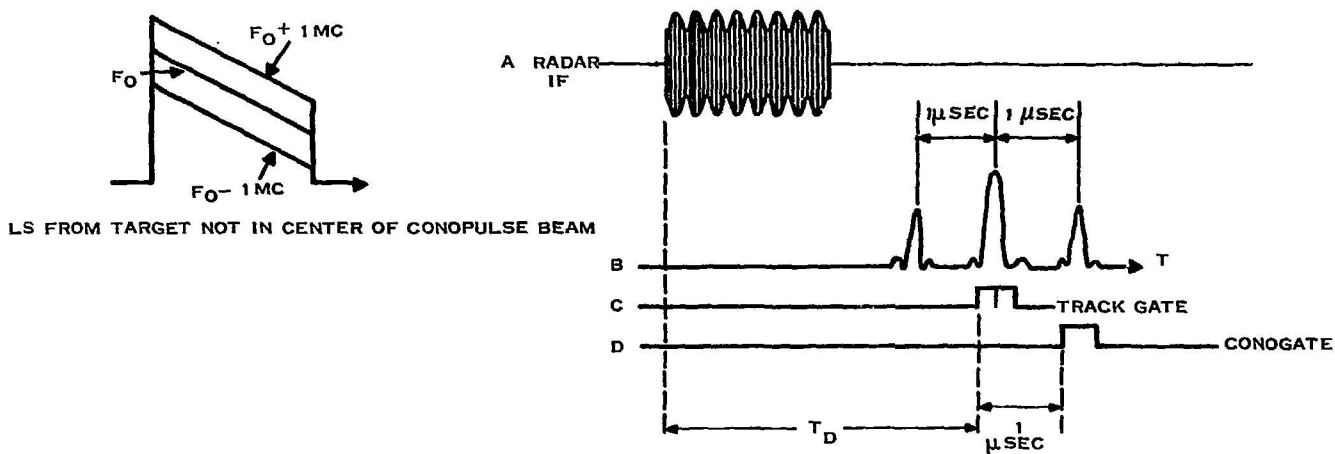


Figure 3-125. Conopulse Transmission and Reception System

FOLDOUT FRAME 2



(A) SIGNALS FROM TARGET NOT IN CENTER OF CONOPULSE BEAM



For purposes of discussion, assume that the transmitted pulse-length is 10 seconds and the conopulse frequency is 1 MHz, resulting in the transmitted beam rotating about the tracking axis ten times during one transmitted pulse-length. Neither of these parameters is inherently constrained to that value.

3.3.2.3.1.1 Transmitter

The first item in the sequence of events is the transmission of a pulse in the conical fashion described. The conical scan of the beam about the tracking axis is produced by altering the phase of the local oscillator signals entering each quadrant of the array at a sinusoidal rate called the conopulse frequency. The 1.0-MHz conopulse oscillator is capable of producing four 90-degree separated phases of the conopulse tone. Two of these components are required in the receiver to detect azimuth and elevation error angles for target tracking.

The expressions for the signals entering each array quadrant are given by

$$M_B(t) = A \cos \left[\omega_o t + \frac{\Phi}{2} \cos (\omega_p t + 0^\circ) \right] \quad (3-81)$$

$$M_D(t) = A \cos \left[\omega_o t + \frac{\Phi}{2} \cos (\omega_p t + 90^\circ) \right] \quad (3-82)$$

$$M_A(t) = A \cos \left[\omega_o t + \frac{\Phi}{2} \cos (\omega_p t + 180^\circ) \right] \quad (3-83)$$

$$M_C(t) = A \cos \left[\omega_o t + \frac{\Phi}{2} \cos (\omega_p t + 270^\circ) \right] \quad (3-84)$$

where Φ is the phase deviation of the carrier which is required to deflect the conopulse beam off the tracking axis by the desired amount. If the separation between two phase centers is L , a wave of frequency, ω_o , will be deflected away from normal by an angle, β , if one introduces an electrical phase difference, Φ , between the waves at the two phase centers where Φ is given by

$$\Phi = \frac{\omega_o}{c} L \sin \beta \quad (3-85)$$

and c is the velocity of light.

Assuming that β is the off-boresight angle by which the digital phase shifters have caused the tracking axis to be deflected, the quantity of interest to the conopulse transmitter is the phase shift which must be added in the phase modulator to deflect the beam an additional angle. This can be determined as follows.

$$\Phi(\beta) = \frac{\omega_0}{c} L \sin \beta \quad (3-86)$$

$$\Phi(\beta + \alpha) = \frac{\omega_0}{c} L \sin (\beta + \alpha) \quad (3-87)$$

$$\Delta\Phi = \Phi(\beta + \alpha) - \Phi(\beta) = \frac{\omega_0}{c} L [\sin (\beta + \alpha) - \sin \beta], \quad (3-88)$$

which reduces to

$$\Delta\Phi = \frac{\omega_0}{c} L [\sin \alpha \cos \beta + (\cos \alpha - 1) \sin \beta] . \quad (3-89)$$

Using the small angle approximations

$$\cos \alpha = 1 \quad (3-90)$$

and

$$\sin \alpha = \alpha , \quad (3-91)$$

the final result is seen to be

$$\Phi \Delta = \Delta\Phi = \frac{\omega_0}{c} L \alpha \cos \beta . \quad (3-92)$$

The importance of Equation (3-92) lies in the dependence of β upon Φ as well as α . The consequence of this dependence is that the four conopulse phase modulators must take into account the value of β as determined by the beam-steering computer in order to determine the required value of Φ which will produce the desired α , or conopulse cone half angle.

If this is not done, that is, if Φ is held constant, there will be a variation in α as the tracking axis moves away from boresight. The seriousness of this variation can be determined from Equation (3-92). Under the assumption that Φ is elected to be held constant, one sees that

$$\alpha \cos \beta = \frac{c\Phi}{\omega_0 L} , \text{ a constant} . \quad (3-93)$$

Further assume that Φ is determined to be that which produces the ideal α at boresight, or $\beta = 0$. The question of interest is the variation of α from ideal as β varies out to the extreme limit. Assuming the maximum value of β is

45 degrees, then $\cos \beta = 0.707$ and α must have increased by a factor of 1.414 or a resultant increase of 41.4 percent from boresight to maximum tracking axis deflection. Possible alternatives include:

1. Phase modulator accepts input from beam-steering computer in determining required phase shift for constant squint angle.
2. Squint angle is optimized at some tracking axis pointing angle, for example, 22.5 degrees, with subsequent deterioration in squint angle as tracking axis varies away from optimum.

It is necessary that the beginning of transmission of the pulse always occur at the same direction with respect to the tracking axis. This is called the reference direction and is the origin of the coordinate system which the receiver uses to establish error angles.

In summary, the transmitter moves the center of the beam away from the tracking axis to the reference point which is the squint angle away from the tracking axis. There transmission begins and the center of the beam rotates about the tracking axis at the conopulse frequency and describes a cone with α being the conical half angle.

3.3.2.3.2 Receiver

At the conclusion of the transmitted pulse, the conopulse phase modulation on the beam is terminated causing the center of the beam to move back so that it is coincident with the tracking axis for the receive portion of the cycle. If the target was on the tracking axis, that is, the axis of rotation of the squinted beam, there will be no amplitude modulation on the received pulse as depicted in (A) of Figure 3-125. Assuming the transmitted pulse was a linear FM as shown in (A), the output of the pulse-compression filter is the compressed pulse in (A) which is shorter than 10 μ sec by a factor equal to the pulse-compression ratio.

In the search mode, the procedure is to transmit the linear FM pulse without conopulse modulation until a target is observed in some angular sector. Upon detection of a target, the angular search is interrupted and the radar reverts to the tracking mode. While the tracking axis remains pointed in the direction where the target was detected, the range-tracking loop is allowed to lock on the compressed pulse. When the loop is locked the conopulse modulation is enabled so that automatic tracking can commence. Assuming that the target is not precisely on the tracking axis, there will be an amplitude modulation of conopulse frequency on the returned signal as shown in (B) of Figure 3-125.

In the next section, how the amplitude and phase of the amplitude modulation on the returned signal can be determined and used to derive error signals will be shown.

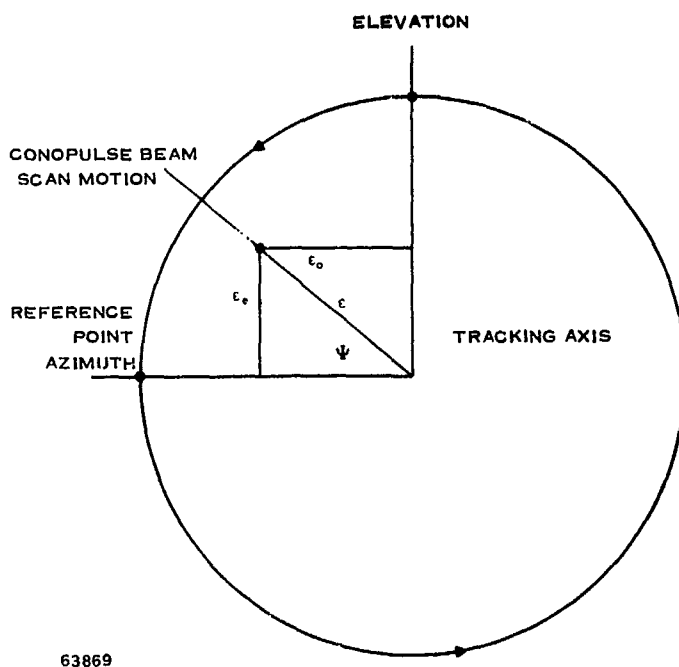


Figure 3-126. Conopulse Scan Coordinate System

3.3.2.3.3 Signal Processing

Recall that the conopulse transmission always began at the reference point which will now be defined as 0 degree where the tracking axis is the origin. Assume that a target is within the cone defined by the squint angle, α , and that the target falls on a line of angle, ψ , in the conopulse scan coordinate system shown in Figure 3-126. As the beam center moves from the reference point in a counterclockwise direction, the intensity of energy falling on the target, hence that reflected, reaches a maximum when the beam center falls on the line defined by ψ since the target is nearest the beam center at that time. As the beam center moves around 180 degrees, the intensity reaches a minimum since the target is furthest from the beam center then. As the beam makes ten revolutions about the tracking axis, the signal intercepted and reflected by the target is amplitude modulated at the conopulse frequency where the modulation is approximately sinusoidal with the phase being defined by ψ . The amplitude increases as the target moves away from the tracking axis.

For proper tracking capability one must determine the two error components, ϵ_a and ϵ_e , so that the tracking axis can be moved to coincide with the target.

From the waveform of Figure 3-125, the received signal has the waveform

$$r(t) = E \left[1 + k_p \epsilon \cos (\omega_p t + \psi) \right] \cos \omega_o t . \quad (3-94)$$

Assume, for mathematical convenience, that the envelope detector is a square-law device, then the square of the receive signal is given by

$$r^2(t) = E^2 \left[1 + 2k_p \epsilon \cos(\omega_p t + \psi) + k_p^2 \epsilon^2 \cos^2(\omega_p t + \psi) \right] \cos^2 \omega_o t. \quad (3-95)$$

The further assumption is made that the modulation is small (so that $k_p^2 \epsilon^2$ is negligible) and that higher frequency terms are filtered out, the output of the detector becomes

$$y(t) = 1/2 E^2 + k_p \epsilon E^2 \cos(\omega_p t + \psi). \quad (3-96)$$

The phase-sensitive detectors are synchronous detectors. For the azimuth channel, the product of the two input functions to the phase-sensitive detector is given by (assuming the conopulse oscillator is gated on by the leading edge of the return uncompressed pulse)

$$\begin{aligned} y(t) \times A \cos \omega_p t &= 1/2 A E^2 \cos \omega_p t + 1/2 k_p \epsilon A E^2 \cos(2\omega_p t + \psi) \\ &+ 1/2 k_p \epsilon A E^2 \cos \psi. \end{aligned} \quad (3-97)$$

Again, assuming suitable filtering, the output of the azimuth phase-sensitive detector is given by

$$z(t) = E^2 \epsilon \cos \psi,$$

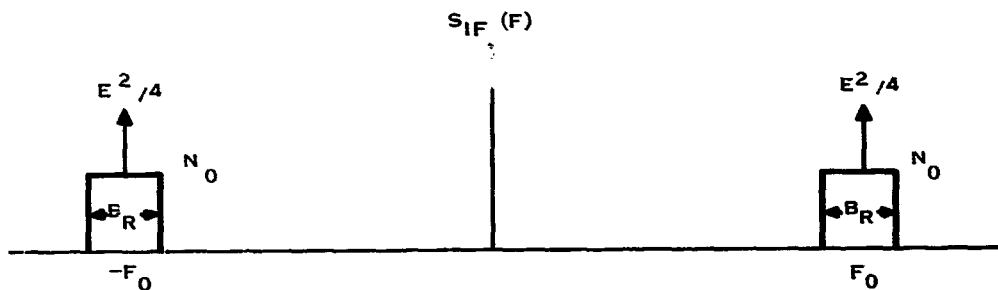
where, for convenience, we have set $A = 2/k_p$. The effect of automatic gain control is to divide all signal amplitude by twice the if. power with no modulation.⁶⁷ Therefore, with AGC, the error signal from the phase-sensitive detector is given by

$$z(t) = \frac{E^2}{2(S + N)} \epsilon \cos \psi. \quad (3-98)$$

Since $E^2 = 2S$, Equation (3-98) becomes

$$z(t) = \frac{S/N}{1 + S/N} \epsilon \cos \psi. \quad (3-99)$$

A similar expression could be developed for the elevation channel.



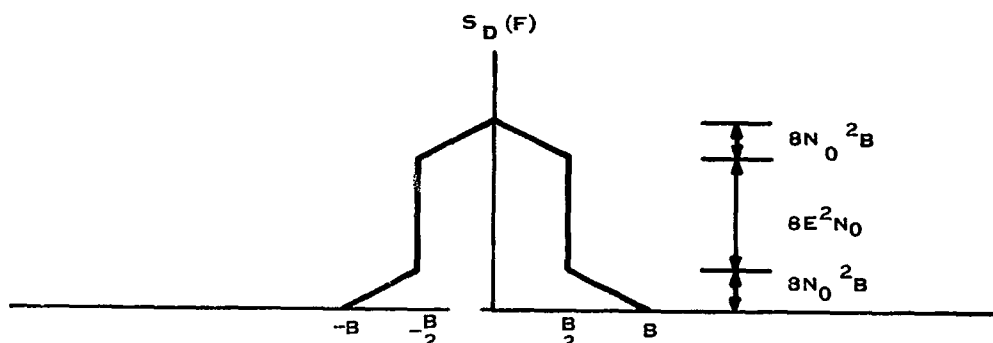
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Figure 3-127. Power Spectral Density of If. Signal (No Modulation)

The noise response of the system will now be determined. It is convenient to assume the modulation to be zero; for this condition, the power spectral density of the if. signal is shown in Figure 3-127, and the power spectral density of the output of the square law detector, assuming suitable filtering, is shown in Figure 3-128. For high signal-to-noise ratio, the principal contribution to the detected power spectral density is due to beating of the noise with the carrier, hence the detected power spectral density is approximately flat and confined to the bandwidth B .⁶⁸ An effective detected power spectral density may then be defined which is flat over the bandwidth and which has the same power content as the spectrum of Figure 3-128. The power content of the spectrum of Figure 3-128 is given by

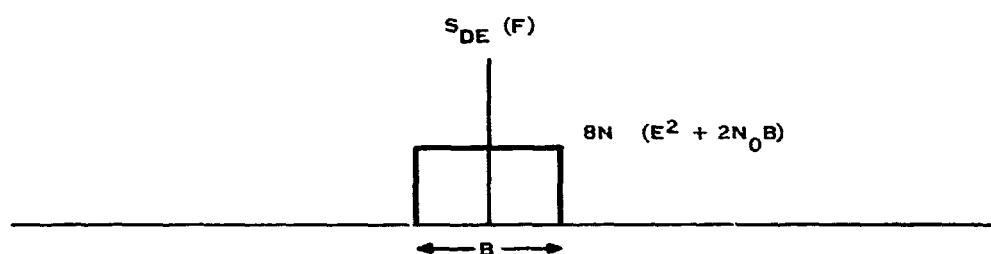
$$P_d = 8N_0 B(E^2 + 2N_0 B_r) , \quad (3-100)$$

so that the effective power spectral density of the detected signal (for high signal-to-noise) is as shown in Figure 3-129. The output of the conopulse oscillator, $2/k_p \cos \omega_{pt}$, has the power spectral density shown in Figure 3-130; hence, the power spectral density of the output of the phase-sensitive detector is given by the convolution of the spectra of Figure 3-129 and is shown in Figure 3-131. As indicated in Figures 3-129 and 3-130, the



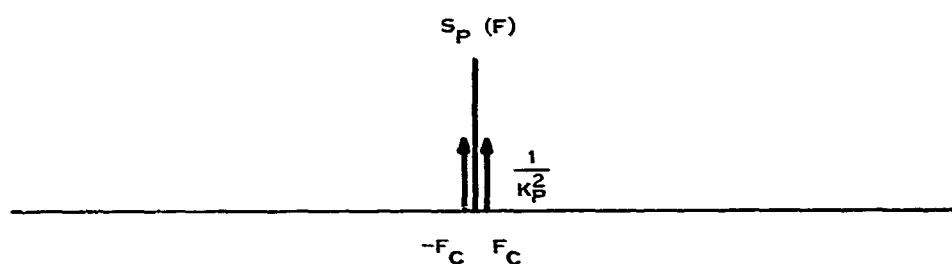
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Figure 3-128. Power Spectral Density of Detected If. Signal



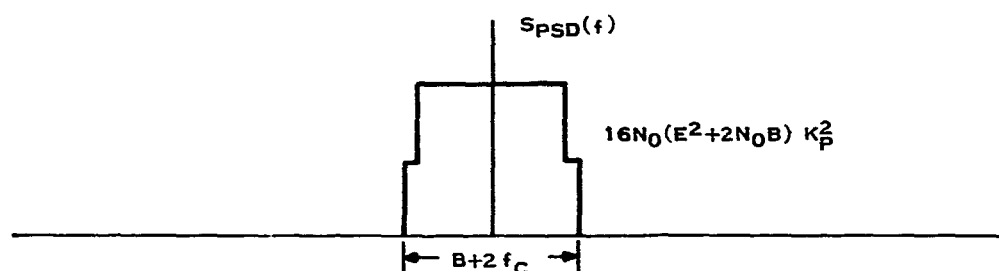
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Figure 3-129. Effective Power Spectral Density of Detected If. Signal



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Figure 3-130. Power Spectral Density of Output of Conopulse Oscillator



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Figure 3-131. Power Spectral Density of Output of Phase-Sensitive Detector

conopulse frequency is considerably less than the if. bandwidth. If we ignore the skirts of the spectrum of Figure 3-131, the output noise power from the phase-sensitive detector is given by (assuming noise power is divided equally between the two channels)

$$P_o = 8N_o B \frac{(E^2 + 2N_o B_r)}{k_p^2} = \frac{16}{2} N_o^2 B_r^2 \left(1 + 2 \frac{S}{N}\right) = \frac{4}{2} N^2 \left(1 + 2 \frac{S}{N}\right). \quad (3-101)$$

The effect of AGC, which has been ignored, may now be considered by dividing the power expression of Equation (3-101) by the square of twice the if. power. Thus with AGC, the noise power from the phase-sensitive detector is given by

$$P_o = \frac{4N^2 \left(1 + 2 \frac{S}{N}\right)}{k_p^2 4(S + N)^2} = \frac{\left(1 + 2 \frac{S}{N}\right)}{k_p^2 \left(1 + \frac{S}{N}\right)^2} , \quad (3-102)$$

and the rms noise voltage is given by

$$E_{no} = \frac{\left(1 + 2 \frac{S}{N}\right)^{1/2}}{k_p \left(1 + \frac{S}{N}\right)} . \quad (3-103)$$

The signal required to balance this noise voltage may now be found, hence the rms thermal-noise error, by equating Equation (3-103) to Equation (3-99). Thus

$$\frac{\frac{S}{N}}{1 + \frac{S}{N}} \sigma_\theta = \frac{\left(1 + 2 \frac{S}{N}\right)^{1/2}}{k_p \left(1 + \frac{S}{N}\right)} , \quad (3-104)$$

so that

$$\sigma_\theta = \frac{\left(1 + 2 \frac{S}{N}\right)^{1/2}}{k_p \frac{S}{N}} . \quad (3-105)$$

For high signal-to-noise ratio, Equation (3-105) reduces to

$$\sigma_\theta = \frac{\sqrt{2}}{k_p \sqrt{\frac{S}{N}}} . \quad (3-106)$$

If, as is usual, k_p is in units percent modulation per fractional beamwidth, then Equation (3-106) becomes

$$\sigma_{\theta} = \frac{\sqrt{2\theta}}{k_P \sqrt{\frac{S}{N}}} \quad , \quad (3-107)$$

where θ is the half-power beamwidth.

Now the noise bandwidth of the receiver is B_r , so that noise samples separated in time by $1/B_r$ are independent, and in one-pulse duration τ , there are $B_r \tau$ such samples. Therefore, the noise error can be reduced by the factor $\sqrt{B_r \tau}$ by following the phase sensitive by a filter matched to the pulse duration. Finally, the response time of the closed-loop system will generally be several interpulse periods in duration. If the bandwidth of the closed-loop system is β_{ℓ} , then the response will be based on an average of f_r/β_{ℓ} pulses, where f_r is the pulse-repetition rate. Equivalently, a digital phase-shift computer averages the error signal over $n = f_r/\beta_{\ell}$ pulses before updating the beam position. In any case, noise samples separated by the interpulse period are independent, so that a further improvement in noise, given by $1/\sqrt{f_r/\beta_{\ell}}$, is obtained. Thus Equation (3-107) can be modified by these two factors to give

$$\sigma_{\theta} = \frac{\sqrt{2\theta}}{k_P \sqrt{\left(\frac{S}{N}\right) \left(\frac{f_r}{\beta_{\ell}}\right) B_r \tau}} \quad . \quad (3-108)$$

The error expression of Equation (3-108) was derived for large signal-to-noise ratios. For low signal-to-noise ratios, two factors must be taken into account. First, there is a small signal suppression effect in the first detector,⁶⁹ so that for small signal-to-noise ratios, the signal-to-noise ratio out of the detector varies as twice the square of the input (if.) signal-to-noise ratio rather than directly as the input signal-to-noise ratio. Thus, the effective signal-to-noise ratio is given by

$$\left(\frac{S}{N}\right)_{\text{eff}} = \frac{1}{c_d} \frac{S}{N} \quad , \quad (3-109)$$

where $c_d = (2S + N)/2S$. Secondly, as shown by Equation (3-99), loop gain is reduced by a factor $S/(S + N)$. Since the closed-loop bandwidth is directly proportional to loop gain,

$$B_{\ell \text{ eff}} = \frac{1}{c_a} \beta_{\ell} \quad , \quad (3-110)$$

where $c_a = (S + N)/S$. Combining these effects we see that Equation (3-108) becomes

$$\sigma_\theta = \frac{\sqrt{2}\theta}{k_p \sqrt{\frac{1}{c_d} \left(\frac{S}{N}\right) c_a \left(\frac{f_r}{\beta_\ell}\right) B_r \tau}} = \frac{\sqrt{2}\theta \sqrt{c_d/c_a}}{k_p \sqrt{\left(\frac{S}{N}\right) \left(\frac{f_r}{\beta_\ell}\right) B_r \tau}}, \quad (3-111)$$

which holds for any value of signal-to-noise ratio, provided it is remembered that β_ℓ is not the actual closed-loop bandwidth, but rather the bandwidth measured at high signal-to-noise ratios. One final modification to the noise expression can be made. Conventionally, signal-to-noise ratio, as determined from the range equation, is that due to a target which is on the maximum of the antenna beam. The signal-to-noise ratio in Equation (3-111) is that for a target on the tracking axis, which does not coincide with the maximum of the antenna beam. We may account for this difference by introducing a loss factor L_p and using the conventional signal-to-noise ratio. Thus, the final expression for rms thermal-noise error for the conopulse range-tracking system is given by

$$\sigma_\theta = \frac{\sqrt{2}\theta \sqrt{c_d/c_a}}{k_p \sqrt{\frac{1}{L_p} \left(\frac{S}{N}\right) \left(\frac{f_r}{\beta_\ell}\right) B_r \tau}}, \quad (3-112)$$

where

- | | |
|---|-----------------------|
| σ_θ = rms thermal-noise error in angle
per channel | } any consistent unit |
| θ = antenna beamwidth | |
| k_p = error slope | |
| L_p = loss factor due to off-beam-center tracking | |
| S/N = signal-to-noise power ratio for an on-beam-center target | |
| f_r = pulse-repetition rate (sec^{-1}) | |
| β_ℓ = nominal (high signal-to-noise) closed-loop bandwidth | |
| B_r = if. bandwidth (Hz) | |
| τ = duration of received uncompressed pulse (sec). | |

The error slope, k_p , and off-beam-center tracking loss, L_p , for the conopulse system, which appear in Equation (3-112), are easily determined. They are defined in the same way as for a conical scan system. However, since the beam is switched back to the tracking axis during the receive portion of the cycle, they are functions of the one-way beam pattern rather than the two-way pattern. For an assumed $\sin x/x$ beam pattern, these two factors are plotted in Figure 3-132.

As mentioned previously, the conopulse system does not suffer a degradation in performance due to either scintillation or glint. However, it does have an additional possibility for error which is not found significant in either the monopulse or conical scan systems, namely crosstalk. For simplicity, let us assume the output of the lower envelop detector of Figure 3-125 to be given by

$$v_1(t) = 2k_p\epsilon \cos(\omega_p t + \psi), \quad (3-113)$$

where t is measured from the time of occurrence of the leading edge of the return pulse. Then, if the conopulse oscillator is synchronized perfectly to the leading edge of the return pulse, the filtered outputs of the synchronous detectors will be given by

$$e_a(t) = k_p\epsilon \cos \psi \quad (3-114)$$

and

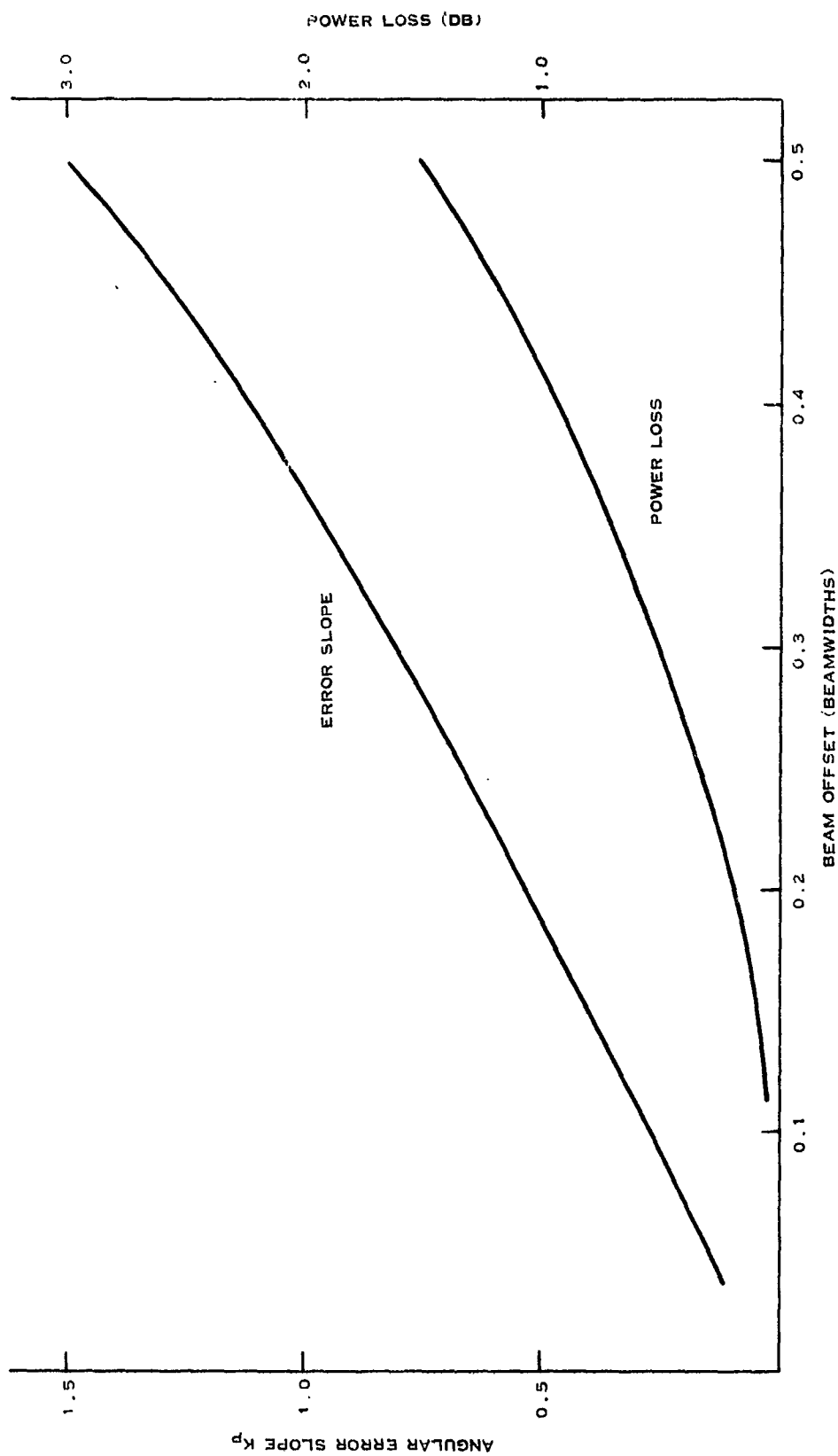
$$e_e(t) = k_p\epsilon \sin \psi, \quad (3-115)$$

which are the desired angular error signals for the azimuth and elevation channels. In the actual system, however, the conopulse oscillator is synchronized by a tracking gate from the range-tracking loop. The range loop does not, in general, track perfectly due to target range rate and/or thermal noise. In fact, the position of a gate pulse in the range tracking, as shown in Subsection 3.3, is a random variable with mean ϵ_t and standard deviation σ_t . ϵ_t is the lag error due to target range rate and σ_t is the rms thermal-noise error. Define t_e to be the error in the tracking gate position, where $t_e = \epsilon_t$ and $t_e^2 = \sigma_t^2$. The effect of this error will be to turn on the conopulse oscillator at the wrong time, so that the outputs of the conopulse oscillator will be

$$r_a(t) = \cos \omega_p(t + t_e) = \cos \omega_p t \cos \omega_c t_e - \sin \omega_p t \sin \omega_c t_e \quad (3-116)$$

and

$$r_e(t) = -\sin \omega_p(t + t_e) = -\sin \omega_p t \cos \omega_c t_e - \cos \omega_p t \sin \omega_c t_e \quad (3-117)$$



63875

Figure 3-132. Angular Error Slope and Power Loss for Conopulse Tracking System

and the outputs of the phase-sensitive detectors will be given by

$$\dot{v}_a(t) = k_p \epsilon \cos \omega_p t_e \cos \psi + k_p \epsilon \sin \omega_p t_e \sin \psi \quad (3-118)$$

and

$$v_e(t) = k_p \epsilon \cos \omega_p t_e \sin \psi - k_p \epsilon \sin \omega_p t_e \cos \psi . \quad (3-119)$$

Assuming $\omega_p t_e$ to be small enough that the small angle approximation applies, Equations (3-118) and (3-119) reduce to

$$v_a(t) = e_a(t) + \omega_p t_e e_e(t) \quad (3-120)$$

and

$$v_e(t) = e_e(t) - \omega_p t_e e_a(t) , \quad (3-121)$$

where $e_a(t)$ and $e_e(t)$ are the desired azimuth and elevation angular error signals. The errors due to crosstalk in the azimuth and elevation angular error signals are then given by

$$\delta_a = e_a - v_a = -\omega_p t_e e_e \quad (3-122)$$

and

$$\delta_e = e_e - v_e = \omega_p t_e e_a . \quad (3-123)$$

Now the signals in Equations (3-122) and (3-123) are all directly related to azimuth and elevation angles. Thus, if γ_a is the azimuth angle by which the target is off the tracking axis, ν_{γ_a} is the error in γ_a , due to crosstalk, and similar quantities are defined for the elevation dimension, then

$$\nu_{\gamma_a} = -\omega_p t_e \gamma_e \quad (3-124)$$

and

$$\nu_{\gamma_e} = \omega_p t_e \gamma_a . \quad (3-125)$$

Now γ_a and γ_e , which are the angular-tracking errors in the absence of crosstalk are, in general, random variables with $\bar{\gamma}_a = \epsilon_a$, where ϵ_a is the lag error of the azimuth tracking loop due to a target with azimuthal angular velocity, and $\overline{\gamma_a^2} = \sigma_a^2$, where σ_a is the rms thermal-noise error defined by Equation (3-111). Similar statements hold for the elevation dimension. If we assume the random variable t_e to be independent of the random variables γ_a and γ_e , then

$$\epsilon_{act} = \overline{\nu \gamma_a} = -\omega_p \overline{t_e \gamma_e} = -\omega_p \epsilon_t \epsilon_e, \quad (3-126)$$

$$\epsilon_{ect} = \overline{\nu \gamma_e} = +\omega_p \overline{t_e \gamma_a} = \omega_p \epsilon_t \epsilon_a, \quad (3-127)$$

$$\sigma_{act} = \left(\overline{\nu \gamma_a^2} \right)^{1/2} = \omega_p \left(\overline{t_e^2 \gamma_e^2} \right)^{1/2} = \omega_p \sigma_t \sigma_e, \quad (3-128)$$

$$\sigma_{ect} = \left(\overline{\nu \gamma_e^2} \right)^{1/2} = \omega_p \left(\overline{t_e^2 \gamma_a^2} \right)^{1/2} = \omega_p \sigma_t \sigma_a, \quad (3-129)$$

where

ϵ_{act} = lag error in azimuth dimension due to crosstalk (rad)

ϵ_t = lag error in range-tracking loop (sec)

ϵ_e = lag error in elevation dimension due to target angular velocity (rad)

σ_{act} = rms jitter in azimuth dimension due to crosstalk (rad)

σ_t = rms jitter in range-tracking loop (sec)

σ_e = rms jitter in elevation dimension due to thermal noise.

ω_p = conopulse scan frequency (rad/sec)

The lag errors due to target angular velocity are given by the azimuth (elevation) component of target angular velocity divided by the actual bandwidth of the azimuth tracking loop.

3.4 Data Processing Analysis

The previous three subsections of Section 3 have considered those aspects of a radar system which are involved in transmitting a pulse in some direction and detecting the presence of a reflected pulse in the event a target was present within the beamwidth. The purpose of this section is to consider the problem of determining the desired direction of pulse emission and how this information is relayed to the phased-array elements. Also, consideration is given to the need for multiple operational modes and multiple target capability.

The data processing area differs materially from the previous three areas in Section 3 in that there are fewer examples of inherent tradeoff relationships where the increased precision in measuring a quantity necessarily costs the precision with which another task is performed. Increased precision is obtained at the expense of increased equipment complexity only. For that reason the analysis in this subsection is concerned with developing the minimum complexity data processor which adequately performs the tasks required of it.

3.4.1 Multiple Mode, Multiple Target Capability

3.4.1.1 Multiple Mode Capability

The requirements of the SOW and the results of discussions indicate that two range modes should be considered; these include (1) far range and (2) near range. Also, two tracking modes should be considered: (1) continuous tracking after acquisition and (2) multiple target combined acquisition and tracking (track-while-scan).

The SOW specifically calls for a range between 5 and 200 nm and that multiple target combined acquisition and tracking should be investigated. To accomplish this, pulse compression may be required. A rough block diagram to accomplish this is drawn in the solid lines on Figure 3-133. The rf portion is similar to all phased-array mechanizations. The prime difference between a tracking and a track-while-scan mechanization is the need for an instantaneous automatic gain control (IAGC), such as a logarithmic intermediate frequency amplifier (LOG. IF). Note that for a linear constant system gain, the output signal level change for a dynamic range change from 5 to 200 nm is $(200/5)^4 = 2.56 \times 10^4$; but for a log amplifier the dynamic change is $k \log 2.56 \times 10^4 = k(4.48)$ where k can be adjusted to accommodate the possible output signal dynamic range.

There is also a loss in system parametric accuracy in track-while-scan. A tracking system can reduce the total noise tracking bandwidth, resulting in more accurate measurements. A track-while-scan system must sample each return pulse and if no one pulse signifies zero error, it must predict where the target should be and its amplitude based upon the error values. Allowing the scan to dwell on each target until zero error is indicated can overcome this fault but results in a reduction of scan volume.

The broken line boxes in the block diagram indicate the functions that would be changed if ranges less than 5 nm are obtained. These include: (1) inhibiting the chirp and reducing the transmit pulse proportional to the desired minimum range (probably 500 ft $\pm$ 1 μ sec) and (2) reducing the if. bandwidth for matched-filter operation. The IAGC would still be required for track-while-scan since $(5 \text{ nm}/0.12 \text{ nm})^4 = 3.0 \times 10^4$ is the total dynamic signal changes. Also, the measurement accuracy of the cross section is severely reduced by using log amplifiers.

To measure the cross section accurately, a predictable, preferably linear, automatic gain control is required. This has never been mechanized at the intermediate frequencies. It can be accomplished at video and audio frequencies. A ready reference to calibrate the system to account for gain changes.

The block diagram in Figure 3-133 is valid in the event that monopulse is the scheme selected for angle tracking. In the event that another angle-tracking scheme was selected involving only a single channel, the block diagram would have the difference channels removed.

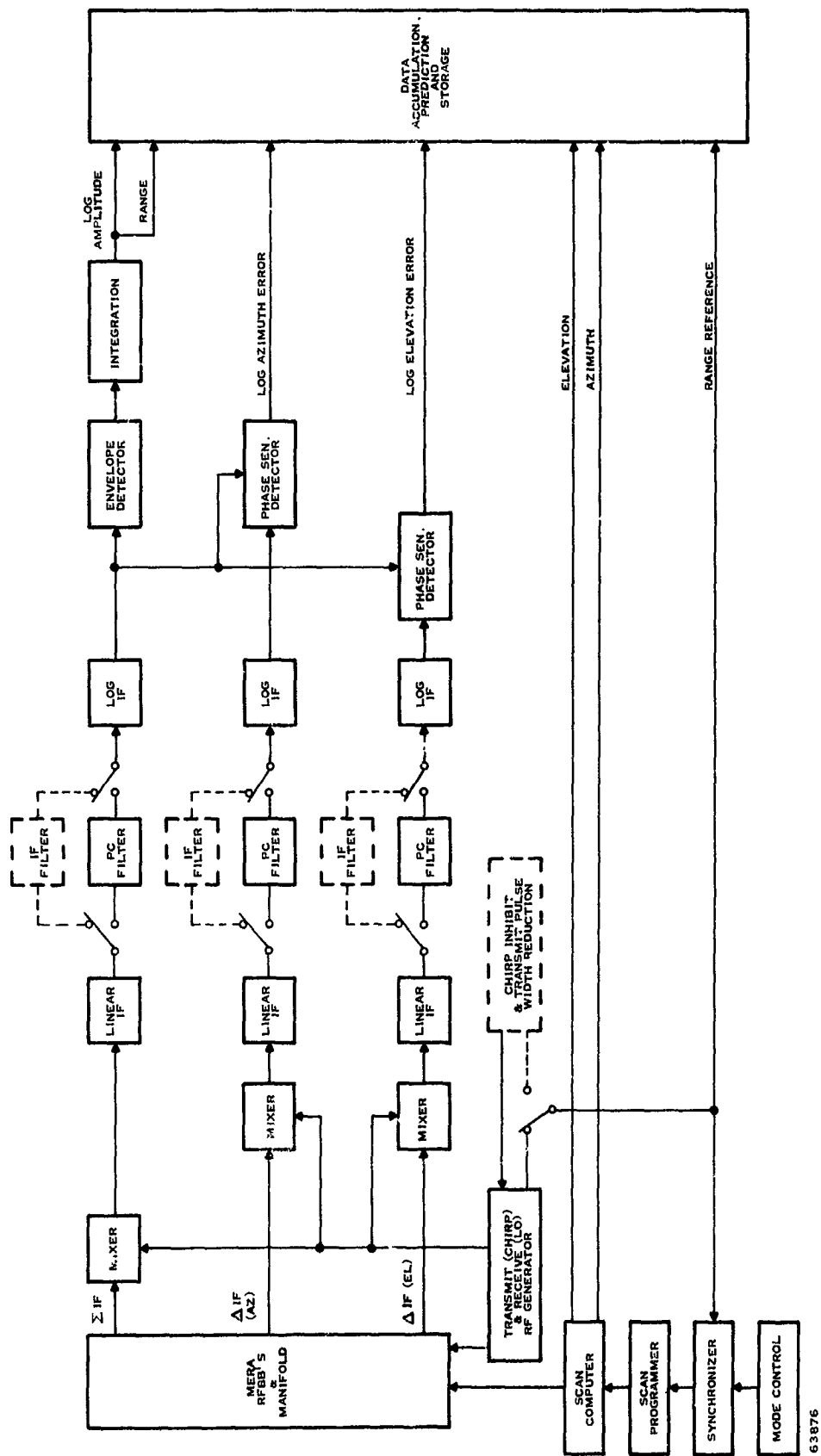


Figure 3-133. Block Diagram for Near and Far Range Modes

3.4.1.2 Multiple Target Capability

The requirements for multitarget capability will be discussed for three cases. The first case will be when known targets are ejected from the S/C that houses the CAT radar; the CAT radar must detect these targets and determine their characteristics for comparison to a predetermined baseline measurement. The second case will be to detect and characterize targets of opportunity. A target of opportunity would be any piece of space traffic that falls in the field of view of the CAT radar and registers as a target. Case 2 could be considered as a follow-on after the completion of Case 1. Case 3 will be when the CAT radar is to detect and locate a specific target; that is, a rescue or reconnaissance mission which cannot be aided by the target satellite.

Case 1: Cross-Section Experiments

The basic experiment would be to determine the change in radar cross section for earth-based and then spaceborne targets. The parameters of interest would be radar cross section versus frequency and polarization (transmit and receive). It is very important that the CAT radar operate in the same manner in the space environment as was used in determining the baseline measurements during ground-based tests. That is, the number of variables should be reduced to a minimum. The best possible accuracy would be obtained when the CAT radar continuously tracked the target. This is especially true for targets that vary in radar cross section as a function of aspect angle.

If multiple targets were ejected from the S/C, the CAT radar must carefully separate and store the characteristics of each target, no confusion between targets can be permitted to exist. This complicates the hardware and reduces the look time at each target by that time which is lost in moving the beam from target to target. The optimum operational mode would be to eject, acquire, and track a target until all measurements have been made and repeat this same mode for each target. This will require that targets be ejected in order of increasing cross sections.

Case 2: Targets of Opportunity

This case will be considered as a follow-on to the space experiments. After completion of the experiments the CAT radar will be commanded to search for all targets; the purpose could be to aid in space traffic surveillance or to aid in the location of a particular target of interest. Thus, it is of interest to determine the number of possible targets that could be detected by the CAT radar.

Information in the article "Space Traffic Surveillance" by Paul G. Thomas, Space/Aeronautics, November 1967, indicates that in June 1966 there were 984 articles of space traffic with apogees and perigees ranging from 75.6 to 1620 nm. Forecasts for the '69 to '71 period indicate

an increase to approximately 2000 pieces of space traffic. This number (2000) will be used to investigate the need for multitarget capability.

The total volume occupied by this space traffic is given by $4\pi/3[(r_e + 1620)^3 - (r_e + 75.6)^3]$, where r_e is the radius of the earth. This gives a total volume of $4.18[(3444 + 1620)^3 - (3444 + 75.6)^3] = 3.5 \times 10^{11} \text{ nm}^3$. The total volume enclosed by one radar beam position is given by $(\pi/12)R^3\theta^2$, where R is the maximum range capability and θ is the beamwidth in radians. Thus, if $R = 200 \text{ nm}$ and $\theta = 2^\circ$ (35×10^{-3} radians) then $\pi/12(2 \times 10^2)^3(3.5 \times 10^{-2})^2 = 2.5 \times 10^3 \text{ nm}^3$. For a uniform density, the probability of finding a target for each beam position is the ratio of these quantities or $2.5 \times 10^3 / 3.5 \times 10^{11} = 7.14 \times 10^{-9}$. Another way of stating this is that there will be one target per 1.4×10^8 beam positions. This does not justify the need for multitarget acquisition and tracking capability. The fault may be in assuming a uniform density, when in fact, high peak densities may be the case.

Case 3: Rescue and Reconnaissance

In this case, a priori knowledge is a must. The S/C must be guided to the vicinity of the intended target by ground direction. That this could be done very accurately was proven on the Gemini missions. Here again, continuous and accurate data is required to perform the final maneuvers to place the two or more S/C on an intercept course. This is most easily accomplished by continuous tracking of the intended target.

No requirement exists for simultaneously tracking more than one target by the CAT radar. Consequently, the most severe case would occur in a track-while-scan situation which had the first target being tracked and a second target detected during the scan cycle. Some criteria is necessary for instructing the CAT radar on which target is to be tracked. Much thought could be devoted to an elaborate criteria. However, that thought is not considered within the scope of this study and it will be simply assumed that the CAT data processor is suitably informed as to which target to track. The CAT radar is only responsible for detecting the presence of other targets and not tracking them.

3.4.2 Beam Steering and Stabilization Capability

The beam-steering computer must calculate the electronic phase delay required for beam steering, the stabilization for each of the rf modules in the array, and generate and distribute appropriate signals to each module for phase control. This must be accomplished with sufficient accuracy to give good array performance and with sufficient speed to provide adequate scanning rates and beam stabilization.

The functional operation of the beam-steering computer naturally divides into two parts:

Computation of the incremental phase shift between each rf module for each beam position. This is the sum of two terms—one dependent upon the row location and one dependent upon the column spacing. This functional circuitry will be called the scan programmer.

Generation of the appropriate digital signals for the correct phase delay of each rf module and parallel distribution of these signals to be modules within the required time interval. This functional circuitry will be called the phase-shift computer.

The phase-shift equations for beam steering will now be derived first; then, the scan programmer and the phase-shift computer will be discussed individually.

3.4.2.1 Phase-Shift Equations for All-Electronic Stabilization and Scanning of Beam Axis

The phase delay which must be added to the rf signals of an rf module for scanning the beam of a phased array is given by

$$\frac{2\pi}{\lambda} (\underline{\rho} \cdot \underline{r}) ,$$

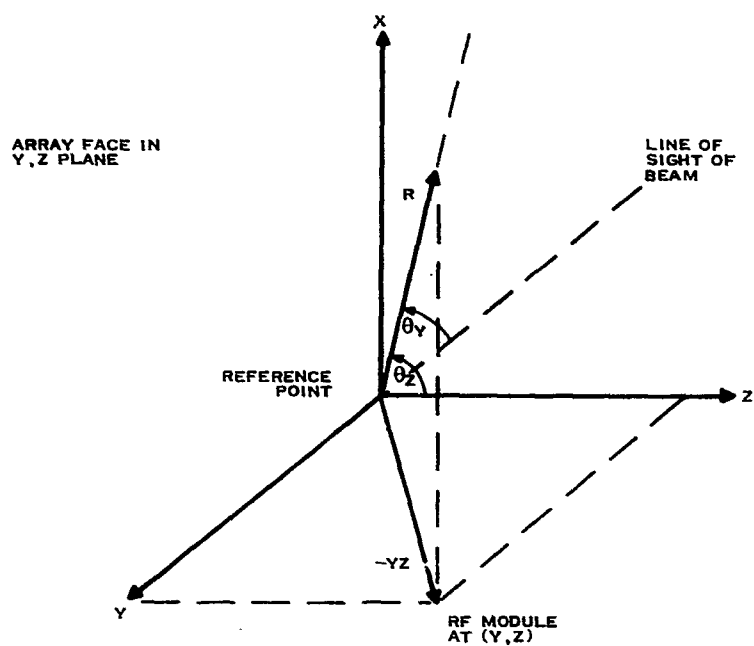
where λ is the wavelength, $\underline{\rho}$ is the vector distance of the module measured from a common reference point, and $\underline{r}$ is a unit vector in the direction in which the beam axis is to point. The dot ($\cdot$) indicates the scalar product operation. For a planar array with coordinate axes of y and z lying in the array face, the phase delay (Γ_{yz}) to be added to an rf module located at point (y, z) in the array face

$$\Gamma_{yz} = \frac{2\pi}{\lambda} \cdot (\underline{r} \cdot \underline{\rho}_{yz}) , \quad (3-130)$$

where $\underline{\rho}_{yz}$ is the vector distance of the module from the origin of the y, z coordinates as shown in Figure 3-134.

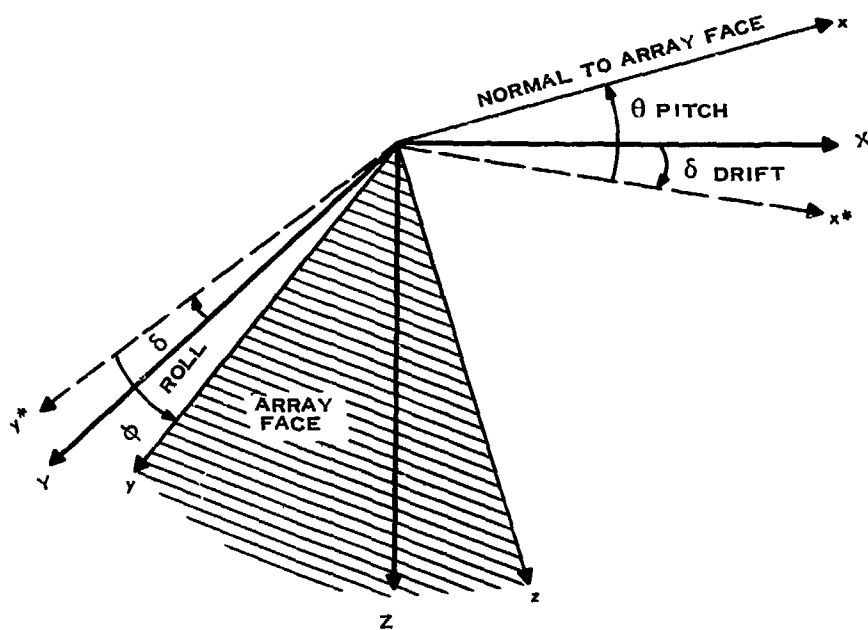
If θ_y and θ_z are defined to be the angles between the unit vector $\underline{r}$ and the y and z axes, respectively, then Equation (3-130) can be written as follows

$$\Gamma_{yz} = \frac{2\pi}{\lambda} (y \cos \theta_y + z \cos \theta_z) . \quad (3-131)$$



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Figure 3-134. Array Face Coordinates



47455

Figure 3-135. Relation of x, y, z Coordinates to the X, Y, Z Coordinates

This equation shows that the phase delay for beam steering decomposes into the sum of two parts—one depending upon its column position y and one depending upon its row position z ; that is,

$$\Gamma_{yz} = \Gamma_y + \Gamma_z \quad (3-132)$$

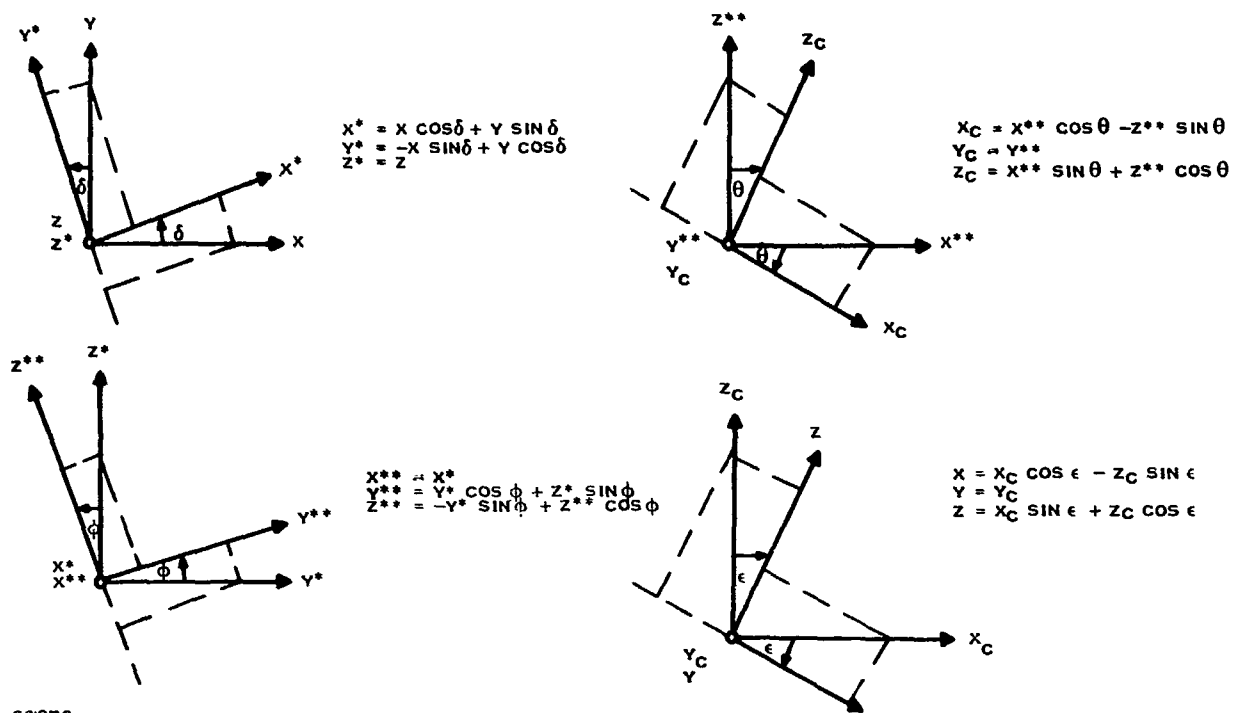
$$\Gamma_y = \frac{2\pi}{\lambda} y \cos \theta_y \quad (3-133)$$

$$\Gamma_z = \frac{2\pi}{\lambda} z \cos \theta_z \quad (3-134)$$

The array scan may be electronically stabilized. The angles to be measured are the azimuth angle, α , measured in the locally horizontal plane, and the elevation angle, β , measured in the vertical plane containing the beam direction vector, $\underline{r}$. In order to write Equations (3-133) and (3-134) in terms of α and β , the orientation of the array face with respect to the local horizontal plane and the flight vector must be known, and the appropriate coordinate transformations utilized. The resulting phase-shift equations provide all-electronic stabilization of the beam axis.

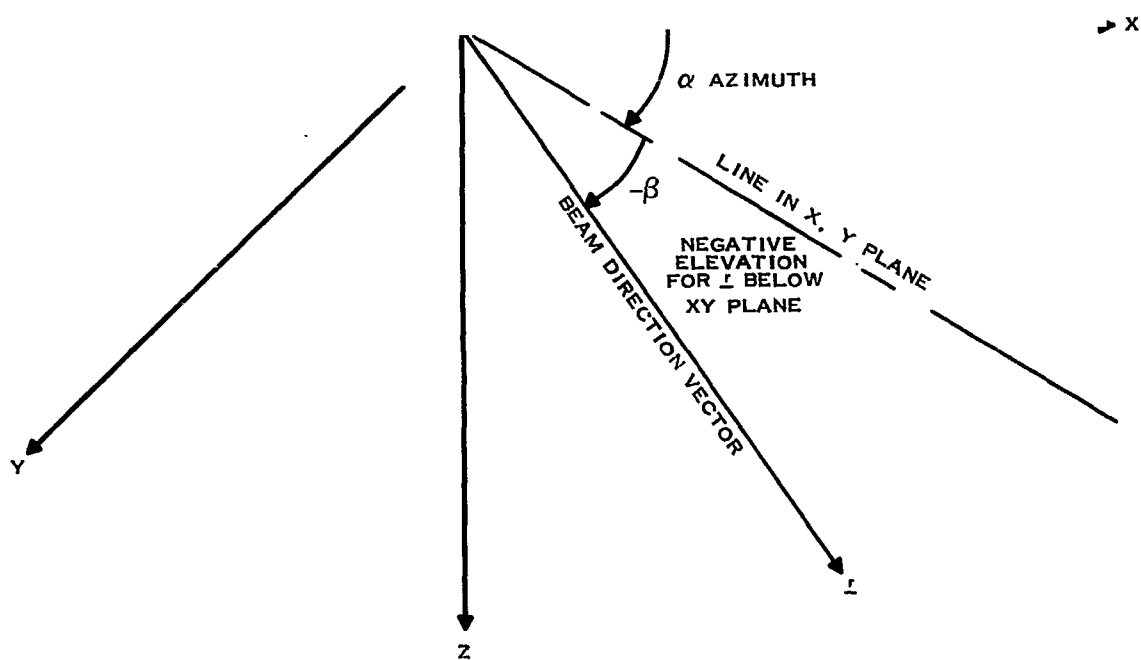
Figure 3-135 shows the geometry involved. The x -axis is normal to the array face with x, y, z forming a right-handed orthogonal coordinate system. The X, Y, Z right-handed coordinate system defines the locally horizontal plane with the Z -axis directed down along the local vertical direction. The X -axis lies along the horizontal component of the flight vector; that is, along-ground track, and the Y -axis is in the across-track direction. The orientation of the x, y, z axes (linear array axes) with respect to the X, Y, Z spacecraft inertial coordinate system is defined by a drift angle, δ , a pitch angle, θ , and a roll angle, ϕ , which will vary as a function of spacecraft attitude stability. The possibility of a cant angle, ϵ , between the rf array coordinates and the spacecraft inertial coordinates must also be considered. Choosing the proper order of rotation about the coordinate axes will simplify the mechanization of the scan programmer. If the cant angle were about the pitch axes, the order of rotation would be as shown in Figure 3-136. Thus, the orientation of the x, y, z axes with respect to the X, Y, Z coordinate system is defined by positive angular rotations about the Z axes, the x^* axes, the y^{**} axes and the y_c axes.

The direction of the beam axis vector, $\underline{r}$, is defined with the azimuth angle, α , and the elevation angle, β . The azimuth angle, α , is measured in the X, Y plane from the X -axis, and the elevation angle, β , is the angle between $\underline{r}$ and the X, Y plane. Each of these angles is shown in Figure 3-137.



63878

Figure 3-136. Rotation of Axes



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Figure 3-137. Definition of α and β

From Figure 3-137 it is clear that the X, Y, Z components of the unit vector, $\underline{r}$, are, respectively,

$$r_X = \cos \alpha \cos \beta \quad (3-135)$$

$$r_Y = \sin \alpha \cos \beta \quad (3-136)$$

$$r_Z = -\sin \beta . \quad (3-137)$$

The transformation from the X, Y, Z coordinates to the x, y, z coordinates is given by the transformation matrix $T(\theta, \epsilon, \phi, \delta)$

$$\begin{bmatrix} x \\ y \\ z \end{bmatrix} = T(\theta, \epsilon, \phi, \delta) \begin{bmatrix} X \\ Y \\ Z \end{bmatrix} \quad (3-138)$$

where

$$T(\theta, \epsilon, \phi, \delta) = [\theta][\epsilon][\phi][\delta] \quad (3-139)$$

and

$$[\delta] = \begin{bmatrix} \cos \delta & \sin \delta & 0 \\ -\sin \delta & \cos \delta & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad (3-140)$$

$$[\phi] = \begin{bmatrix} 1 & 0 & 0 \\ 0 & \cos \phi & \sin \phi \\ 0 & -\sin \phi & \cos \phi \end{bmatrix} \quad (3-141)$$

$$[\epsilon] = \begin{bmatrix} \cos \epsilon & 0 & -\sin \epsilon \\ 0 & 1 & 0 \\ \sin \epsilon & 0 & \cos \epsilon \end{bmatrix} \quad (3-142)$$

$$[\theta] = \begin{bmatrix} \cos \theta & 0 & -\sin \theta \\ 0 & 1 & 0 \\ \sin \theta & 0 & \cos \theta \end{bmatrix} \quad (3-143)$$

Application of Equation (3-138) to the X, Y, Z components of $\underline{r}$, given by Equations (3-135), (3-136), and (3-137), yields the x, y, z components of $\underline{r}$.

$$\begin{bmatrix} r_x \\ r_y \\ r_z \end{bmatrix} = T(\theta, \epsilon, \phi, \delta) \begin{bmatrix} \cos \alpha \cos \beta \\ \sin \alpha \cos \beta \\ -\sin \beta \end{bmatrix} \quad (3-144)$$

Since the positive vector ρ_{yz} of each module is coplanar with the array and has no x component, Equation (3-130) can be written as

$$\Gamma_{yz} = \frac{2\pi}{\lambda} (y r_y + z r_z) \quad (3-145)$$

$$\cos \theta_y = r_y \quad (3-146)$$

$$\cos \theta_z = r_z, \quad (3-147)$$

as expected from the definition of θ_y , θ_z , and $\underline{r}$.

Using Equations (3-139) through (3-144), r_y and r_z and, consequently, $\cos \theta_y$ and θ_z may be expressed as follows

$$r_y = \cos \theta_y = \sin (\alpha - \delta) \cos \phi \cos \beta - \sin \phi \sin \beta \quad (3-148)$$

$$\begin{aligned} r_z = \cos \theta_z &= [\sin (\delta - \alpha) \sin \phi \cos \beta - \cos \phi \sin \beta] \cos (\epsilon + \theta) \\ &+ \cos (\alpha - \delta) \cos \beta \sin (\epsilon + \theta). \end{aligned} \quad (3-149)$$

For completeness, r_x is

$$\begin{aligned} r_x = \cos \theta_x &= [\sin (\delta - \alpha) \sin \phi \cos \beta - \cos \phi \sin \beta] \cos (\epsilon + \theta) \\ &+ \cos (\alpha - \delta) \cos \beta \sin (\epsilon + \theta). \end{aligned} \quad (3-150)$$

For equal spacing of the rf modules d_z in the rows and d_y in the columns, one can write

$$y = m d_y \quad (3-151)$$

$$z = n d_z, \quad (3-152)$$

with m and n varying over the number of columns and rows, respectively. Then Equations (3-133) and (3-134) become

$$\Gamma_y = m \left[\frac{2\pi}{\lambda} d_y \cos \theta_y \right] \quad (3-153)$$

$$\Gamma_z = n \left[\frac{2\pi}{\lambda} d_z \cos \theta_z \right]. \quad (3-154)$$

These two equations with Equations (3-132), (3-148), and (3-149) give the phase delay required for each of the rf modules located at (md_y, nd_z) in the array coordinates.

3.4.2.2 Scan Programmer

The scan programmer must compute $\cos \theta_y$ and $\cos \theta_z$ given by Equations (3-148) and (3-149) for steering the beam in the inertial coordinate system of the spacecraft (X, Y, Z) to the azimuth angle, α , and the elevation angle, β . It must also contain the necessary function generators and control logic for selecting the proper position of the beam and generate the corresponding time functions for scanning the beam in azimuth and elevation. The control logic must also accept externally generated signals for α and β from either the operator control panel or tracking circuits.

The functional block diagram of the scan programmer is shown in Figure 3-138. The spacecraft attitude angles, ϕ , θ and δ , are inputs from the spacecraft navigation system. The cant angle, ϵ , is a positive fixed rotation about the pitch axes, θ .

Two additional functions are generated in the scan programmer. Since only rf electrical phase delays can be achieved in the antenna array, the lines to the array elements must be interchanges to produce the phase delays from the opposite side of the array when the beam passes through zero degrees. A zero crossing detector is connected to each of the computed values of $\cos \theta_y$ and $\cos \theta_z$ to detect the zero degree point. This circuit then generates the gates to switch the lines to the elements of the array.

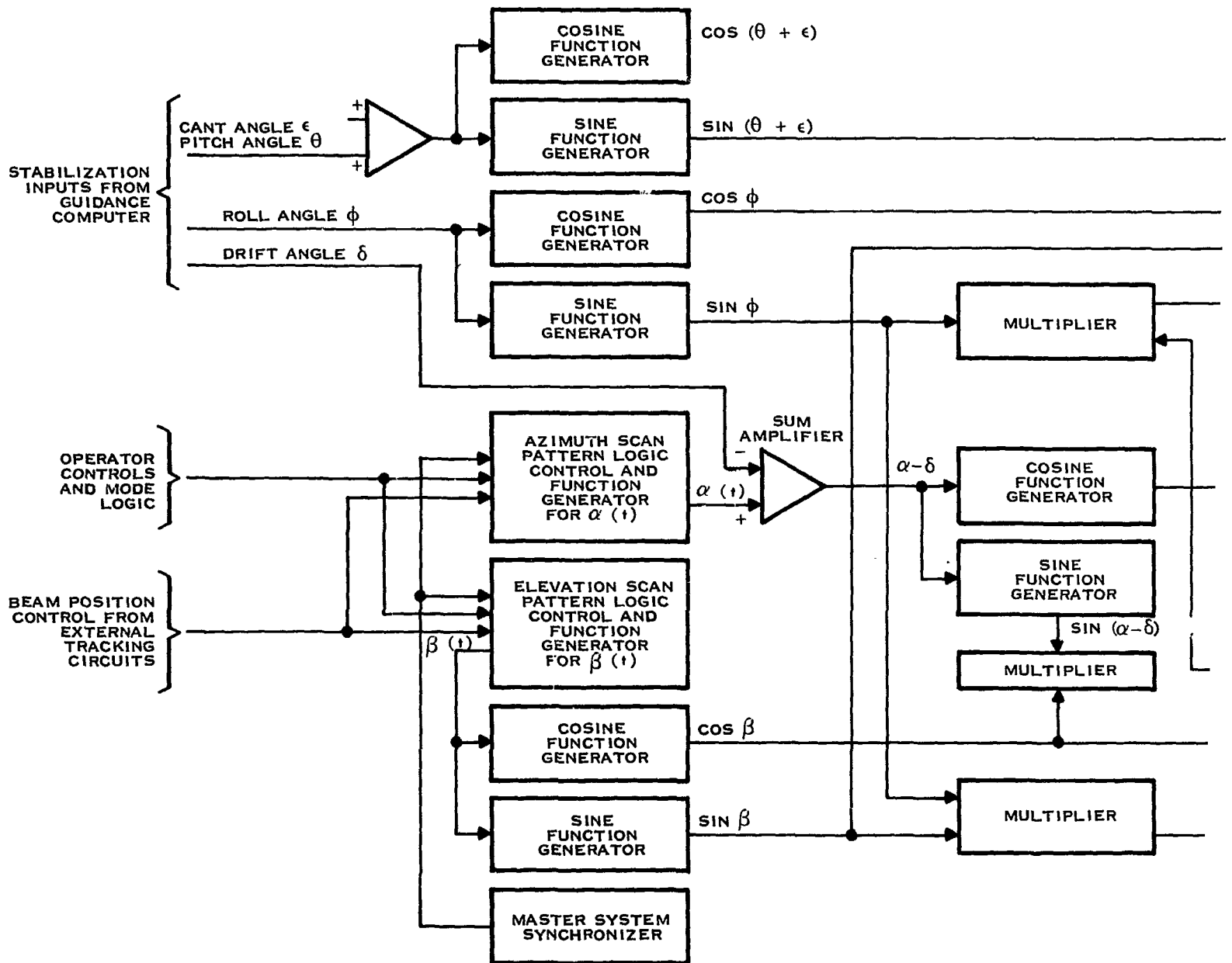
The second additional function is that of converting the analog value of $\cos \theta_y$ and $\cos \theta_z$ to pulsewidth variations. The pulsewidth is then proportional to the amount of required electrical phase delay required to scan to the desired azimuth and elevation angles.

3.4.2.3 Phase-Shift Computer

The phase-shift computer must generate the appropriate digital signals for the correct phase shift required in each rf module and distribute these signals to the modules at the correct times. The phase-shift computer block diagram is shown in Figure 3-139.

The pulsewidth modulated $\cos \theta_y$ and $\cos \theta_z$ signals from the scan programmer are used to generate the phase delay required in each rf module. $\cos \theta_y$ signals are converted to row signals and $\cos \theta_z$ signals are

FOLDOUT FRAME



FOLDOUT FRAME

2

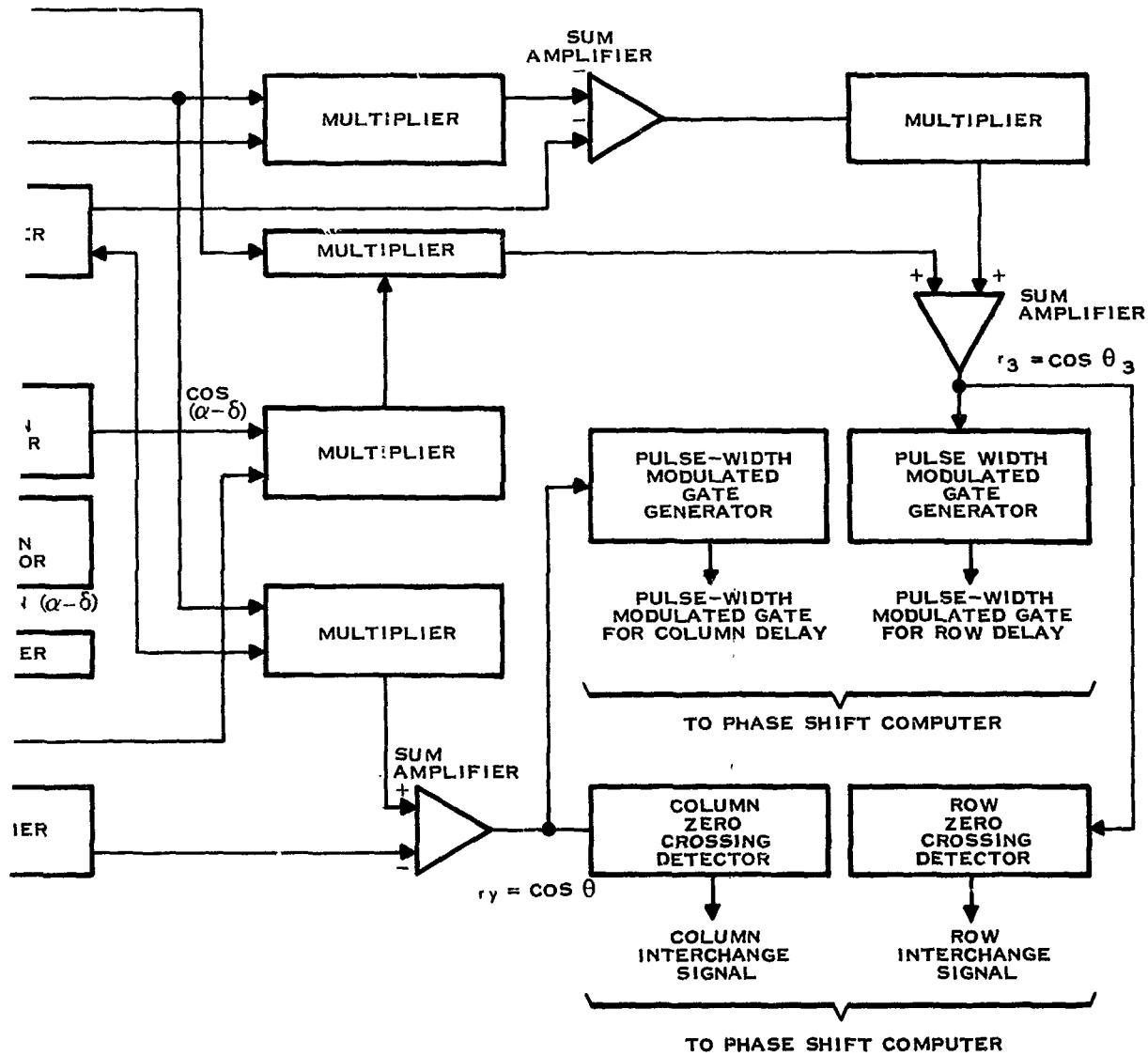
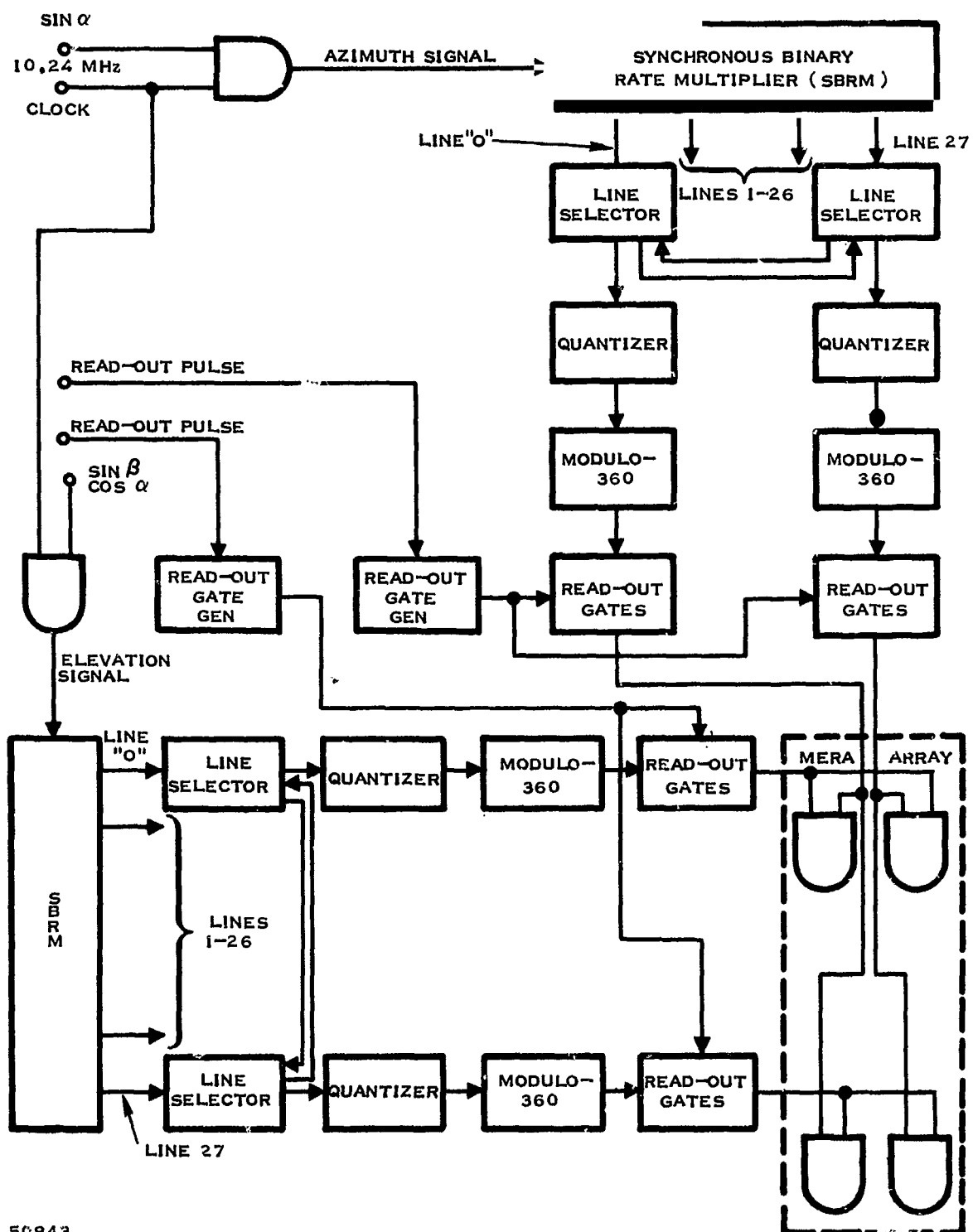
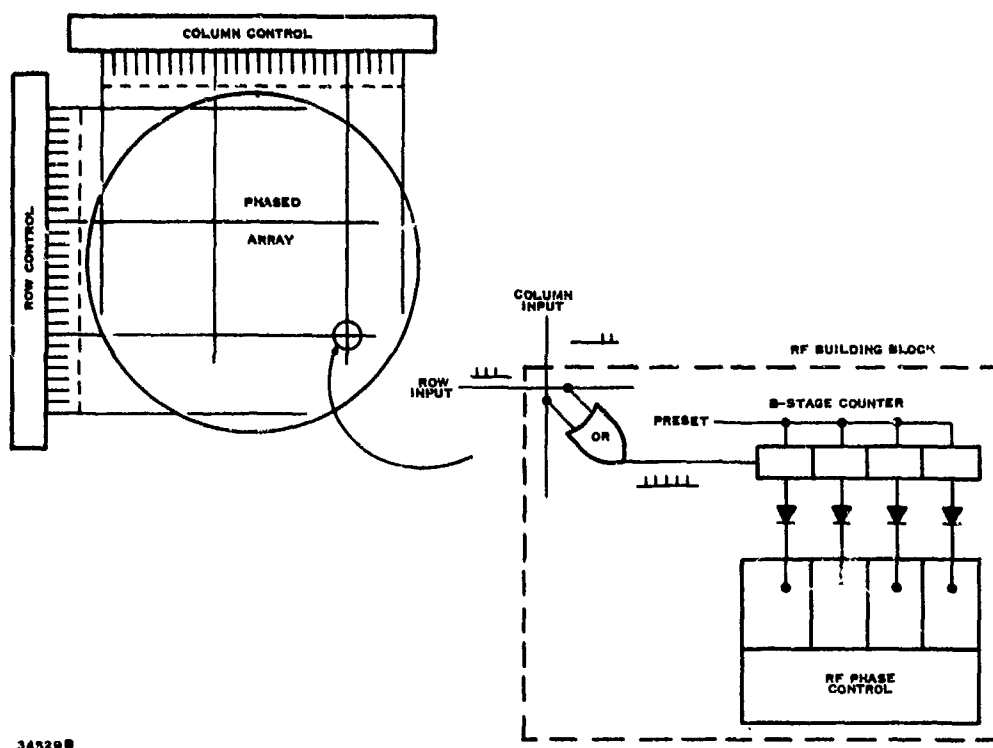


Figure 3-138. Scan Programmer



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Figure 3-139. Phase-Shift Computer



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Figure 3-140. Phase Control in RF Module

converted to column signals. These angle signals are received as single pulses whose widths are proportional to $\cos \theta_y$ and $\cos \theta_z$ and are used to enable gates to allow the clock signal to reach the row and column synchronous binary rate multipliers (SBRM).

Starting at the individual rf module and working back to the SBRM's, the composition of the phase-shift computers is as follows. There are N^2 rf modules arranged in an $N \times N$ grid. Each of these modules contain a four-stage switch. These switches either by pass or insert a fixed phase shift, according to the state of the associated flip-flop. Each module contains an "or" gate to sum the inputs from the rows and columns. This is illustrated in Figure 3-140. Each row and column of the array has an associated modulo 360 generator and quantizer (Figure 3-139). There are two line selectors and SBRM's, one for the row signals and one for the column signals.

The pulsewidth proportional to the desired steering angle allows a specific number of f_0 clock pulses to be received by the SBRM, which is essentially a frequency divider. The output of the row SBRM would be N individual trains of pulses, one for each row. The number of pulses in a train would be proportional to the desired scan angle and to the distances from the row to the reference edge of the array. Each bit out of the SBRM would represent a specific number of degrees of electrical phase shift. The quantizer then divides the number of pulses by a specific number to be compatible with the phase-shift steps which are used in the module. The amount of phase shift

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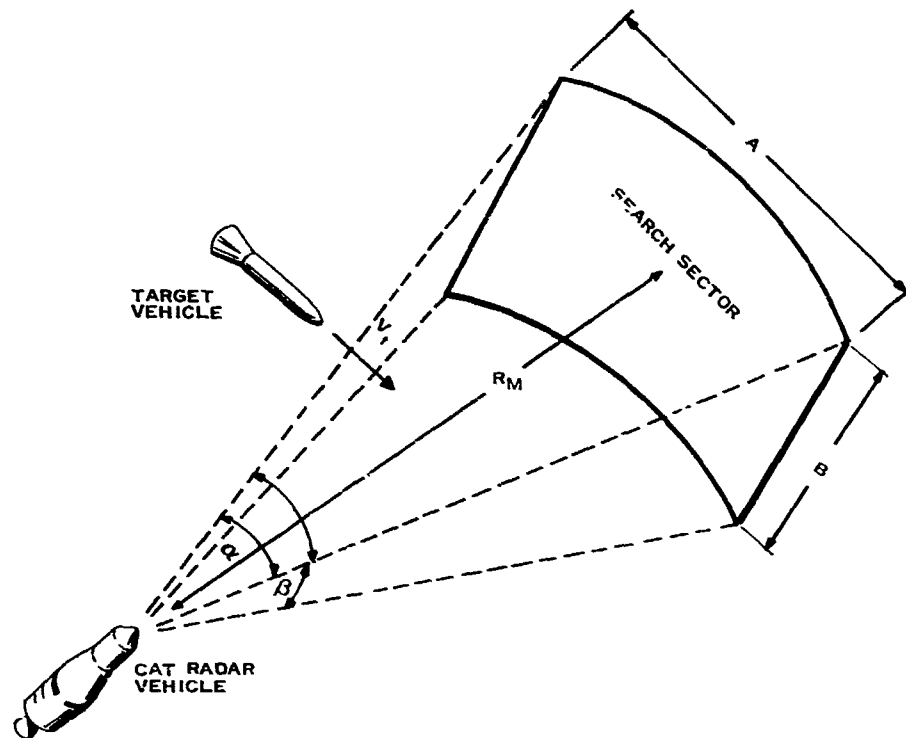


Figure 3-141. Search Geometry

commanded to the farthest element from zero reference, during the extreme scan angle, is measured in thousands of degrees. To relieve the module counter from time-consuming operation of cycling and recycling, until the final amount of phase shift is attained, the train of pulses to each row and column is subjected to a modulo 360 operation, and only the remainders are distributed to the modules. The readout generator gates the modulo 360 remainders to the modules at the time when the beam scan angle is to change.

3.4.3 Beam-Pattern Scan-Control Capability

3.4.3.1 Search Time

The mission of the CAT radar requires search for target acquisition and track for target identification and parameter estimation. The maximum allowable time available to search, acquire, and identify a target will now be investigated. The geometry of the situation is shown in Figure 3-141. In this figure, α and β are the maximum search angles in azimuth and in elevation, respectively. The velocity of the CAT radar spacecraft is v_c and v_t in the tangential velocity of the target spacecraft. The range between the CAT radar and the target vehicle is R_m .

If the CAT antenna beamwidth is less than the total search angle, there is a prescribed time available to search this total sector. This time varies with the total search angle, velocity and range between the satellites. From Figure 3- 41, the length of the chord A is seen to be

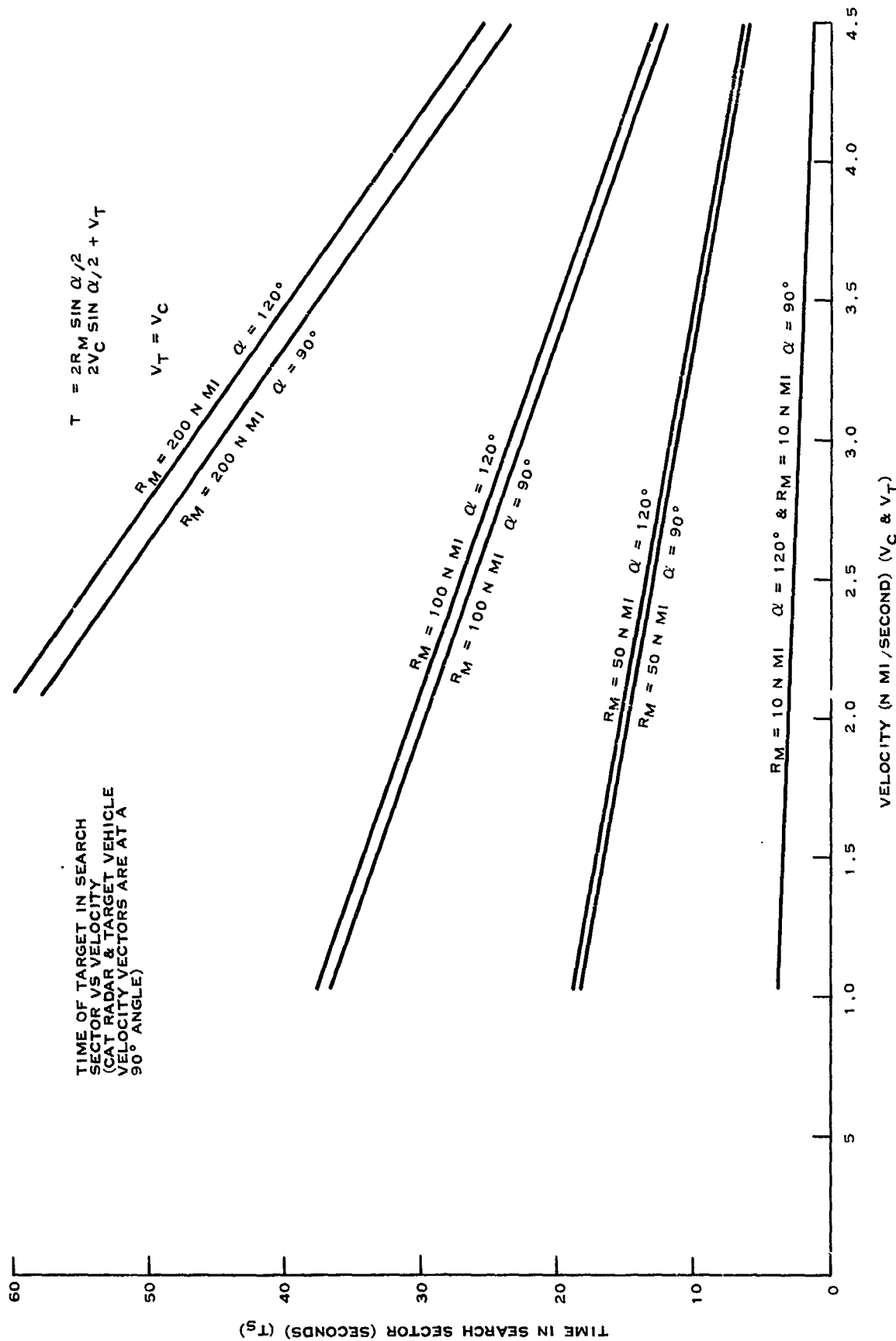


Figure 3-142. Allowable Search Time Versus Velocity

$$A = 2(R_m - v_c t_s) \sin \alpha / 2 , \quad (3-155)$$

where t_s represents the time that the target spacecraft will be in the search sector. Also, t_s must equal A/v_t , or

$$t_s = \frac{2(R_m - v_c t_s) \sin \alpha / 2}{v_t} , \quad (3-156)$$

$$t_s = \frac{2R_m \sin \alpha / 2}{2v_c \sin \frac{\alpha}{2} + v_t}$$

This equation shows that the time the target vehicle will be in the search sector is directly proportional to the range, inversely proportional to the velocity and very insensitive to the maximum search angle when α is a reasonable value; that is, greater than 60 degrees. This is illustrated by Figure 3-142. The minimum time that the target will be in the search sector will occur at the required minimum detection range. As an example, if the circular orbit altitude is 100 nm for each vehicle, the velocity of each spacecraft would be about 4.2 nm/sec. Thus, from the graph, the time the target vehicles will be in the search sector, if the range is 10 nm, would be 1.5 seconds. Therefore, the CAT radar must search, acquire, and identify the target vehicle in 1.5 seconds with the above-stated conditions.

Acquiring targets at long ranges requires reasonably long search times. As the search time is increased to detect targets at long ranges, the ability to detect targets at short ranges is decreased. This occurs at short ranges because the target moves through the search sector more rapidly than at long ranges. The impact of this is illustrated in Figure 3-143. This graph shows the range where the radar search time is equal to the target time in the search sector as a function of the satellite velocities for given radar search times of 5 and 10 seconds. As an illustration of the usefulness of the graph, if the radar search time were 5 seconds, the total search angle, α , were 90 degrees, and the altitude were 100 nm ($v_c = v_t = 4.2$ nm/sec), then the range between the radar and target vehicle, where it is possible for the target vehicle to move through the search sector undetected, would be 35.8 nm. This is for the worst possible condition, or when the target and radar velocity vectors are at 90 degrees.

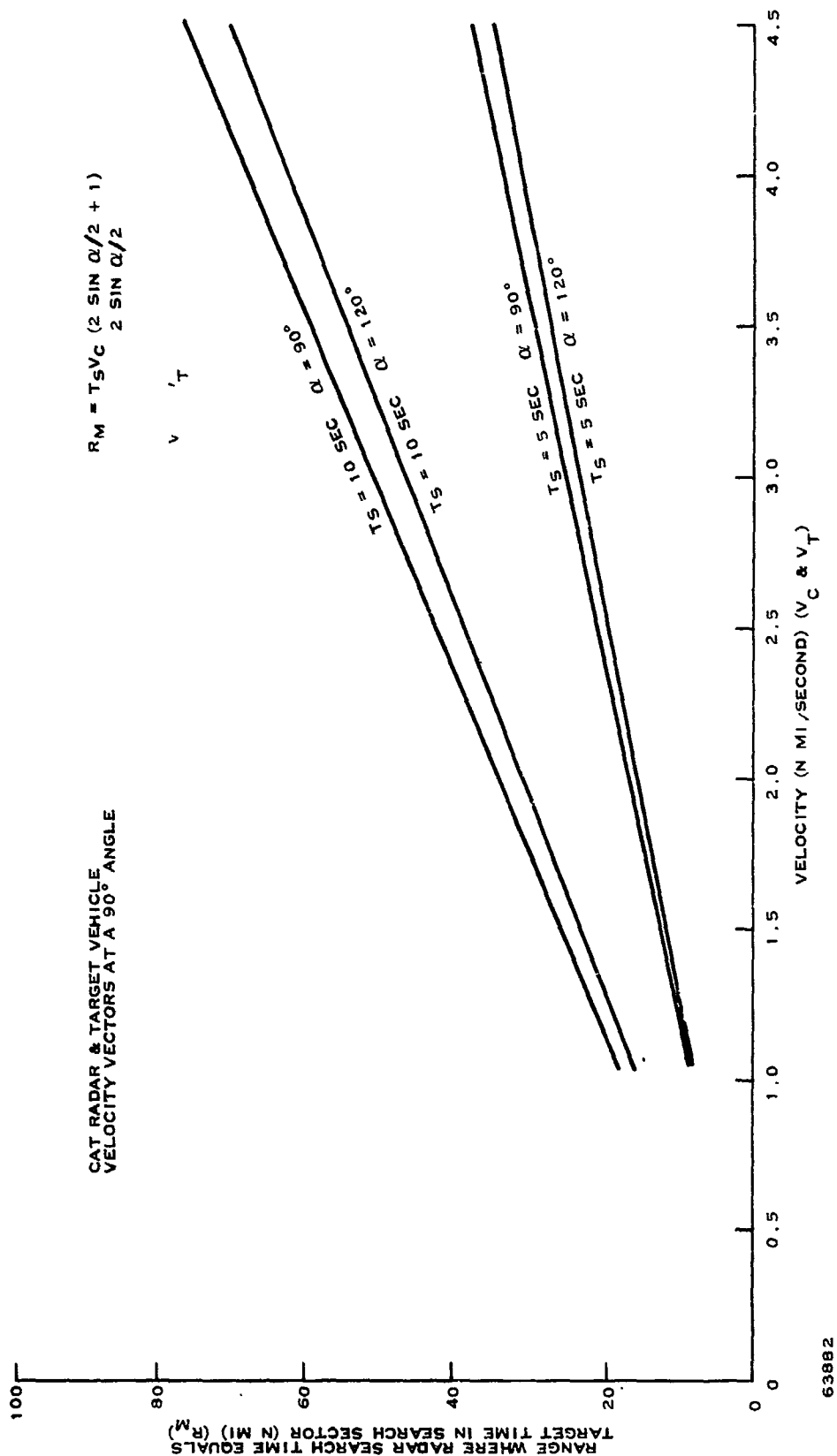
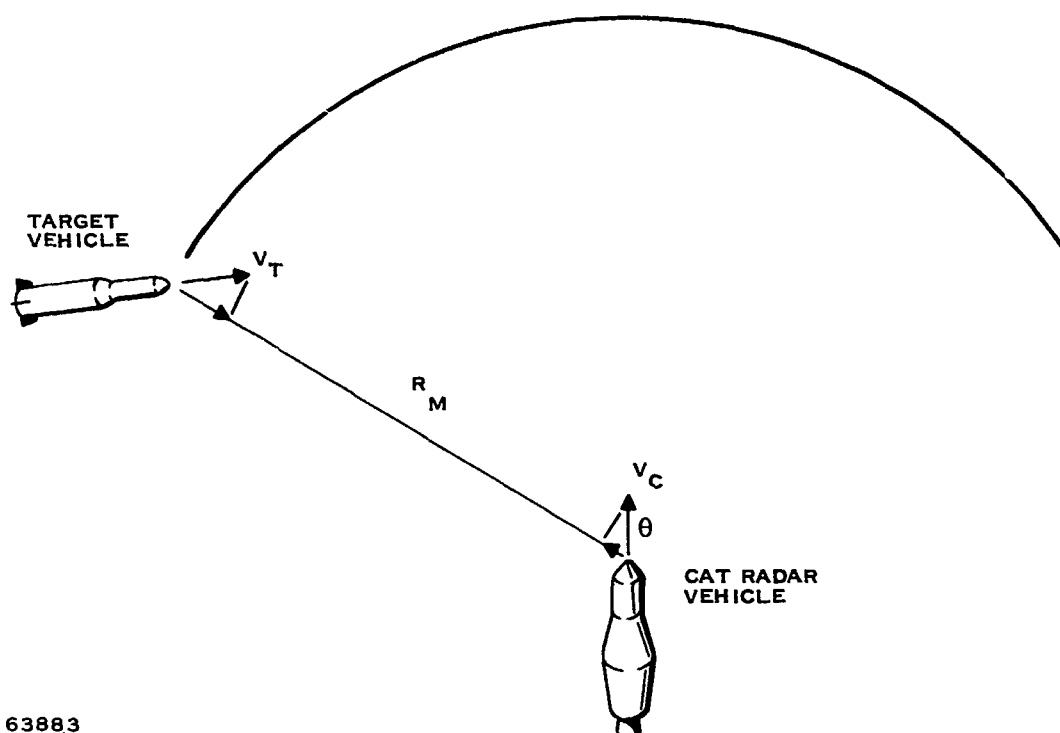


Figure 3-143. Range for Equal Radar Search Time and Target Time in Search Sector



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Figure 3-144. Two-Dimensional Velocity Representation

3.4.3.2 Angular Rate

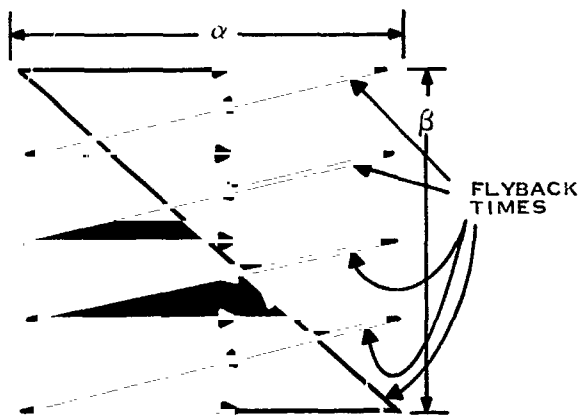
The worst case angular-tracking rate would also occur when the target vehicle and CAT radar velocity vectors are at a right angle. The maximum relative velocity will occur at the minimum altitude; therefore, the two satellites will be assumed at or near the same altitude. This allows the two-dimensional representation illustrated in Figure 3-144, where θ is the angle between the true and radial velocity vector of the CAT radar and v_t is the velocity of the target vehicle. This diagram shows that the tangential velocity is

$$v_a = v_t \cos \theta - v_c \sin \theta . \quad (3-157)$$

The angular rate (ω_t) is equal to v_a/R_m , therefore

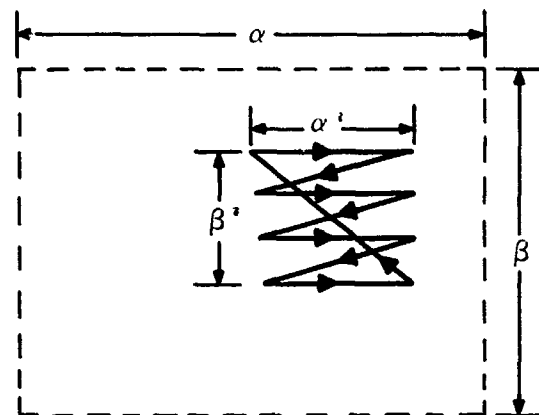
$$\omega_t = \frac{v_t \cos \theta - v_c \sin \theta}{R_m} . \quad (3-158)$$

As might be expected, the angular rate (ω_t) varies directly as the velocity of the two spacecraft and inversely as the range between them. The minimum angular rate occurs when $\theta = 45$ degrees and is equal to zero when $v_t = v_c$. The maximum angular rate occurs when $\theta = -45$ degrees and is equal to the following equation.



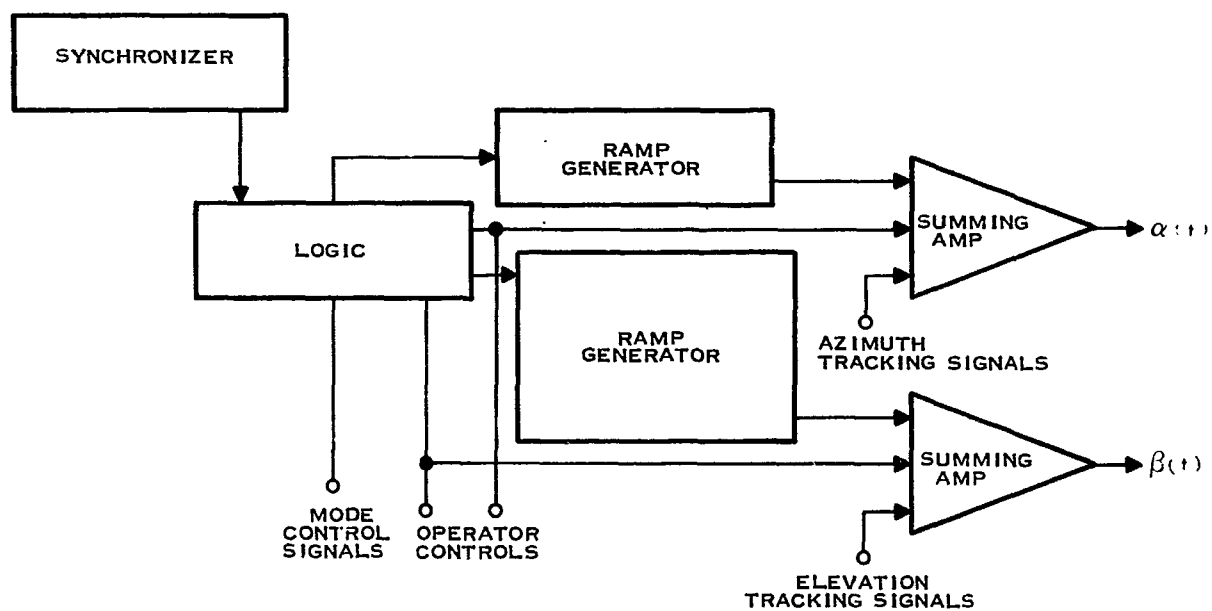
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Figure 3-145. Multibar Raster Scan



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Figure 3-146. Search Sector Reduction



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Figure 3-147. Beam-Pattern Scan Controller

$$t_{(\max)} = \frac{0.707(v_c + v_t)}{R_m} \frac{\text{radians}}{\text{second}}, \quad (3-159)$$

if $v_t = v_c$. The maximum angular rate the CAT radar must be expected to follow can be found by letting $v_t = v_c = 4.2 \text{ nm/sec}$ and $R_m = 5.0 \text{ nm}$. This yields a maximum rate of $1.18 \text{ radians/second} = 68 \text{ degrees/second}$.

3.4.3.3 Beam-Scan Mode Control

The previous two paragraphs have shown the strict requirements that will be placed upon the CAT radar angular control circuitry if the radar must search for target acquisition and track for target identification when the target vehicle and CAT radar spacecraft velocity vectors are at right angles. Certainly, the probability of occurrence of this condition is small. Therefore, configuring the beam scan controller to search only for the worst case conditions would be wasteful. The desirable situation would be for the CAT radar search beam to spend more time at angles where a target is suspected and less time where targets are not suspected. The ability to accomplish this depends upon the mission constraints.

In the case of a vectored intercept, there will be a priori knowledge of suspected target location relative to the interceptor. This knowledge can be used to position the CAT radar beam and reduce the total required search sector. The fly by inspector and noncooperative rendezvous are two examples of this type of mission.

Several types of beam-scan patterns have been employed, in the past, with pencil-beam systems. The most common has been a simple constant-velocity raster scan with a flyback at the bottom of the pattern. The all-electronic-scan capability of the phased-array reduces the flyback time to a negligible quantity. Thus, the optimum pencil-beam scan pattern would be that illustrated in Figure 3-145. The beam is always scanned from left to right and the flyback is always right to left. The total search sector is covered in a uniform manner, ensuring rapid detection of targets. The search sector reduction to α' and β' with a priori knowledge of the location of the target is illustrated in Figure 3-146. This reduction can be accomplished through the spacecraft computer or by the spacecraft astronauts.

The system beam-pattern scan control will be accomplished by the two functional blocks labeled "azimuth (elevation) scan pattern logic control and function generator for $\alpha(t)$, $[\beta(t)]$ " in Figure 3-138. The beam-scan pattern produced by these two blocks may be automatically programmed, controlled by the astronauts, or controlled by the external tracking circuits when target acquisition has occurred. The functional block diagram incorporating these features is shown in Figure 3-147.

3.4.4 Data Accumulation

Assuming successful target acquisition and tracking, the target data is conveniently available in either digital or analog form. The angle data in both array and inertial coordinates can be obtained from the scan programmer (Figure 3-138). The angle data in array coordinates is available in analog form at the output of the "elevation (azimuth) scan pattern logic control and function generator for $\beta(t)[\alpha(t)]$." This analog output is the commanded scan angle in array coordinates.

The angle data in spacecraft inertial coordinates is more interesting and useful. The four outputs of the scan programmer to the phase-shift computer include two pulsewidth modulated gates and two sign indicators. The width of the pulsewidth modulated gates is proportional to the direction cosines ($r_y = \cos \theta_y$, $r_z = \cos \theta_z$) as defined in Figure 3-134. The state of the two-sign indicators gives the direction of the angle (up or down, right or left). Thus, it is a simple matter to use the pulsewidth modulated gates to gate a clock, obtaining a digital representation of the angle data and sensing the state of the zero crossing detectors to obtain the sign. This angular data would be with respect to the spacecraft inertial axis as explained in Subsection 3.4.2. The possible data readouts are illustrated in Figure 3-148.

Also illustrated in Figure 3-148 is the range and range rate data that is readily available. The range-tracking loop is the one described in Subsection 3.3.1.2. The versatility of this loop allows analog range to be sampled directly at the output of the integrator and analog range rate is obtained by amplifying the error signal to the integrator. Digital range is found by clocking the time lapsed between the transmitted pulse and the position of the early gate/late gate tracking gates. Digital rate data, both angle and range, can best be evaluated by the system digital computer by calculating the rate of change of the digital information.

3.5 Thermal Analysis

Since any contemplated radar system will be imperfectly efficient, some waste heat must be rejected. A limited thermal study was performed for the purpose of determining the basic limitations imposed by the heat rejection requirement, and to aid in system selection. This was not intended to be a detailed study of a selected design, and in particular, transient calculations appropriate to an orbital mission were not performed. The array is the unique component of the radar and the only one considered.

The first step in the heat-rejection study is to recognize the relevant aspects of the physical situation and the constraints imposed upon any selected system. The first of these constraints is the operating temperature limit. In the interest of reliable and stable operation, the temperatures of such devices as transistors, diodes, etc., should be limited to no more than about 125°C (398°K).

The second fact which must be recognized is that the array modules will be layered structures. The circuits of the modules will necessarily be on dielectric substrates, which in turn will be bonded through metal alloys to the module cases. Therefore, a temperature drop will be associated with heat dissipation from the circuit devices. This topic has been studied in other programs and the range of thermal resistances to be expected within the module may be considered known. The lower limit is of the order of 20°C/watt.

Third, the array itself will most probably be built up from individual modules or small groups of modules, connected at the rear to electrical manifolds. Under these conditions, minimizing the thermal resistances to more than one exterior surface, as might be desirable for radiation cooling, would be a difficult design and fabrication goal. A second consequence of this construction is that the modules cannot a priori be considered attached to any "massive" heatsink. Thus, the module temperatures would respond quite rapidly to environmental changes caused by the orbit path. This second consequence would be mitigated, if a liquid loop and separate radiator were employed for cooling the array.

Finally, an array standing off the spacecraft is envisioned in order to detect targets forward of the craft. Therefore, use of the spacecraft structure and skin as a heat dissipator for the array was not considered.

Since the intended mission is not closely defined, the possibility must be considered, that the heat rejection surface will be insulated for a period significant with respect to the response time of the modules. Under these conditions a thermal control coating, with solar-absorptance/total-emittance ratio less than unity, is required. Candidates for this coating include: paints developed for such a purpose; flame-sprayed alumina; and use of an anodized aluminum surface. For the latter two possibilities, the literature indicates some uncertainty as to their merits. The control coating would in any event be a dielectric, which is an undesirable ground plane if a dipole antenna array is used. However, since the coating thickness would be small compared to the radar wavelength, the effect on the radar operation should be quite small.

With the above physical considerations in mind, two brief analyses establish the range of heat rejection from the array modules. The first considers heat rejection directly from the forward-facing plane of the antenna. We assume this plane to be fully insulated (solar distance 1 AU) for a period great enough to achieve nearly steady state.

To further describe the problem, it is usually possible to identify a particular device which is the greatest source of concern. If this device dissipates fraction F of the total dissipation of an array element; if the internal thermal resistance of the element is θ_1 , and the case-to-array surface resistance is θ_2 ; then the array surface temperature must be maintained at a value

$$T_s = 400^\circ\text{K} - q\theta', \quad (3-160)$$

where

$$\theta' = F\theta_1 + \theta_2 .$$

Furthermore, if the array surface is isothermal, faces the sun, has a solar absorptance, α_s , and total emittance, ϵ_T , and has effective surface area per element of $\phi\lambda^2/4$, where λ is the radar wavelength, then by a heat balance the array surface will attain an equilibrium temperature given by,

$$T'_s = \sqrt[4]{\frac{S\alpha_s}{\sigma\epsilon_T} + \frac{4q}{\phi\lambda^2\sigma\epsilon_T}} , \quad (3-161)$$

where

- S = solar constant, 442 Btu/(hr/ft²)
- σ = Stefan-Boltzmann constant
- ϵ_T = total emittance of surface
- α_s = solar absorptance of surface
- $4q$ = heat dissipated per array element
- λ = radar wavelength
- ϕ = fraction of surface actually available.

The factor ϕ may be near unity for a dipole-and-conducting-plane antenna array, or of the order of 1/3 for a cavity-in-dielectric array. The array may be cooled by radiation alone provided

$$T'_s \leq T_s . \quad (3-162)$$

In the accompanying Figure 3-149, T_s and T'_s are plotted against the array element heat dissipation. To the right of the gray area, operation is not probable; to the left, quite probable. The boundaries of the gray area are defined by Equation (3-160) with θ' varying from 25°C/watt, which is considered an improbable value, to 100°C/watt, which should be well within the realm of possibility. The other curves represent Equation (3-161) for four values of radar frequency, with ϕ equal 1/3 or 1 for each frequency. The emittance is constant at 0.8.

Part (a) of Figure 3-149 is for the case α_s/ϵ_T equal to zero (array does not view the sun). At the left boundary of the gray area, the curves intersect at heat dissipation values ranging from 0.25 watt to 1.8 watts, depending on frequency.

Figure 3-149. Presentation of Operating Envelope for Array Cooled From Surface

Part (b) shows the case $\alpha_s/\epsilon_T = 0.25$. As expected, this shows that less heat can be rejected when the array is insulated.

Part (c) shows the case α_s/ϵ_T equal to 0.5. The curves may seem to intersect at the left boundary at allowable heat values ranging from 0.1 to 1.0 watt. As α_s/ϵ_T approaches 1.0, a case not shown, the allowable heat dissipation for this cooling mode is reduced to milliwatts. For α_s/ϵ_T greater than 1.0, the array elements must be cooled by other means, and perhaps deliberately isolated thermally from the array surface.

In each of the three cases, if the antenna configuration is such that less than the full frontal area can be used, the heat dissipation must be further restricted. Thus, a cavity-in-dielectric antenna array would be restricted to approximately the same heat dissipation as a dipole array operating at twice the frequency of the cavity array.

If the array cannot be directly cooled from its outward-facing surface, a heat-transfer fluid loop and separate radiator would be necessary.

A liquid-cooled design would be judged by at least four criteria. It must:

- Maintain device temperatures at safe levels
- Allow assembly of the array
- Draw minimum pumping power
- Have minimum mass.

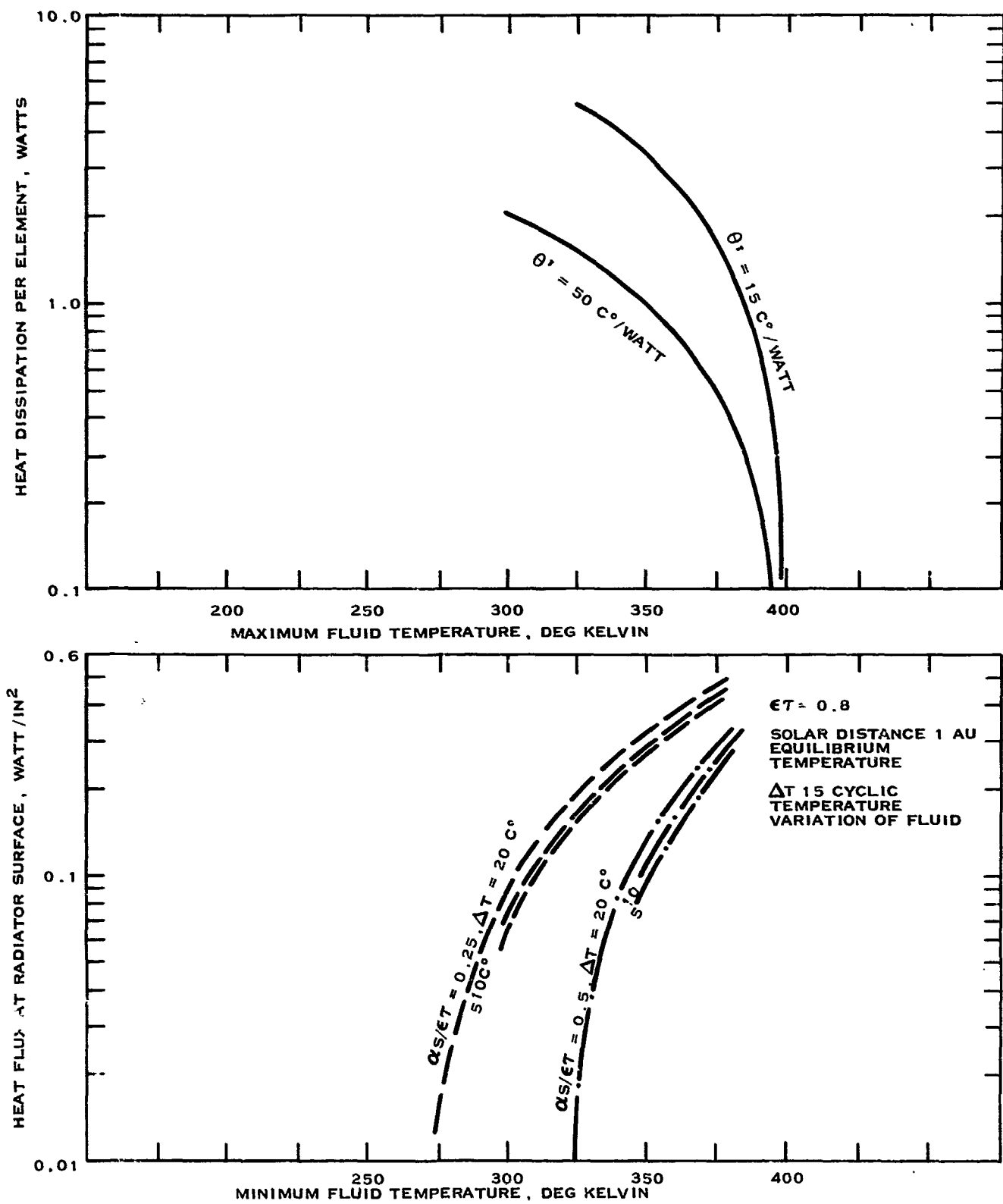
The added mass is expected to be affected by the pumping power and radiator area requirements. The pumping power can be shown to be strongly influenced by the geometry of the heat-exchange surfaces and by the fluid properties—particularly the fluid viscosity.

Although it is not possible to specify a cooling-loop design without a definite statement of the spacecraft thermal environment, by retreating to some of the conditions assumed earlier, we can obtain an estimate of the limitations of this cooling mass.

Assume, for example, that the radiator faces the sun, which is 1 AU distant; that its emittance is ϵ_T and its solar absorptance is α_s . Assume further that its surface temperature varies linearly with respect to distance across one dimension, as the fluid traveling through it cools. If the cooler end is at temperature T_0 and the warmer end at temperature $T_0 + \Delta T$, the heat flux q/A which the radiator may reject in the steady state is given by,

$$q/A = \frac{\epsilon_T \sigma}{5\Delta T} \left[(T_0 + \Delta T)^5 - T_0^5 \right] - \alpha_s S, \quad (3-163)$$

where S is the solar constant and σ is the Stefan-Boltzmann constant.



63889

Figure 3-150. Estimate of Fluid-Loop Heat-Rejection Performance

If, for this analysis, we further assume the fluid-to-radiator temperature difference is negligible, then the junction temperature of the crucial electronic device is,

$$T_j = T_o + T + q\theta' , \quad (3-164)$$

where q is the heat dissipation per array element and θ' is now the junction-to-fluid thermal resistance (defined in the same fashion as the θ' of the previous problem).

Since T_j is fixed, the maximum fluid temperature is defined by Equation (3-164) and the fluid radiator area required per array element may be inferred from Equation (3-163).

Figure 3-150 shows results obtained for the cases $\epsilon_T = 0.8$, $\alpha_s/\epsilon_T = 0.25$ or 0.50 , $\theta' = 15^\circ\text{C/watt}$ or 50°C/watt . The former value is regarded as improbable, the latter readily possible. The curves show, as an example, that for $q = 1.0$ watt and $\theta' = 50^\circ\text{C/watt}$, the maximum allowable fluid outlet temperature is 350°K . The radiator flux would be about 0.3 watt/in² for the case $\alpha_s/\epsilon_T = 0.25$, or about 0.1 watt/in² for the case $\alpha_s/\epsilon_T = 0.5$. The exact value depends slightly on the fluid cyclic temperature variation, ΔT . The radiator heat flux diminishes very rapidly as the fluid temperature decreases, and a minimum temperature exists because of the absorbed solar heat.

Under the conditions assumed, the limit of performance for a fluid loop would be a heat removal of from 1 to (at the most) 3 watts per array element for which a radiator surface of 3 to 10 square inches per watt would be required.

In conclusion, the heat rejection capability of the system under conditions of full insulation is of the order of 1 to 3 watts per array element. It may be possible to provide more heat-rejection capability if more detailed definition of the constraints is provided.

The above remarks cover only the cases in which steady temperatures are attained in the insulated and noninsulated cases. Certainly, in the design of a selected system, an analysis of the expected actual cyclic temperature variations encountered in an orbital mission would be required. This is particularly true since such temperature variations are recognized as a cause of circuit failures.

4.0 SYSTEM PERFORMANCE ANALYSIS

4.1 Introduction

In this section the subsystem analyses of preceding sections will be integrated in order to obtain parametric performance measures for the overall system. The threshold detection technique discussed in Subsection 3.3 will first be analyzed. Then, for a nominal set of parameters, a comparison will be made of the performance of the various tracking techniques discussed in Subsection 3.3. Combining these results, a parametric analysis of the overall signal processing system for various possible system configurations will be found.

4.2 System Parametric Analysis

4.2.1 Target Detection Analysis

Detection performance of the CAT radar can be determined in terms of the range equation. In one of its fundamental forms, the range equation is

$$S/N = \frac{P_t \tau G^2 \lambda^2 \sigma n}{(4\pi)^3 kT F_n L_s R^4} \quad (4-1)$$

where

P_t = Peak transmitted power (W)

τ = Duration of transmitted pulse (sec)

G = Antenna gain

λ = Transmitted wavelength (m)

σ = Target radar cross section (m^2)

n = Number of pulses integrated

k = Boltzmann's constant = 1.38×10^{-23} (joules/degree)

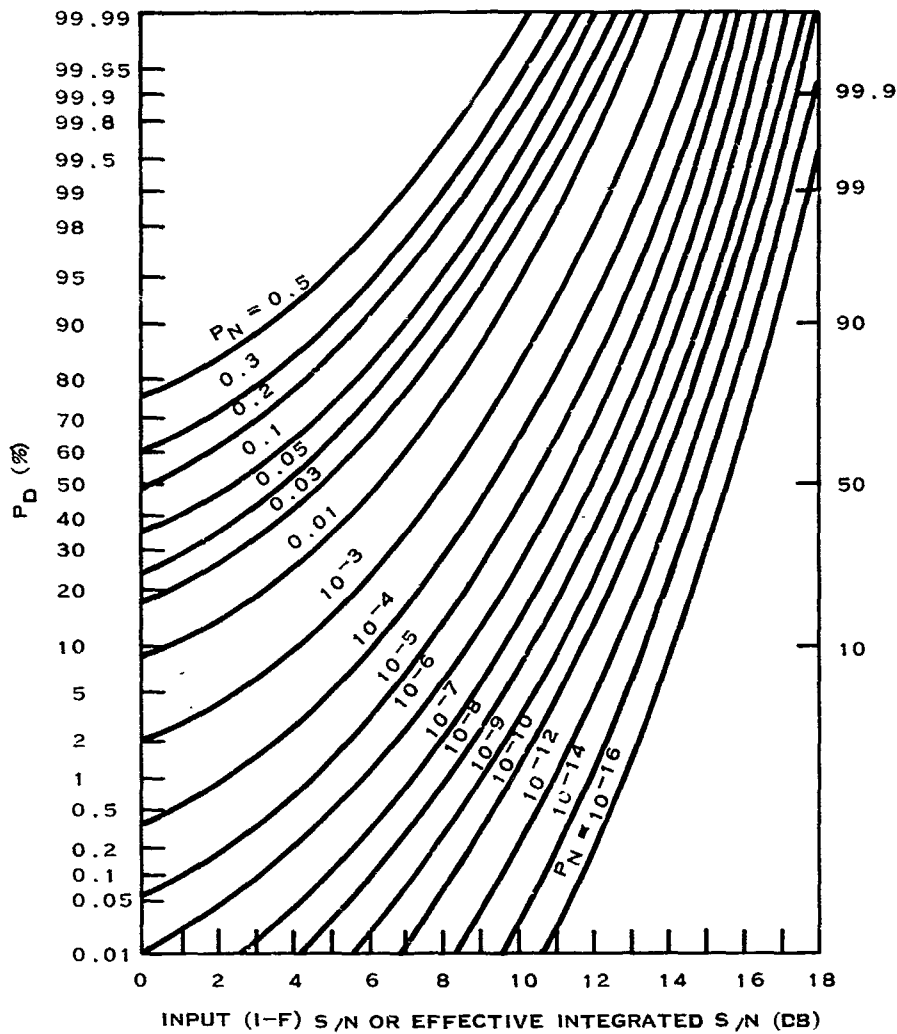
T = Receiver noise temperature (degrees K)

F_n = Receiver noise figure

L_s = System loss factor

R = Range to target (m)

S/N = Effective peak if. signal-to-noise (power ratio for a receiver matched to the signal bandwidth.



63890

Figure 4-1. Probability of Detection (P_D) as a Function of Signal-to-Noise Ratio (S/N) and Probability of a False Alarm (P_N)

Thus, one may determine the S/N ratio for a particular set of parameters. The S/N ratio is related to probability of detection and probability of false alarm shown by Figure 4-1 for a simple threshold detection technique like that discussed in Subsection 3.3.

Equation (4-1) can be put into a more convenient form by assuming a standard noise temperature, $T = T_0 = 290^\circ\text{K}$, and converting the range parameter to nautical miles. For a phased-array system, such as envisioned for the CAT radar, assume

$$G = \pi N_m, \quad (4-2)$$

where N_m is the number of modules comprising the array. Making these substitutions in (4-1) gives

$$S/N = 1.06 \times 10^5 \frac{P_t \tau N_m^2 \lambda^2 \sigma_n}{F_n L_s R^4} \quad (R \text{ in nm}) . \quad (4-3)$$

For an all solid-state corporate-fed array,

$$P_t = N_m P_{tm} , \quad (4-4)$$

where P_{tm} is the peak power transmitted per module, thus (4-3) becomes

$$S/N = 1.06 \times 10^5 \frac{P_{tm} \tau N_m^3 \lambda^2 \sigma_n}{F_n L_s R^4} . \quad (4-5)$$

Alternatively, one may determine the minimum signal-to-noise ratio, $(S/N)_{\min}$, for given detection and false alarm criteria and rewrite (4-5) in terms of the maximum range for which these conditions can be satisfied. This gives

$$R_{\max}^4 = 1.06 \times 10^5 \frac{P_{tm} \tau N_m^3 \lambda^2 \sigma_n}{F_n L_s (S/N)_{\min}} . \quad (4-6)$$

The number of pulses integrated can be found as follows. Let ψ_s be the solid angle to be searched. If the search sector is defined by an azimuth angle $\pm\alpha$ and an elevation angle $\pm\beta$, as shown in Figure 4-2, it is shown that

$$\psi_s = 4\alpha \sin \beta . \quad (4-7)$$

By definition, the solid angle of the antenna beam is given by

$$\psi_b = 4\pi/G . \quad (4-8)$$

Thus, the number of beam positions to be searched is given by

$$n_b = \frac{\psi_s}{\psi_b} = \frac{G\alpha \sin \beta}{\pi} . \quad (4-9)$$

If t_s is the time available to search the sector, then the dwell time per beam position is given by

$$t_b = \frac{t_s}{n_b} = \frac{\pi t_s}{G\alpha \sin \beta} . \quad (4-10)$$

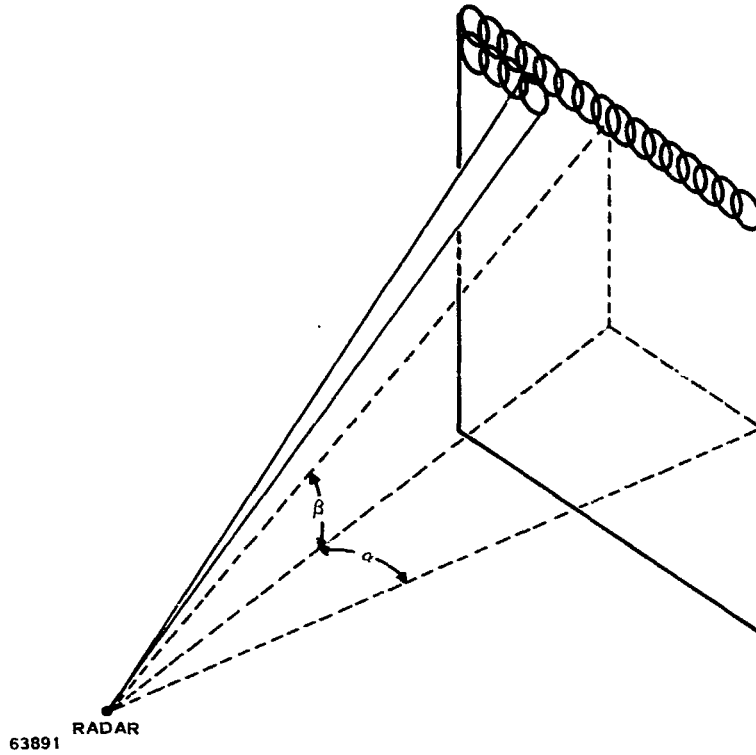


Figure 4-2. Search Sector and Search Scan Pattern for CAT Radar

If f_r is the pulse repetition rate of the radar, then the number of pulses exchanged with each beam position

$$n = t_b f_r = \frac{\pi f_r t_s}{G \alpha \sin \beta} . \quad (4-11)$$

Substituting (4-2) into (4-11) gives

$$n = \frac{f_r t_s}{N_m \alpha \sin \beta} . \quad (4-12)$$

We can now justify selection of a pencil beam, rather than a fan beam for the search mode. Substituting (4-12) into (4-6),

$$R_{max}^4 = 1.06 \times 10^5 \frac{P_{tm} \tau N_m^2 \lambda^2 \sigma f_r t_s}{F_n L_s (S/N)_{min} \alpha \sin \beta} . \quad (4-13)$$

The total power transmitted is given by (4-4) and the aperture of the array is given by

$$A_e = \frac{G \lambda^2}{4\pi} = \frac{N_m \lambda^2}{4} . \quad (4-14)$$

Substituting (4-4) and (4-14) into (4-13) gives

$$R_{\max}^4 = 4.24 \times 10^5 \frac{P_t \tau A_e \sigma_r^f t_s}{F_n L_s (S/N)_{\min} \sigma \sin \beta} \quad (4-15)$$

Assume an array capable of generating a pencil beam; that is, a symmetrical array, which has both dimensions of the array limited by interface considerations. Such an array can be used to form a fan beam by use of defocusing techniques. The effect is to cause one dimension of the effective aperture to be smaller than the other. Since both dimensions of the assumed symmetrical array are taken to be limited, the only way the fan beam can be formed is by reducing one of the dimensions of the effective aperture. The effective aperture area decreases and the maximum range of (4-15) is decreased. Using only a portion of the aperture to form a fan beam is detrimental to detection-range performance if an array is capable of forming a pencil beam when the full aperture is used.

4.2.1.1 System Loss Factor

The overall system loss factor, which appears in (4-1) and subsequent equations, is composed of several terms, any of which may be significant for a particular system configuration. The individual components of the system loss factor are now discussed in detail.

4.2.1.2 Antenna Losses

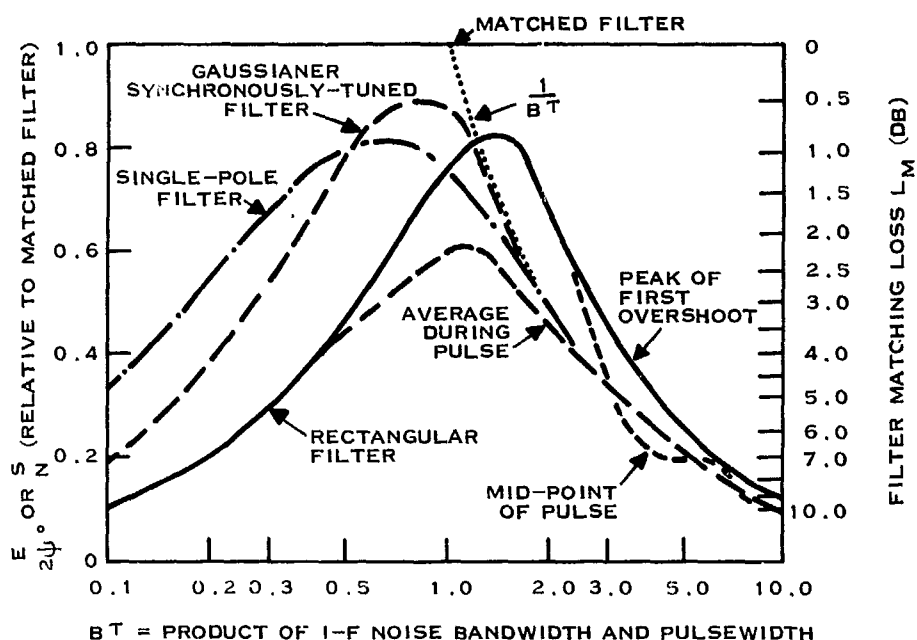
From Subsection 3.1 antenna losses are assumed to be 1.94 dB for a corporate-fed array and 5.2 dB for a space-fed array.

4.2.1.3 Ionospheric Attenuation

The attenuation produced by propagation through the ionosphere was studied by Millman,⁷⁰ whose results show that radars operating above 100 MHz will not normally encounter losses greater than 1 dB, even under daytime conditions when the ionospheric density is greatest. Similarly, results obtained by Blake⁷¹ show that ionospheric loss is negligible. Assume no ionospheric attenuation.

4.2.1.4 Filter Mismatch Loss

Equation (4-2) assumes a matched receiver; that is, $B\tau = 1$ for a system not using pulse compression or $B\tau_c = 1$ for a pulse compression system, where τ_c is the duration of the compressed pulse. Figure 4-3⁷² gives the mismatch loss for various types of filters. For the commonly used rectangular filter, $B\tau = 1.2$. A loss of 0.8 dB occurs relative to a matched filter system. For $B\tau \gg 1$ (as might be used for a leading-edge tracker), the mismatch loss is considerable. For purposes of this study, a mismatch loss of 0.8 dB will be assumed.



63892

Figure 4-3. Filter Mismatch Loss Relative to a Matched Filter

4.2.1.5 Integration Loss

Equation (4-1) assumes coherent pulse-to-pulse integration. Use of post-detection integration entails a loss due to small signal suppression effects in the detector. Figures 4-4 and 4-5 present curves of integration loss as obtained by Marcum.⁷³ As shown on the figures, integration loss is a function of probability of detection, false alarm numbers, and number of pulses integrated.

4.2.1.6 Collapsing Loss

Use of a delay-line integrator provides a means of avoiding collapsing loss due to an ungated analog integrator. As mentioned above, a BT product greater than optimum may be used, resulting in a large mismatch loss. A portion of this loss may be recovered if a low-pass video filter is used after the envelope detector. The filter would have a bandwidth of about half of optimum bandwidth.

Such a technique introduces a collapsing loss. The collapsing ratio is defined by

$$p = \frac{2B_v + B}{2B_v},$$

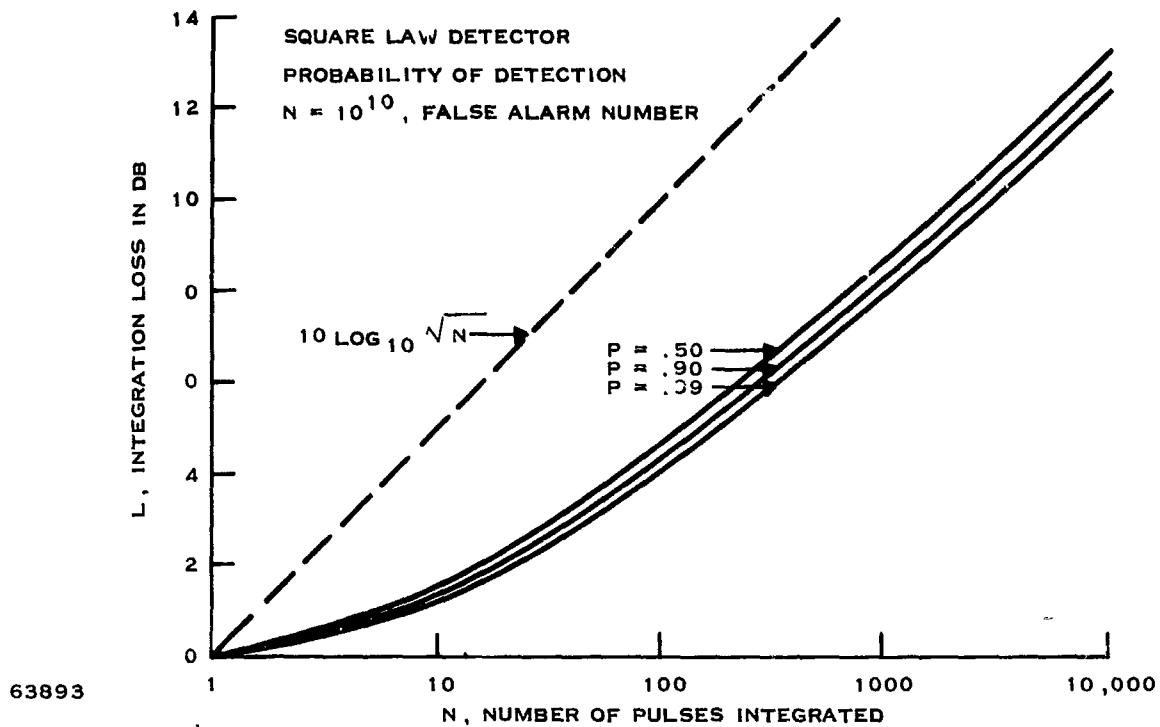


Figure 4-4. Post-Detection Integration Loss Relative to Coherent Integration

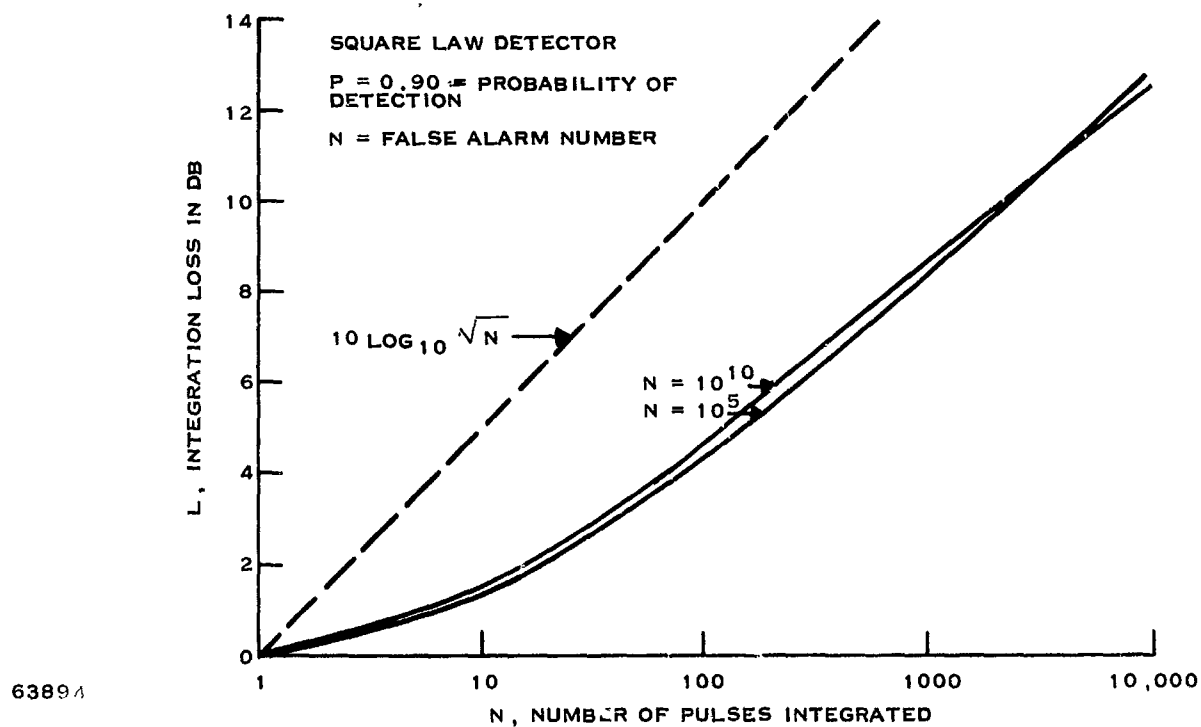


Figure 4-5. Post-Detection Integration Loss Relative to Coherent Integration

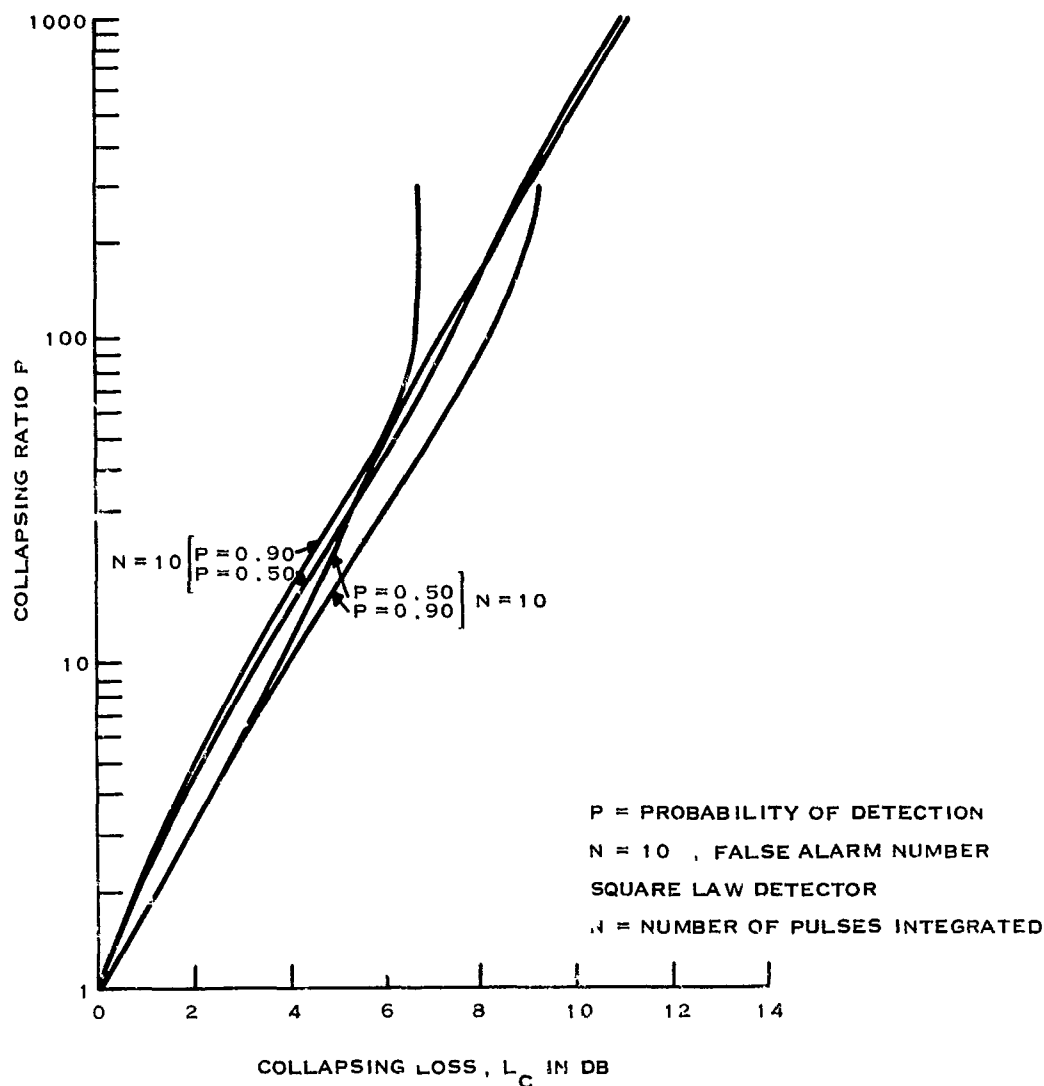


Figure 4-6. Collapsing Loss

where B_v is the video filter bandwidth and B is the if. bandwidth, and the collapsing loss is obtained from the collapsing ratio through use of Figure 4-6.⁷⁷ Generally the collapsing loss in this situation will be less than the filter mismatch loss due to not correcting the video bandwidth. We will assume collapsing loss to be negligible unless high $B\tau$ products are involved.

4.2.1.7 Crossover Loss

As shown in later discussions, systems employing monopulse tracking technique suffer no degradation of detection performance for on-tracking-axis targets due to crossover loss. The conical scan system entails a crossover loss of approximately 2.1 dB for an error slope of 1.57.

4.2.1.8 Fluctuation Loss

A nonfluctuating signal return from the target is assumed.

4.2.1.9 Beam Shape Loss

The antenna gain which appears in the range equation is assumed to be constantly equal to the maximum value. If the target has a crossing velocity with respect to the radar, it will not always be in the position of maximum gain (assume that the antenna beam is stationary for any beam position and steps discretely between beam positions). Thus, there is a degradation in system detection performance due to the nonuniformity of the integrated pulses. Skolnik⁷⁹ has derived an expression for beam-shape loss for a Gaussian antenna pattern, and calculates that a loss of 1.96 dB is associated with integrating 11 pulses uniformly spaced across the 3-dB beamwidth. A target crossing at right angles to the radar, at a range of 50 nm, will have an angular velocity, with respect to the target, of about 0.082 rad/sec. Assume a dwell time in each beam position of 100 ms (in order to search the entire sector in a reasonable time), and a prf of 400, 40 pulses would be exchanged with the target as the target angle varies over 0.0082 radian. The effect of this target motion is negligible for a 3-dB beamwidth of the order of 0.1 radian.

Consequently, a target which is on the 3-dB point of the beam when the beam is switched to a particular beam position will remain on the 3-dB point and never fall within the maximum beam gain. Thus, if the target is in such a position, the signal will be down 3 dB on the two-way power pattern. If the beam positions are defined by the two-way 3-dB beamwidth, the S/N ratio for a particular beam position may be degraded by as much as 3 dB. Counterbalancing this effect is the fact that the target is equally likely to be detected in the adjacent beam position where it would also be on the 3-dB point. Thus, for an overall probability of detection of 0.9, for example, the probability of detection for either of the adjacent beam positions would have to be only 0.684. This implies a lower S/N in each beam position than would ordinarily be required. Figure 4-1 shows that if S/N = 15 dB gives a probability of detection of 0.9, a S/N of only 13.9 dB is required for a probability of detection of 0.7, thus in effect recovering 1.1 dB of the loss due to beam shape. It seems reasonable to assume 2-dB beam shape loss.

4.2.1.10 Scanning Loss

Since the antenna scans discretely rather than continuously, scanning loss in the usual sense is not a factor in this system. However, an electronically scanned phased array suffers a broadening of the beam as it is scanned off boresight. This loss is discussed in more detail in Subsection 3.1. It shows that a 4-dB two-way power loss in maximum antenna gain is to be expected for an antenna scanned ± 45 degrees off boresight.

To summarize the preceding discussion of losses, those which are considered significant for this study are listed with their assumed values:

Antenna Loss	-	1.9 dB (corporate-fed array)
	-	5.2 dB (space-fed array)
Filter Mismatch Loss	-	0.8 dB (rectangular filter)
Integration Loss	-	3.5 dB (50 pulses integrated)
	-	1.5 dB (10 pulses integrated)
Crossover Loss	-	2.1 dB (conical scan-angle tracker)
	-	1.5 dB (conopulse-angle tracker)
Beam Shape Loss	-	2.0 dB
Scanning Loss	-	3.0 dB.

The loss factor of the system may vary from approximately 10 dB to 17 dB, depending on the details of the system.

4.2.2 Target Tracking Analysis

In Subsection 3.3, three different range tracking systems and three different angle tracking systems were discussed, and performance equations derived. The results are summarized in Table IV-1 and Table IV-2. The following is a comparison of the various systems in order to determine their relative merits. First, a guideline system is selected. A system which appears compatible with the restrictions imposed on the CAT radar will have the parameters set forth in Table IV-3 (see Subsection 3-2). The fixed system losses above include antenna loss, beam-shape loss, and scanning. Generally these will not change for different detection and tracking techniques. Other losses must be determined for particular system configurations and added to this figure. For example, in the detection mode there will generally be integration loss and filter mismatch loss. Designating the added loss in the detection mode by L_{ad} , and substituting this with the above parameters into (4-5), we obtain for the single-pulse signal-to-noise ratio,

$$(S/N)_1 = \frac{3.26 \times 10^8}{L_{ad} R^4} \text{ (R in nm) .} \tag{4-16}$$

During the search mode of operation (before detection of a target), pulse integration will be used. The maximum number of pulses which can be integrated is given by (4-12); for $f_r = 400$ (corresponding to a 200-nm unambiguous range), a search time of 10 seconds, and a 45×45 degrees search sector, the number of pulses integrated is $n = 8$. From Figure 4-4, integration loss

Table IV-1. Summary of Performance Equations of Range Tracking Systems*

	Open Loop	Split Gate	Separated Gate
RMS Thermal Noise in Location of Received Pulse (Sec)	$\sigma_t = \frac{1}{\sqrt{2} B_r \sqrt{(S/N)_n}}$ $(S/N)_n \gg 1$	$\sigma_t = \frac{1}{1.785 B_r \sqrt{(S/N) (f_r/\beta_l)}}$ $B_r \tau = 1.4 \quad \tau_g = 1.4 \tau$	$\sigma_t = \frac{1}{1.4 B_r \sqrt{(S/N) (f_r/\beta_l)}}$
Maximum Target Velocity Which Can Be Tracked (M/Sec)	$ \dot{R}_{\max} = \frac{c^2 f_r}{4n}$	$ \dot{R}_{\max} = \frac{c^2}{2} \beta_l' \left[1.558 - \left(\frac{f_r}{\beta_l} \right) \frac{1}{n_a} \right]$ $(f_r/\beta_l') \frac{1}{n_a} = 0 \text{ for disabled sweep}$	$ \dot{R}_{\max} = \frac{c^2}{2} \beta_l' \left[1 + \frac{1}{B_r \tau} - \left(\frac{f_r}{\beta_l} \right) \frac{1}{n_a} \right]$ $(f_r/\beta_l') \frac{1}{n_a} = 0 \text{ for disabled sweep}$
Lag Error in Location of Received Pulse (Sec)	$\epsilon_t = \frac{1}{2} \tau$	$\epsilon_t = \frac{1}{\beta_l} \left[\frac{2\dot{R}}{c} - \frac{\tau f_r}{n_a} \right]$ $\tau f_r/n_a = 0 \text{ for disabled sweep}$	$\epsilon_t = \frac{1}{\beta_l'} \left[\frac{2\dot{R}}{c} - \frac{\tau f_r}{n_a} \right]$ $\tau f_r/n_a = 0 \text{ for disabled sweep}$
Probability of Loss of Track	$\sqrt{2\pi} P_{LT} =$ $\int_{-\infty}^{-(1.558\tau + \epsilon_t)/\sigma_t} \exp\left(-\frac{x^2}{2}\right) dx$ $+ \int_{-\infty}^{(1.558\tau - \epsilon_t)/\sigma_t} \exp\left(-\frac{x^2}{2}\right) dx$	$\sqrt{2\pi} P_{LT} =$ $\int_{-\infty}^{-(\tau + 1/B_r + \epsilon_t)/\sigma_t} \exp\left(-\frac{x^2}{2}\right) dx$ $+ \int_{-\infty}^{(\tau + 1/B_r - \epsilon_t)/\sigma_t} \exp\left(-\frac{x^2}{2}\right) dx$	

* See next page for definition of symbols.

Definition of Symbols for Table IV-1

- σ_t = RMS thermal noise in measured location or return pulse (sec)
- B_r = If. bandwidth of the receiver (Hz)
- $(S/N)_n$ = Effective integrated if. signal-to-noise (power) ratio for n pulses
- S/N = If. signal-to-noise power ratio for a single pulse
- f_r = Pulse repetition rate (sec^{-1})
- β_ℓ = Nominal tracking loop bandwidth (rad/sec)
- τ = Duration of transmitted pulse when no pulse compression is used; equivalent compressed pulse duration when pulse compression is used (sec)
- τ_g = Duration of tracking gate (sec)
- $|\dot{R}_{\max}|$ = Maximum trackable target relative velocity, assuming closing and separating targets equally likely (m/sec)
- c = Speed of light (m/sec)
- n = Number of pulses integrated
- β'_ℓ = Actual tracking loop bandwidth $\cong [S/N/(1 + S/N)] \beta_\ell$
- ϵ_t = Lag error in measured location of return pulse (sec)
- $\dot{R}$ = Target relative velocity in range (m/sec)
- n_a = Number of pulses exchanged with each range bin during range loop acquisition
- P_{LT} = Probability of loss of track

Table IV-2. Summary of Performance Equations of Angle Tracking Systems*

	Monopulse	Conical Scan	Conopulse
RMS Thermal Noise Error in Angle of Tracking Axis, Per Channel	$\sigma_\theta = \frac{\theta}{k_m \sqrt{(B_r \tau)(S/N)(f_r/\beta_l)}}$	$\sigma_\theta = \frac{1.4 \theta \sqrt[4]{c_d/c_a}}{k_c \sqrt{(B_r \tau)(S/N)(f_r/\beta_l) 1/L_c}}$	$\sigma_\theta = \frac{1.4 \theta \sqrt{c_d/c_a}}{k_p \sqrt{(B_r \tau)(S/N)(f_r/\beta_l) 1/L_p}}$
RMS Noise Due to Target Glint, Per Channel (Rad)	$\sigma_g = L_t/4R$	$\sigma_g = 0$	$\sigma_g = 0$
RMS Noise Due to Target Scintillation, Per Channel	$\sigma_s = 0$	$\sigma_s = \frac{0.27 \theta}{f_c} \sqrt{f_g \beta_l'}$	$\sigma_s = 0$
Crosstalk Error (Rad)	Crosstalk Error = 0	Crosstalk Error = 0	$\epsilon_{act} = \omega_p \epsilon_t \epsilon_e$ $\sigma_{act} = \omega_p \sigma_t \sigma_e$ $\epsilon_{ect} = \omega_p \epsilon_t \epsilon_a$ $\sigma_{ect} = \omega_p \sigma_t \sigma_a$
Dynamic Error (Rad)	$\epsilon_a = \omega_a/\beta_l'$ $\epsilon_e = \omega_e/\beta_l'$	$\epsilon_a = \omega_a/\beta_l'$ $\epsilon_e = \omega_e/\beta_l'$	$\epsilon_a = \omega_a/\beta_l'$ $\epsilon_e = \omega_e/\beta_l'$

* See next page for definition of symbols

Definition of Symbols for Table IV-2

σ_θ	RMS thermal noise error in tracking axis angle, per channel	} any consistent units
θ	Half-power beamwidth	
k_m, k_c, k_p	Error slope for monopulse, conical scan, and conopulse, respectively	
B_r	Receiver if. bandwidth (Hz)	
τ	Duration of transmitted pulse, if pulse compression is not used; equivalent duration of compressed pulse if pulse compression is used	
(S/N)	If. signal-to-noise (power) ratio for a single pulse for a target on the beam maximum	
f_r	Pulse repetition rate (sec ⁻¹)	
β_ℓ	Nominal tracking loop bandwidth (rad/sec)	
L_c, L_p	Off-beam center power loss factors for conical scan and conopulse scan, respectively	
σ_g	RMS glint error in tracking axis angle, per channel (rad)	
L_t	Target extent	} any consistent units
R	Target range	
σ_s	RMS scintillation error in tracking axis angle (units consistent with θ)	
f_c	Conical scan frequency (Hz)	
f_g	Scintillation frequency (Hz) = $2f_a L_t / \lambda$	
f_a	Rate of change of target aspect angle (Hz)	
λ	Wavelength (units consistent with L_t)	
β'_ℓ	Actual tracking loop bandwidth $\cong [S/N / (1 + S/N)] \beta_\ell$ (rad/sec)	
ϵ_{act}	Lag crosstalk error in azimuth channel angle due to crosstalk (rad)	
ω_p	Conopulse angular scan frequency (rad/sec)	
ϵ_t	Lag error of range tracking loop (sec)	
ϵ_e	Lag Error in elevation channel angle due to target velocity (rad)	

Definition of Symbols for Table IV-2 (Continued)

σ_{act}	= RMS noise error in azimuth channel angle due to crosstalk (rad)
σ_t	= RMS noise error in range tracking loop (sec)
σ_e	= RMS thermal noise error in elevation channel angle (rad)
ω_a	= Azimuth component of target relative angular velocity (rad/sec)
C_d	= Detector loss factor
C_a	= AGC loss factor

Table IV-3. CAT Radar Parameters

Frequency	$f = 3.0 \text{ GHz}$
Number of modules	$N_m = 1000$
Peak power per module	$P_{tm} = 8.3 \text{ W}$
Transmitted pulse duration	$\tau = 20 \text{ } \mu\text{seconds}$
Noise figure	$F_n = 7.3 \text{ dB}$
Target cross section	$\sigma = 50 \text{ m}^2$
Fixed system losses	$L_s = 7 \text{ dB}$

associated with integration of 8 pulses is approximately 1 dB. Thus, for an ideal system with a matched if. filter, the effective integrated signal-to-noise ratio is given by

$$(S/N)_n = \frac{3.26 \times 10^8 \times 8}{1.11 R^4} = \frac{2.37 \times 10^9}{R^4} . \quad (4-17)$$

For a nonmatched ideal system, Figure 4-3 indicates that there is a filter mismatch loss which is the reciprocal of the $B_r \tau$ product. For an ideal non-matched system, Equation (4-17) becomes

$$(S/N)_n = \frac{2.37 \times 10^9}{B_r \tau R^4} , \quad (4-18)$$

which can be used to determine the detection performance of the radar. Figure 4-7 is a plot of $(S/N)_n$ of Equation (4-18) as a function of target range. For probability of detection of 0.9 and probability of false alarm of 10^{-9} , Figure 4-1 shows a signal-to-noise ratio of about 15 dB. The maximum detection ranges for the three cases shown are approximately 94 nm, 78 nm, and 62 nm.

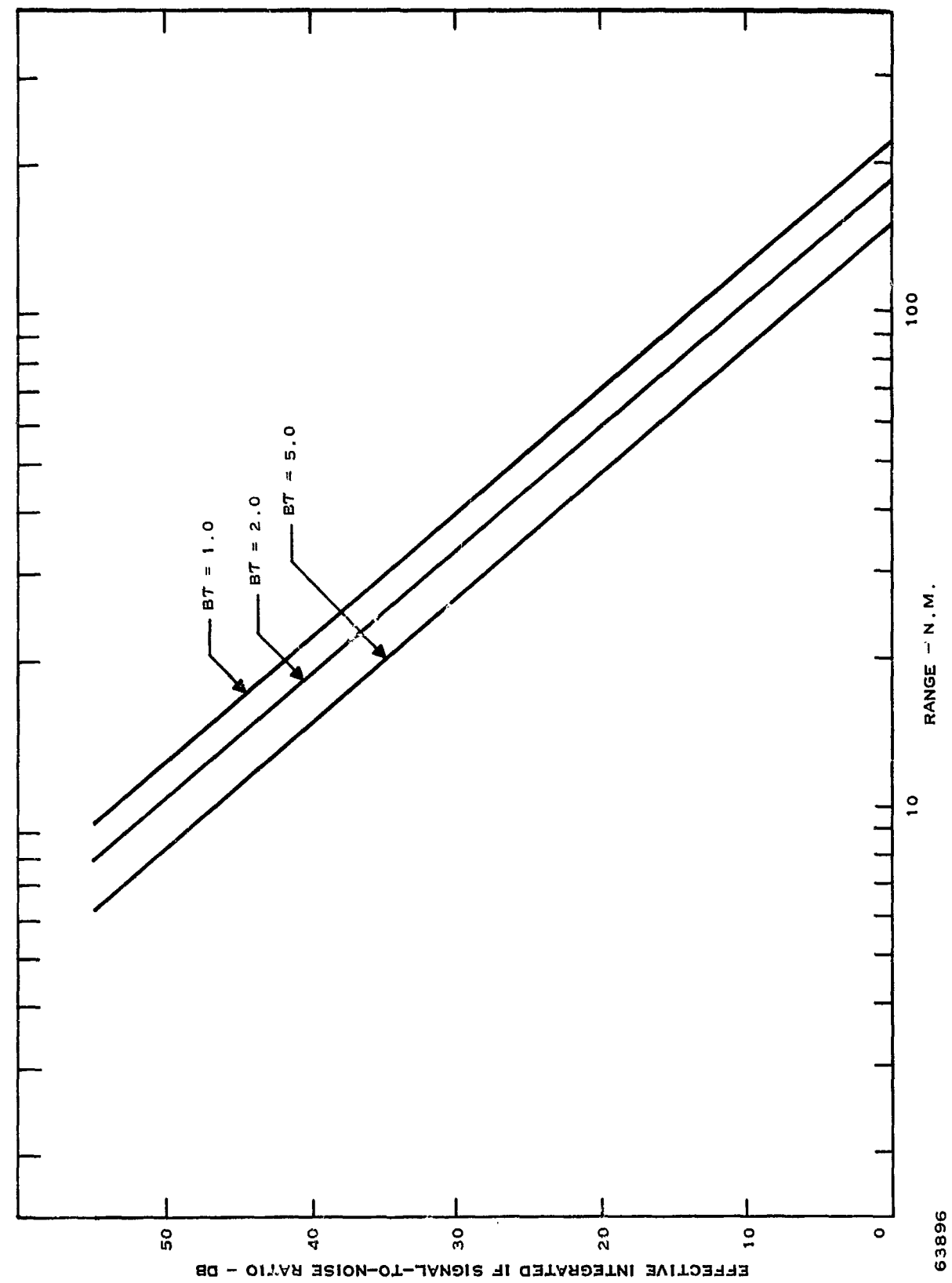


Figure 4-7. Effective Integrated If. Signal-to-Noise Ratio for System Parameters Defined in Table IV-3; Eight Pulses Integrated

63896

Figure 4-8 is a plot of single-pulse signal-to-noise ratio of Equation (4-16) versus target range, when the only contribution of the loss term L_{ad} comes from filter mismatch loss.

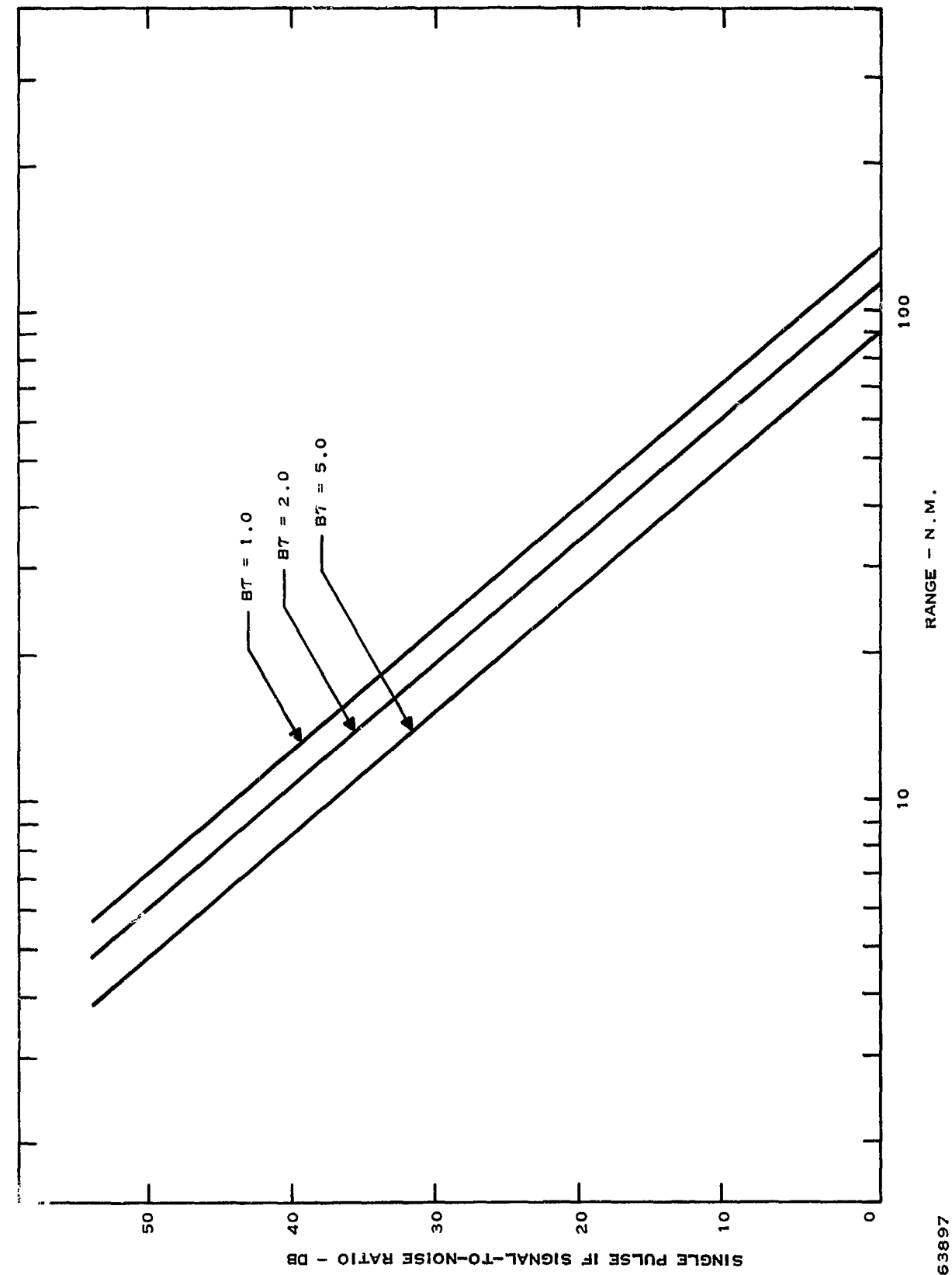
From the curves of Figure 4-7 and Figure 4-8, the rms thermal noise range error is determined for the range tracking loops of Table IV-1. Figure 4-9 is a plot of rms error in the open-loop measurement for various $B_r\tau$ products. These curves are valid for a pulse compression system if the ordinate scale is divided by the pulse compression ratio. Figure 4-10 is a plot of rms error in the split gate tracker for the optimum $B_r\tau$ product of 1.4, for various values of the ratio f_r/β_l . Figure 4-11 is a plot of the rms error in the separated gate tracker for various values of B_r and various values of f_r/β_l . A comparison of thermal noise error of the three systems for equal $B_r\tau$ products (which correspond to equal detection ranges) and for the factor f_r/β_l of the closed-loop systems equal to the number of pulses integrated in the open-loop system leads to $f_r/\beta_l = n = 8$. Select $B_r\tau = 1.4$, which corresponds to the optimum split gate tracker. The results are shown in Figure 4-12.

Figure 4-13 gives a comparison of the maximum trackable target velocity for the three systems. For the open-loop system, 8 pulses are integrated. For the closed-loop systems, $B_r\tau = 1.4$. It is assumed that the sweep is disabled during tracking. Again, the results are applicable to a pulse compression system if the velocity scale is divided by the pulse compression ratio.

Figure 4-14 gives a comparison of the lag errors due to target velocity (assuming that the target velocity does not exceed the maximum trackable velocity) for the three range tracking techniques. Table IV-1 shows that both closed-loop systems have the same lag error. In Figure 4-14 the tracking loop sweep is assumed to be disabled during track.

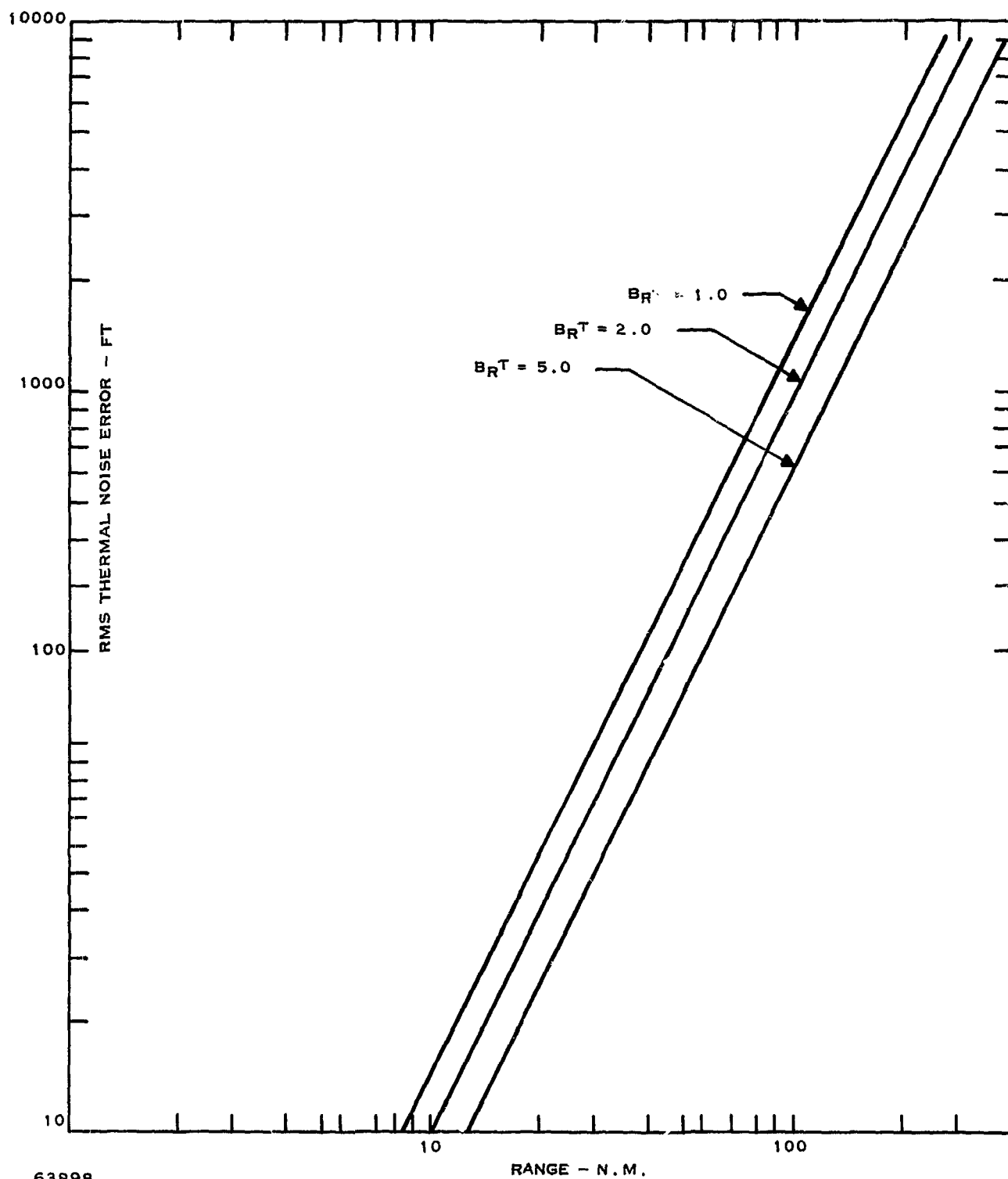
The preceding set of performance curves for the various range tracking systems, in conjunction with specification of accuracy and target velocity requirements for a particular mission, permit a rational choice of range tracking techniques to be used. Next, similar performance curves are derived for the three types of angle trackers considered, based on the guideline system parameters of Table IV-3.

Figure 4-15 presents a comparison of the rms thermal noise error for the three types of angle tracking loops. Angular error slopes of 1.57, 1.5, and 1.5 were assumed for monopulse, conical scan, and conopulse, respectively. Note that for monopulse and conical scan, the $B_r\tau$ product which appears in the thermal noise expressions will be approximately unity, even if pulse compression is used, for the angular error information is derived from the compressed pulses. For conopulse, error information is derived from the wide-bandwidth uncompressed pulses. Therefore, $B_r\tau$ for the conopulse system will be approximately equal to the pulse compression ratio. $B_r\tau = 100$ was assumed for the curves of Figure 4-15.



63897

Figure 4-8. Single Pulse IF. Signal-to-Noise Ratio for System Parameters Defined in Table IV-3



63898

Figure 4-9. RMS Thermal Noise Error in Open-Loop Range Measurement for System Parameters Defined in Table IV-3. Eight Pulses Integrated

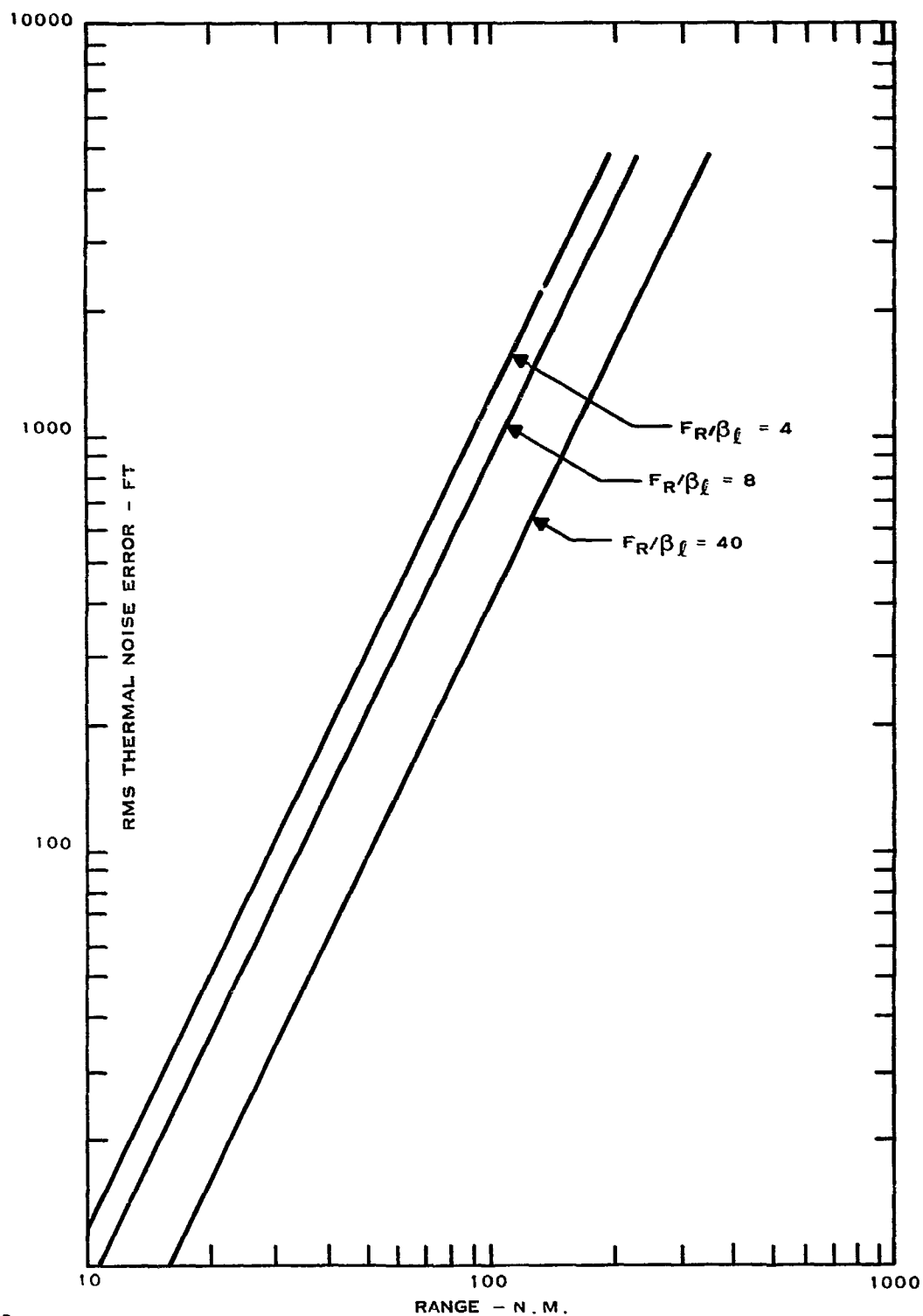


Figure 4-10. RMS Thermal Noise Error in Split Gate Range Tracker for System Parameters Defined in Table IV-3

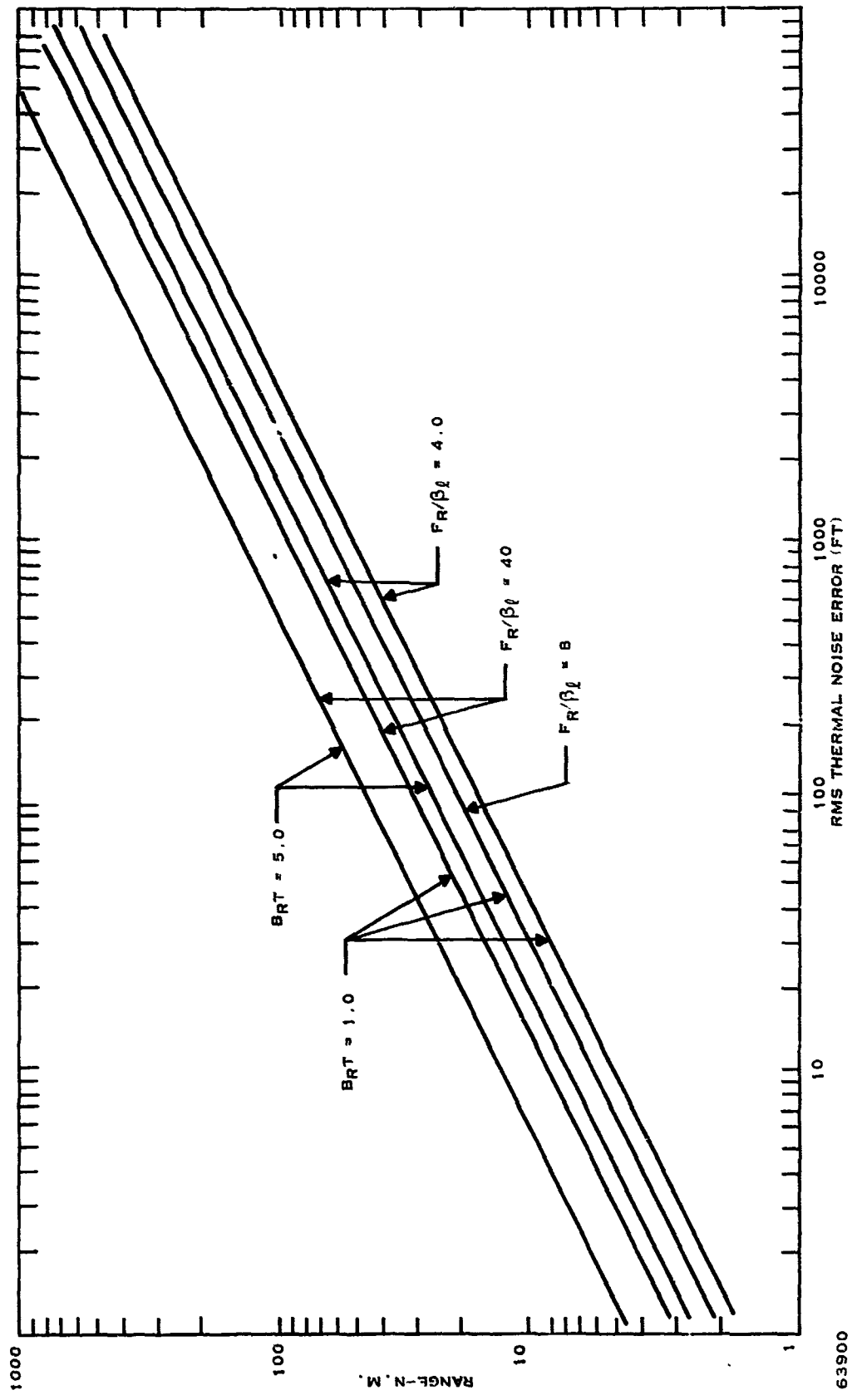


Figure 4-11. RMS Thermal Noise Error in Separated Gate Tracker for System Parameters Defined in Table IV-3

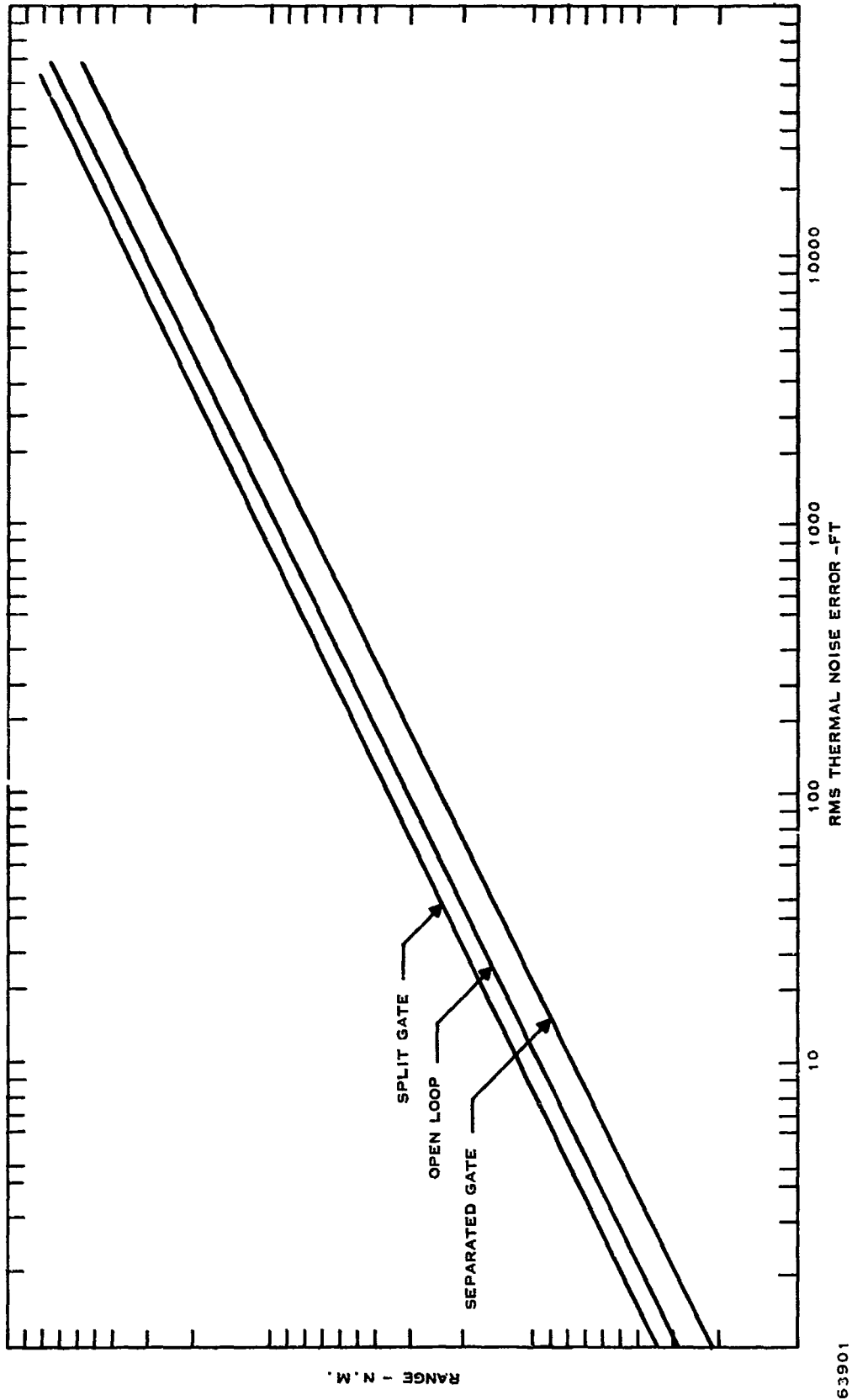


Figure 4-12. Comparison of Thermal Noise Response of Range Trackers for System Parameters Defined in Table IV-3

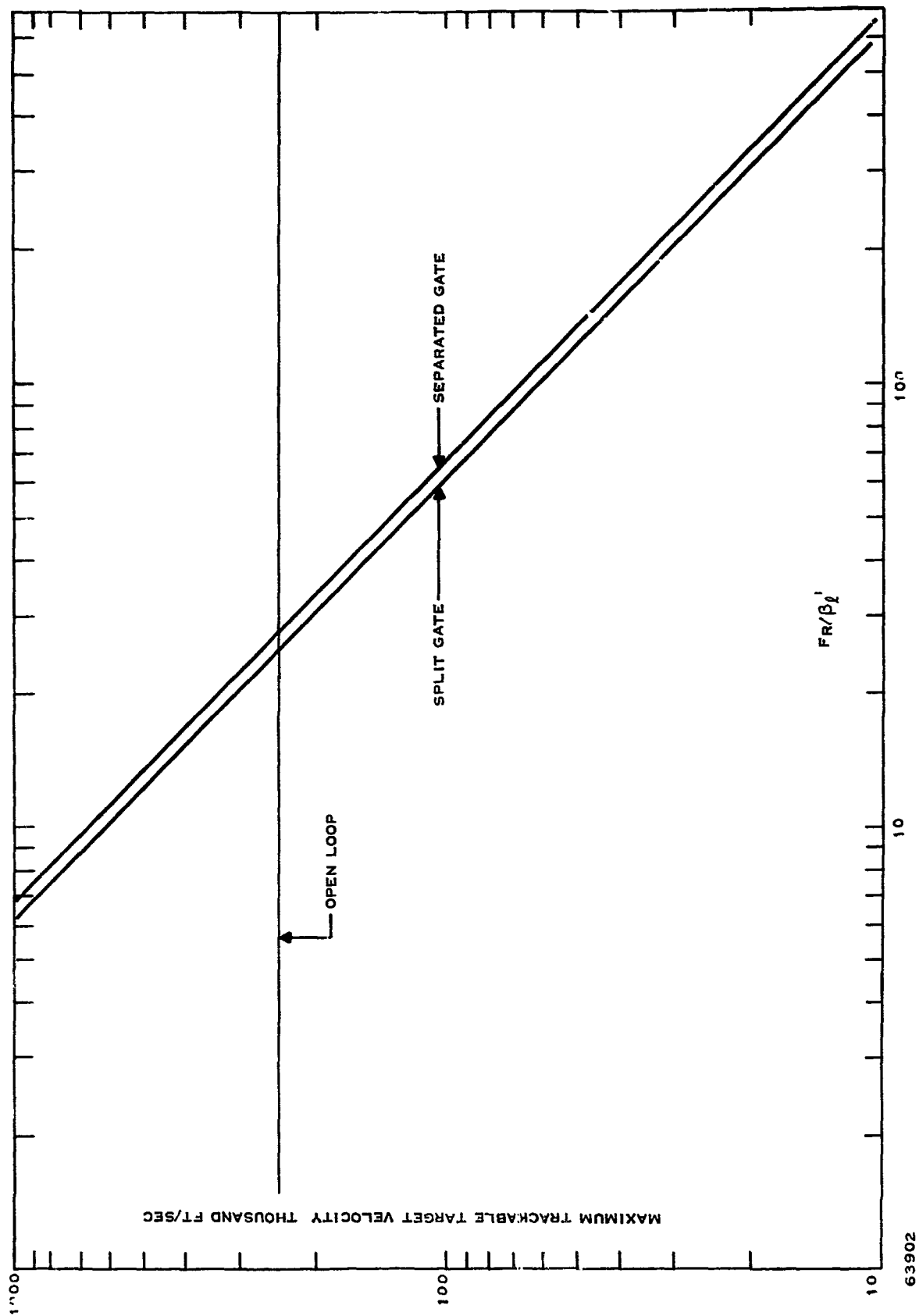
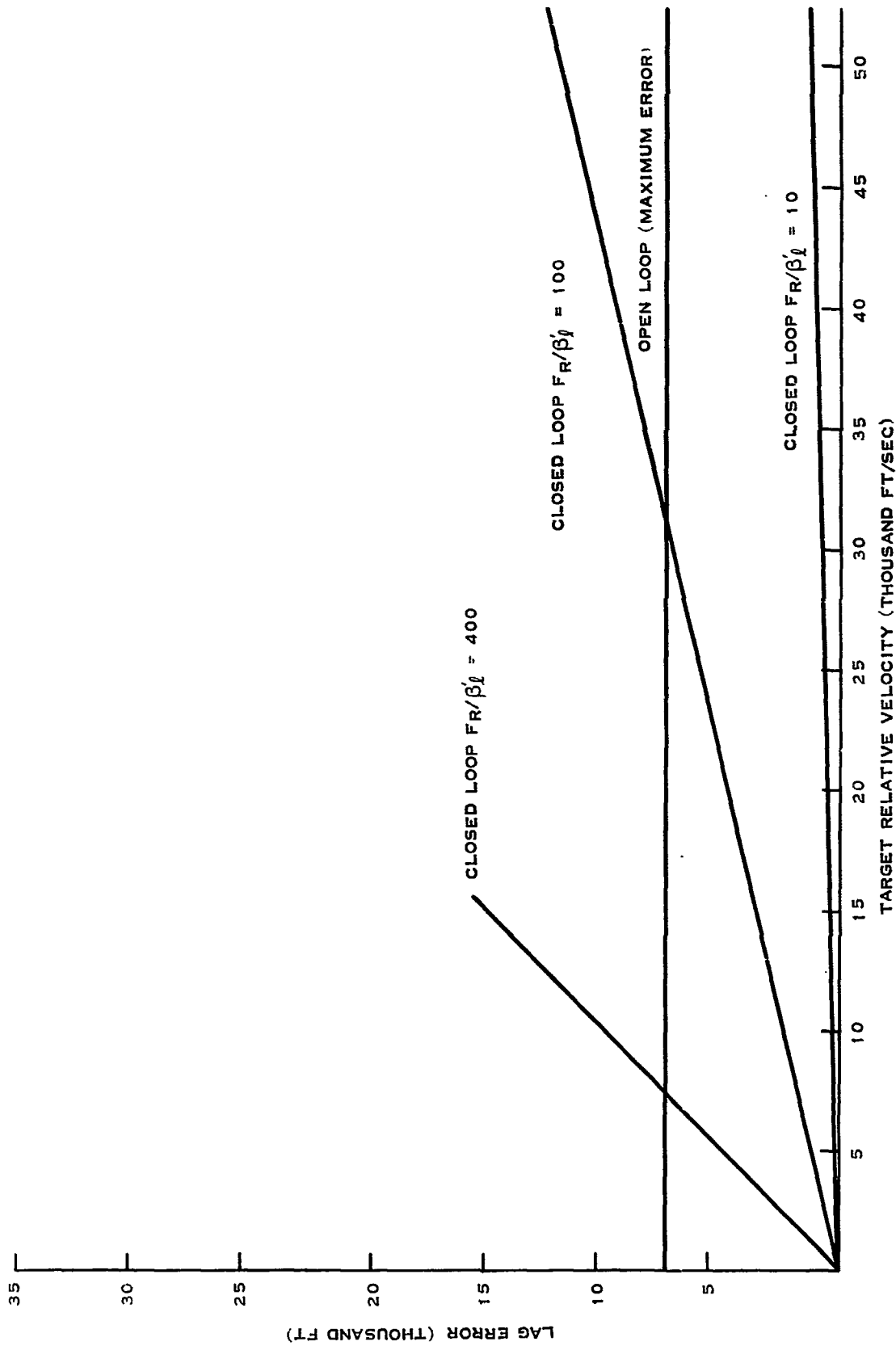


Figure 4-13. Comparison of Maximum Trackable Target Velocities for Range Trackers
with System Parameters Defined in Table IV-3



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Figure 4-14. Comparison of Lag Errors of Range Trackers for System Parameters Defined in Table IV-3

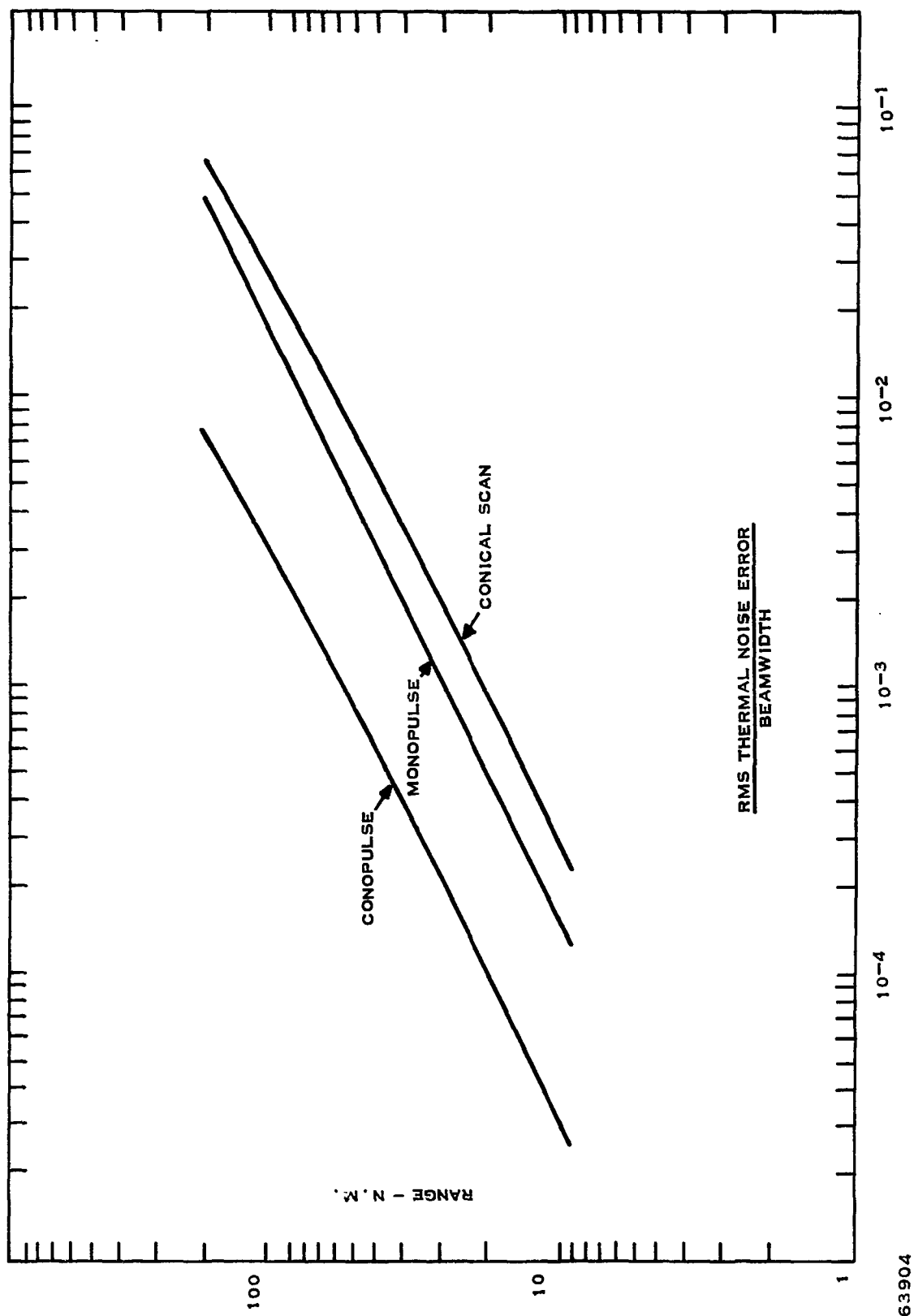


Figure 4-15. RMS Thermal Noise Error in Angle Tracking Loops for System Parameters Defined in Table IV-3

Only the conical-scan system, of the three, suffers degradation due to scintillation. Assuming $f_g = 5$ Hz, $f_c = 30$ Hz, and $\beta'_l = 8$ Hz, a typical value for σ_s can be calculated to be $\sigma_s = 0.0570$.

Only the conopulse system suffers degradation due to crosstalk. For example, assume the thermal noise error in each of the azimuth and elevation channels is 0.33 mils, this corresponds to a range of 50 nm in Figure 4-15 for a 3-degree beam. Further assume the range-tracking loop thermal noise at 50 nm is 2.2 feet, which is obtained from Figure 4-12 assuming 100:1 pulse compression. Thus, the rms jitter in the range-gate position, σ_t , is 4.46×10^{-9} . If the conopulse frequency is 1 MHz, the expression of Table IV-2 gives the rms crosstalk per channel as $\sigma_{ct} = (6.28 \times 10^6) (4.46 \times 10^{-9}) (0.33 \text{ mils}) = 0.00924$ mils or 2.8 percent.

This discussion of the relative performance of the various angle trackers, together with the previous similar discussion for range trackers, forms a basis for selecting a combination of these subsystems in order to synthesize an overall system which will satisfy particular mission objectives.

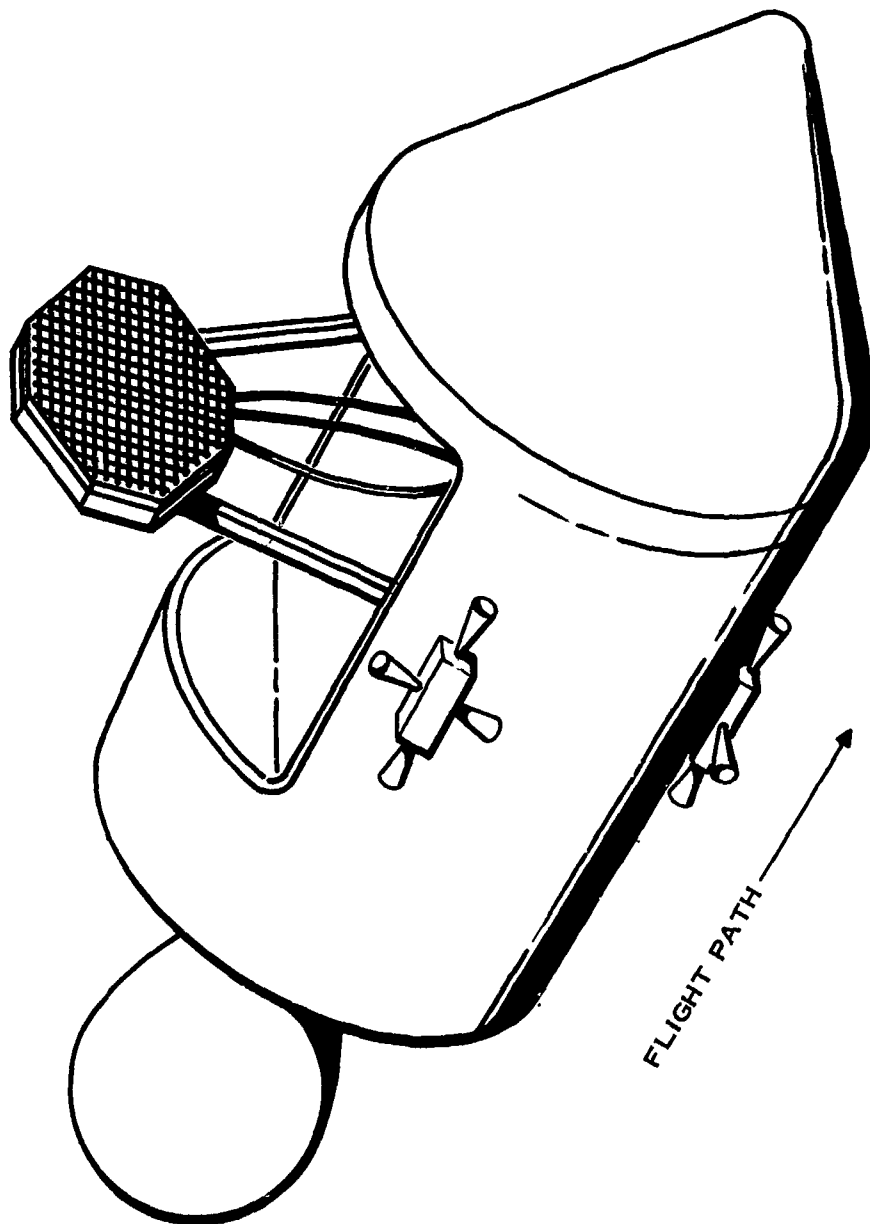
5.0 VEHICLE INTERFACE

The vehicle interface that is important to the operation of the CAT radar includes the installation or mounting of the radar on the spacecraft, power conditioning necessary to satisfy the requirement of the CAT radar system, command and control of the radar and data display.

In order to consider the broadest set of possible radar configurations, a study was made to determine the size of the largest aperture that would be practical to mount on the Apollo Service Module with the information that was available through general familiarization reports on the Apollo Spacecraft. This aperture was estimated to be a six-foot octagon with an area of approximately 31.5 square feet. Figures 5-1 and 5-2 illustrate two typical mounting positions for the CAT radar on the service module. Figure 5-1 shows the active-element corporate-fed array pivoted into place where it can electronically scan a 45-degree cone without interference from the command module. The array would be stored, folded back into the service module, when not in use. A similar configuration is shown in Figure 5-3 except a space-fed array with a conventional transmitter tube is substituted for the active-type array.

Figure 5-2 illustrates a conformal array which would require yawing the spacecraft 90 degrees in order to search the area in the flight path as defined by the mission requirements. The active-element corporate-fed array and the space-fed array could also be mounted in the same location as the conformal array shown in Figure 5-2. A protective cover that protects the array during launch into orbit would be jettisoned before radar operation. The problems introduced by yawing the spacecraft 90 degrees are significant if it is necessary to change the orbit altitude at frequent intervals. The mechanical simplicity of the array may make this approach attractive if the experiments with the CAT radar can be run without making frequent changes in the orbit.

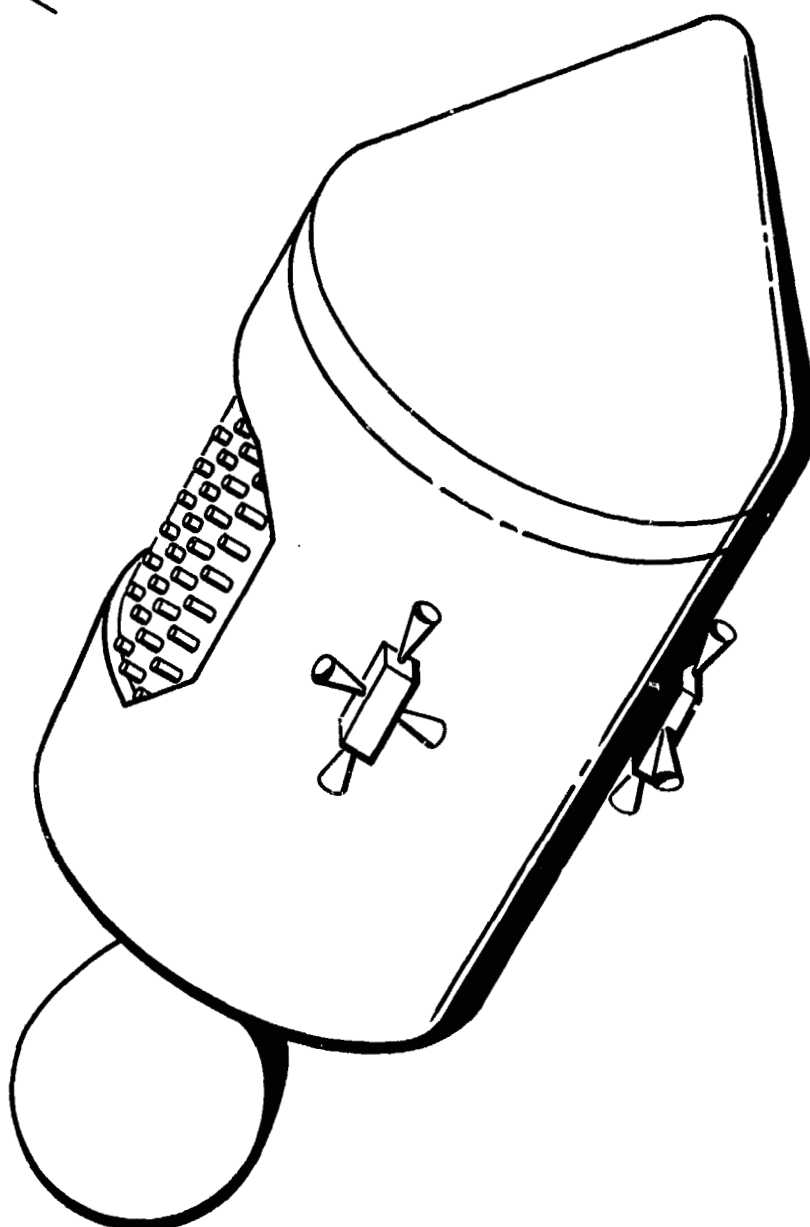
The folding array concept offers the greatest flexibility since it does not cause restrictions to be made on the spacecraft during radar operation. The array would be covered and reinforced by a shroud during lift-off and flight into orbit. The array cover would be jettisoned after entering into orbit, however, at the lower altitudes it probably would not be desirable to erect the array into operating position until just before performing radar experiments or otherwise utilizing the radar system. The support structure for the array would utilize high-strength alloy materials in conjunction with honeycomb construction in order to maintain a high strength-to-weight ratio consistent with good design for highest reliability and minimum weight. The CAT radar mechanism for positioning the array in its extended position must accurately align the array boresight with the spacecraft axis.



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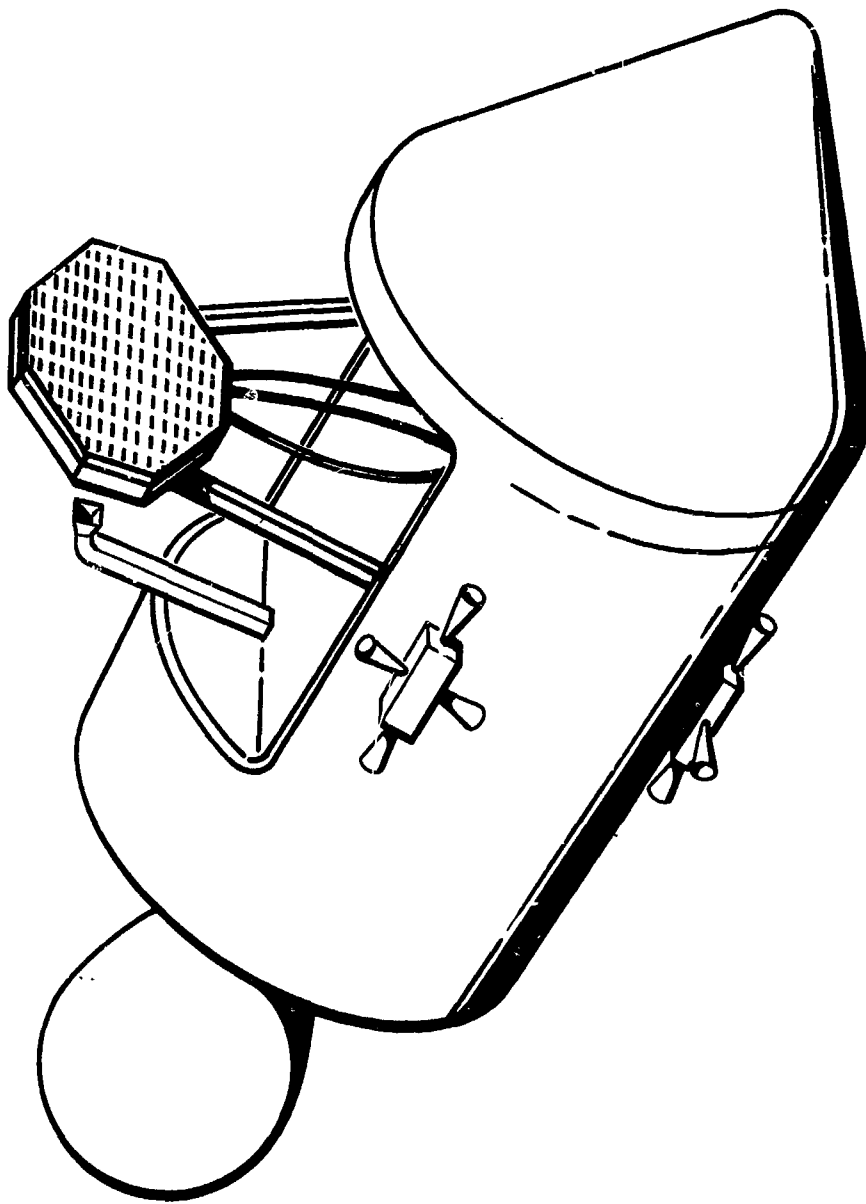
Figure 5-1. MERA Type Corporate-Fed Array

FLIGHT
PATH



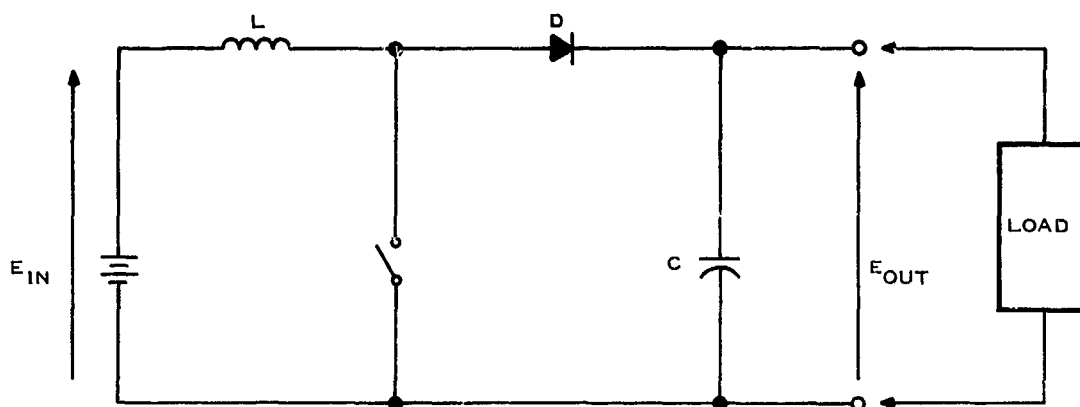
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Figure 5-2. Conformal Array



63907

Figure 5-3. Electronically Scanned, Rear-Fed
Array With Tube-Type Transmitter



$$\text{APPROXIMATE: } E_{OUT} = E_{IN} \left[\frac{F T_{ON}}{1 - F T_{ON}} + 1 \right]$$

$$\text{EXACT: } E_{OUT} = \left[\frac{T_{ON}}{T_{OFF}} + 1 \right] \left[E_1 - L_{IN} (R_O + R_C) \right] + E_D - \left[\frac{R_{CS} L_{IN} T_{ON}}{T_{OFF}} \right]$$

63908

Figure 5-4. TRC Basic Power Circuit

There is a variety of mechanism types that could be adapted for extending the array into position. Expandable trusses, gear motors, recirculating ball screws, and a variety of linkage combinations are a few of the concepts that should be considered for positioning the array. The mechanism could be powered by either electric motors, gas motors, hydraulic motors, gas cylinders or hydraulic cylinders. A detailed analysis of the different design approaches for folding out the array is not considered within the scope of the CAT radar study.

The interface of the CAT radar with the space vehicle will include the generation and regulation of the required voltages for the radar from the space vehicle primary power supply which is assumed to be 25 to 31 Vdc.

The corporate-fed array for the CAT radar which utilizes all solid-state rf modules requires the following power conditioning:

1. 40 Vdc $\pm$ 1-percent regulation
2. 100 Watts to 10 kW depending on required range and mission definition to be determined at final tradeoff study.

The 40 Vdc is required for the optimization of the semiconductor circuits used in the rf amplifiers. The step-up in voltage can be achieved with efficiencies of approximately 90 percent, by using a relatively new circuit referred to as a time ratio control (TRC)⁸⁰ shown in Figure 5-4. The circuit

consists of an inductor, diode, capacitor, and a transistor switch or SCR which is used only in its switching capacity and not as a modulator. Therefore, if lossless components could be obtained, the circuit would be 100-percent efficient. The output voltage is the sum of an inductor voltage plus the input voltage. Therefore, the inductor is not required to store the full output energy.

The equations for the output voltage are also shown in this figure. For a constant switch "on" time, the output voltage is a function of frequency. Also the voltage may be a function of duty cycle or time "on" if the frequency is held constant. The efficiency of the TRC circuit is plotted on Figure 5-5. The weight and volume of the circuit is plotted on Figure 5-6.

The voltage output will have to be further regulated in order to respond to the fast switching of relatively heavy currents in the antenna array. The overall power conditioning efficiency should approach 80 percent.

The possibility of requiring a larger amount of power for the CAT radar for short periods of time, depending on the definition of the mission and the radar range required, makes it practical to consider the use of secondary rechargeable batteries.⁸¹ Space batteries may be utilized to great advantage in many applications and should be considered for CAT radar trade-off study. Figures 5-7 and 5-8 show a plot of the power density and power-volume relationships for typical batteries used in space applications.

A detailed analysis of the mission requirements would indicate the type of power required. Also should a space-fed array be used, the power conditioning requirements would be more complex since more voltages would be required.

Command and control of the CAT radar could be located in the Command Module so the astronauts could select either manual or automatic control of the system. The command and control would also interface with other Apollo Systems like the guidance computer, and stabilization control systems. Other control functions could be incorporated into the CAT radar as required to achieve the desired mission results.

The CAT radar will be expected to produce target data for use by the guidance computer and the astronaut.

Ideally, the data displayed would include the target range, range rate, azimuth, elevation, and possibly cross section. The astronaut would have full knowledge of the target parameters as the target is acquired and tracked.

The first question, which must be answered in regard to the functions of data display, concerns the astronaut's use of the target information displayed for him. In space situations, where the target is approaching near

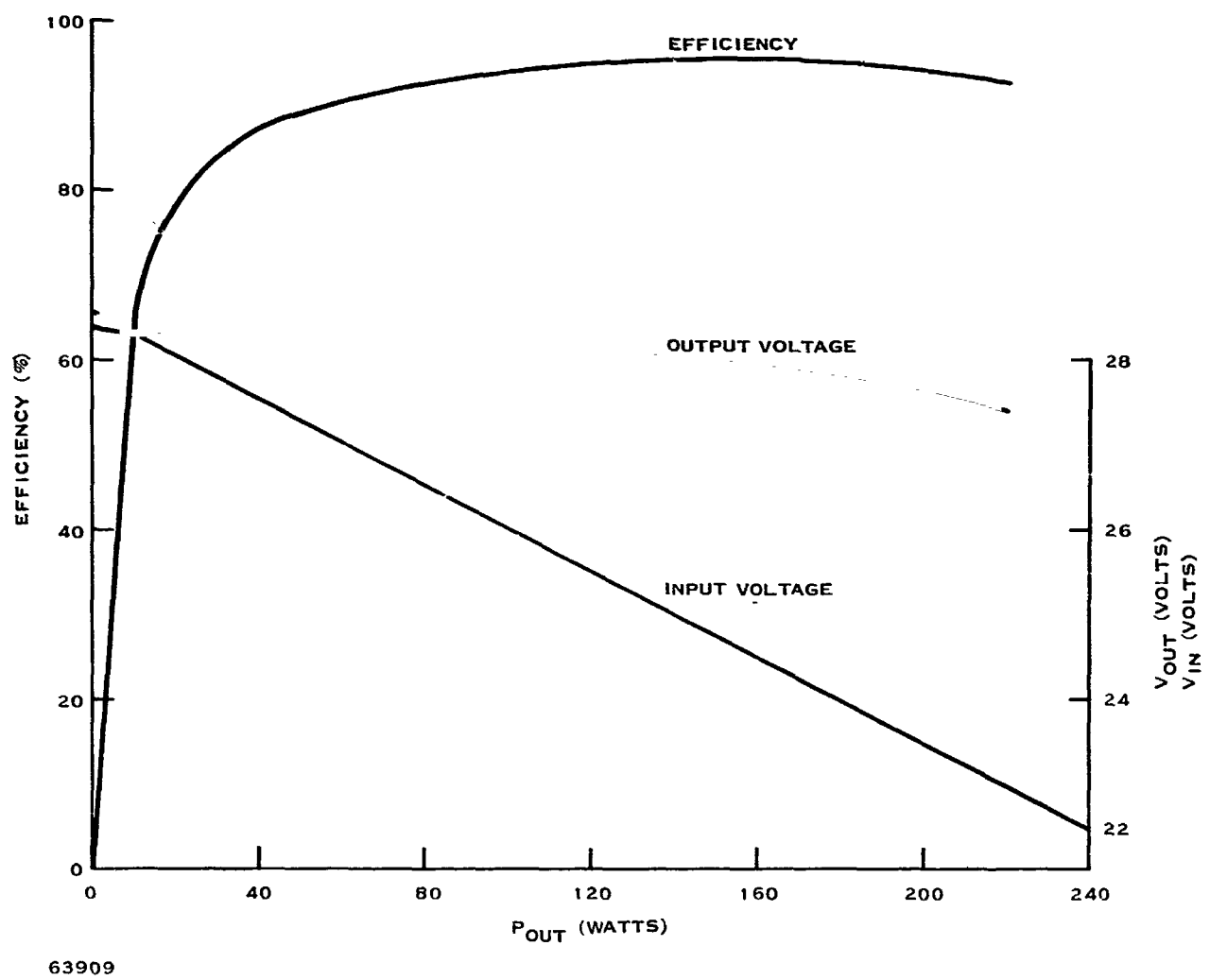


Figure 5-5. Efficiency and Regulation Time Ratio Control

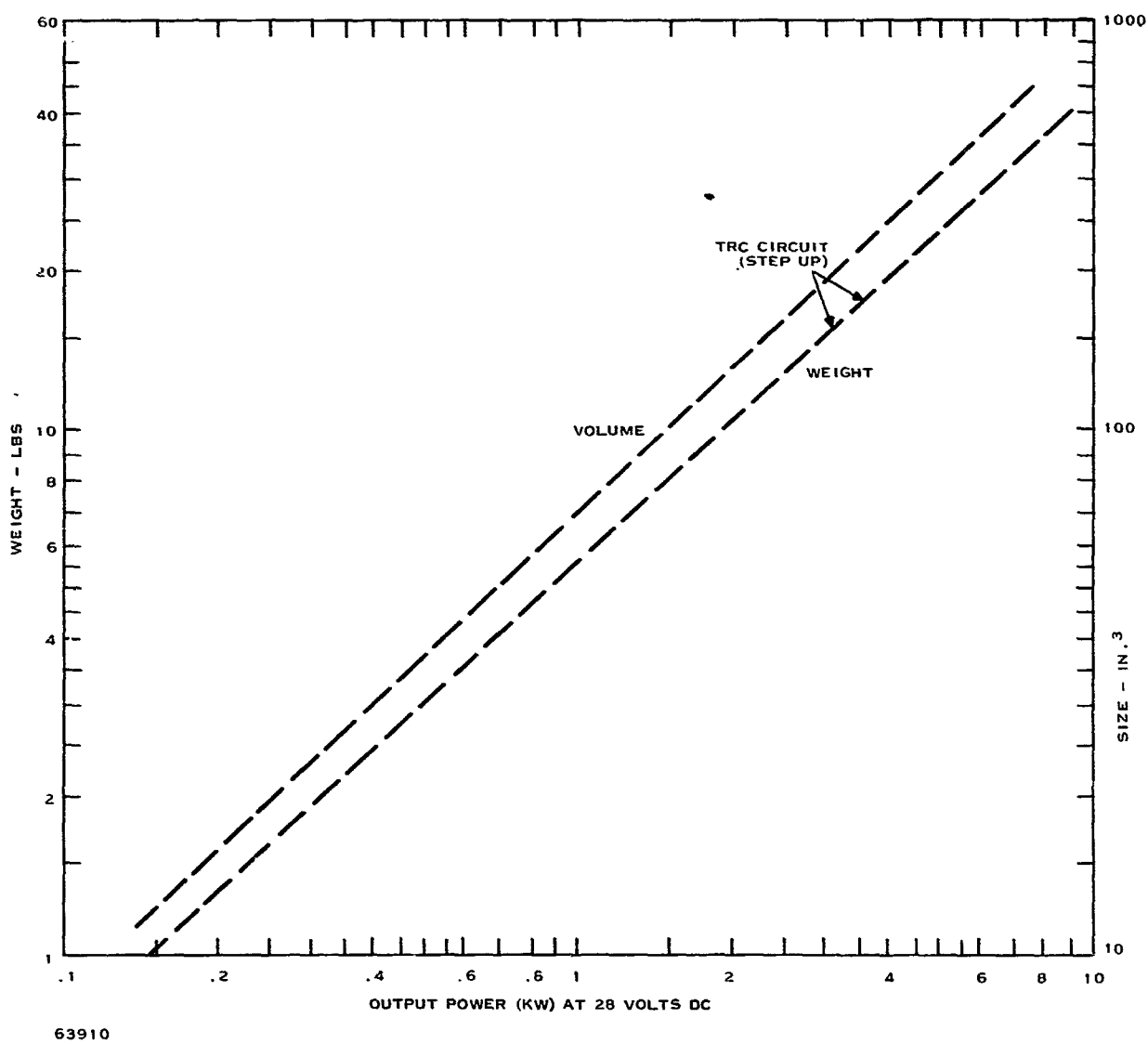


Figure 5-6. TRC Circuit at 50-Percent Input Voltage Regulation

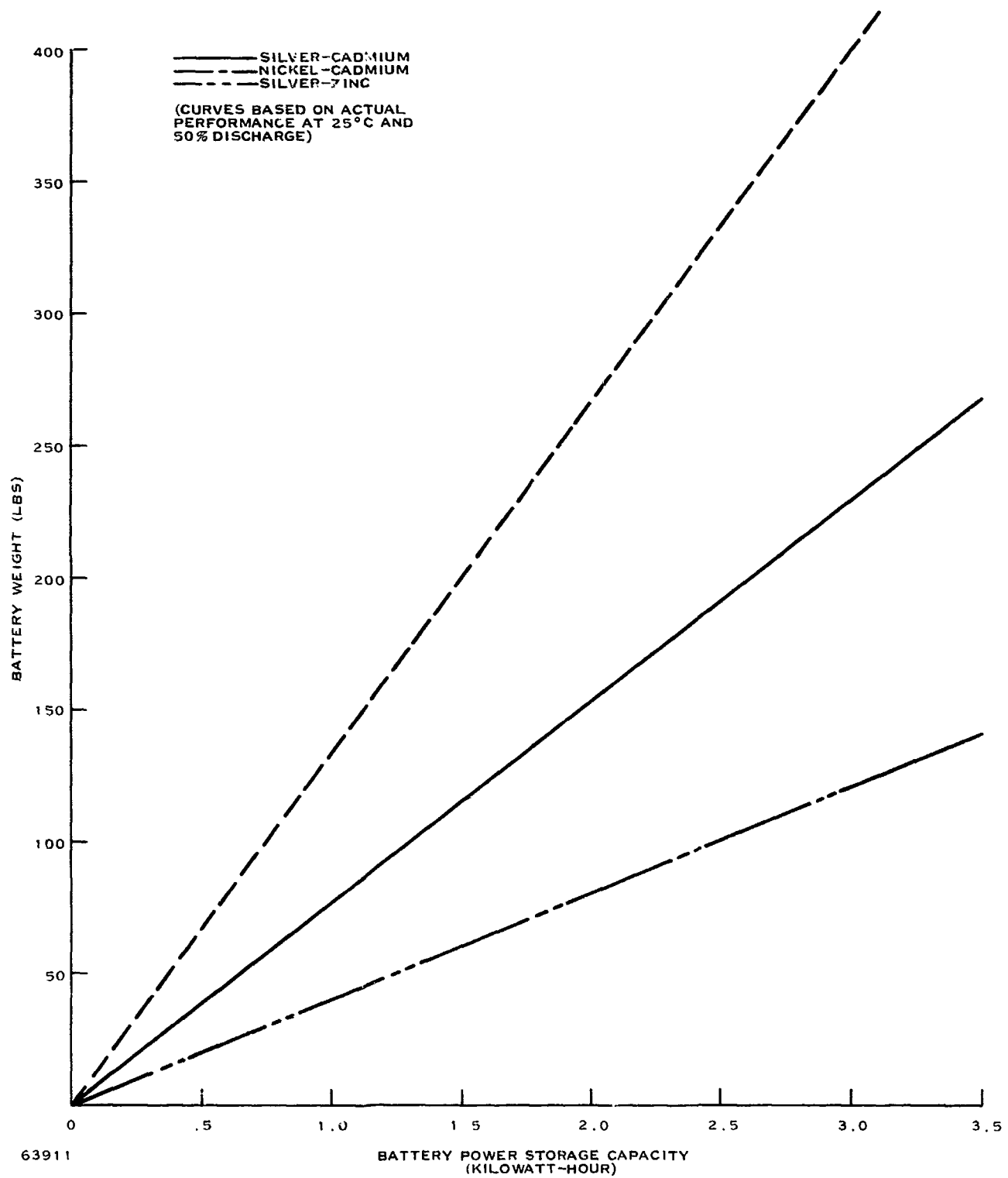
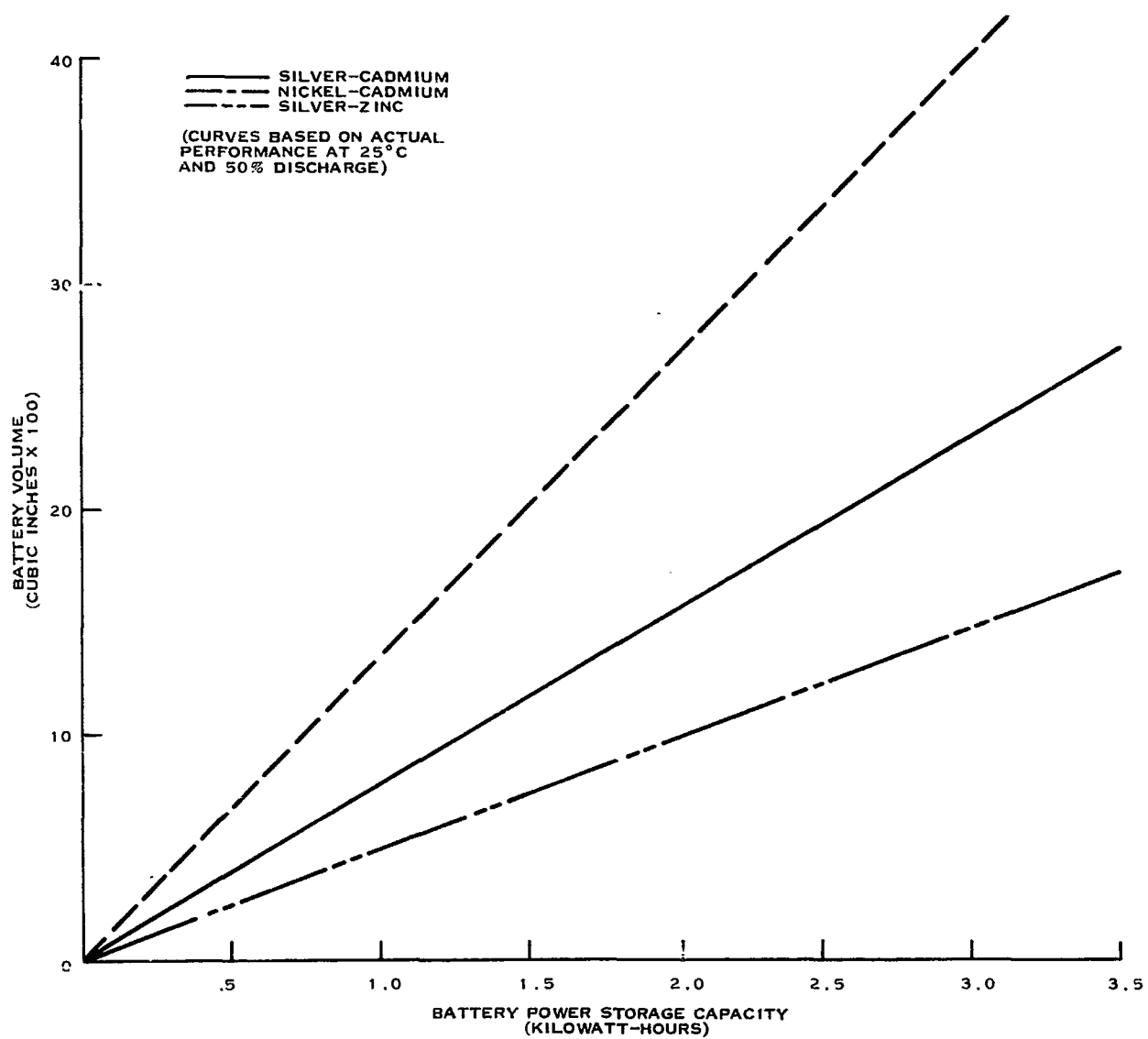


Figure 5-7. Battery Power Storage Capacity (Kilowatt-Hour)



63912

Figure 5-8. Battery Power Storage Capacity (Kilowatt-Hour)

right angles to the radar flight path, the target data is of scientific use only. The time the target is within the field of view is too short for the astronaut to perform a significant task. For example, a radar having ± 45 -degree field of view and 100 nm range would have a 6 nm-per-second target within its field of view less than 20 seconds, which is not sufficient time to permit interaction with the astronaut, considering that the range and angles undergo large changes during this time. One may reach the conclusion that real-time display of target data for the astronaut is not required for noncoplanar orbits.

For coplanar (or near coplanar) orbits the target can spend considerable time within the radar field of view and conceivably the astronaut may want to take some course of action depending upon the parameters of the target being tracked. As an example, consider a rendezvous situation. If the guidance computer is functioning properly, it most likely would be expected to exert the commands required to navigate the craft to the target. In the event of guidance computer failure there may be a need to complete the rendezvous under manual control. In this case it would be essential for the astronaut to have access to some target parameters via a data display. Consequently, we will assume that a requirement exists to display some minimum set of target parameters for the astronaut.

5.1 Types of Radar Data Displays

There are at least five devices which are potentially applicable to the problem of displaying radar data to the astronaut:

- Galvanometers
- Servo-driven tapes
- Cathode-ray tubes
- Light emitting diode arrays
- Electroluminescent displays.

5.1.1 Galvanometers

The galvanometer type of indicator requires an analog signal level which is used to physically deflect an indicator needle across a fixed scale. The scale position indicates the magnitude of the quantity being observed. This approach was successfully used in the Gemini rendezvous radar system.

5.1.2 Servo-Driven Tapes

This type of indicator is similar to the galvanometer except that the needle is not fixed and the background graduations are printed on a roll of tape which is driven by a servo until the proper graduation falls under the

needle. The servomotor represents additional complexity over the simple galvanometer, but in exchange offers higher resolution since the tape can be several feet in length.

5.1.3 Cathode-Ray Tubes

The most common type of display for search and tracking radars is the cathode-ray tube. The electron beam is deflected to indicate some combination of the parameters range, azimuth, elevation, and amplitude. The CRT is probably the most versatile radar display device but is not a strong candidate for manned space applications because of high voltage, vacuum, and space limitations.

5.1.4 Light Emitting Diode Arrays

A step in the direction of the CRT, but without some of the hazards, is the array of light emitting diodes. GaAs and GaP diodes exist which emit light in the visible spectrum when biased properly. Assuming that a large array of such diodes could be fabricated, and that each diode occupied a relatively small area, data could be displayed in a manner much like the CRT, where the spot would be selected digitally and biased for light emission. The data format could be identical to that of a CRT display, or additional items could be displayed since the diode array would have greater flexibility.

The diode array requires considerable digital circuitry for gating the diodes on and off. In fact, for a large array approaching the size of a CRT the amount of digital logic required becomes prohibitive to the present state-of-the-art.

5.1.5 Electroluminescent Displays

The last ten years have seen the development of the electroluminescent (EL) phenomena to the point that EL displays are beginning to be used in special applications. These applications have taken a number of forms. One method of employing the EL phenomena is in the thermometer or bar graph type of display. A segmented EL lampon the edge of a glass substrate constitutes the display. Analog or binary inputs are scaled to the binary representation of the number of segments in the entire bar. The instantaneous value of the input data then is decoded so that the display shows a lighted column indicating that value.

A second method of employing EL phenomena is area lighting of panels which represent GO/NO-GO quantities or alphanumeric displays. An example of this is an array of specially shaped panels which can form any letter of the alphabet or numeral depending on which subsections of the array are activated.

A third application of EL phenomena is the crossed-grid display which can function similarly to the CRT. Narrow parallel electrode strips are arranged at right angles to each other on opposite sides of an inner layer of EL material and nonlinear resistive material. When horizontal and vertical grids are activated sufficient voltage occurs across the EL material at the intersection point so that the cell emits light. Scan circuitry can be used to move the illuminating point around as in a CRT. Claims have been made for EL displays in spacecraft applications that EL displays offer advantages over competing methods in the areas of size, weight, power consumption, and reliability.

5.2 CAT Radar Display Requirements

Considerable insight into the problems related to displaying the proper data to the astronaut for a satisfactory orbital rendezvous can be gained from the Holthaus and Spink article.⁸² They describe five approaches to the solution of this problem. Under simulated conditions where an analog computer provided a real-time test interface with the astronaut, rendezvous was attempted from an initial range separation of 100,000 feet and a closing velocity of 500 feet-per-second with the objective being to alter the chase vehicle's velocity in successive steps until the range and range rate reached zero simultaneously.

The result of this experimental investigation indicated that the best configuration was that used in the Gemini vehicle for its rendezvous missions. The Gemini display consisted of two concentric galvanometer pointers indicating range and range rate. The scales for the two pointers were aligned opposite each other so that the proper rendezvous trajectory required keeping the pointers coincident while they each decreased to zero.

At this time it seems likely that this instrument is capable of handling the display requirement for the rendezvous aspects of the CAT radar system.

Other aspects of the CAT radar mission are necessarily general at this point, so the data display must be one which is sufficiently flexible that it can adapt to the various applications of the CAT radar with a minimum of change. Considering the undesirability of CRT's for manned-space applications and the state of development of large arrays of point sources of light, it is recommended that the additional features of the CAT radar display consist of alphanumeric registers similar to those of the Apollo guidance computer on the main display console. Space permitting there should be registers for range, azimuth, elevation, and range rate. If space limitations are severe the registers should have the flexibility of displaying any of the target parameters on manual selection.

6.0 MECHANICAL ANALYSIS AND DESIGN

The configuration of the antenna is a deployable array which could be up to six feet in diameter. This array will have to be strongly supported and cushioned in order to withstand the stresses imposed at lift-off. The array will be protected by a cover which secures it to the service module frame. The cover would be jettisoned before deploying the array into operating position. Another factor to be considered is the maintainance of boresight accuracy following deployment.

The supporting mechanism must be designed for maximum rigidity and minimum weight by utilizing high strength to weight ratio alloys.

Another consideration is the drag imposed by the array to the spacecraft. Let us analyze a six-foot octagon-shaped array mounted on the service module as shown in Figure 6-1. The drag and the resulting moment it exerts on the spacecraft is analyzed as follows:

LOW ORBIT ALTITUDE=100 nm (Minimum perigee)

∴ $V_1 = 25,567$ ft/sec (Velocity)

$E_1 = 1.52 \times 10^{-12}$ gm/cm³ (Atmospheric density)

HIGH ORBIT ALTITUDE=1469 nm

∴ $V_2 = 21,713$ ft/sec (Velocity)

$E_2 = 6.0 \times 10^{-21}$ gm/cm³ (Atmospheric density).

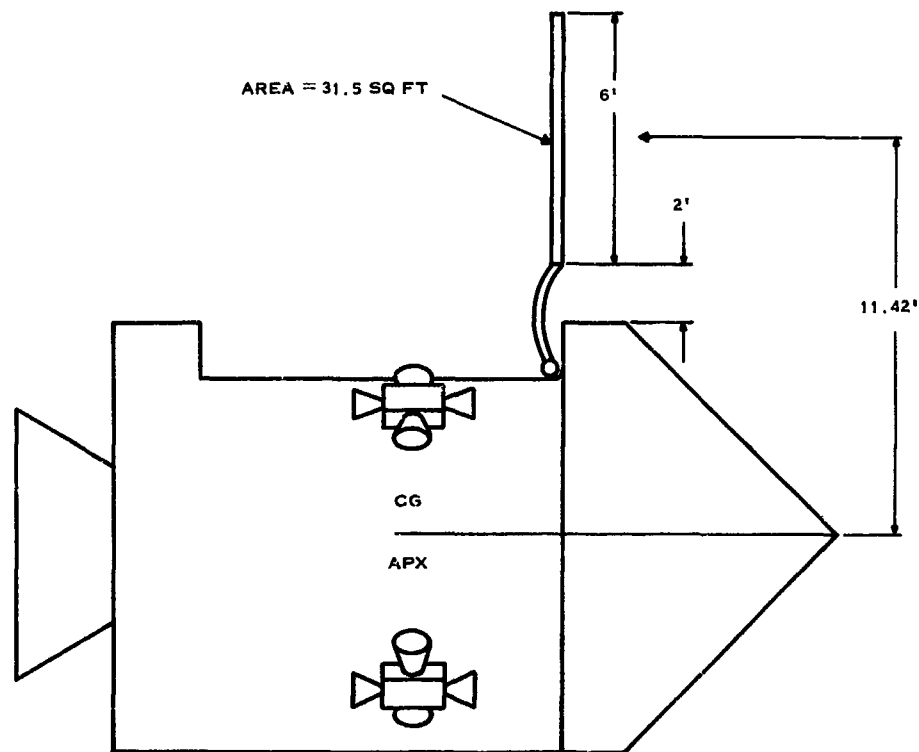
The moment exerted on the spacecraft is the product of the force exerted by the drag of the antenna and the lever arm presented by distance from the center of drag to the center of gravity of the spacecraft.

$M = D \times L$, $L = 11.42$ feet approximately

Drag $D = C_d \cdot q \cdot S_p$ where C_d = Drag coefficient

S_p = Area of array

For a Flat Plate $C_d = 1.28$



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Figure 6-1. Six-Foot Octagon Shaped Array Mounted On Service Module

From Geometry of the Array $S_p = 31.5 \text{ ft}^2$ and $q = 1/2 \text{ EV}^2$ is dynamic pressure lb/ft^2

$$q_1 = 1/2 \times 1.52 \times 10^{-12} \frac{\text{gm}}{\text{cm}^3} \frac{\text{lb}}{453.6 \text{ gm}} \frac{28400 \text{ cm}^3}{\text{ft}^3} \times (25,567)^2$$

$$q_1 = \underline{3.1 \times 10^{-4} \text{ lb/ft}^2} \quad q_2 = \underline{8.9 \times 10^{-11} \text{ lb/ft}^2}$$

$$D_1 = C_d q S_p$$

$$D_1 = 1.28 \times 3.1 \times 10^{-4} \text{ lb/ft}^2 \times 31.5 \text{ ft}^2$$

$$\underline{D_1 = 0.0125 \text{ lb at 100 nm altitude}}$$

$$\underline{D_2 = 3.59 \times 10^{-5} \text{ lb at 1469 nm altitude}}$$

The resulting moment —

$$M_1 = D_1 L = 0.0125 \text{ lb} \times 11.42 \text{ ft}$$

$$M_1 = 0.143 \text{ lb/ft}$$

$$\underline{M_2 = 4.11 \times 10^{-8} \text{ lb/ft}}$$

The attitude control system must furnish a compensating moment to the spacecraft for maintaining spacecraft stability. The final analysis will consider the specific details of the experiment performed by the CAT radar. If a long duration experiment was to be performed at a low altitude of 100 nm, it could be desirable to fold out a symmetrical platform on the opposite side of the vehicle. The main thruster could then compensate for the added drag. At the higher altitudes there would be no problem presented by the minute moment exerted on the spacecraft by the array.

The weight analysis of the active element phased arrays considers all arrays up to a maximum aperture of 31.5 square feet and at each of the six frequencies under consideration. The typical active element array consists of the following major components:

1. Rf Modules
2. Rf Manifold
3. Dc Manifold
4. Energy Storage (capacitors)
5. Cooling Manifold
6. Support Structure.

The rf modules are the most important and complex part of the corporate-fed phased array. The proposed modules are described by Figures 6-2 thru 6-7. A detailed analysis of the circuit weights is included in Figure 6-8 along with its corresponding dipole and housing. The module package is patterned after the existing 9-GHz active module with several refinements incorporated into the package. The antenna dipole, rf connector housings and module body are made of aluminum alloy materials in order to reduce the weight of the module. The original 9-GHz MERA module weighs 0.082 pounds and is constructed of brass while the same module configuration made of aluminum weighs 0.052 pounds. Another small weight reduction could be made by packaging several modules together in a cell. There would be some decrease in the case of maintainability with the addition

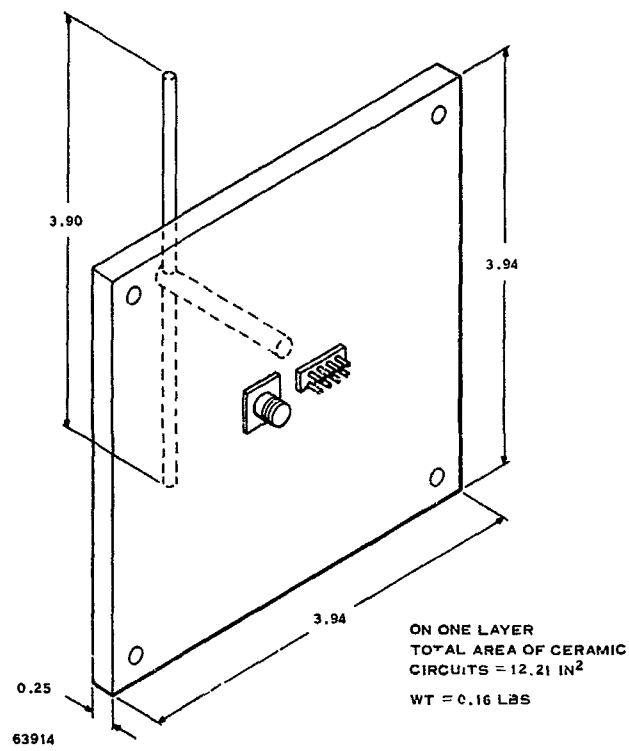


Figure 6-2. 1.5-GHz Module

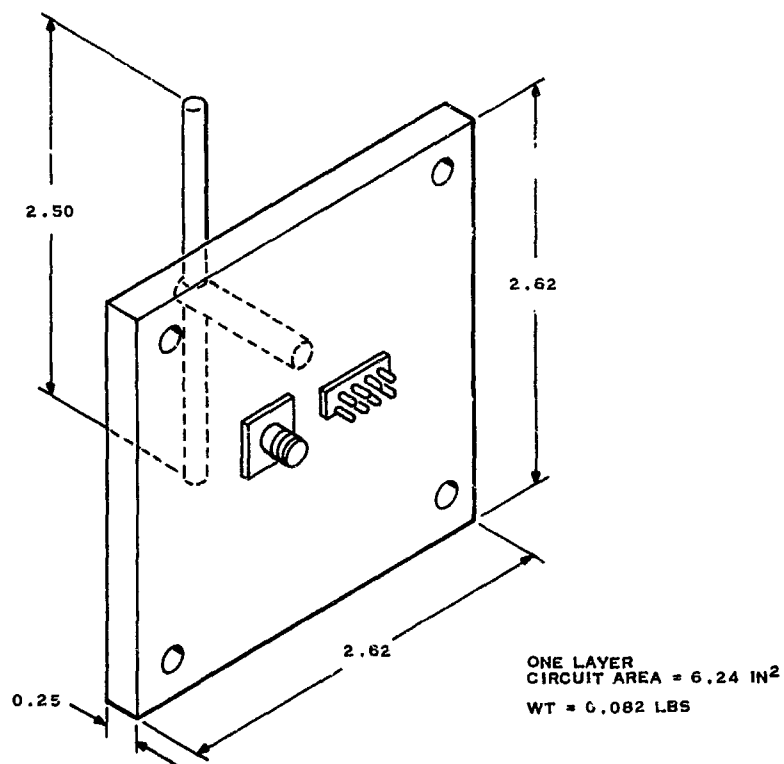


Figure 6-3. 2.25-GHz Module

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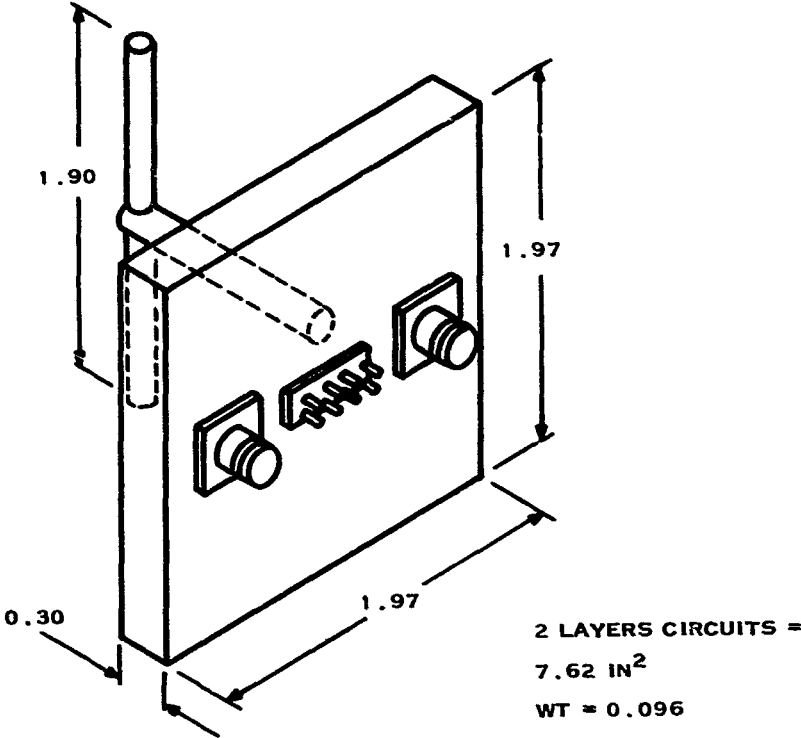
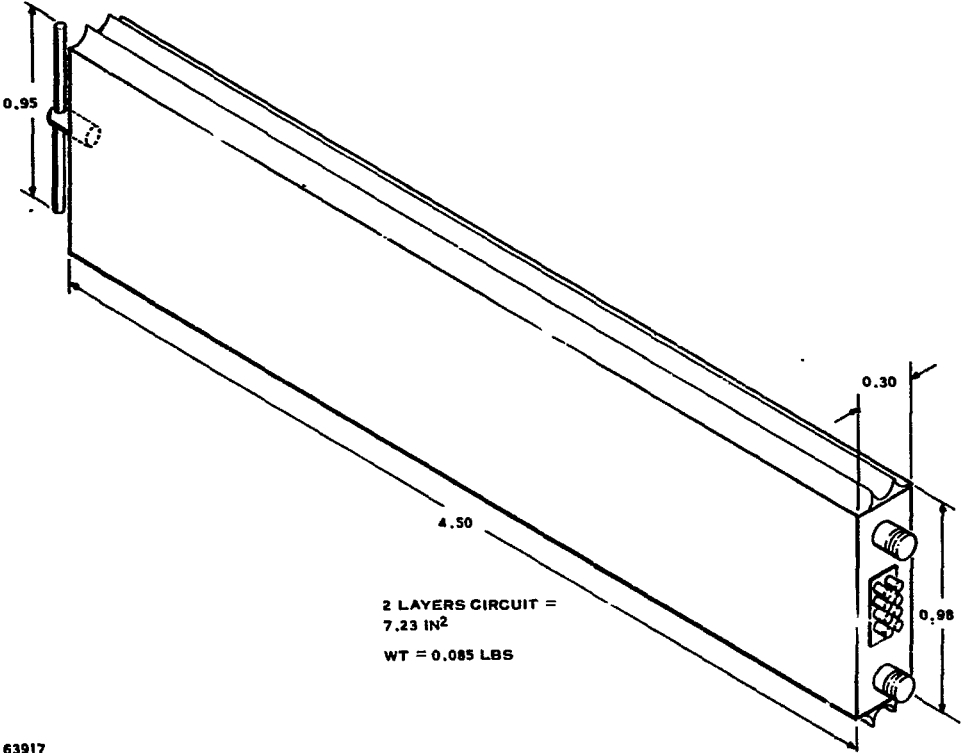
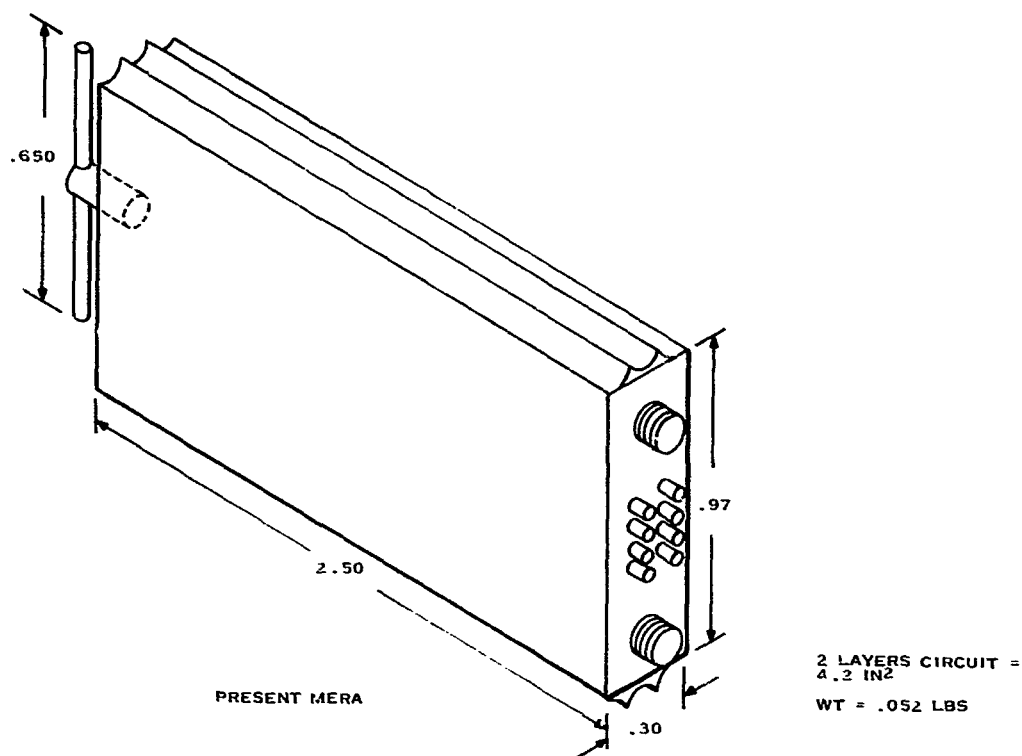


Figure 6-4. 3.0-GHz Module



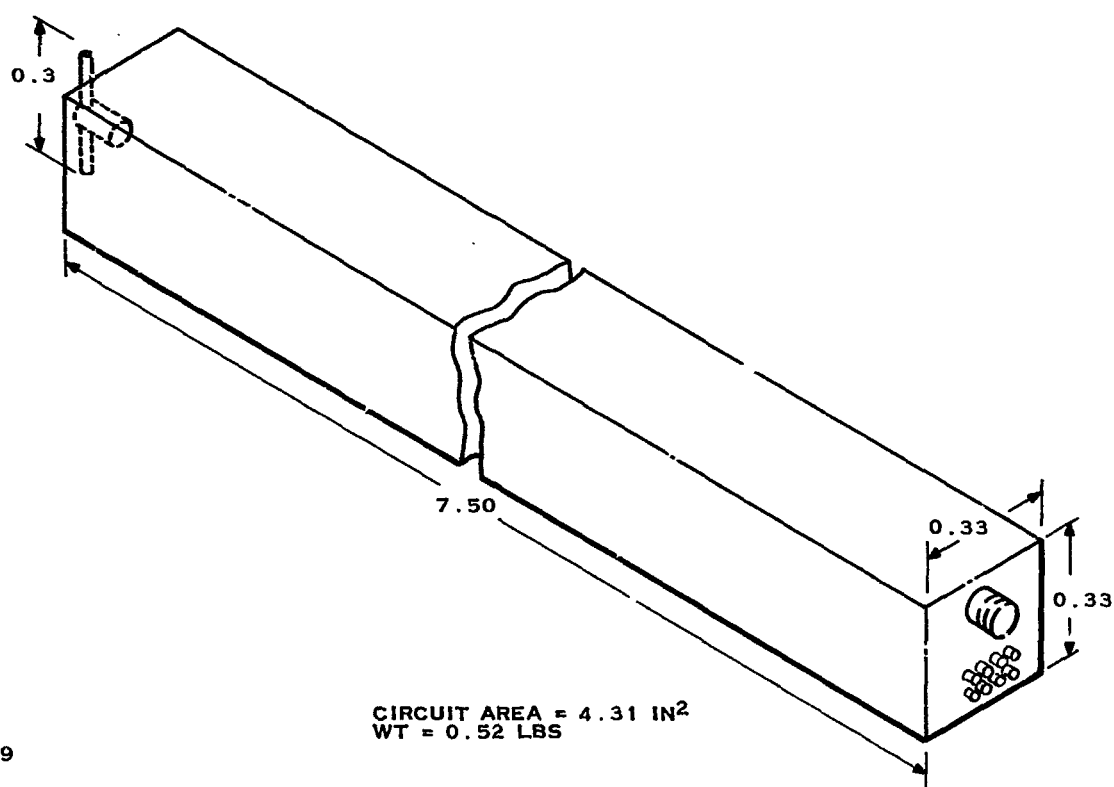
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Figure 6-5. 6.0-GHz Module



63918

Figure 6-6. 9.0-GHz Module



63919

Figure 6-7. 18.0-GHz Module

MODULE ELEMENT DESCRIPTION	CIRCUIT BOARD AREA IN IN ²					
	1.5GHz	2.25GHz	3.0GHz	6.0GHz	9.0GHz	18.0GHz
Pre Amp	---	---	---	---	---	.20
T-R	.21	.18	---	.21	.18	.18
Phase Shifter	.88	.63	1.17	1.75	.69	.53
Power Amplifier	3.14	1.40	3.14	3.14	1.40	1.20
Multipliers, Power	---	---	.75	.60	.33	.71
Circulator	7.0	3.36	2.10	.60	.42	---
Limiter	.10	.07	.06	.04	.08	---
T-R	---	---	---	---	---	.34
Mixer or Mixer Pre-Amp	---	---	---	.29	.42	.44
Multipliers, LO	---	---	---	.60	.50	.71
Pre-Amp	.88	.60	.40	---	---	---
TOTALS Area IN ²	12.21	6.24	7.62	7.23	4.02	4.31
Weight of Circuits .020 thick .2 ceramic .0028 #/in ²	.034lb	.0175 lb	.0213 lb	.0208 lb	.0113 lb	.012 lb
Weight of Al Package Including Dipole Antenna, Solder, Ballbonds, Connectors, and Lids	.12 lb	.075	.075	.065	.040	.040
TOTAL Module Weight	.154 lb	.0825 lb	.096 lb	.085 lb	.052 lb	.052 lb
Antenna Dipole Weight	.035 lb	.024 lb	.016 lb	.008 lb	.0018 lb	.0018 lb
Total Number of RF Manifold Layers Required	2.0	2.0	3.0	4.5	5.0	8.0
RF Manifold Frequencies Required	1.5GHz	2.25GHz	1.5GHz 3.0GHz	1.50GHz 100MHz	2.25GHz 100MHz	2.25GHz 100MHz
Estimated Manifold Wt./RF Module	.15 lb	.067 lb	.059 lb	.032 lb	.026 lb	.014 lb

Figure 6-8. Module Weight Analysis

of greater complexity. The circuits within the rf modules consist of 0.020 thick ceramic which has a density of 0.14 pounds per cubic inch or 0.0028 pounds per square inch.

The rf manifold weight summary can be found on Figure 6-8 which includes data on the number of manifold frequencies, number of manifold layers required for the complete corporate feed, and the estimated weight of the manifolds per rf module at each of the proposed frequencies. The rf manifolds or corporate feed consists of a series of power dividers fabricated on a polyphenylene oxide stripline material. The advantage of using polyphenylene oxide over conventional Teflon glass is that the density of polyphenylene oxide is less than 50 percent of Teflon glass and that polyphenylene oxide is easily bonded to itself thus eliminating the necessity of having bulky clamping plates on the manifold. Aluminum rf connectors are used on both the input and output connections to help reduce weight. Aluminum plates are considered to support the rf connector on the rf manifold. The 75-MHz manifold will probably utilize ferrite toroidal power dividers since conventional stripline power dividers would require an addition of extra layers to the rf manifold. The toroidal power dividers weigh 2 grams each and the quantity required is equal to the number of modules minus one assuming all two-way power dividers are used.

The dc manifold consists of conventional multilayer G-10 circuit material. The estimated dc manifold weights are based on a similar MERA dc manifold weight. The energy storage capacitors are required to meet the demand of high speed switching currents that are required by the modules. The cooling manifold and support structure would be closely related and could be combined into the same structure. However the rf modules can dissipate such a large range of heat values depending on the module duty cycle that the weight of the cooling manifold used in the weight analysis is based on a nominal 1 percent duty cycle. The structure weight would also be a variable since it would be a function of the detail environmental specifications imposed on the system. The weight of the structure used in the analysis assumes a relatively calm environment during actual operation of the radar.

Figure 6-9 is a summary of the weight analysis for the 31.5-square-foot array at each of six frequencies.

The weight analysis of the space-fed array is similar to the active-element array except the modules include only the weight of the ground planes, dipoles, and a fraction of the circuit. The two different types of space-fed arrays that are being analyzed are transmissive and reflective. The typical space-fed array consists of the following components:

DESCRIPTION	FREQUENCY GHz					
	1.5	2.25	3.0	6.0	9.0	18.0
Modules	45.0	54.6	112.5	398.0	545.0	1260.0
Manifold (RF)	43.1	44.0	69.1	147.5	276.0	585.0
Support Structure	50.0	50.0	80.0	100.0	125.0	400.0
DC Manifold	15.0	20.0	25.0	30.0	38.6	60.0
Energy Storage (Capacitors)	10.0	20.0	35.0	50.0	100.0	200.0
Cooling Manifold	10.0	15.0	35.0	45.0	60.0	100.0
TOTAL WEIGHT (Lbs.)	173.1	203.6	356.6	770.5	1144.6	2605.0
TOTAL NO. MODULES/ 6' DIA APERTURE	292	662	1170	4680	10,600	42,000
TOTAL NO. SQ. WAVELENGTHS	73	165	292	1170	2650	10,500

WEIGHT AS A FUNCTION OF FREQUENCY

	1.5 GHz	2.25 GHz	3.0 GHz	6.0 GHz	9.0 GHz	18.0 GHz
Weight per Sq. Wavelength lbs/ λ^2	2.37	1.23	1.22	.66	.44	.248
Total Weight per Module lbs/module	.59	.31	.30	.165	.11	.062
Total Weight per Unit Area lbs/in ² of Aperture	.038	.045	.079	.170	.252	.575

Figure 6-9. Weight Estimate of 31.5-Ft² Aperture (Approx. 6-ft Octagon)

1. Phase-shift modules
2. Dc manifold
3. Support structure
4. Transmitter and feedhorn assembly.

The estimated weights of both the transmissive and reflective array modules are shown on Figure 6-10 for each of five frequencies. Figure 6-11 is a summary of the weights for a 31.5-square foot aperture at five frequencies for both a transmissive array and a reflective array. Figure 6-12 presents the specific weights of the space-fed array as a function of array area at each frequency. Figures 6-13 and 6-14 are two representative curves showing how weight varies as a function of frequency for both space-fed arrays and active-element arrays.

The electronics in the CAT radar system could be packaged utilizing recent developments in large scale integration (LSI) and medium scale integration (MSI) for high-speed switching circuits. The use of MSI or LSI will enable all computer type circuitry to be packaged at a higher density and with higher reliability by reducing the number of components and the associated interconnections. LSI packages have been developed which have over 200 gates per package while MSI packages have from 15 to 50 gates per package. With the exception of possibly incorporating MSI or LSI in the system, the packaging will be of the conventional high-density type used in most space radar applications. It will probably be necessary to use a cold plate or other heat transfer mechanism to cool the individual electronic packages since a minimum of approximately 100 watts will be dissipated by the synchronizer-scan computer, receiver processor, data processor, and signal generator unit. Figure 6-15 shows the power required for each of the electronic assemblies in a typical 292-element 1.5-GHz CAT radar. Active element arrays with more elements will require proportionally more power in the signal generator unit and scan computer. Likewise larger and more power space-fed arrays will require more power in the scan computer in the same proportion.

The most severe environmental conditions imposed on the CAT radar while operating are the thermal shock imposed by the sun and the sun's shadow and the high vacuum surrounding the spacecraft. The effect of the thermal shock may be minimized by proper coating on exposed surfaces and the selection and use of reliable qualified electronic components. The high vacuum imposes some requirements on power handling capability of the space-fed array. It is desirable to pressurize the high-voltage transmitter

Transmissive System (Module includes 2 Antennas, 2 gnd planes,
Phase Shifters, housing and connector)

18 GHz	.015 lb	W/G Coupler Antenna
9 GHz	.02 lb	Dipole
6 GHz	.04 lb	Dipole
3 GHz	.07 lb	Dipole
1.5 GHz	.13 lb	Dipole

Reflective System (Module includes 1 Antenna, 1 gnd plane,
phase shifter, housing and connector)

18 GHz	.01 lb	W/G Coupler Antenna
9 GHz	.015 lb	Dipole Antenna
6 GHz	.025 lb	Dipole
3 GHz	.04 lb	Dipole
1.5 GHz	.08 lb	Dipole

Ground plane area, circuit area, and dipole weights are based on
corporate fed module analysis.

Figure 6-10. Space-Fed Module Weight Analysis

Description		Frequency GHz				
		1.5	3.0	6.0	9.0	18.0
Modules	1b	38.0	82.0	187.0	212.0	630.00
DC Manifold	1b	12.0	16.0	20.0	30.0	50.00
Support Structure	1b	25.0	45.0	70.0	90.0	120.0
Array WT Total	1b	75.0	143.0	277.0	332.0	800.0
10 KW Transmitter including TWT, Coax and/or W/G, Duplexer, Modulator and Housing	1b	40.0	40.0	40.0	40.0	40.0
TOTAL lbs		115.0	183.0	317.0	372.0	820.0

**WEIGHT ESTIMATE FOR 31.5 Ft²
REFLECTIVE SPACE FED ARRAY**

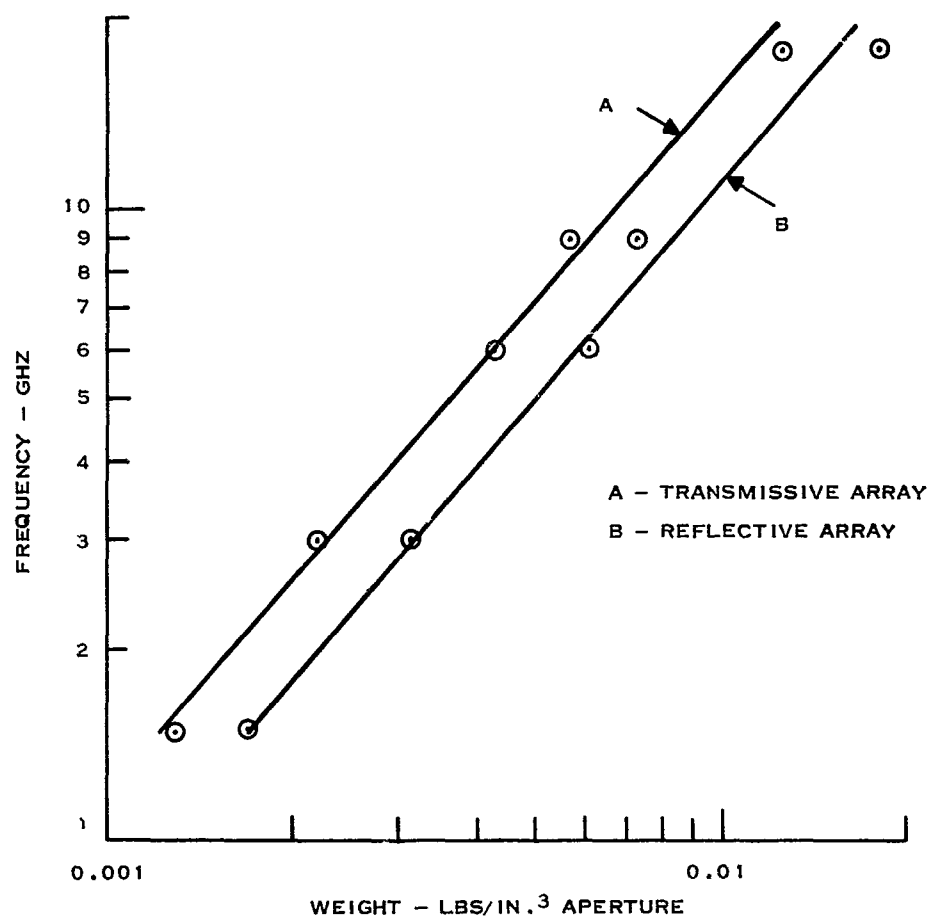
Description		Frequency GHz				
		1.5	3.0	6.0	9.0	18.0
Modules	1b	23.4	46.8	117.0	159.0	420.0
DC Manifold	1b	12.0	16.0	20.0	30.0	50.0
Support Structure	1b	25.0	35.0	55.0	70.0	100.0
Array Weight Total	1b	60.4	97.8	192.0	259.0	570.0
10 KW Transmitter	1b	40.0	40.0	40.0	40.0	40.0
TOTAL lbs		100.4	137.8	232.0	299.0	610.0

Figure 6-11. Weight Estimate for 31.5-Ft² Transmissive Space-Fed Array

Transmissive Array		Frequency GHz				
		1.5	3.0	6.0	9.0	18.0
Array Weight per Sq. Wavelength lb / λ^2		1.03	.49	.24	.126	.076
Total Weight per Module of Array lbs/module		.26	.12	.06	.031	.019
Total Weight of Array per Unit Area lb /in ² of Aperture		.0017	.0031	.0061	.0073	.018

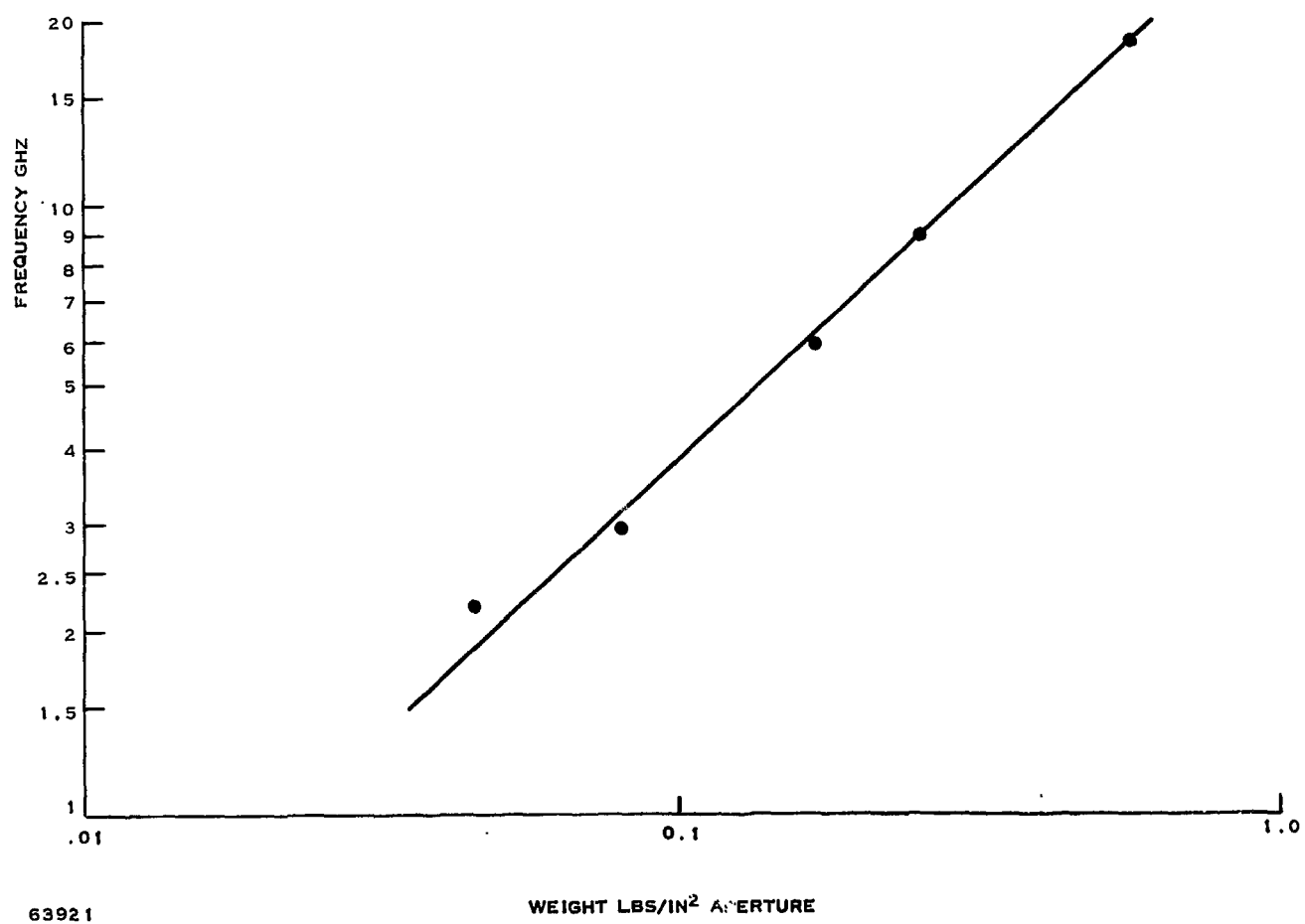
Reflective Array		Frequency GHz				
		1.5	3.0	6.0	9.0	18.0
Array Weight per Sq. Wavelength lb / λ		.83	.35	.164	.098	.054
Total Weight per Module of Array lb /module		.21	.087	.041	.025	.014
Total Weight of Array per Unit Area lb /in ² Aperture		.0013	.0022	.0043	.0057	.0126

Figure 6-12. Space-Fed Array Weight Analysis



63920

Figure 6-13. Weight as a Function of Frequency for a Space-Fed Array



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Figure 6-14. Weight as A Function of Frequency for an Active-Element Array

CORPORATE FED ACTIVE ARRAY

<u>DESCRIPTION</u>	<u>WEIGHT</u>	<u>VOLUME</u>	<u>POWER</u>
1.5 GHz Array	173. lb	10 ft ³	1000 watts
Synchronizer and Scan Computer	15. lb	350 in ³	30 watts
Receiver Processor	25. lb	900 in ³	30 watts
Signal Generator Unit	15. lb	450 in ³	30 watts
Data Processor	<u>6. lb</u>	<u>200 in³</u>	<u>10 watts</u>
TOTAL	234. lbs	11.1 ft ³	1100 watts

SPACE FED ARRAY

<u>DESCRIPTION</u>	<u>WEIGHT</u>	<u>VOLUME</u>	<u>POWER</u>
1.5 GHz Array Reflective and Transmissive	60 & 75 lb	10 & 16 lbs	650 watts
10 KW Transmitter	40 lb	800 in ³	350 watts
Receiver Processor	25.0 lb	900 in ³	30 watts
Synchronizer and Scan Computer	15.0 lb	350 in ³	30 watts
Data Processor	<u>6.0 lb</u>	<u>200 in³</u>	<u>10 watts</u>
TOTAL	146 & 161 lbs	11.1 & 17.4 ft ³	1070 watts

Figure 6-15. 1-kW System Weight Power And Volume Estimate

and waveguide with an inert gas to prevent the possibility of arcing. Other precautions must be taken with lubricants in bearings and materials generally to avoid excessive evaporation. Since the missions will be of relatively short durations the evaporation of materials should not be a difficult problem if proper guidelines are used in the selection of materials.

7.0 RELIABILITY ANALYSIS

This section discusses the index to be used for measuring reliability performance and the maintainability concept for the subject mission class. Equations for calculating mission reliability from basic failure rates are reviewed. System reliability is estimated for a specific system, and the functional relations of system reliability to system, mission and environmental parameters are presented.

7.1 Reliability Indices

The three most commonly used reliability indices are: failure rate; mean-time-between-failures (MTBF); and probability of mission success (also referred to as reliability or mission reliability). In many applications, the three are equally meaningful, and one can be calculated from another by a simple algebraic relation. The CAT radar contains major functional blocks which have nonconstant failure rates, however, and the simple interrelation among MTBF, failure rate and reliability does not hold. The basic interrelation, valid for all cases, is given in Equations (7-1) and (7-2).

$$\text{MTBF} = \int_{t=0}^{\infty} R(t) dt, \text{ and} \quad (7-1)$$

$$\text{Failure Rate } \lambda(t) = \frac{-1}{R(t)} \frac{d}{dt} R(t), \quad (7-2)$$

where $R(t)$ is probability of equipment survival, as a function of equipment operating time.

Notice that MTBF is single-valued, while λ and R are functions of time. Therefore the MTBF of a system is significant only when either failure rate λ is constant (and reliability R thus negative exponential) or when equipment is to be used for operational periods of several MTBF's (thus implying maintenance). When a large number of equipments are being used for an operational time period less than an MTBF, the logistics of replacing or repairing failed units are of prime concern. Then failure rate as a function of time is the most significant form for expressing reliability behavior.

When an important mission or task is to be performed by an equipment or system, the significant index is probability of mission success, or reliability. When used in a quantitative connotation, reliability means probability of success. In a qualitative context, reliability is a more general term. The reliability time function $R(t)$ can be calculated from failure rate $\lambda(t)$ by manipulating Equation (7-2), but $R(t)$ cannot be retrieved from MTBF Equation (7-1). The constraints of the CAT mission make mission reliability

most meaningful index. While failure rate can also be used as an index for CAT, MTBF is not a valid measure for evaluating reliability performance.

The need for careful differentiation between MTBF, reliability, and failure rate is brought about by the partial redundancy inherent in phased arrays, and by the partial redundancy which may be designed into the dc power supplies. Calculation of reliability for the case of partial redundancy and for the more common constant failure rate case is presented in Subsection 7.2.

7.2 Calculation of Reliability

7.2.1 Constant Failure Rate Case

If failure rate $\lambda(t)$ is constant, reliability is calculated by

$$R(t) = e^{-\lambda t} . \quad (7-3)$$

The assumption of constant failure rate is a valid approximation for most parts and circuits.

7.2.2 Case of Partial Redundancy

If J -elements of an array of N -elements may fail without causing array failure, then the reliability of the array is given by

$$R_a(t) = \sum_{i=0}^J \binom{N}{i} R_e^{(N-i)} (1 - R_e)^i, \quad (7-4)$$

where R_e is the reliability of a single element of the array, and is a function of time. Usually, element failure rate is constant, and $R_e(t)$ is given by Equation (7-3) above. Equation (7-4) then can be written as

$$R_a(t) = \sum_{i=0}^J \binom{N}{i} e^{-(N-i)\lambda_e t} \left(1 - e^{-\lambda_e t}\right)^i . \quad (7-5)$$

Equation (7-5) is a satisfactory relation for calculating the reliability of a small, simple functional block, such as an array of power supplies feeding a power bus through isolation diodes with J extra supplies. Two problems arise when calculating the reliability of a partially redundant phased-antenna array, however. First, for large N and J , the computation becomes unwieldy, even for a computer, and therefore expensive. Second, in some cases the critical number of failures, J , cannot be explicitly determined. J is implicitly set in all cases, by defining the allowable signal-to-noise degradation. But if some failures have a different effect than others,

J is not uniquely determined, and differs according to which type failures actually happen. Examples of difference in failure effect appear in both space-fed and corporate-fed arrays. In active module corporate-fed antennas, a failure may occur in a part affecting only transmit or only receive. This does not decrease received signal-to-noise ratio as seriously as does the failure of a part affecting both. A space-fed array is ordinarily tapered by the feed-horn pattern, and thus the failure of an edge module is less serious than failure of a center module.

For a large array, ($N \geq 40$ has been used as an approximate definition of large for antenna arrays) the true failure behavior does not vary as greatly (percentage-wise) from the average behavior as it does with small arrays. Thus, when array size becomes too large for easy reliability calculation by Equation (7-5), an approximate solution becomes acceptably accurate. This solution consists of treating the failures of modules as a Gaussian random process. Then the reliability at a given time is the probability (from a table of the Gaussian integral) that the deviation from expected behavior is great enough that the critical number of failures is exceeded.

If the expected number of events in a random process is M , the standard deviation from the expected is $\sqrt{M}$. The expected number of failures in any array of N modules would be

$$M = N \left(1 - e^{-\lambda_e t} \right). \quad (7-6)$$

If the critical number of failures, J , is known, reliability is calculated by

$$R = \int_{x=-\infty}^{(J-M)/\sqrt{M}} (2\pi)^{-1/2} e^{-x^2/2} dx, \quad (7-7)$$

which is the integrated Gaussian distribution. Tables of this integral are readily available.

If all failures do not have the same effect, the average number of malfunctions necessary for array failure can be determined and used for J . Let λ_1 be the failure rate of the portion of a module that affects both transmit and receive, λ_2 the portion that affects transmit only, and λ_3 the part that affects receive only, so that $\lambda_e = \lambda_1 + \lambda_2 + \lambda_3$. Let G_{TR} be the round-trip normalized antenna gain (that is, the product of transmit gain and receive gain, which can be different for active module arrays, divided by maximum available gain). Let this relative transmit gain include the loss of transmitted power as well as loss due to pattern degradation. Then transmit gain G_T is the square of the ratio of transmitting modules N_T to N . The number of transmitting modules as a function of time is

$$N_T = N e^{-(\lambda_1 + \lambda_2)t}. \quad (7-8)$$

Relative receive gain G_R varies as $(N_R/N)^2$ due to losses in power combiners for a corporate-fed array, and due to superposition of the electric fields for a space-fed array, where

$$N_R = N e^{-(\lambda_1 + \lambda_3)t} \quad (7-9)$$

Therefore, on the average, relative transceive gain G_{TR} varies as a function of time according to

$$G_{TR} = \frac{N_T^2 N_R^2}{N^4} = e^{-2(2\lambda_1 + \lambda_2 + \lambda_3)t} \quad (7-10)$$

If a minimum useful relative antenna transceive gain G_{min} is established from the signal-to-noise ratio margin available (or equivalently, from the minimum useful range), an average time to failure θ_a can be obtained from Equation (7-10).

$$G_{min} = e^{-2(2\lambda_1 + \lambda_2 + \lambda_3)\theta_a}$$

Solving for θ_a gives

$$\theta_a = \ln(1/G_{min}) / (4\lambda_1 + 2\lambda_2 + 2\lambda_3) \quad (7-11)$$

The critical number of failures J is then found by evaluating Equation (7-6) at time θ_a .

$$\begin{aligned} J &= N(1 - e^{-\lambda_e \theta_a}) \\ J &= N \left\{ 1 - \left[e^{-\ln(1/G_{min})} \right]^{\lambda_e / (4\lambda_1 + 2\lambda_2 + 2\lambda_3)} \right\} \\ J &= N \left[1 - G_{min}^{(\lambda_1 + \lambda_2 + \lambda_3) / (4\lambda_1 + 2\lambda_2 + 2\lambda_3)} \right] \end{aligned} \quad (7-12)$$

This value of J can then be used in Equation (7-7) to calculate the reliability of the antenna for any given time span.

Equation (7-10) has an interesting implication with respect to radar range. For Equation (7-10) and the radar range equation, it is established that range decreases gradually. The average behavior is given by

$$\text{Range} = \text{max range} \times e^{-(2\lambda_1 + \lambda_2 + \lambda_3)t/2} \quad (7-13)$$

for corporate-fed antennas, and is

$$\text{Range} = \text{max range} \times e^{-\lambda_e t} \quad (7-14)$$

for space-fed antennas.

The reliability of a system R_s is the product of the reliability of the system parts,

$$R_s(t) = \prod_{i=1}^n R_i(t), \quad (7-15)$$

where the $R_i(t)$ are given by Equations (7-3), (7-4), or (7-7) as appropriate, and n is the number of system parts (or functional blocks).

7.3 System Reliability

System reliability will be discussed in terms of the functional blocks of Figure 7-1.

7.3.1 Conditions Assumed

To estimate the potential reliability of a CAT radar system, a heat transfer system is assumed which will maintain part temperatures below 125°C for high-power parts, and near 80°C for low-power parts. The effect on reliability of other temperatures is presented in this subsection.

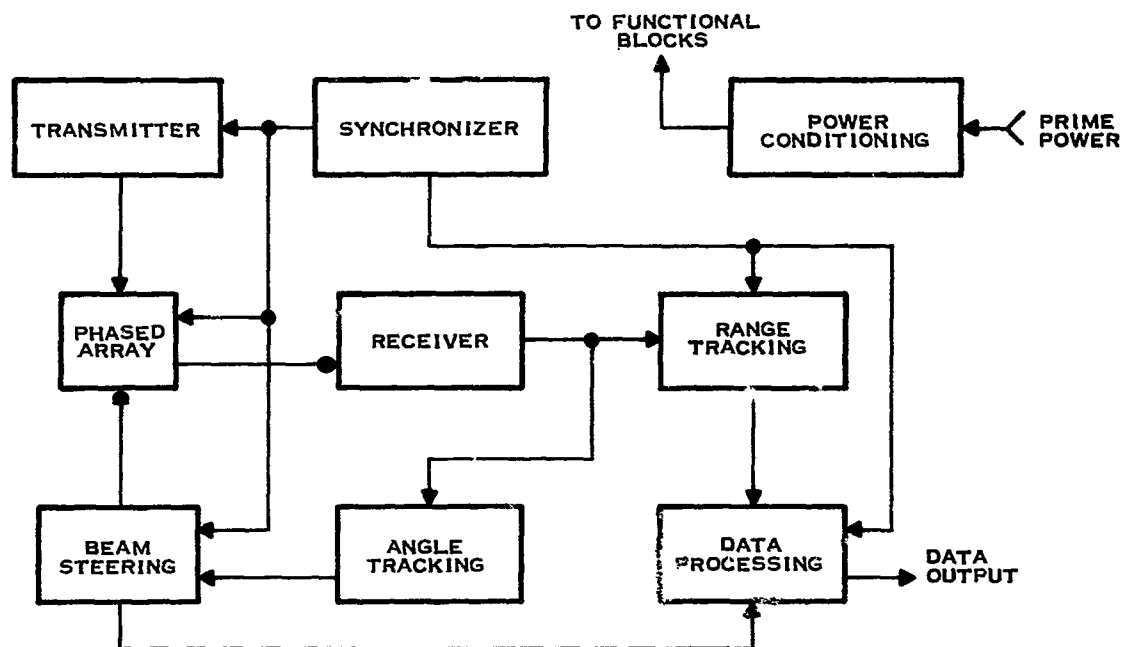
Mission reliability is calculated for a 1000-hour radar experiment. This represents 14-percent radar usage during a 10-month mission, or 100-percent usage during a 6-week mission. Shorter operating time results in higher reliability.

The liberal use of large-scale, medium-scale, and monolithic and hybrid integrated circuits is assumed. The use of high-reliability parts and of equipment burn-in is considered standard procedure for the CAT effort. Meteoroid and radiation damage are insignificant (see Subsections 7.4.2 and 7.4.5). Thermal cycles and mechanical vibration are assumed to be moderate (see Subsection 7.4.3).

7.3.2 Power Conditioner

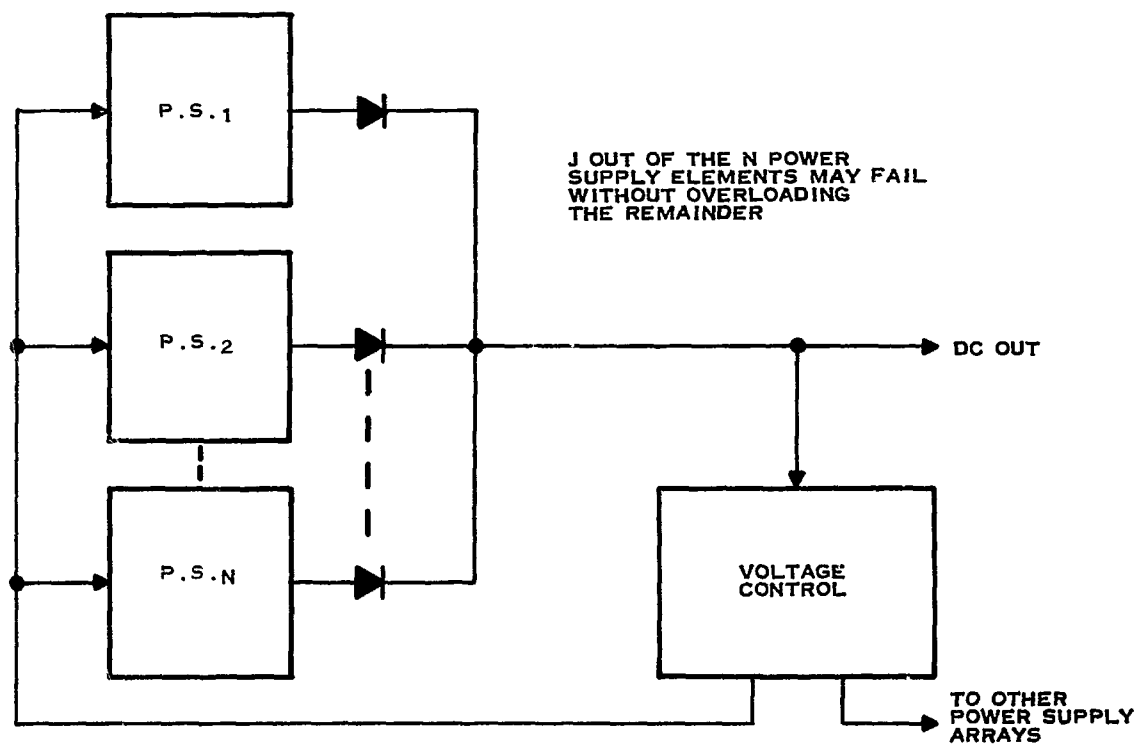
Both the active module and the diode-switched passive module arrays require relatively high direct current at moderate to low voltages. Partial redundancy may be used to advantage in this situation. Figure 7-2 illustrates a group of moderate-current power supplies feeding a power bus through isolation diodes. If more individual supplies are provided than are needed, a high-mission reliability can be obtained for the power conditioner.

Figure 7-3 illustrates that a reliability of nearly 100 percent can be obtained from an array of 12 power supplies, of which three are surplus. This is based on a 1000-hour operating time, and an individual module failure rate of 40 per 10^6 hours, which is conservative for a moderate current, low-voltage supply. Figure 7-3 is a computer generated plot following Equation (7-5) for several different numbers of allowed failures J .



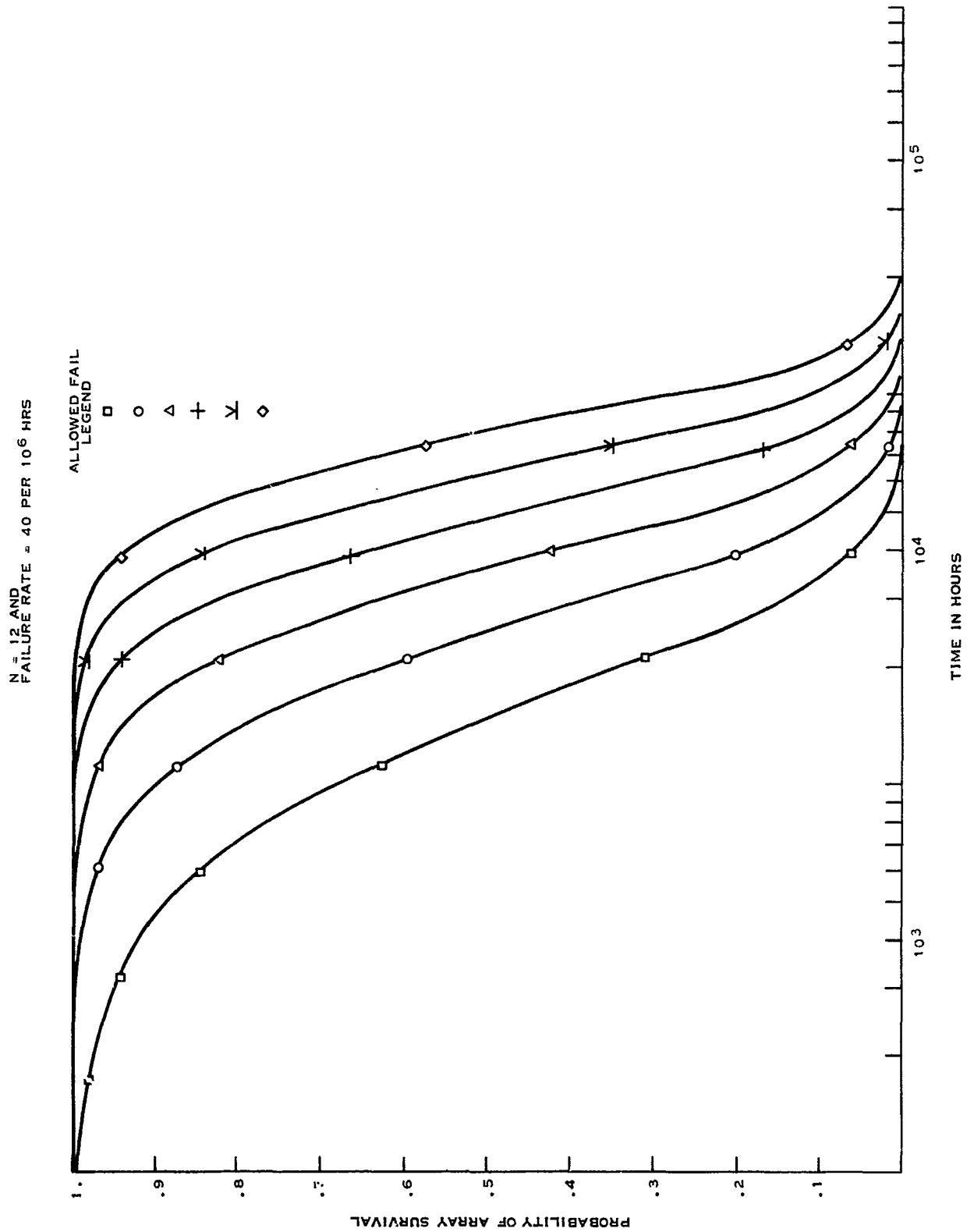
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Figure 7-1. Basic System Functional Block Diagram



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Figure 7-2. Power Conditioner



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Figure 7-3. Array Survival Probability Versus Time for J-Allowed Failures

Since the voltage control does not necessarily benefit from redundancy, some allowance must be made for its failure during the mission. A controller handling all the low voltages used, made of digital and linear integrated circuits, might have a failure rate of 10 per 10^6 hours, or a 99-percent reliability for the 1000-hour operation.

A high-voltage supply would be needed if a vacuum-tube rf source were used. Partial redundancy does not appear as naturally adaptable to a high-voltage supply as to a high-current supply. Since reasonable reliability can be obtained without resorting to full redundancy, full redundancy is not recommended. A high-voltage power supply failure rate of 50 per 10^6 hours would have a 95-percent mission reliability.

7.3.3 Phased Array

Both space-fed and corporate-fed phased arrays benefit from partial redundancy. Equation (7-7), with auxiliary Equations (7-6) and (7-12), can be used to calculate array reliability versus time, if the number of array modules N , the module failure rate λ_e , and the allowable range decrease are established. For example, let a 10-percent range decrease be allowed. The relative transceive gain G_{TR} can be considered a portion of the loss term in the radar range equations. Therefore, $G_{min} = 0.9^4$. Let $N = 300$, which is approximately the smallest number of modules which would be used in a 6-foot diameter antenna. Failure rates for antenna array elements are listed in Table VII-1, and are based on the module configurations of Subsection 3.2. The portion of the active module failure rate which affects both transmit and receive is about 20 percent of the total failure rate. That is $\lambda_1 = 0.2\lambda_e$.

Table VII-1. Antenna Module Failure Rates

Type Module	λ_e (in failures per 10^6 hours)
Diode phase-shifter module	2
1.5 GHz active module	18
2.25 GHz active module	16
3.0 GHz active module	20
6.0 GHz active module	24
9.0 GHz active module	24
18.0 GHz active module	30

Using the 1.5-GHz active module for example, the expected number of failures from Equation (7-6) is 5.4 at 1000 hours. From Equation (7-12), the critical number of failures is 48. Forty-eight is 18 standard deviations greater than 5.4, so that array reliability is 100 percent, to the accuracy of common tables of the normal function. The time to 90-percent antenna reliability is about 7,800 hours. Without the benefit of partial redundancy, an antenna system would need a 74,000-hour MTBF to match that reliability.

The reliability of the array for a space-fed system is similarly 100 percent. The portion of manifolding which does not benefit from partial redundancy contributes a constant failure rate of about 40 failures per 10^6 hours for the corporate-fed system, and about 25 failures per 10^6 hours for the phase shifters in a space-fed system.

7.3.4 Transmitter

The constraints of the CAT mission and present technology limit consideration of rf power sources to a vacuum-tube source, space-fed array, and a solid-state active corporate array. Thus, the vacuum-tube, solid-state tradeoff cannot be treated independently of the space-fed, corporate-fed tradeoff, for the CAT study. The reliability of the vacuum-tube power generation technology, to a large degree, is inversely related to the required average power output and frequency.

Significant reliability progress has been made in the area of low-powered TWT's. As an example, the helix-type TWT has been used in the satellite transmitter application with excellent reliability results. Some of these transmitters have been functioning continuously for more than three years without failure. All have been subjected to severe vibration and shock with no detectable effects on electrical performance. At microwave frequencies and in the 1 to 20 watt CW class, which the subjected devices were in, the TWT is without serious competition from other vacuum tubes, in terms of the more critical parameters such as efficiency, gain, bandwidth, linearity, noise, size, reliability, and weight. Although other devices may possess advantages in one or more of these areas, none manifest the combined attributes of the TWT.

Unfortunately, it seems that the only significant reliability progress has been made in the area of the lower powered TWT's, basically, because they are inherently more reliable and easier to work with. It should be pointed out that at higher power levels the TWT reliability is more sensitive to higher frequencies because the structures at these frequencies (upper X-band and K_u -band) are too small to handle safely the intercepted beam-power coupled with the rf losses. In addition, reliability has been noted to drop

significantly in going from a few kilovolts to the 50-kilovolt region. This is particularly evident in the electron gun region where the electrode spacings are determined by electron optics and not breakdown consideration.

A general comparison of TWT, klystron, and crossed-field amplifier (CFA) reliability indicates that the TWT and klystron are similar with respect to life and reliability. CFA's are considered to be further behind in these areas, although qualified personnel feel that continued development will change this. One of the significant shortcomings of the CFA with respect to TWT's and klystrons is that it generates considerably more internal heat (runs hotter) for a given power level.

As a result of this evaluation and in light of the required parameter ranges of the required system, the conclusion is that the TWT will offer the greatest reliability advantages of the vacuum-tube class. Therefore, a system comprised of a TWT space-fed array will be used in the numerical comparison with the solid-state corporate-fed system.

For simplicity, this analysis will be restricted to circuitry unique to each system. Circuitry similar in function and part count will be considered identical for both systems and will be excluded from the comparison.

The same environmental conditions will be assumed to exist for both systems. The systems are assumed to be operating in their optimum power and frequency range for best reliability results. The power and frequency ranges are assumed to be comparable to each other.

For purposes of the analysis, both arrays will be considered to have a failure rate of zero on the basis that the system operational life does not approach the life of the arrays. Table VII-2 lists failure rates for the functions which the two systems accomplish in a different manner. The active module system shows an order of magnitude lower failure rate than the space-fed system during system operational life. This is because the active module brings the functions of power amplification, modulation, and duplexing into the partially redundant array, and because high voltage is not needed.

7.3.5 Other Functional Blocks

The other functional blocks do not enter into the corporate-fed, space-fed tradeoff. The tradeoff of complexity versus function would be too involved without a narrower mission definition. Therefore, a comparison to other radar systems has been made as to the failure rate of the other functions. Representative values are listed in Table VII-3.

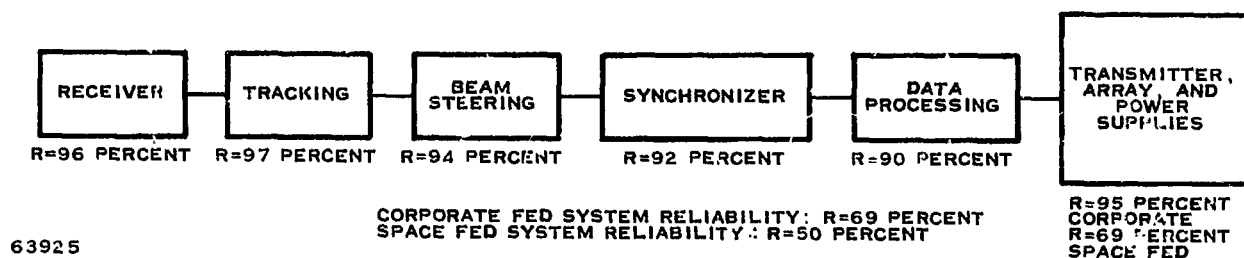


Figure 7-4. Combined System Reliability

Table VII-2. Solid-State Vacuum-Tube Power-Chain Comparison

Vacuum-Tube Power		Solid-State Power	
TWT*	143×10^{-6}		
Modulator	100×10^{-6}		
High-Voltage Supply	50×10^{-6}		
Duplexer	50×10^{-6}		
Manifold	25×10^{-6}	Manifold	40×10^{-6}
	368×10^{-6} failures per hour		40×10^{-6} failures per hour

*Failure rate demonstrated by US Air Force in the field for 5 kilowatts power, TWT. Data supplied by Varian.

Table VII-3. Functional Block Failure Rates

Function	Failure Rate (per 10^6 hours)
Receiver	40
Tracking	30
Beam Steering	60
Synchronizer	80
Data Processing	100

7.3.6 Combined System Reliability

The 1000-hour reliabilities of the system functions are indicated on Figure 7-4. Both system alternatives have estimated 1000-hour reliabilities

of 50-percent or above. Since most mission operating times will be much shorter, no maintenance or additional redundancy appears necessary.

The above estimate is for an example system under a specific set of conditions. Many factors will vary the actual system reliability. Several factors were presented in mathematical form in Subsection 7.2. Factors of an environmental nature are presented in Subsection 7.4.

7.4 Environmental Factors

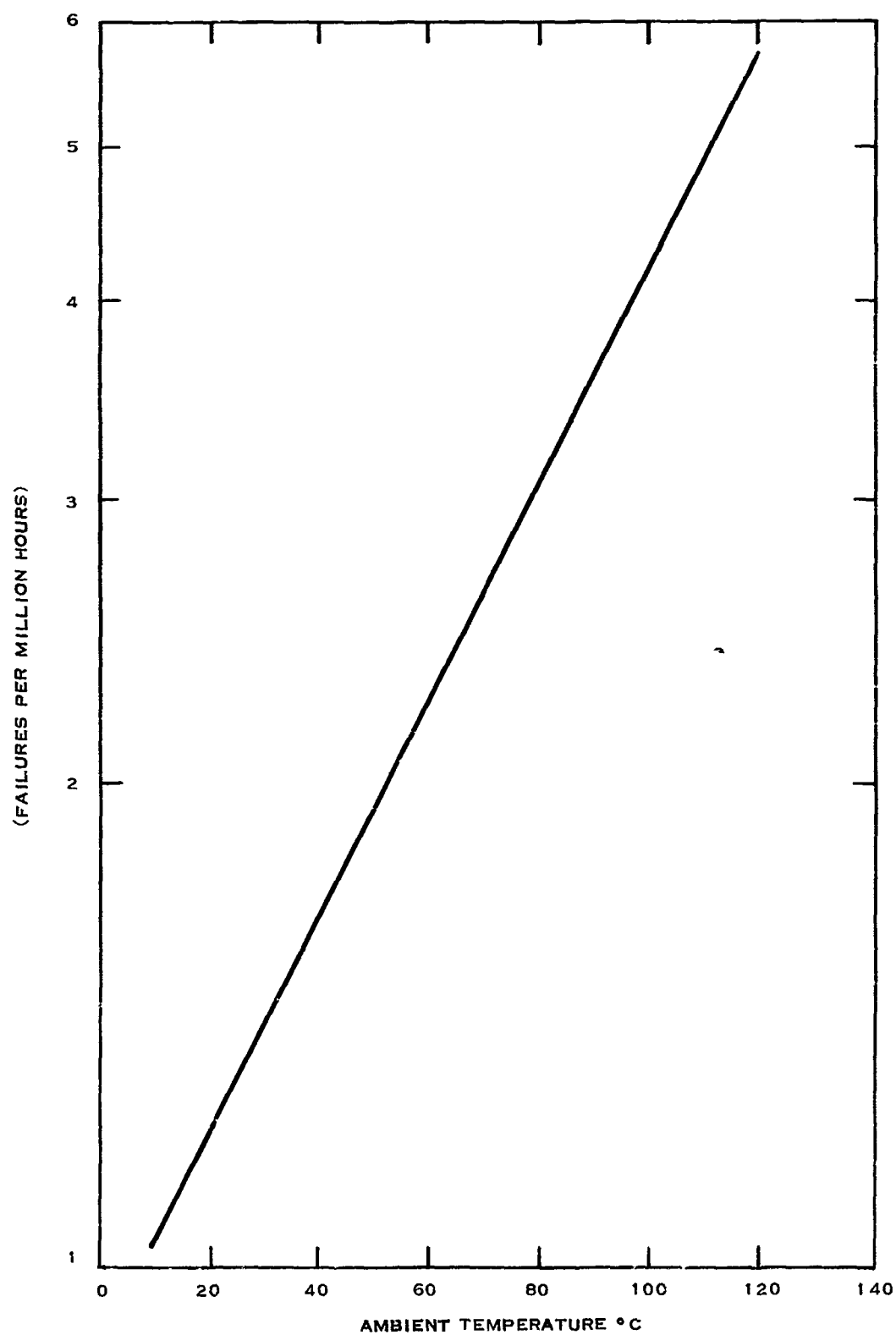
7.4.1 Temperature

An estimate of the effects of temperature on part reliability is depicted in Figure 7-5. This relation is calculated from MIL-HDBK-217A temperature relations. Part failure rates at several temperatures are weighted according to typical frequency of usage in a radar. The sum of the weighted failure rates for high-population parts is plotted versus temperature. The plot shows that failure rate may be increased by a factor of two for a 45°C rise. This is in close agreement with Texas Instruments data on transistors, which indicates 50°C for a factor of two in the temperature region between 25°C and 125°C.

The above information should be tempered with the thought that it is based on standard parts, rather than high-reliability burned-in parts as should be used for the CAT radar. The temperature effect on such parts may not be as pronounced, because many failure modes having activation energies below 125°C may have been eliminated by reliability screening.

7.4.2 Radiation

Radiation in the space environment varies from point-to-point. With reference to earth orbits, specific points must be associated with specific times to define the true radiation environment. Since the danger regions and their locations are defined and periods of exposure can, to some degree, be controlled by proper selection of orbit or space trajectory, it is assumed that exposure to radiation regions will be minimized. The thresholds for radiation effects sufficient to degrade electronic and electromechanical parts are relatively high when compared to the radiation levels experienced in the normal space environment. However, should the orbit or penetration angle into space be of such a nature as to prolong exposure in the denser radiation regions, such as the Van Allen and artificial belts, thresholds could be exceeded in a relatively short period of time. In addition to exposure, the time to exceed the threshold would also be a function of the amount and type of shielding employed in the vehicle and the antenna array. Table VII-4 presents the types of degradation which could occur should radiation thresholds be exceeded as a result of over-exposure.



63926

Figure 7-5. Average Component Failure Rate

Table VII-4. Effects of Excessive Radiation

Device Name	Effects
Antennas	Surfaces annealed, possible crystal growth, impurities migrated.
Capacitors	Capacitance changed, dissipation factor and leakage increased
Coaxial connectors	Conductivity and leakage increased
Coaxial relays	Conductivity and leakage increased
Couplers, directional	VSWR and insertion loss changed
Diodes	
Switching	Highly susceptible to radiation, leakage current increased, minority carriers degraded
Varactor	Leakage current increased, minority carriers degraded
Zener	Leakage current increased, minority carriers degraded
Inductors	Possible insulation damaged, remanence and permeability reduced, coercive force increased
Junction boxes	Conductivity increased (ceramic insulator assumed)
Mixers	(Same effects as for semiconductors)
Resistors	
Composition	Possible resistivity changed, probably negative (resistance increased)
Film	Similar to aging, resistivity changed, probably positive (resistance reduced)
Wire-wound	Possible insulation damage
Solder joints	Possible whisker growth and crystalization
Transformers	Possible insulation damaged, remanence and permeability reduced, coercive force increased

Table VII-4. Effects of Excessive Radiation
(Continued)

Device Name	Effects
Transistors	Leakage current induced, gain decreased
Traveling-wave tubes	Possible detuning and seal degradation
Solar cells	Transmissivity of cover glass changed, available power reduced, specular response changed, adhesive damaged
Mechanical components	Film layers including oxides removed, possible cold-welding if under pressure, some diffusion of metal to metal where intimate contact exists, surface annealed, possible crystal growth, some migration of impurities in parent metals, possible hydrogen embrittlement.*

* Molecular hydrogen is generated as a product of radiation when protons and electrons (primary or secondary) combine on a surface after losing their radiation energy.

Tests have shown that semiconductors are the most susceptible component to radiation degradation. Their radiation threshold is reached at approximately 10^2 ergs/g. This is approximately four times greater than the presently accepted tolerance rate or radiation dose for human beings. Other component radiation thresholds range from three to four magnitudes greater than the semiconductor.

As was previously mentioned, the equipment exposure is not anticipated for periods of time sufficient to exceed the minimum threshold. For those periods that it is exposed, the protection offered by the vehicle skin, structure, and system packaging will normally be sufficient to prolong the longevity of the equipment far beyond that required by the mission life. In conclusion, the effects of radiation on the system component parts can be considered insignificant.

7.4.3 Shock, Vibration, and Thermal Cycling

Conditions associated with shock and vibration are not considered to be significant contributors to system unreliability for space equipment. Exposure to these conditions is normally encountered only during the launch phase. Good design practices developed as a result of extensive vibration testing have proven themselves to be adequate in controlling mechanical stressing during the critical launch period.

Thermal design practices of space equipment have now developed to the extent that they are considered generally successful. Telemetry data from thermal sensors in spacecraft have indicated that thermal design objectives have been satisfied. Further telemetry data has demonstrated that the space environment thermal gradients are controllable.

Improper application of thermal techniques may result in far more serious problems in space than would be encountered in a suborbital environment. To be more specific, improper matching of electronic and electromechanical thermal coefficients, coupled with the continual thermal cycling experienced in the space environment can result in continual internal and external stressing of components. In addition, improper control of temperature excursions may result in the following effects: work hardening of material, sodium ion migration, chemical reaction due to impurities, and surface tracking. In conclusion, adequate design and fabrication controls based on present space technology will prevent significant degradation of system reliability which may result from vibration, shock, and thermal effects in the space environment.

7.4.4 On-Off Cycling

The type mission under consideration may or may not call for continuous operation, consequently an effort will be made to evaluate possible influences resulting from on-off cycling that have the potential to degrade reliability.

Basically, two areas exist that should be considered: effects resulting from thermal cycling; and circuit degradation which can result from transients and parametric drift during circuit stabilization. Controls such as good design practices and extensive circuit testing have proved adequate to control the latter.

Effects of thermal cycling, discussed in the previous paragraph, to some degree will be accelerated as a result of on-off cycling. As was pointed out, previously proper selection of materials, components, and good design and manufacturing practices can maintain the rate of degradation to a level that is considered insignificant, even when considering the maximum mission time. The preceding comments do not go without sound basis as evidenced by information obtained from a report prepared for FARADA pertaining to the subject of space vehicle equipment reliability. The report provides actual data which supports the concept of minimum degradation of reliability resulting from on-off cycling. The report disclosed that a data encoder designed and built by Texas Instruments has operated in the space environment for approximately 27,664 hours (or 3 years) without failure or apparent degradation. This particular system has operated continuously, however it is coupled with a transmitter which has been on-off cycled over the three-year period with no apparent reliability degradation.

7.4.5 Meteoroids

As a result of the prolonged mission time in the space environment and the frontal size of the exposed phased-array antenna, it is felt that careful consideration should be given to the potential reliability degradation of the array and substituent elements resulting from meteoroid collision. Therefore, the approach that will be employed will be directed toward determining the probability of collision with meteoroids of sufficient size (that is, energy) to cause degradation or permanent damage.

Consider the face of the antenna array to be a continuous plate. The criteria for penetration of such a plate by a spherical particle at meteoroid velocities is the Herman Jones penetration criteria.

$$t = 0.6K \left(\frac{6m\zeta}{\pi \zeta_+^2} \right)^{1/3} \log_e \left(1 + \frac{K_o^{2/3} B}{4} \right). \quad (7-16)$$

Since spalling of the back surface of the plate may occur without penetration, we will assume conservatively that $K = 4$ where:

- t = material thickness
- K = penetration constant, normally 1.5
- m = particle mass
- ζ = particle density = 0.44 gm/cm³
- ζ_+ = plate density = 2.7 gm/cm³ for Al
- $K_o = \zeta/\zeta_+$
- $B = \zeta_+ V^2/H$, where H = Brinell hardness
= 40 for Al
- V = velocity of meteoroid in units of km/sec

units are cgs unless otherwise specified. The cumulative number (N) of meteoroids with mass (m) or greater which strike a randomly oriented surface of one square meter in one second near earth is given by the Whipple equation

$$\log_{10} N = -14.48 + 2.68 \log_{10} \left(\frac{0.44}{\zeta} \right) - 1.35 \log_{10} m$$

for

$$\begin{aligned} \zeta &= 0.44 \text{ gm/cm}^3 \\ \log_{10} N &= -14.48 - 1.34 \log_{10} m \end{aligned} \quad (7-17)$$

$$N = \frac{\text{impacts}}{\text{meter}^2 \text{ sec}} \text{ of meteoroids of mass } \geq m.$$

The probability of n failures during time increment T for a system having a failure rate λ is expressed by a Poisson probability distribution as

$$P(n) = \frac{(\lambda T)^n e^{-\lambda T}}{n!} .$$

Let n = zero failures

$$P(0) = e^{-\lambda T} .$$

The system failure rate contribution (module failures resulting from collision with meteoroids) can be obtained from the results of Equations (7-16) and (7-17) which give the number of damaging meteoroids per m^2 sec and A (the area of the array surface). The factors NA are multiplied by 3.6×10^3 to convert from failures per sec to failures per hour.

$$\lambda = NA (3.6 \times 10^3) .$$

To determine the failure-rate contribution λ_m to the rf module, the area of the array is divided by the number of modules in the array. It should be recognized that this is a conservative estimate since a significant percentage of the array surface is comprised of the H clamps used to secure the modules to the array.

Since the number of rf module failures resulting from meteoroid collision is a function of module material thickness [reference Equation (7-16)] and frontal area of the rf module which is inversely proportional to the frequency of the module, a family of curves has been developed to relate the λ_m as a function of material thickness for a series of different frequencies (Figure 7-6).

An example, to demonstrate the λ_m contribution to the overall rf module, of 1.5 GHz will be used. Based on present technology a failure rate λ_{RF} of 18×10^{-6} parts per hour for a comparable rf module is considered to be reasonable.

$$\frac{\lambda_m}{\lambda_{RF}} = \frac{0.022 \times 100}{18} = 0.122 \text{ percent.}$$

Relating λ_m for a material thickness 1 mm (now employed for rf modules) and the rf module failure rate λ_{RF} indicates a 0.122-percent contribution to the overall rf-module failure rate. Since the percentage of the failure rate contribution resulting from meteoroid collision is much lower than the uncertainty of the rf module failure rate, it can be considered insignificant.

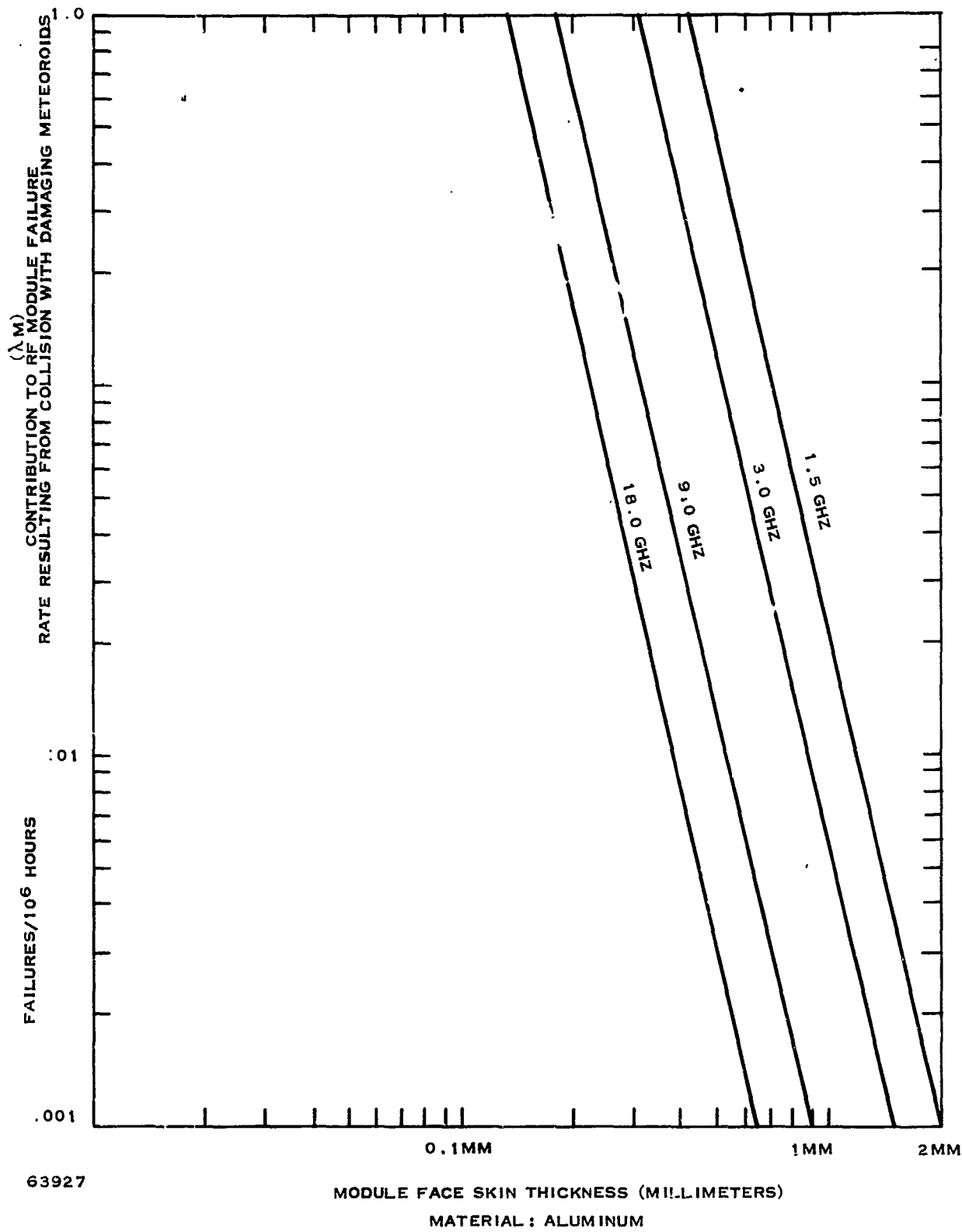


Figure 7-6. Module Failure Rate Due to Meteoroids

7.4.6 Off-Duty Reliability

Review of RADC-TR-67-307, entitled Dormant Operating and Storage Effects on Electronic Equipment and Part Reliability, indicates that the average catastrophic part failure rate in the average system environment under nonoperating conditions was found to be six failures per billion hours.

The seriousness of this off-duty failure rate is proportional to the ratio of off-to-on time. If a failure rate λ_x is the apparent increase in operational part failure rate due to off-time failures it is related to duty factor (d) by

$$\lambda_x = 6 \times 10^{-9} \frac{1-d}{d} \text{ failures per hour.} \quad (7-18)$$

If the average CAT part failure is around 0.3 failures per 10^6 hours, the effect of λ_x is not serious for duty factors greater than 10 percent.

7.5 Reliability Summary

Both space-fed and corporate-fed radar systems are capable of providing moderate-to-high reliability for a CAT mission. Use of maintenance or extensive redundancy appears unnecessary unless extremely high mission reliability is needed. The solid-state active module system has an order of magnitude reliability advantage over the vacuum-tube space-fed system (unless the latter resorts to full redundancy for several functions) in the rf power chain.

The functional blocks of the radar system which adapt to partial redundancy without severe weight and size penalties are the phased-array and the high-current dc power converter. Antenna array, useful life is directly proportional to the decibels of signal-to-noise margin available, inversely proportional to the antenna module failure rate, and nearly independent of the actual number of modules in the antenna. With reasonable care, the partially redundant functions will have near-perfect reliability during the CAT mission.

Outside the rf power chain, the most significant reliability factor is complexity. Temperature, temperature cycling, and vibration are important environmental factors, while radiation and meteoroid flux are not expected to be of concern.

8.0 SYSTEM SYNTHESIS AND OPTIMIZATION

8.1 Introduction

The purpose of this section is to use the results of the preceding parametric studies and evolve systems representative of the best attainable under limitations mentioned in previous sections concerning size, weight, and power. As a minimum requirement, two systems will be synthesized. The first such system will be a corporate-fed array which has the best performance characteristics that are attainable under conditions that will be defined during the development. The second system will be the best possible space-fed array under the same limiting conditions. Since much of the radar system is independent of the antenna being space-fed or corporate-fed, the corporate-fed array development will be accomplished first, the space-fed array next, and finally the development of the remainder of the radar system which is common to either type of array. The development of these systems is carried out under the assumption that the best system is one which maximizes the signal-to-noise ratio while not exceeding those limitations placed on the radar system by the spacecraft.

8.2 System Synthesis

8.2.1 Array Synthesis

8.2.1.1 Corporate-Fed Array

As a point of departure, consider the range equation from Berkowitz⁸³ for the case of a radar where the receiver bandwidth is matched to the transmitter pulsewidth

$$(S/N)_{db} = (P_t)_{dbw} + (\tau)_{dbsec} + 2(G)_{db} + 2(\lambda)_{dbcm} + (\sigma)_{dbm}^2 - 4(R)_{dbnmi} - (NF_o)_{db} - (L)_{db} \quad (8-1)$$

where the terms have the usual definitions given earlier in the report.

Recall from subsection 3.2 that peak power per module and noise figure were determined for each of the six frequencies of interest. Similarly, the steady-state power consumption of a module and its efficiency in converting dc to rf power were given. Because this data is highly relevant to the present development, it is summarized again in Table VIII-1 parametric in frequency.

The module input power can be expressed as

$$P_{im} = P_{ss} + \frac{P_{pm}}{\eta} \tau f_r, \quad (8-2)$$

where

P_{ss} = steady-state power consumption of module when not transmitting

P_{pm} = peak radiated power per module

η = efficiency in converting dc to rf

τf_r = duty cycle (pulse length—prf product).

Equation (8-2) shows the dependence of module power consumption upon P_{ss} , P_{pm} , and η which are seen in Table VIII-1 to be determined by frequency and by duty cycle which is a parameter that can be traded to maximize system performance.

To a large extent, the relative performance of a system is defined by the parameter $(\Sigma)_{db}$ where

$$(\Sigma)_{db} = (P_t)_{dbw} + (\tau)_{dbsec} + 2(G)_{db} + 2(\lambda)_{dbcm} - (\overline{NF_o})_{db}. \quad (8-3)$$

A large part of system optimization is concerned with maximizing this parameter. This importance is demonstrated when the radar range equation is derived in terms of $(\Sigma)_{db}$ from Equation (8-1) and Equation (8-3) in a form that will be used later in this section.

$$\{(S/N)_{db} + (L)_{db}\} = \{(\Sigma)_{db} + (\sigma)_{dbm^2}\} - 4(R)_{dbnmi}. \quad (8-4)$$

Table VIII-1. Solid-State Module Parametric Dependence on Frequency

f	λ	NF_o	P_{pm}	η	P_{ss}
GHz	CM	db	W		W
1.5	20	4.7	17.5	0.306	0.46
2.25	13.3	5.9	8.7	0.150	0.46
3	10	7.3	8.3	0.145	0.41
6	5	8.8	5.3	0.0927	0.335
9	3.33	8.9	2.6	0.0447	0.335
18	57	13.0	0.63	0.0108	0.760

Equation (8-4) is useful in that inherent trades between S/N and system losses are demonstrated, trades between target cross section and Σ are indicated, and both trades are shown in perspective relative to the range dependence.

The system loss term is not included in Σ because several constituents of L depend on whether the mode is acquisition or tracking. In the interest of a general derivation, this should not be held against one candidate system over another. One needs to remember that the loss term will be the last term considered in the optimization procedure.

The goal is to obtain those parameters which result in the maximum value for Σ as defined in Equation (8-3). The plan of attack is to first maximize Σ for each frequency under consideration, then selecting the frequency, hence system, which has the greatest Σ .

In order to maximize Σ , it is necessary to place some restrictions on the system, else the solution would be the trivial case where the array had infinitely many elements. The reality of the situation dictates that there are at least three upper bounds on system parameters.

(a) Weight Limit. The number of modules is bounded above by that number which exceeds the weight limit.

$$N_{WL} = \frac{W_{avail}}{W_m}, \quad (8-5)$$

where

W_{avail} = available weight

W_m = weight per module

N_{WL} = maximum number of modules based on weight.

(b) Size Limit. The number of modules is bounded above by that number which exceeds the size limit.

$$N_{SL} = \frac{A_{avail}}{A_m}, \quad (8-6)$$

where

A_{avail} = available area

A_m = area occupied per module (approx. $\frac{\lambda^2}{4}$)

N_{SL} = maximum number of modules based on size.

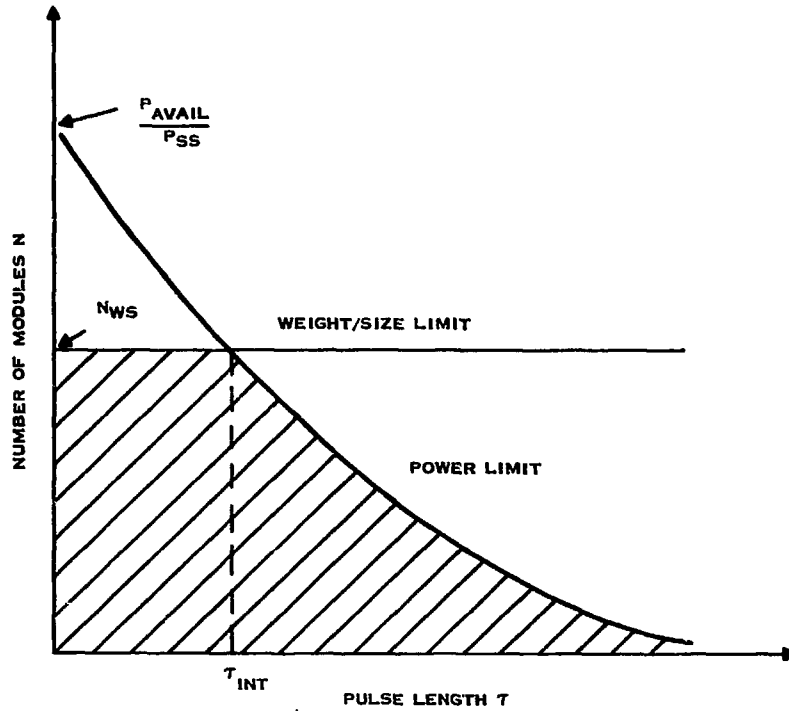
(c) Power Limit. The number of modules is bounded above by the amount of power available to the modules. The dependence of this bound is not as straightforward as (a) and (b), because the power consumed per module varies with duty cycle. The upper bound based on power consumption is given by

$$N_{PL} = \frac{P_{avail}}{P_{im}} = \frac{P_{avail}}{P_{ss} + \frac{P_{pm} \tau f_r}{\eta}}, \quad (8-7)$$

where

P_{avail} = available power

N_{PL} = maximum number of modules based on power.



63928

Figure 8-1. Array Limitation Dependence on Pulse Length

It is useful at this point to show the cumulative effect of these bounds as a function of pulse length (assuming a constant f_r) which is depicted in Figure 8-1. In this figure, N_{WS} is the smaller of the weight and size limits and is assumed to be less than P_{avail}/P_{ss} . If it is larger, it then plays no part in the optimization procedure.

The shaded area in Figure 8-1 constitutes the permissible region for an optimum solution to occur. The task is to define the parameter which is to be maximized so that the search for the optimum point can begin.

Given that the frequency, hence wavelength, is to be held constant, one can see in Equation (8-3) that Σ is maximized when $E = P_t \tau G^2$ is maximized.

Since

$$P_t = NP_{pm} \quad (8-8)$$

and

$$G = 0.8\pi N, \quad (8-9)$$

one finds that

$$E(n, \tau) = 6.32 P_{pm} N^3 \tau \quad (8-10)$$

is the parameter to be maximized.

Obviously, the maximum value of $E(N, \tau)$ occurs somewhere on the upper boundary of the shaded area in Figure 8-1. Also, the maximum obviously

does not occur on the line to the left of the point of intersection of the N_{WS} line with the hyperbola because moving to the right increased the value of $E(N, \tau)$ linearly with τ .

Consequently, the optimum point is either the point of intersection of the horizontal line with the hyperbola, or some point on the hyperbola to the right of that intersection point. The problem now is to determine the circumstances under which the latter event could occur.

To do this, a point must be found on the hyperbola where $E(N, \tau)$ reaches a maximum. When the point is found, observe if the peak point is to the left of the intersection point (in which case the intersection point is optimum) or to the right of the intersection point (in which case the peak point is optimum).

Combining Equations (8-7) and (8-10), one can eliminate N and determine an expression for $E[N(\tau), \tau]$ in terms of τ alone which is valid everywhere on the hyperbola of Figure 8-1.

$$E[N(\tau), \tau] = \frac{6.32 P_{pm} P_{avail}^3 \tau}{(P_{ss} + \frac{P_{pm} f_r \tau}{\eta})^3} \quad (8-11)$$

To determine the maximum value of E , wrt τ must be differentiated, equated to zero, and solved for τ .

$$E_{peak} = 0.935 \frac{\eta P_{avail}^3}{f_r P_{ss}^2} \quad (8-12)$$

which occurs when

$$\tau_{peak} = \frac{\eta P_{ss}}{2 f_r P_{pm}} \quad (8-13)$$

and the array would consist of

$$N_{peak} = \frac{2}{3} \frac{P_{avail}}{P_{ss}} \quad (8-14)$$

modules.

The previous three equations give the conditions for maximizing E , hence Σ , independent of the size/weight constraint. In order to factor this constraint into the procedure, it is necessary to determine the parameters at the point of intersection of the horizontal line and hyperbola in Figure 8-1.

Given N_{WS} (the lesser of the weight and size limits) which is less than P_{avail}/P_{ss} one finds that

$$\tau_{int} = \frac{\eta}{P_{pm} f_r} \left(\frac{P_{avail} - P_{ss}}{N_{WS}} \right), \quad (8-15)$$

which corresponds to N_{WS} elements in the array and

$$E_{int} = 6.32 \frac{\eta}{f_r} N_{WS}^2 (P_{avail} - N_{WS} P_{ss}), \quad (8-16)$$

being the value of the optimized parameter E at the point of intersection.

Once the optimum value for E is determined, Σ is found by

$$(\Sigma)_{db} = (E_{opt})_{db} + 2(\lambda)_{dbcm} - (NF_o)_{db}. \quad (8-17)$$

In order to summarize the development just discussed, a flow chart is included as Figure 8-2 which depicts each step in the optimization procedure. This flow chart was applied to the data given in Table VIII-1 in determining the value of τ for each of the frequencies under consideration. This data is tabulated in Table VIII-2. Several assumptions were necessary in order to begin the procedure. It was assumed that 1000 watts of power was available to the array, that the area available was the 4536 square inches mentioned in Sections 5.0 and 6.0 and that the array weight was limited at 350 pounds.

Figure 8-3 is included as an aid for computing the radar system performance as a function of range when Σ , σ , and L are known.

For the optimum value of Σ in Table VIII-2, namely 76.81 dB, Figure 8-4 is included to show achievable system performance for the optimum corporate-fed array under consideration.

8.2.1.2 Space-Fed Array

In the space-fed array, the concern is with deciding how much power to allot to the phase-shift modules and to the transmitter. That is, a tradeoff exists between antenna gain and transmitter power. The relationship between these parameters may be expressed by the following equation

$$N = \left(P_T - P_{PC} - \frac{P_{OA}}{\eta} \right) \frac{1}{P_n}, \quad (8-18)$$

where N = number of phase-shift modules

P_T = total power to transmitter and array, watts

P_{DC} = power to transmitter tube heater, watts

P_{OA} = average power output from transmitter, watts

η = efficiency of transmitter

P_n = average power required by phase-shift module, watts.

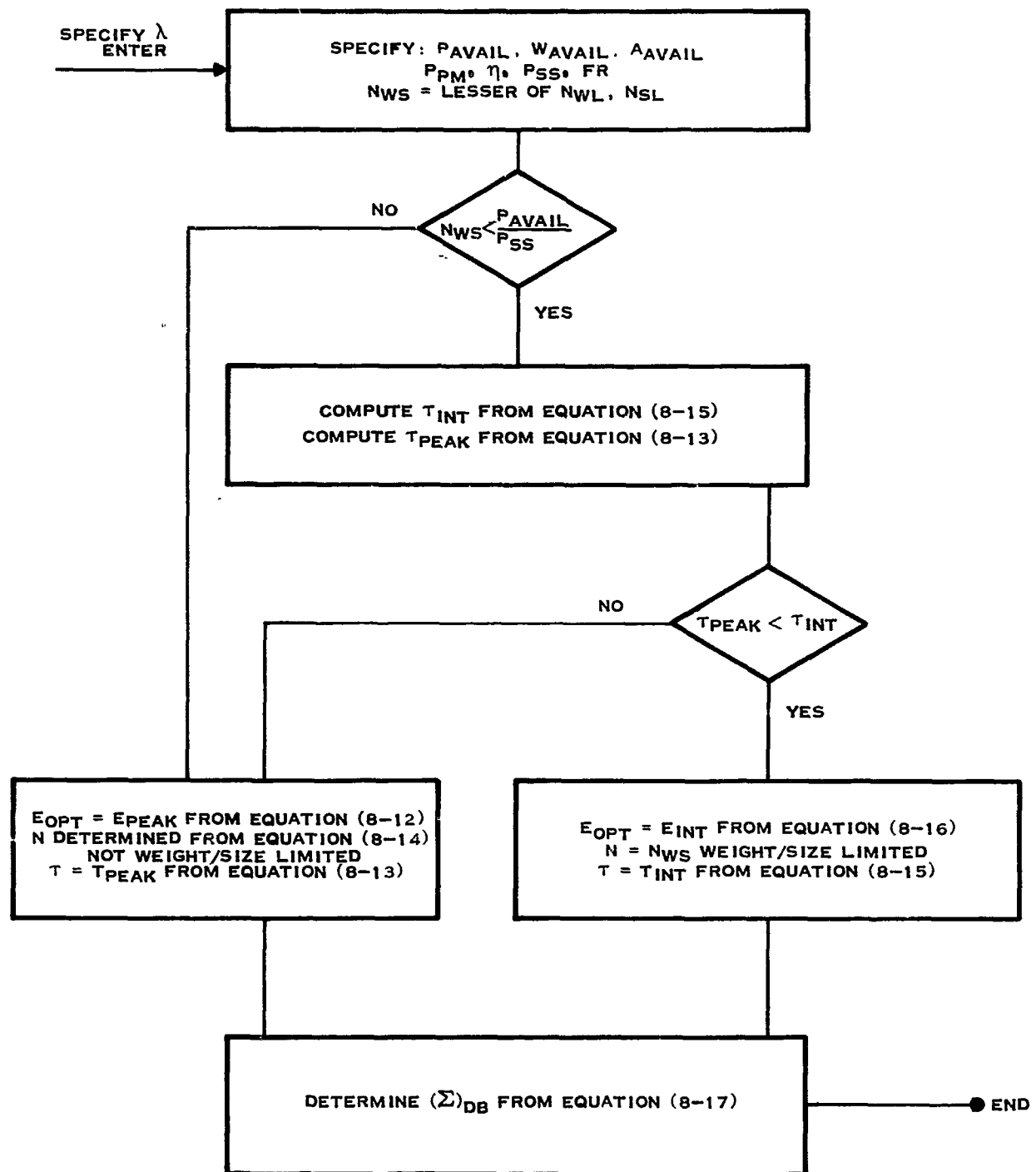
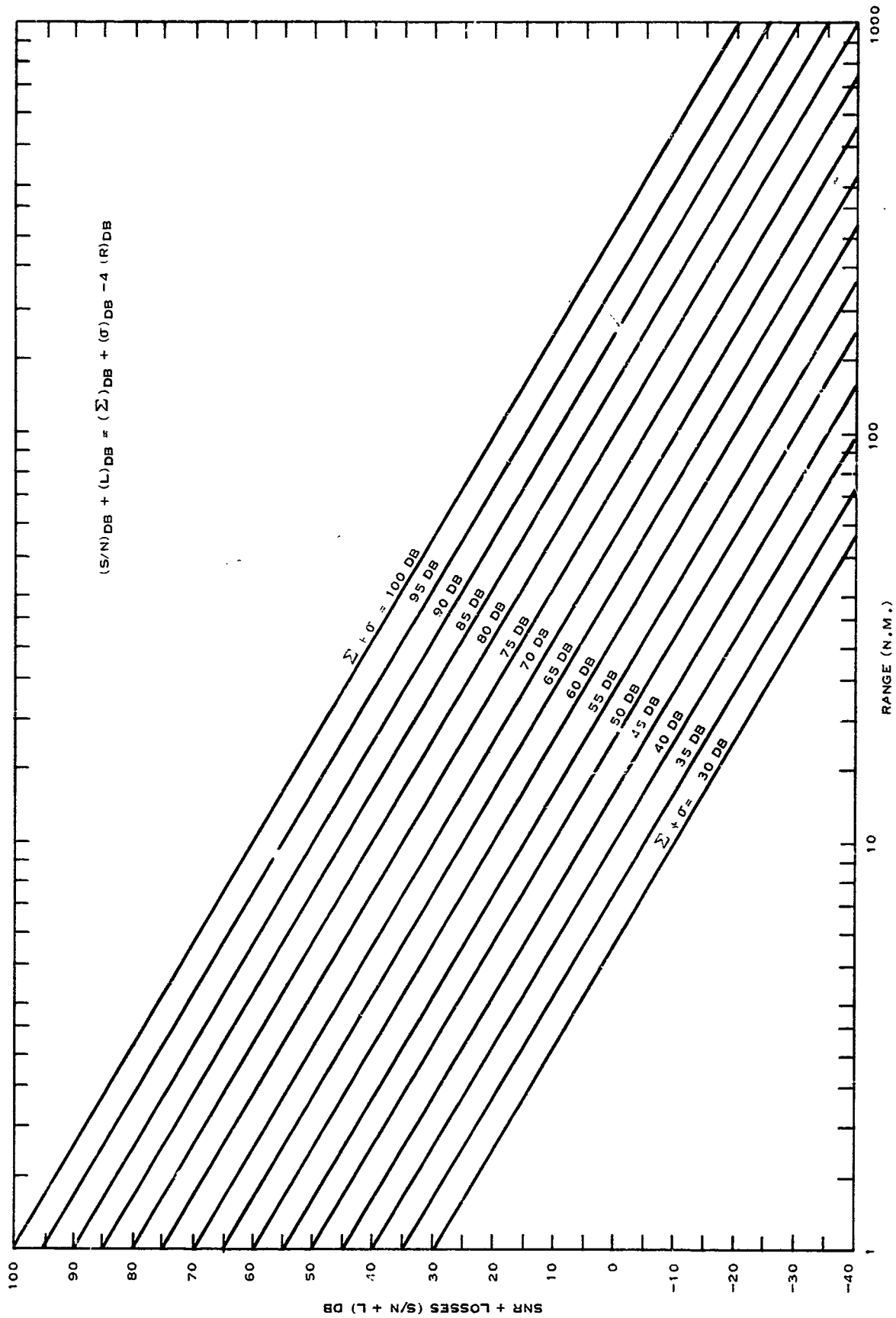


Figure 8-2. Optimization Flow Chart



63930

Figure 8-3. Range Dependence on Σ , σ , and L

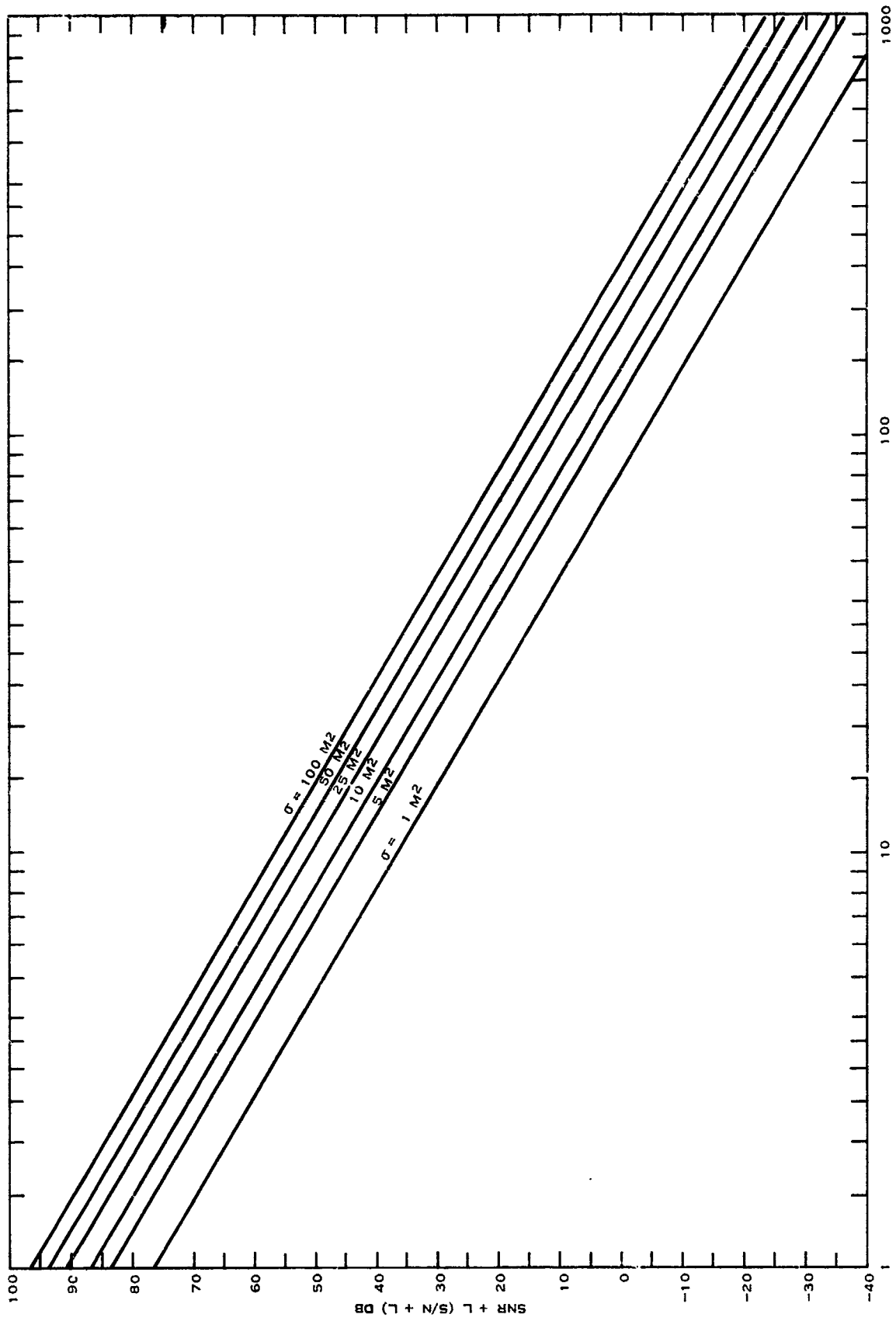


Figure 8-4. Optimum Corporate-Fed System Performance

63931

Table VIII-2. Tabulation of Optimum Σ for Each Frequency

f GHz	N _{SL}	N _{WL}	N _{WS}	τ_{int} μsec	τ_{peak} μsec	E _{int}	E _{peak}	(E _{opt}) _{db} dB	(Σ) _{db} dB	N Elements	W lbs
1.5	292	593	292	129.4	10.05	3.56×10^5	N.A.	55.51	76.81	292	173
2.25	662	1130	662	45.3	9.90	7.22×10^5	N.A.	58.58	75.16	662	204
3.0	1170	1167	1167	19.5	8.95	1.626×10^6	N.A.	62.11	74.81	1167	350
6.0	4680	2120	2120	6.44	7.32	N.A.	1.927×10^6	62.85	68.05	1990	328
9.0	10,600	3180	3180	none	7.20	N.A.	9.30×10^5	59.68	61.22	1990	219
18.0	42,000	5650	5650	none	16.3	N.A.	4.37×10^4	46.40	37.84	877	54

The system design problem is to maximize radar range. Which means maximize the expression

$$R^{1/4} \propto G^2 P_{OA} , \quad (8-19)$$

or

$$\frac{1}{K} R^{1/4} \propto N^2 P_{OA} . \quad (8-20)$$

This is an example of a constrained maximum. Where maximizing the expression in Equation (8-20) is desirable and where the variables are constrained by Equation (8-18). This may be solved by use of the Lagrange multiplier. Let

$$f(N, P_{OA}) = N^2 P_{OA} \quad (8-21)$$

and

$$\phi(N, P_{OA}) \equiv N - K_1 - K_2 P_{OA} = 0, \quad (8-22)$$

where the fixed quantities of Equation (8-18) are treated as constants,

$$K_1 = \frac{P_T - P_{DC}}{P_n}, \quad (8-23)$$

$$K_2 = \frac{1}{P_n \eta} . \quad (8-24)$$

Now, if λ is so chosen that

$$\left. \begin{aligned} \frac{\partial f}{\partial N} + \lambda \frac{\partial \phi}{\partial N} &= 0 \\ \frac{\partial f}{\partial P_{OA}} + \lambda \frac{\partial \phi}{\partial P_{OA}} &= 0 \\ \phi(N, P_{OA}) &= 0 \end{aligned} \right\} , \quad (8-25)$$

then the necessary condition for maximum will be satisfied.

Solving Equation (8-25),

$$\left. \begin{aligned} 2N P_{OA} + \lambda &= 0 \\ N^2 + \lambda K_2 &= 0 \\ N - K_1 + K_2 P_{OA} &= 0 \end{aligned} \right\} . \quad (8-26)$$

Solving these three simultaneous equations,

$$N = \frac{2}{3} K_1 \quad (8-27)$$

$$P_{OA} = \frac{1}{3} \frac{K_1}{K_2} . \quad (8-28)$$

Substituting Equations (8-23) and (8-24) into (8-27) and (8-28) and letting the variables be noted as N' and P'_{OA} to signify these are the values of N and P_{OA} at which $R^{1/4}$ is maximized,

$$N' = \frac{2}{3} \left(\frac{P_T - P_{DC}}{P_n} \right) \quad (8-29)$$

and

$$P'_{OA} = \frac{1}{3} \eta (P_T - P_{DC}) . \quad (8-30)$$

The relationship between these parameters is graphically shown in Figure 8-5. Assuming an antenna efficiency of 55 percent for a space-fed array, 1 kilowatt of power to the array and transmitter, 20 watts of continuous power to the transmitter tube, and transmitter efficiency of 20 percent (30 percent for the tube and 65 percent for the high-voltage modulator), the curves are plotted for fixed phase-shifter module power requirements. The reflective array module requires 0.6-watt average power. Thus, the number of modules used for maximum radar range, or S/N , is 1088. The transmissive array module requires 1.2 watts and N' here is 544 modules. The reflective array is four times better than the transmissive array if operated at maximum performance. Also note, that for less than N' modules, the arrays approach the same performance. A curve for a P_n of 0.1 watt is shown to evaluate array improvement if a low-power module were available. A likely candidate would be a lightweight ferrite-latching phase shifter which has low-average power requirements.

Further tradeoffs can be made between transmitted peak power, P'_p , and pulse length, τ . Since

$$P'_{OA} = P'_p (\tau f_r) = \frac{1}{3} \eta (P_T - P_{DC}) , \quad (8-31)$$

then

$$P'_p \tau = \frac{1}{3 f_r} \eta (P_T - P_{DC}) . \quad (8-32)$$

Taking the case of the reflective array using 0.6 watt per phase shifter and the parameters given in Figure 8-6, a set of curves are drawn for fixed transmitter efficiency. Using a transmitter efficiency of 20 percent, an 83-kilowatt pulse of length 2 μ sec or an 8.3-kilowatt pulse of length 20 μ sec could be used.

The optimum number of modules, N' , needs to be further tested to stay within specified array size and weight limits. This test may be made by comparing N' to N_{WL} [Equation (8-5)] and N_{SL} [Equation (8-6)].

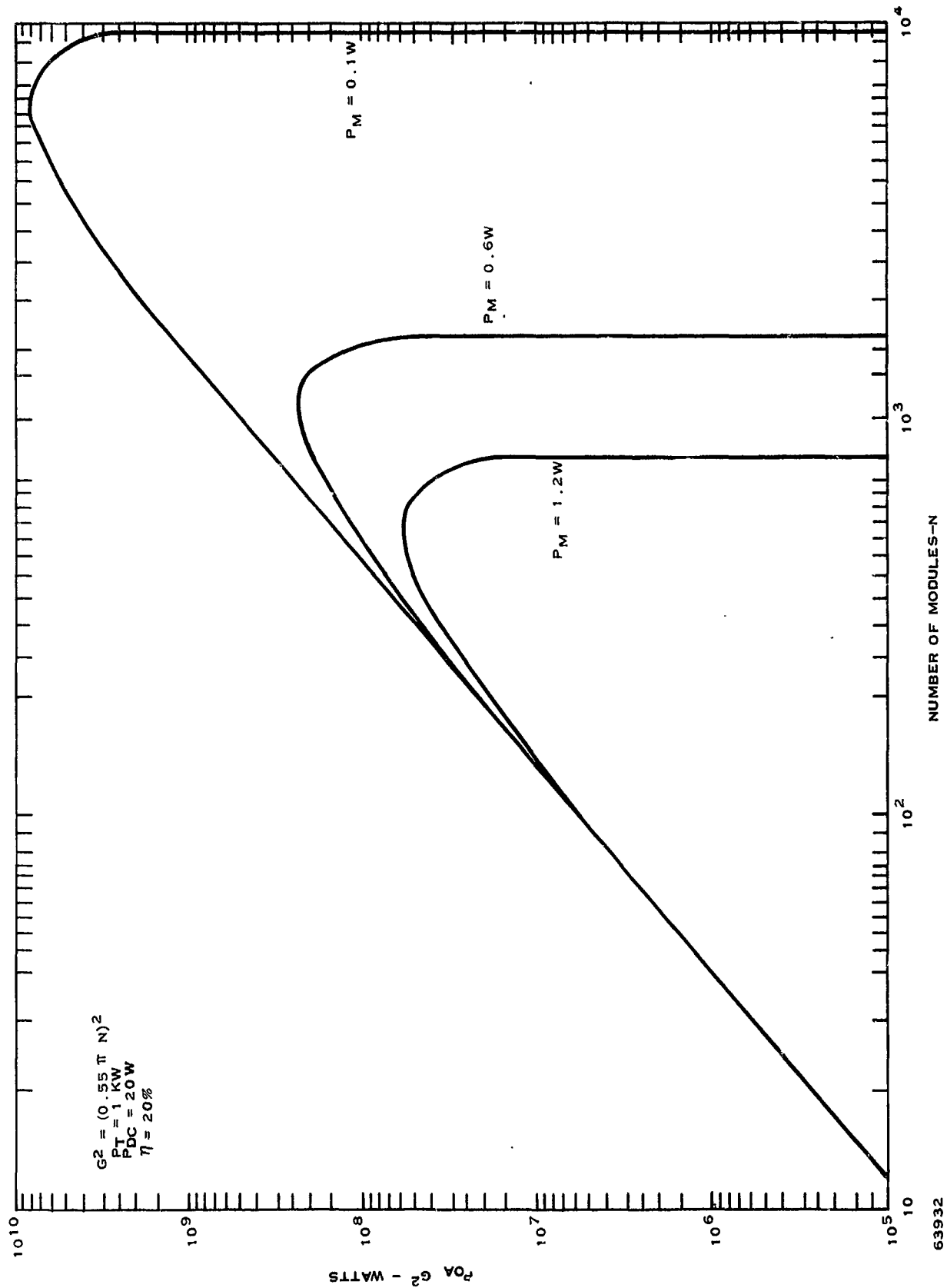


Figure 8-5. Average Tx Power-Gain² Product Versus Number of Modules for Space-Fed Array

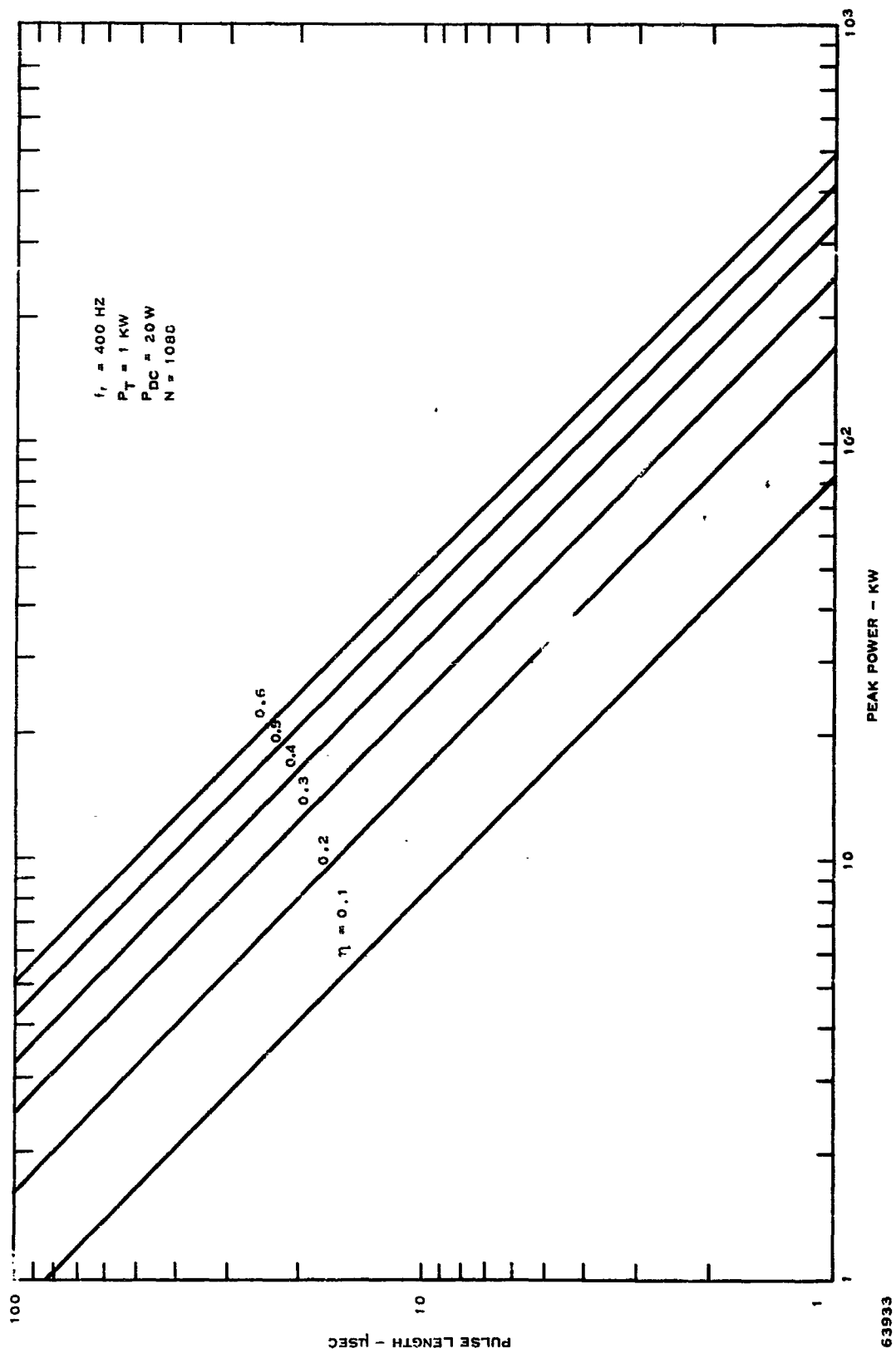


Figure 8-6. Peak-Power and Pulse-length Tradeoff for Space-Fed Array

Table VIII-3 summarizes the optimum space-fed array parameters at each frequency using the assumptions shown.

The parameter $(\Sigma)_{db}$ may be developed for the space-fed arrays at each frequency. Assuming an antenna efficiency of 55 percent and using Table VIII-3 then $(\Sigma)_{dB}$ is tabulated in Table VIII-4. These values will be used in the next section to compare all systems performance.

8.2.2 Radar System Synthesis

In synthesizing a signal processing system for the radar, one must consider the two aspects of the problem: detection and tracking. A range-tracking loop can perform both operations. For each beam position, the tracking gate can be swept over the full range sector; if it locks on a target, a detection is constituted. For a tracking loop of reasonable bandwidth; however, the sweeping rate would have to be slow enough so that the gate dwells on each range bin for several interpulse periods. Clearly, searching each of the beam positions over a 200-nm range would require much time in order to search the entire volume sector. Similarly, a threshold device can furnish range information, but does not furnish a convenient method of applying range gating to the angle-track loop. Thus, the system which seems most suitable would consist of a conventional threshold device (such as that of Figure 3-112) to allow search of the entire volume sector in a reasonable time, followed by a range-tracking loop which would be activated by the detection decision and would search, at most, the entire range sector of only one beam position. Lockup of the range-tracking loop would then activate the angle-track loop.

Selection of a particular closed-loop range tracker depends, to a certain extent, upon the mission envisioned. Figure 4-12 shows the split-gate tracker somewhat superior to the separated-gate tracker for equal $B_r \tau$ products (which implies equal detection performance) and equal f_r/β_ℓ' ratios (which implies equal dynamic lag errors for equal target velocities). However, Figure 4-14 shows the split-gate tracker somewhat inferior to the separated-gate tracker with respect to maximum permissible target velocity. The selection of one or the other of the loops depends on which of these aspects is more important. Assuming that these aspects are equally important, then the thermal noise improvement of split-gate over separated-gate is approximately 60 percent while the improvement in maximum trackable velocity of the separated-gate over the split-gate is approximately 9 percent. Based on these considerations, the split-gate system appears to be the more attractive of the range trackers. Selection of the f_r/β_ℓ' ratio for the split-gate tracker involves again a tradeoff between maximum trackable velocity and lag error in one case, and thermal noise error in the other case. Rather than assigning an arbitrary degree of importance, which will change with different missions, to each of these criteria, it seems more reasonable to examine the requirements of a particular mission, and base the selection of the f_r/β_ℓ' ratio on these requirements.

Table VIII-4. Tabulation of Optimum for Each Frequency for the Reflective and Transmissive Arrays

Reflective Array									
Frequency (GHz)	P _{TA} W	P _{TA} dB W	N	N dB	2(N) dB	2(G) dB	2(λ) dB	NF _o dB	Σ dB
1.5	95	19.78	292	24.65	49.30	54.06	26	6.9	65.94
3	38	15.80	1088	30.36	60.72	65.48	20	7.4	67.88
6	36	15.56	1088	30.36	60.72	65.48	14	8.2	60.84
9	33	15.18	1088	30.36	60.72	65.48	10.44	9.0	56.10
18	29	14.62	1088	30.36	60.72	65.48	4.44	10.8	47.74
Transmissive Array									
Frequency (GHz)	P _{TA} W	P _{TA} dB W	N	N dB	2(N) dB	2(G) dB	2(λ) dB	NF _o dB	Σ dB
1.5	74	18.69	292	24.65	49.30	54.06	26	6.9	64.85
3	38	15.80	544	27.36	54.72	59.48	20	7.4	61.35
6	36	15.56	544	27.36	54.72	59.48	14	8.2	54.84
9	33	15.18	544	27.36	54.72	59.48	10.44	9.0	50.10
18	29	14.62	544	27.36	54.72	59.48	4.44	10.8	41.74

Figure 4-15 shows the conopulse angle tracker to be significantly superior to either monopulse or conical scan with respect to thermal noise error, counterbalancing this advantage to a certain extent by crosstalk. As shown in Subsection 4.2.1, the rms crosstalk is not significant for nominal values of thermal noise error in the range- and angle-track loops. The lag error due to crosstalk can be made nonsignificant by restricting the range of permissible target velocities so that the lag error in the range loop is sufficiently small. Thus, it appears that the conopulse angle tracker is more attractive than either monopulse or conical scan, at least for a rendezvous mission where there will be suitable restrictions on target relative velocity.

In summary, the selected system will consist of a threshold detection device, a split-gate range tracker which is enabled by the detection decision, and a conopulse angle tracker which is enabled after the range tracker achieves lock. The system is shown in block diagram form in Figure 8-7.

8.3 System Optimization and Specification

The process of optimizing radar system performance at each discrete frequency under consideration has already been performed in Subsection 8.2 for both the corporate-fed active-element array and the space-fed array. The task remaining for this section is to choose the type of array and frequency which will be the preferred system.

The system performance is summarized to a large extent by the parameter Σ tabulated in Table VIII-2 for the corporate-fed arrays and in Table VIII-4 for the space-fed arrays. Since all the arrays have been constrained to use approximately 1100 watts of prime power, the system requirements are largely defined by the weight requirement which is shown in Table VIII-2 for the corporate-fed array portion and in Table VIII-3 for the space-fed array portion of the radar system. The radar system is heavier than these figures by approximately 60 lbs for the corporate-fed array and 50 lbs for the space-fed array case.

From the tables, one observes that the corporate-fed array is approximately 80 lb heavier at 1.5 GHz, 250 lb heavier at 6 GHz, and only 10 lb heavier at 18 GHz. Consequently, the weight penalty influences the decision in favor of the space-fed array.

The system performance index Σ is seen to have a maximum for the case of a 1.5-GHz corporate-fed array and is 9-dB worse for the best performing space-fed array which is the 3-GHz reflective array. Since S/N ratio is directly influenced by Σ for the same external conditions, this would show up as a 9-dB difference in S/N ratio at the receiver. Because the corporate-fed array weight is sufficiently near that of the space-fed array weight at the frequency of 1.5 GHz, the 9-dB advantage in S/N ratio is considered to more than offset the fact that the corporate-fed array radar system outweighs the 3-GHz reflective array radar system 234 pounds to 177 pounds. For that reason, the recommended system is the 1.5-GHz corporate-fed active-element array radar system described in Table VIII-1.

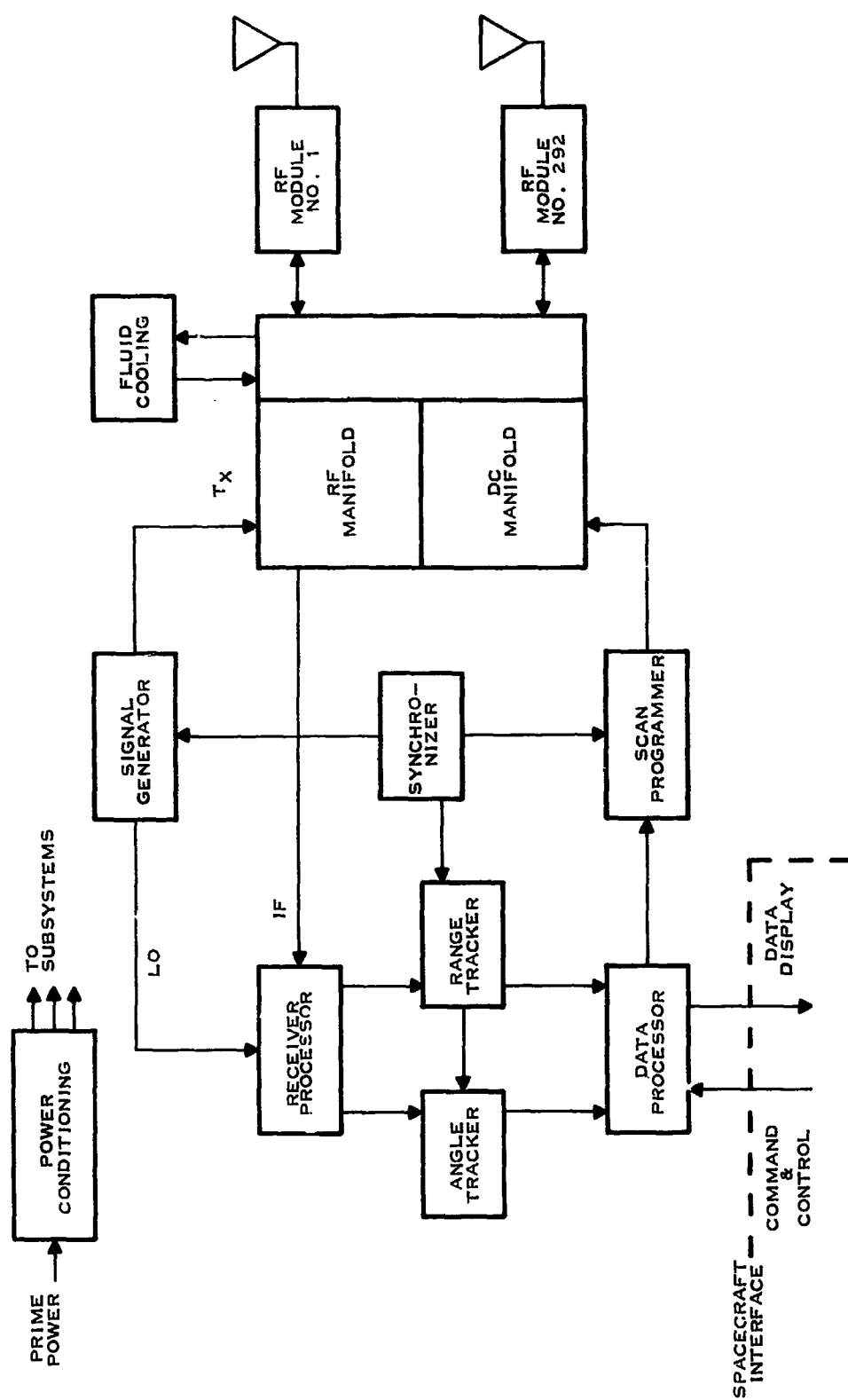


Figure 8-7. CAT Corporate-Fed Radar System

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The comparatively long pulse length of 129.4 μ sec does not prevent adequate tracking accuracy from being obtained when a pulse compression ratio of 100:1 is attained by linearly sweeping the carrier in a chirp fashion. Similarly, adequate angle tracking is possible with the 292-element array which achieves a beamwidth of approximately 6.6 degrees and conopulse processing is employed.

In tracking situations, the 129.4 μ sec pulse is too long to permit tracking within a 10-mile range so that it will be necessary to have a second pulse length option which is commanded by the range tracker when the range approaches 20 nm. At this range, the S/N ratio is sufficiently great to permit shortening the pulse length by a factor of five which will then permit tracking until the target closes within a 3-nm range.

It is somewhat surprising that this study has shown the relative desirability of the lower frequencies. This result is due to two facts. First, solid-state power generation is much more efficient at lower frequencies and second, the weight, size, and power limitations assumed for the system were sufficiently generous to allow the efficiency advantage at low frequencies to more than offset the antenna gain advantage at high frequencies.

If the weight and size limitations had been more severe, the optimum system frequency would shift to the higher frequencies, but with a corresponding loss in system performance.

Similarly, if the prime power available is increased sufficiently, it seems likely that higher frequencies would benefit more because the 1.5-GHz array is already using the full 6-foot aperture while the higher frequencies would greatly benefit both by more power per element and more elements.

Table VIII-5. CAT Radar System Specification

Frequency	1.5 GHz
Number of Array Elements	292
Array Type	Corporate-Fed, Active Element
Pulse Length (Uncompressed)	129.4 μ sec
Pulse Compression Ratio	100:1
Pulse Compression Technique	Chirp
Peak Power Per Module	17.5 watts
Total Peak Power	5.1 kW
System Noise Figure	4.7 dB
Antenna Beamwidth	6.6 degrees
Antenna Gain	28.7 dB
Range Tracking Scheme	Split Gate
Pulse Repetition Frequency	400 pps

Table VIII-5. CAT Radar System Specification (Continued)

Number of Pulses Integrated	25
Search Time (± 45 degrees $\times$ ± 45 degrees)	10 sec
Integration Loss	2.5 dB
Fixed System Losses	7.5 dB
Total Losses	10 dB
Range ($S/N = 15$ dB for $\sigma = 50$ m ²) $B_r \tau = 1.4$, $f_r/B_1 = 25$	110 nm
Single Pulse S/N Ratio (110-nm range)	3.5 dB
RMS Range Jitter	34 ft (0.005 percent)
Maximum Trackable Velocity (110-nm Range)	10,900 ft/sec
Range Lag Error (500 ft/sec Target)	45 ft
Angle Tracking Scheme	Conopulse
RMS Angle Tracking Error per channel (110-nm Range)	0.135 degree
Conopulse Frequency	100 kHz
Crosstalk Error (110-nm Range)	0.00585 degree
Reliability (1000 operating hours)	69 percent
Array Weight	173 lbs
Total System Weight (Exclusive of Power Conditioning)	234 lbs
Total System Volume (Exclusive of Power Conditioning)	11.1 ft ³
Total System Power (Exclusive of Power Conditioning)	1100 watts
Cooling Method	Fluid Loop

For these reasons, it should be obvious that the optimum system depends not only on what performance is attainable from subsystems but on the particular constraints applied to the system design.

9.0 RECOMMENDATIONS FOR FURTHER STUDY

During the study period there were several occasions when interesting problems appeared which were not within the scope of the study contract, but do relate to the CAT radar problem. It is the purpose of this section to list these problem areas in the order in which they would affect the CAT radar program.

9.1 Higher Duty Cycle Radar Systems

Because of time limitations, imposed by the relatively short duration of this study, the most significant contribution was considered to be one by constraining the radar system to be of a low duty cycle, 10 percent or less. When the duty cycle increases appreciably above this, the module configuration changes considerably. Because the target detection probability is determined by the ratio of E/N_0 , the energy in a received pulse over the noise spectral density, Texas Instruments thought that a great deal would be learned about the CAT radar effectiveness by determining the system configuration which maximized E/N_0 for the general class of low-duty-cycle radars; since, increasing the duty cycle does not increase E/N_0 without attendant complications.

However, now that the results are in for the low-duty-cycle system, launching an investigation into higher-duty-cycle approaches seems desirable. If the attendant complications can be handled, the potential payoff would be made possible because of one reason. To attain a high E , peak power must be high or pulse length must be long. Low-duty-cycle radars optimize for the former while high duty cycle radars optimize for the latter. Because of the high peak power requirement, low-duty-cycle radars pay the penalty of decreased efficiency in conversion of power from dc to rf. High-duty-cycle radars pay a penalty in the form of increased pulse compression requirements and increased data processing complexity as well as transmitter/receiver isolation problems in the case of CW radar. If these processing problems are manageable, the peak power requirement could be relaxed in the same ratio that the duty cycle is increased. Thus, for the case of 5 percent duty cycle, which is representative of a good system previously discussed, a potential exists for increasing the duty cycle by a factor of 20, consequently lowering the peak power by the same factor to retain the same performance. Just what efficiency improvement could be achieved by this relaxation in requirement is not known in all cases but certainly it would help the X-band systems more than the L-band which are already relatively efficient.

Representative of techniques which make good use of high duty cycle is the wideband random-signal radar system discussed by Gassner⁸⁴ and the wideband digital correlator discussed by Gassner et al.⁸⁵ This scheme involves transmitting a wideband random noise process and employing a correlation receiver to estimate range and range rate. Motivation results from the fact that for the random signal, time duration and bandwidth are essentially independent

with very large products easily generated. Resolution capabilities of this radar system are good because the ambiguity function is shown to be separable into range and range-rate resolution functions, mutually independent, permitting great freedom in selection of the desired ambiguity function.

This approach to radar signal synthesis is thought to offer fertile ground for an investigation into methods yielding increased power efficiency.

9.2 Experimental Array Verification Measurements

A great amount of analytical array analysis has been published in the literature, but only a few reports on actual array measurements have appeared. Much of this published data on actual measurements has been incomplete in that only the measured data was reported. In many cases, no attempt was made to correlate the experimental data with array theory. Other workers have discovered unpredicted experimental results and have later explained these results.

Texas Instruments recommends that an experimental phased-array measurement study be performed with a goal of analytical verification of measured experimental results. These phased-array measurements should consider the following: (1) quantization effects in the mutual coupling environment; (2) definition of array gain as a function of scan position; (3) element pattern variation and element impedance matching; (4) pointing accuracy in the mutual coupling and quantization environment; (5) possible plasma effects on array performance; (6) polarization characteristics of the multipolarized phased array.

Texas Instruments has both the facilities and the qualified personnel at the Dallas Central Antenna Laboratory facility to perform the suggested measurements study. In addition, many of the array components for such a program are already available from previous measurements programs at both L-band and S-band frequencies.

9.3 Conopulse-Radar Analysis

The new method of angle tracking discussed in Section 3, called conopulse, was analyzed to an extent necessary to obtain a comparison to monopulse and conical-scan methods of angle tracking. However, because of the newness of this method, and the favorable comparison to date, continuing the analysis of this scheme appears to be extremely desirable. A study alone would be of merit. Most value, however, would be obtained by a combination effort involving further analytical investigation followed by an experimental investigation to verify the theory.

Because of time limitations, a complete conopulse analysis could not be completed. One point which was not fully developed is the interrelationship

between angle tracking and range tracking. For example, when the range tracking loop has error, this error is transferred into the angle tracking loop as was discussed in the study. However, an error in the angle tracking loop adversely affects the range tracking loop, but the extent is not fully known.

9.4 Latching Ferrite Thin-Film Phase Shifters

The microstrip latching ferrite phase shifters were discussed in subsection 3.2.3.6(d). These phase shifters require much less dc power than the diode phase shifters and can handle larger values of peak rf power. If they were available, then more prime power would be available for the transmitter. Conversely, for the same transmitted power a larger array could be used; or, for the same transmitted power and antenna size, less input power would be required.

Consideration should be given to a program of designing, developing, and evaluating a small latching ferrite phase shifter, specifically for use in a spaceborne phased array. In doing this, consideration should be given to the design of the driver and control circuits to ensure compatibility as an element in a phased array.

9.5 Additional Study with Different Restrictions

As was mentioned in Section 8 certain assumptions were made concerning size, weight, and power limitations on the CAT radar system configuration. Those assumptions led to an optimum system which maximized the signal-to-noise ratio and, coincidentally, was also the lightest corporate-fed array under consideration. The study report data at hand makes it possible to determine optimum system configurations for other sets of limitations on the radar system physical parameters. For this reason one could expect an additional study of that nature to produce meaningful results on a short turn-around basis.

9.6 Surface-Wave Device Development

Appendix B discusses a new component which should greatly simplify the construction of matched filters for radar and other applications. As the study now stands, there is no urgent need for this device for the low-duty-cycle options discussed. In the event that serious consideration should be given long-duty-cycle options it will be very important to investigate in depth the possibilities offered by the SWD as they pertain to space radar. Power consumption, size, and weight of the device are among the attractive features which should make this device valuable to the CAT radar program. Also the reliability aspects of a small highly integrated, batch-fabricated device are most important as is the low-cost likelihood.

APPENDIX A

CHARACTERISTICS OF PHASED ARRAYS

Phased-array antennas are rapidly finding their way into advanced radar and communication systems. They offer several advantages over their mechanically scanned counterparts—the foremost of which are increased reliability, high scan rates, and the capability to provide simultaneous multimode operation. Although much of phased-array theory is common to that of reflector systems, there are many characteristics (e.g., mutual coupling) unique to electronically scanned antennas. Some of the more important of these characteristics are discussed in the following paragraphs. Included are paragraphs on mutual coupling in phased arrays, array simulation, array grid considerations, fan beam generation and phased-array radiating elements.

I. MUTUAL COUPLING IN PHASED ARRAYS

Mutual coupling is an interaction of rf fields at the array radiating surface and can cause loss of gain, increased sidelobes, and impedance matching problems. This section discusses mutual coupling and impedance variation in finite arrays as well as theoretical coupling in an infinite array.

A. Mutual Coupling in Finite Arrays

The problem associated with mutual coupling in finite arrays may be solved for simple array elements such as dipoles or slots.^{87,88,89} The solution may be obtained by solving the impedance matrix of the array. This method would be rather straightforward except for the fact that it requires the solution of an extremely large set of simultaneous equations ($N \times N$ set of equations for an N -element array). Thus, to obtain valid results without using an excessive amount of computer time one must employ all of the symmetry properties existing in the impedance matrix.

Coupling between elements is described by the following matrix equation.

$$[V] = [Z] [I] , \quad (A-1)$$

where

$[V]$ = driving voltages

$[I]$ = antenna currents

$[Z]$ = mutual impedance matrix.

The mutual impedance matrix consists of all the mutual impedances occurring in the array. For an $M \times N$ element array with the geometry of Figure A-1 the impedance matrix has the form

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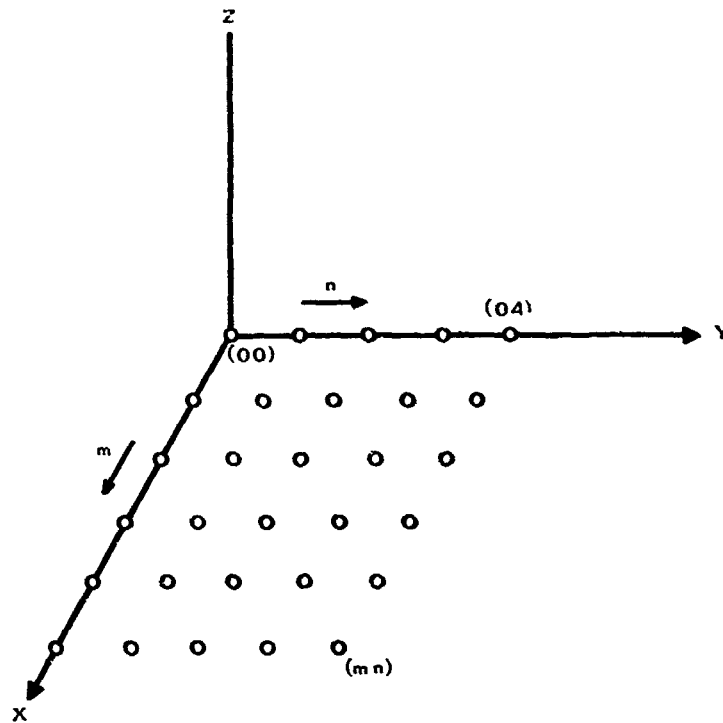
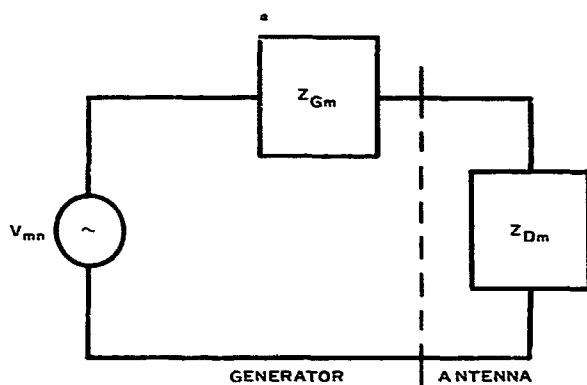


Figure A-1. Uniformly Spaced Array

$$Z = \begin{bmatrix} Z_{00,00} & Z_{00,01} & \dots & Z_{00,mn} \\ \cdot & & & \cdot \\ \cdot & & & \cdot \\ \cdot & & & \cdot \\ Z_{mn,00} & Z_{mn,01} & \dots & Z_{mn,mn} \end{bmatrix}, \quad (A-2)$$

where $Z_{kl,pq}$ is the mutual impedance between the kl^{th} element and the pq^{th} element. These mutual impedance terms may be readily computer by numerical integration for array elements such as dipoles or slots. The diagonal terms in $[Z]$ are the sums of the antenna self-impedance and the generator impedance shown in Figure A-2. For a uniformly spaced $M \times N$ planar array, the impedance matrix has the symmetry properties such that

$$[Z] = \begin{bmatrix} Z_0 & Z_1 & Z_2 & Z_3 & \dots & Z_m \\ Z_1 & Z_0 & Z_1 & Z_2 & \dots & Z_{m-1} \\ Z_2 & Z_1 & Z_0 & Z_1 & \dots & Z_{m-2} \\ \vdots & & & & & \\ Z_m & Z_{m-1} & Z_{m-2} & & & \end{bmatrix} = M \times M \text{ Symmetric Toeplitz Matrix,} \quad (A-3)$$



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Figure A-2. Generator-Antenna
Equivalent Circuit

achieving that result. For large arrays, the impedance variation with scan and extent of edge effects are of prime concern. The far-field amplitude pattern of such an array is of interest; however, it would be very difficult and time consuming to compute. Also, the field pattern of a very large array is affected very little by mutual coupling if the phase shift is assumed to be produced by ideal phase shifters. Thus, the far-field amplitude pattern is not considered to be of primary concern in the direct solution of the mutual coupling problem.

The active impedance of the m th element in an array is

$$Z_{D_{mn}} = \frac{V_{mn}}{I_{mn}} - Z_{g_{mn}} \quad (A-4)$$

As the beam is scanned, the voltage V_{mn} will change in a known way because of the linear variation necessary to scan the beam. The current, I_{mn} , may be calculated by inverting the impedance matrix for each scan position of the array. This would result in the impedance versus scan for each element of the array and the data could be used to plot the far-field amplitude pattern of the array. Obviously, this approach yields the entire solution to the problem, but it requires an excessive amount of computer time because of the large number of matrix inversions.

where Z_0, Z_1 , etc., are $N \times N$ symmetric Toeplitz matrices. A Toeplitz matrix is a symmetrical matrix for which all the elements in any one diagonal are equal. These properties allow the entire matrix to be generated from its first row. By utilizing that fact, the matrix can be solved without requiring the entire matrix to be stored in the computer memory. (For large arrays, it may be impossible to store the entire matrix because of limited computer memory.) The Gauss-Seidel iteration technique works on one row of the matrix at a time and has been found to yield a rapidly convergent solution to the set of simultaneous equations.

Ability to obtain a solution to the coupling problem for large arrays has been established; however, consideration should be given to the desired result and the most efficient means of

A more realistic approach is to generate one row of the inverted impedance matrix so that the impedance versus scan for a one-array element can be obtained with only one matrix inversion. The coupled-matrix equation may be written as

$$[Z]^{-1} [V] = [I] . \quad (A-5)$$

Let $V_{mn} = 1$ and all other voltages equal zero. The solution to the equations yields the mn^{th} column of the inverted impedance matrix. Because of symmetry we also have the mn^{th} row of the impedance matrix and, hence, the current is

$$I_{mn} = \sum_{j=1}^M \sum_{k=1}^N Z_{mn}^{-1, jk} V_{jk}, \quad (A-6)$$

where Z^{-1} represents the terms of the inverted impedance matrix. Thus, the impedance versus scan for any element in an array may be obtained with only one matrix inversion and given as shown in the following equation.

$$Z_{D_{mn}} = \frac{V_{mn}}{\sum_{j=1}^M \sum_{k=1}^N Z_{mn}^{-1, jk} V_{jk}} - Z_{g_{mn}} . \quad (A-7)$$

Although this approach does not result in the calculation of the far-field amplitude pattern of the array, it does give a most efficient means for calculating edge effects and the impedance variations of elements in the array.

B. Impedance Variation in Finite Arrays

The impedance variations in a 15×20 -element rectangular array has been analyzed. The analytical array element was a half-wavelength dipole mounted on a quarter wavelength above a ground plane and the elements were arranged on a square grid of 0.534λ (MERA spacing). The elements shown in Figure 4-3 were analyzed to yield the impedance variation of a central element and a sample of edge elements. The results are presented as VSWR versus scan curves and are assumed to be matched to the generator impedance ($Z_g = 113-j76$). The generator impedance was chosen to minimize the VSWR variations with scan, and the method used in choosing this impedance will be discussed in detail in Paragraph IV. C.

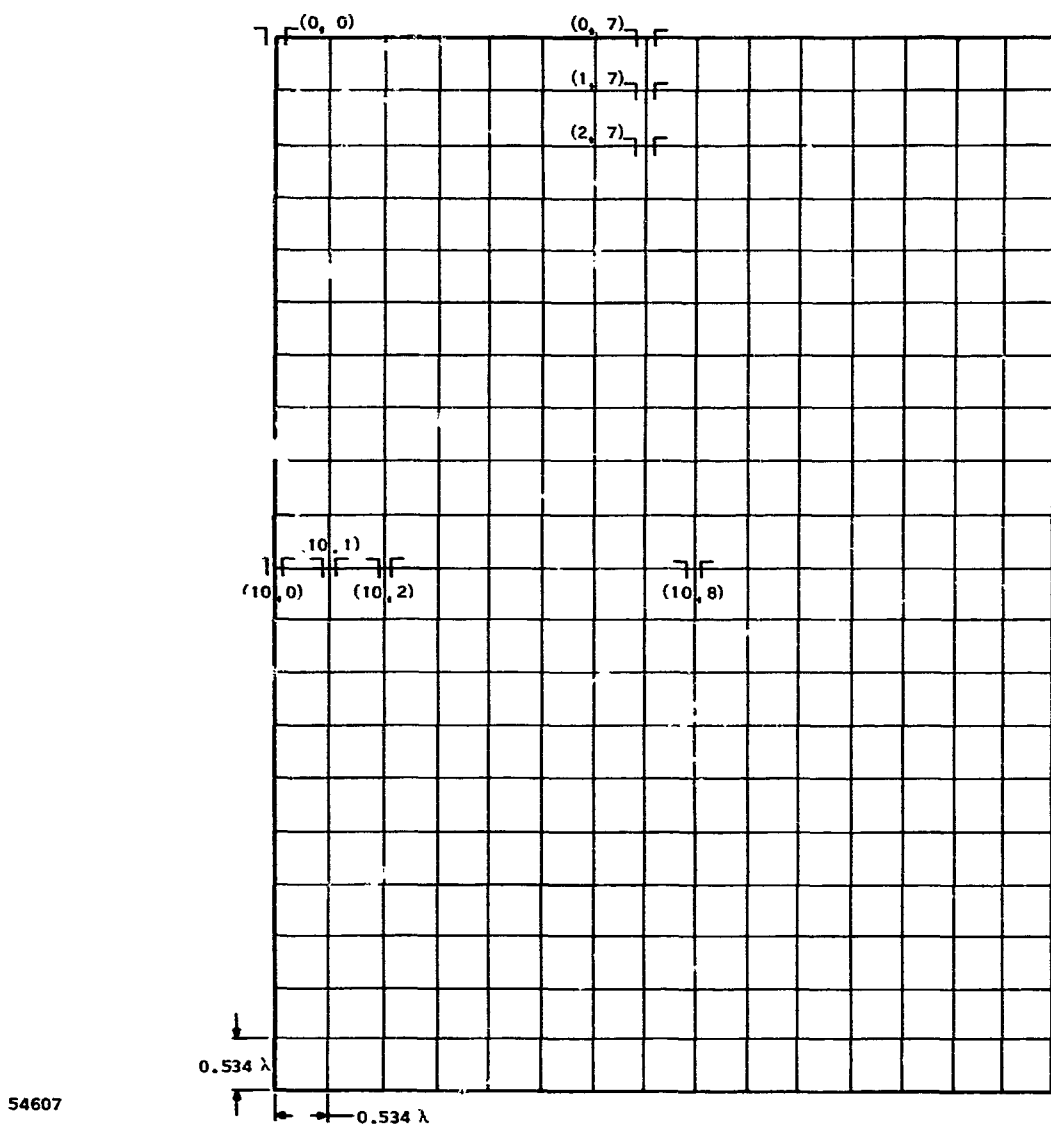


Figure A-3. Fifteen by Twenty-Element Square Grid Configuration

Figure A-4 gives the VSWR variation of a central array element. The VSWR increases sharply at scan angles approaching 60 degrees. This increase is expected because a grating lobe appears in visible space at a 60-degree scan angle in the E- and H-planes. The diagonal-plane scan is observed to exhibit better VSWR characteristics than those of the E- and H-plane scan. This is expected because a grating lobe occurs in visible space at a larger scan angle in the diagonal plane. For this reason, the E- and H-planes represent the extreme cases of VSWR variations with scan. The actual shape of the curve is determined by the generator impedance, but the impedance of the central element will always be symmetrical with respect to positive and negative scan angles.

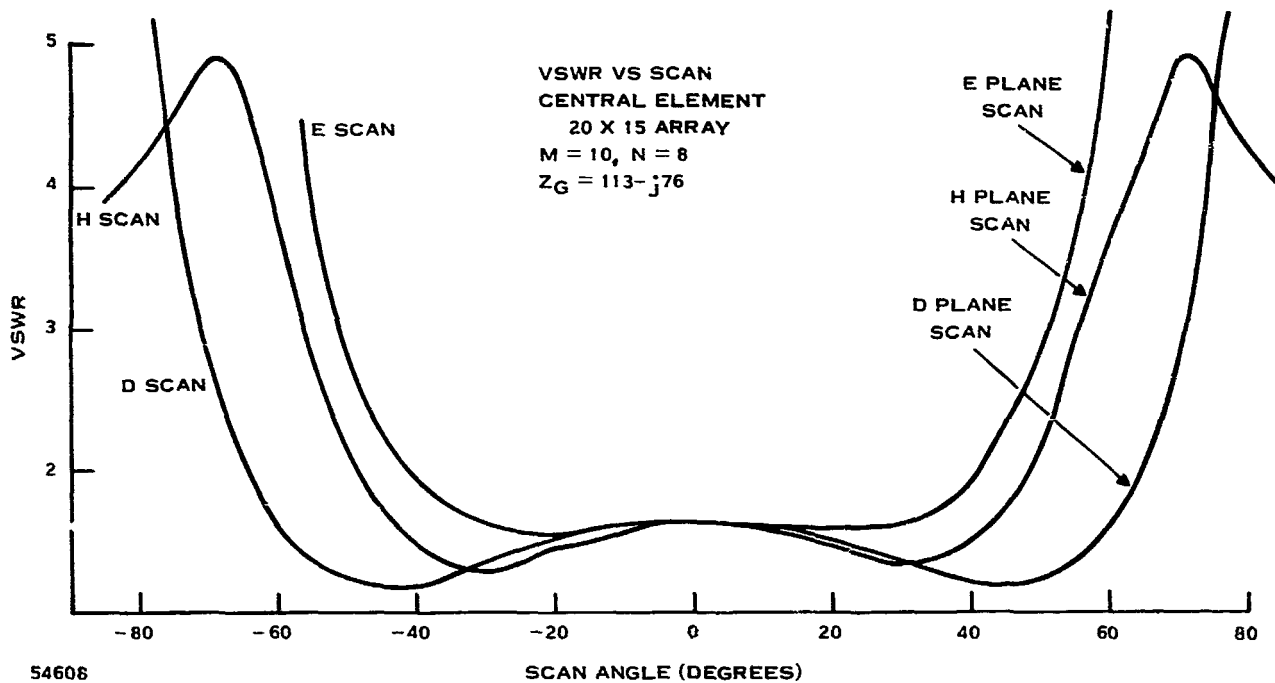


Figure A-4. VSWR Variation of Central Array Element

Figures A-5 through A-7 give the VSWR variations for a sample of edge elements. Unlike the central elements, the edge elements do not produce symmetrical VSWR variations with scan. This trait is produced by the asymmetrical environment of the edge elements. The shape of the VSWR curve is significantly altered and actually is an improvement over those of the central elements. The extent of the edge effects are illustrated in Figures A-8 and A-9. These curves indicate that the majority of the edge effects are confined to the three outermost rings of elements in the array.

For practical phased-array radars, the beam is scanned by employing digital phase shifters to produce the linear phase variation across the array. This results in a quantization of the phase required for each element such that the phase distribution is no longer linear over the array. This phase quantization has a significant effect on the mutual coupling resulting in a distortion of the VSWR versus scan curve as indicated by Figures A-10 and A-11. The phase quantization VSWR oscillates about the VSWR produced by the linear phase distribution. As the number of bits is increased, the accuracy with which the phase may be implemented increases and the curves approach the curve of the linear phase shifter. Phase quantization also influences the far-field amplitude pattern of the array and is discussed in Paragraph II, F.

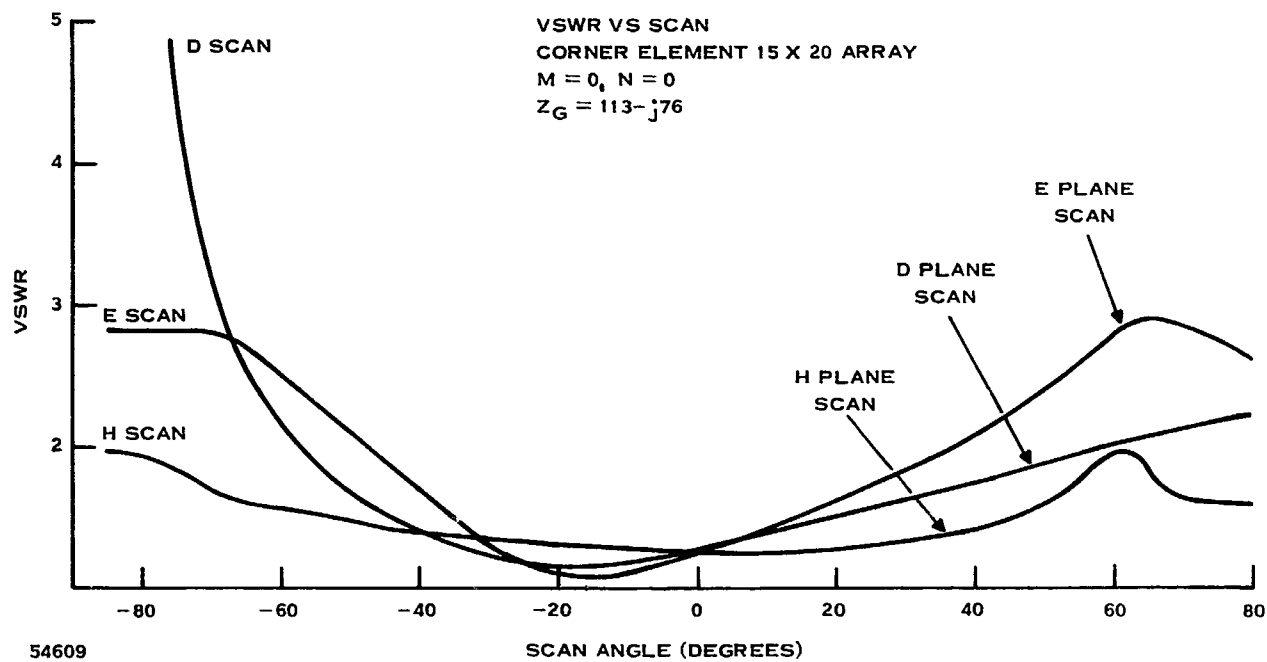


Figure A-5. VSWR Variation of Corner Array Element

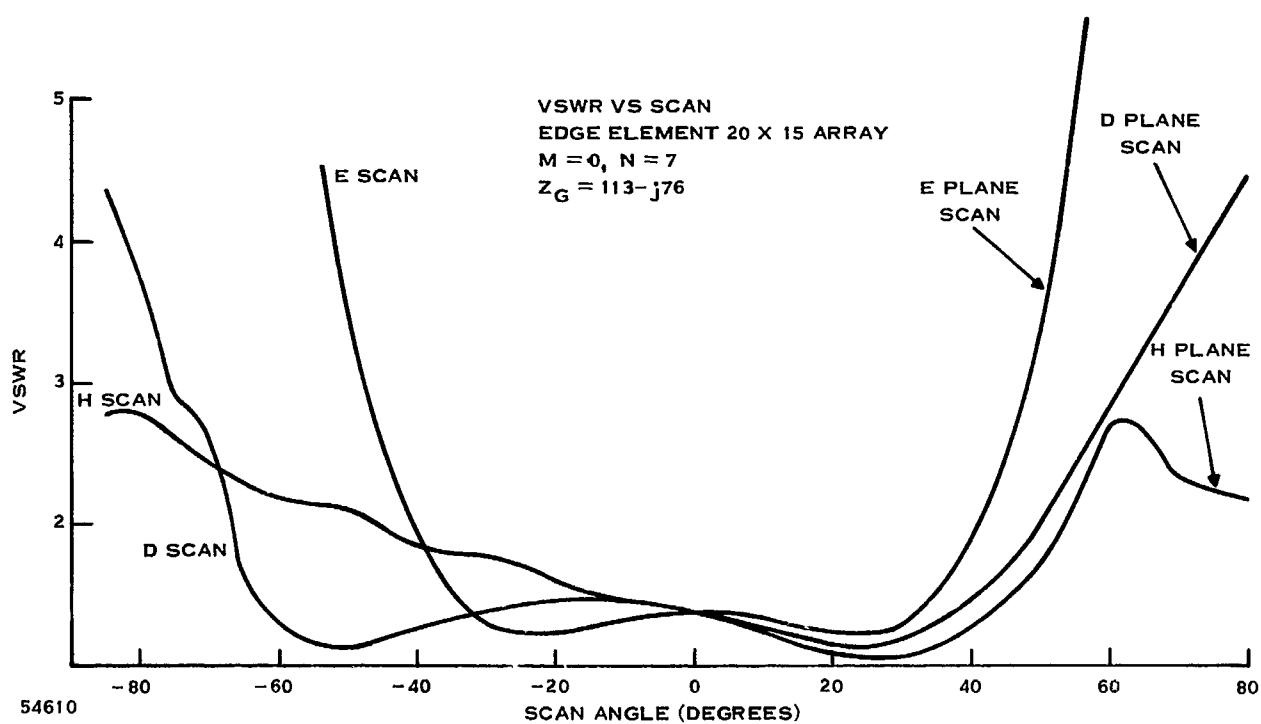


Figure A-6. VSWR Variation of H-Plane Edge Element

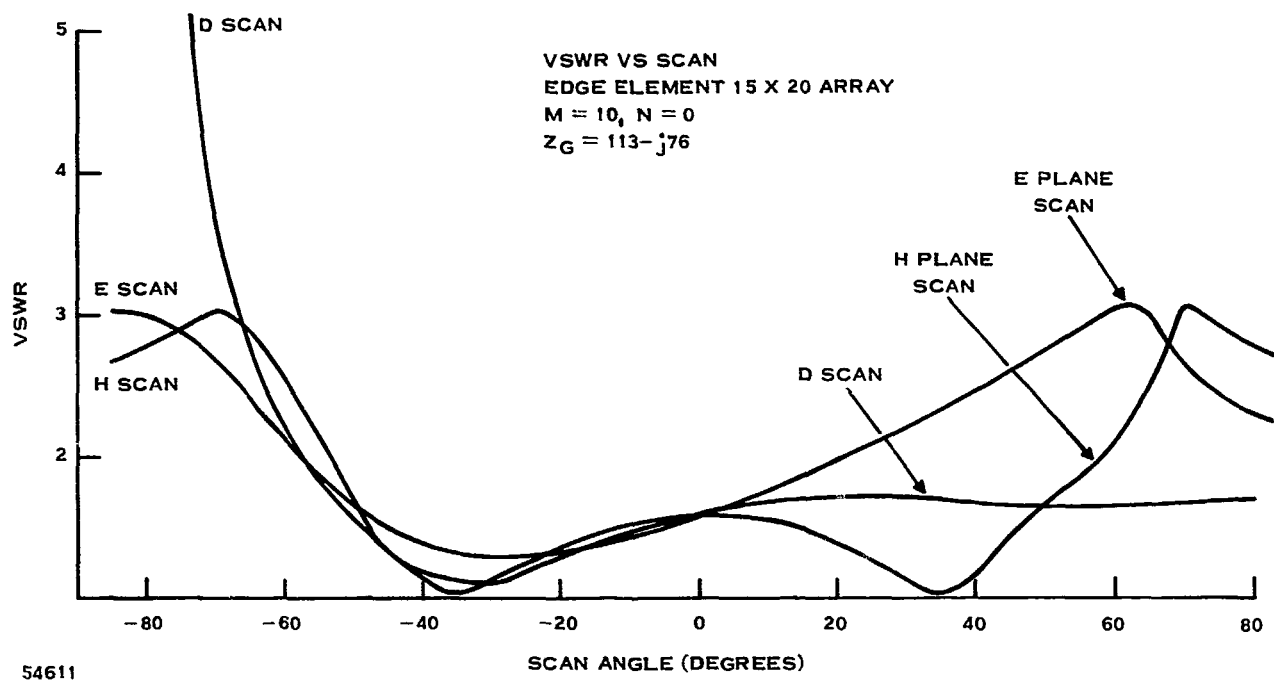


Figure A-7. VSWR Variation of E-Plane Edge Element

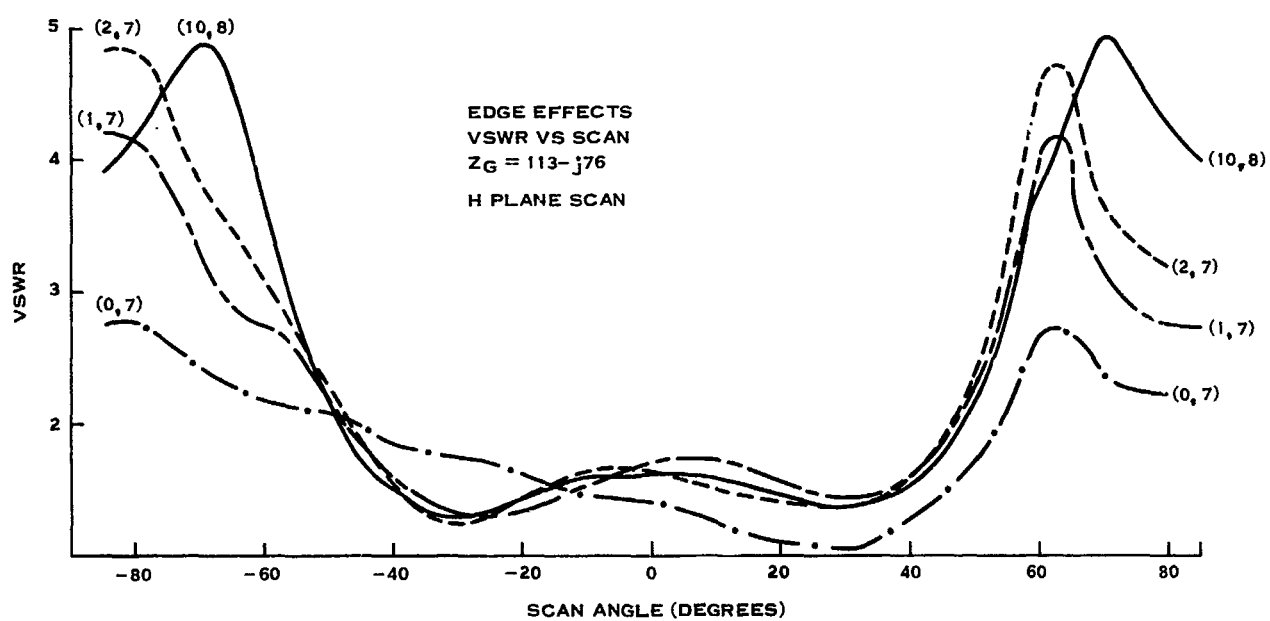
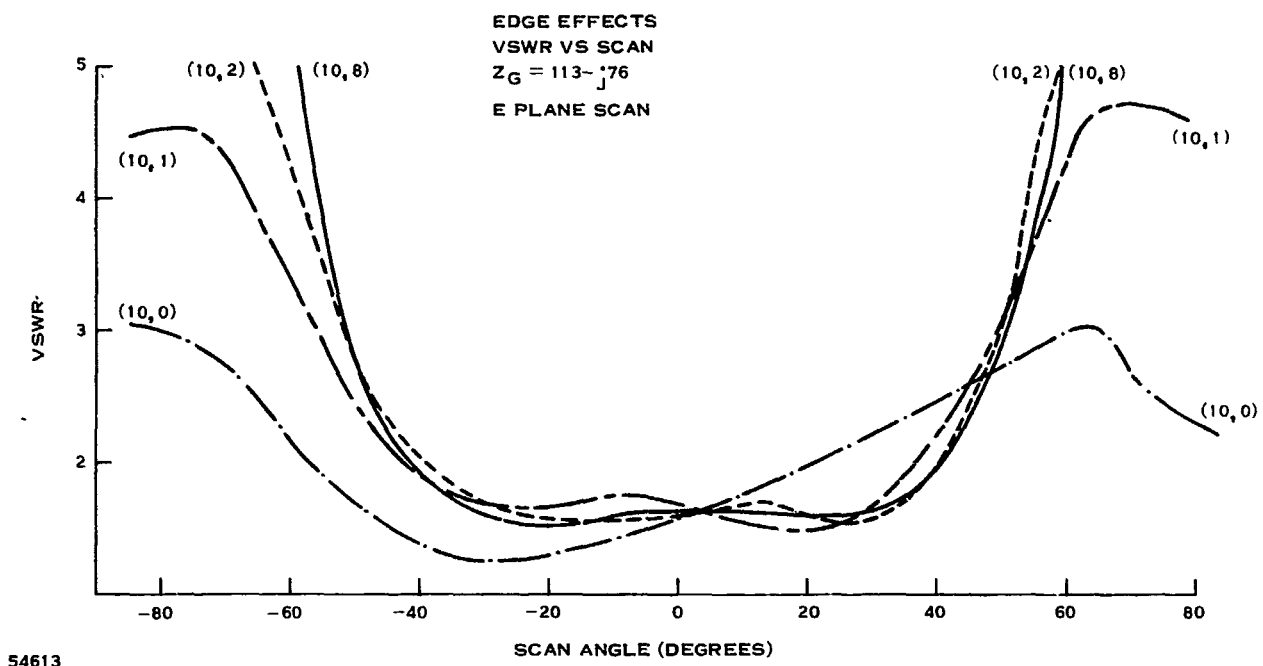
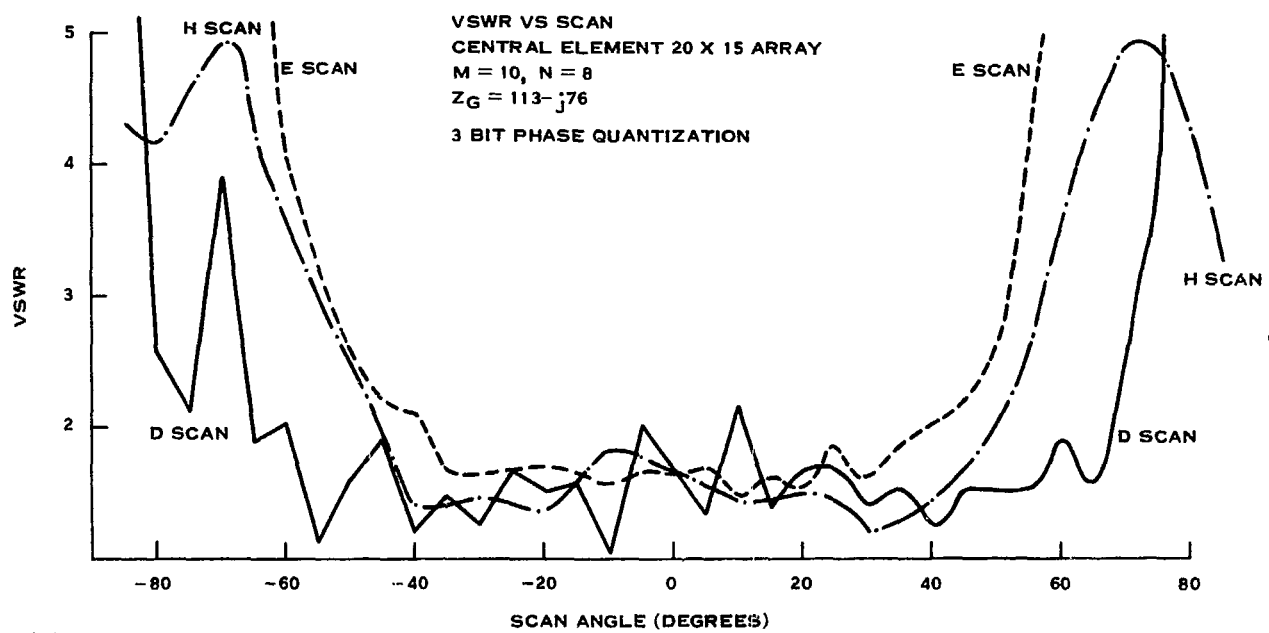


Figure A-8. VSWR Variation of H-Plane Edge Elements



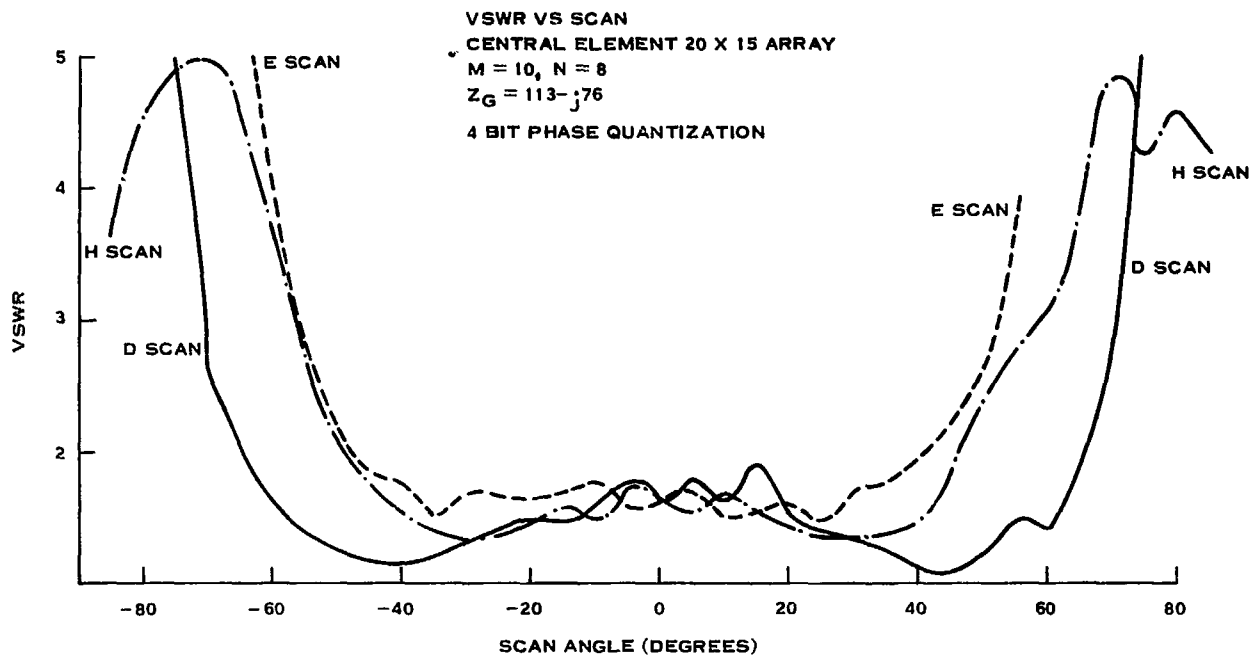
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Figure A-9. VSWR Variation of E-Plane Edge Elements



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Figure A-10. Three-Bit Phase Quantization Coupling Effects



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Figure A-11. Four-Bit Phase Quantization Coupling Effects

C. Mutual Coupling in Infinite Arrays

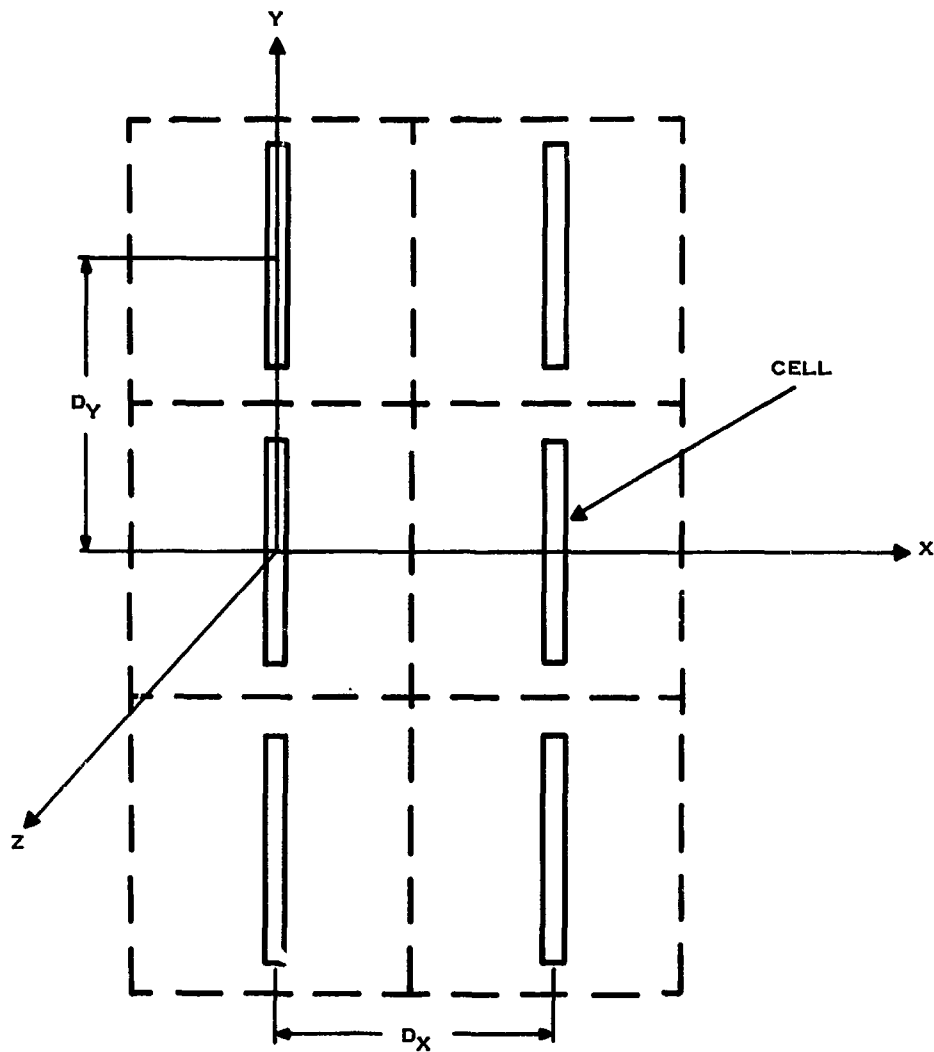
The distinction between finite and infinite arrays is that the infinite array does not exhibit edge effects while the finite array does. In other words, the impedance is the same for each element of the infinite array. This provides the basis for grating-lobe series formulation for computing the impedance versus scan for an element in an infinite array.

The problem is solved by taking advantage of the periodicity occurring in an infinite uniform phased array.^{90, 91, 92} The cells shown in Figure A-12 are repetitive in magnitude with a linear phase variation as given in the following equation.

$$\phi(x, y) = (\alpha/D_x x + \beta/D_y y). \quad (A-8)$$

where $K_0(x, y)$ is the magnitude of the current density. Since $K_0(x, y)$ is repetitive, it may be expressed as a double Fourier series. Thus, the current density over the surface of the array is

$$K(x, y) = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} K_{m,n} e^{-j(\alpha + 2\pi m) x/D_x} e^{-j(\beta + 2\pi n) y/D_y}, \quad (A-9)$$



54616

Figure A-12. Geometry of Infinite Array

where

$$K_{mn} = \frac{1}{D_x D_y} \iint_{\text{cell}} K(x, y) e^{j(\alpha + 2\pi m)x/D_x} e^{j(\beta + 2\pi n)y/D_y} dx dy. \quad (\text{A-10})$$

Utilizing the Fourier expansion of the current, the boundary-value problem may be solved for the pertinent fields in the vicinity of the array. Then, conventional techniques being used, the impedance of a cell may be obtained by calculating the complex power radiated from the cell.

By assuming a sinusoidal current distribution, the active impedance of a dipole near resonance (half wavelength) may be derived. In a scan region where no grating lobes are imminent, the impedance of a half-wavelength dipole, located a distance S over a ground plane, is

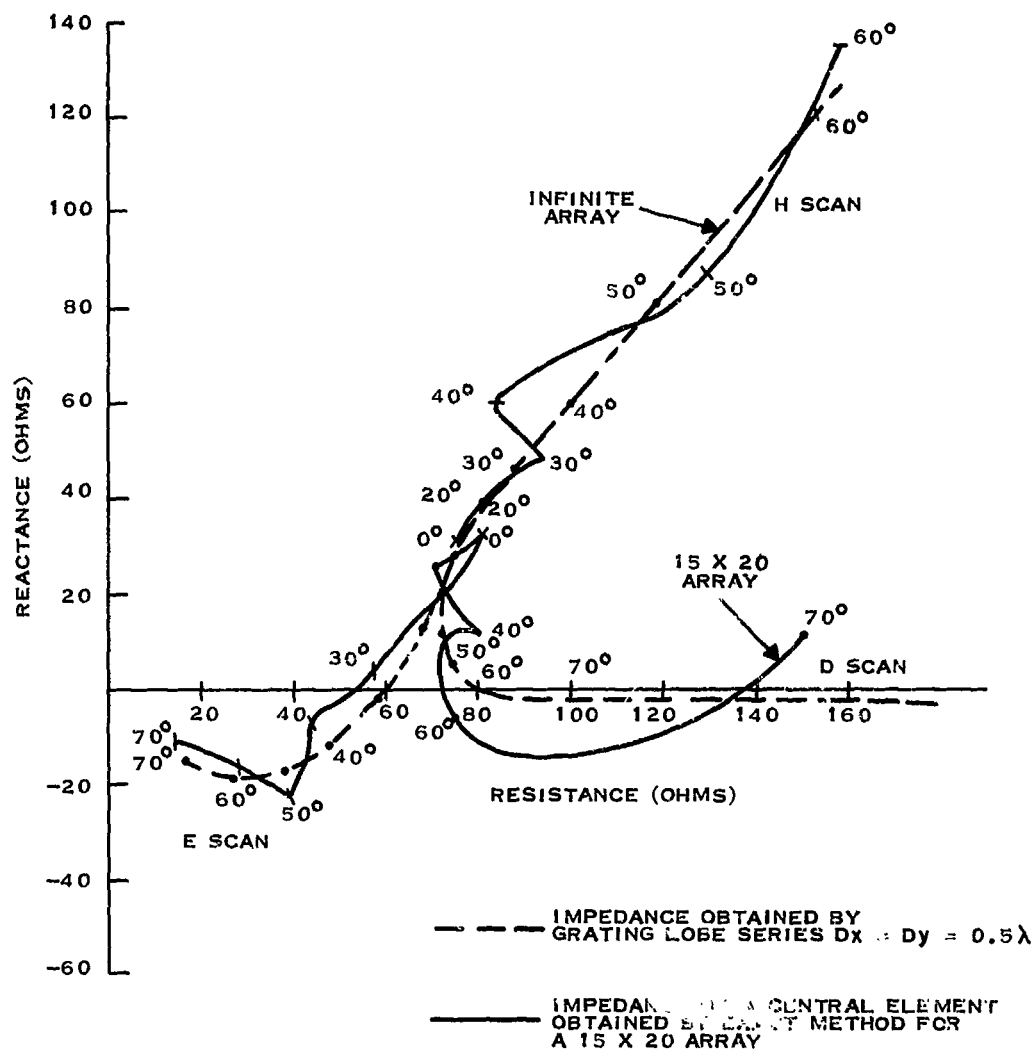
$$R = \sqrt{\frac{\mu}{\epsilon}} \left(\frac{1}{\pi}\right)^2 \frac{1}{D_x D_y} \left[\frac{\cos\left(\frac{\pi}{2} \cos \theta_y\right)}{\sin \theta_y} \right]^2 \left[\frac{\sin^2(KS \cos \theta_z)}{\cos \theta_z} \right], \quad (A-11)$$

and

$$X = \sqrt{\frac{\mu}{\epsilon}} \left(\frac{1}{\pi}\right)^2 \frac{1}{2D_x D_y} \left\{ \left[\frac{\cos\left(\frac{\pi}{2} \cos \theta_y\right)}{\sin \theta_y} \right]^2 \left[\frac{\sin(2KS \cos \theta_z)}{\cos \theta_z} \right] \right. \\ \left. - \sum_{\substack{m=-\infty \\ n=n \neq 0}}^{\infty} \sum_{n=-\infty}^{\infty} \left[\frac{\cos\left[\frac{\pi}{2} \left(\cos \theta_y + \frac{n}{D_y}\right)\right]}{1 - \left(\cos \theta_y + \frac{n}{D_y}\right)^2} \right]^2 \left[\frac{\cos \theta_y + \frac{n}{D_y}}{\sqrt{\left(\cos \theta_y + \frac{n}{D_y}\right)^2 \left(\cos \theta_x + \frac{m}{D_x}\right)^2 - 1}} \right] \right\} \quad (A-12)$$

The angles θ_x , θ_y , and θ_z are the angles measured from the beam-scan position to the x, y, and z axes, respectively, and $k = 2\pi/\lambda$.

Similar expressions may be derived for other elements or configurations, such as the dipole in free space or triangular grids. The criterion for utilizing this technique is simply that the current density may be approximated over the cell and that the element configuration is periodic. The results are presented for an array of dipoles in free space (exact dual of a slot in a ground plane) arranged in a square configuration with 0.5λ spacing. Figure A-13 illustrates that the infinite array gives a good approximation to the central element of a large finite array. Hence, it is feasible to use the infinite array as a model for determining the pertinent design features, such as grid configuration and spacing and the optimum match, of a large finite array. The advantage of such a model is apparent from the form of the solution obtained. The resistance of an array element is given in closed form and the reactance may be computed by a simple summation, except for some instances where the summation does not converge. However, for practical cases such as resonant dipoles or slots, the grating-lobe series does yield satisfactory data in very little computer time. Another advantage of the infinite array model is that it may be experimentally simulated at certain discrete scan angles by constructing only a few elements and testing them in a particular sized waveguide. This simulation method is discussed in the next subsection.



54617

Figure A-13. Impedance of Infinite and Finite Dipole Arrays in Free Space

II. ARRAY SIMULATION

The design of a phased-array antenna element is basically an analytical task because a single isolated element does not perform in the same manner as one in the array environment. For many potential elements the impedance in the array environment can only be roughly approximated. In such cases, it is desirable to have some way to verify experimentally the design. The array simulator, first proposed by Wheeler⁹³ and later used by Hannan and Balfour⁹⁴ in the design of phased-array elements, provides a means for experimentally substantiating an element design. Because the simulator actually represents the performance of an element in an infinite array, it may be employed in the element design for large phased arrays.

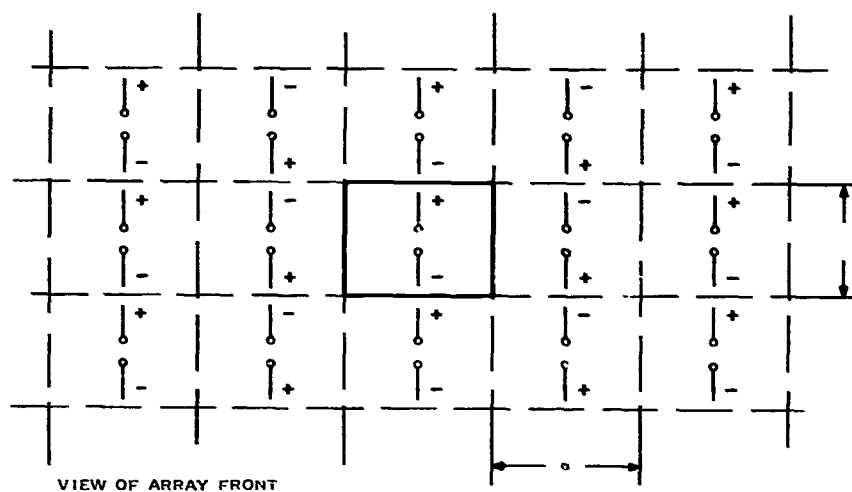
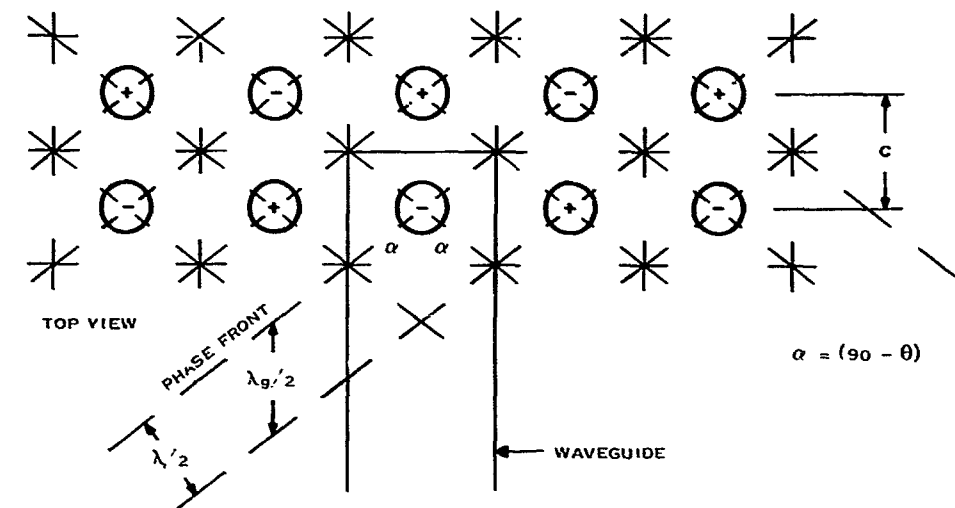
The advantages of the array simulator are distinct because it does allow actual experimental impedance measurements, which include the effects of mutual coupling, without constructing a large array. Using the simulator, one may readily measure the bandwidth of radiating elements at given scan conditions. Also, it can be used to experimentally design and match elements at a certain scan angle.

The simulator is based on the principle of multiple images. By placing array elements in a waveguide, an infinite planar array is imaged in the walls of the guide. For example, consider a vertical dipole excited in a waveguide as shown in Figure A-14a. The walls of the waveguide image element as illustrated in Figure A-14b. In the vertical plane, the element is directly imaged to represent an infinite linear array of elements with spacing (b). However, in the horizontal plane, the element is imaged with the phase reversal indicated. This reversal produces constant-phase fronts as shown and, hence, launches the two-plane waves traveling at the angle $\alpha = (90 - \theta)$ as illustrated in Figure A-14b. The angle of propagation (θ) can be geometrically derived from Figure A-15 and is found to be

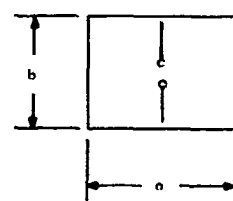
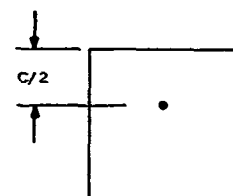
$$\theta = \sin^{-1} \lambda/2a , \quad (\text{A-13})$$

where θ = the angle of propagation for the TE₁₀ mode. The oblique phase fronts obtained in the guide correspond to the phase front of an infinite array with linear phase distribution necessary to point the beam at an angle in the H-plane. Hence, the element in the waveguide exactly simulates an element in an infinite array with rectangular element spacings ($a \times b$) at an H-plane scan angle, θ . The same approach is applicable to other waveguide modes and provides the foundation for the waveguide simulation technique. In practice, it is generally easier to simulate the element as seen looking in from free space. This is done by exciting the waveguide and placing the appropriate terminated elements in the guide. Simulators can be constructed with several elements designed to yield various scan conditions.

To apply the principles of the waveguide simulator, one must keep in mind that waveguide walls may be placed only at planes of symmetry in the array. For a symmetrical element capable of supporting any polarization, the planes of symmetry are shown in Figure A-16. In this case, symmetrical planes exist along the horizontal and vertical (cardinal) planes and diagonal (intercardinal) planes. By choosing any desired set of planes and by exciting the proper mode in the waveguide, an element in an infinite array environment can be simulated. H-plane scan simulation for the linear polarized element can be accomplished by exciting a TE mode in the simulator guide section. Examples of such are shown in Figure A-17. One of the simulators shown contains two half elements which cannot propagate. Nevertheless, the simulator is still valid and exact representation of the infinite array is obtained. By exciting the TE₁₀ mode, the elements will represent the H-plane cardinal-scan angle shown.



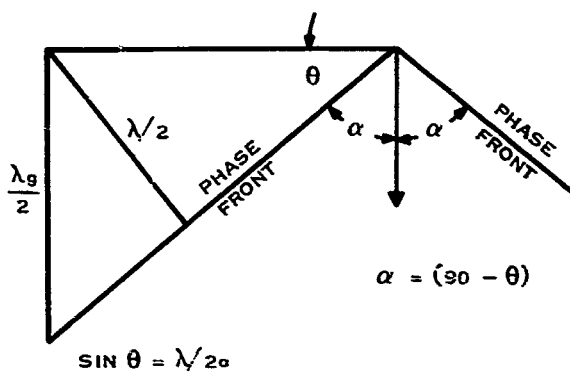
(B) ELEMENT IMAGES



(A) ELEMENT GEOMETRY

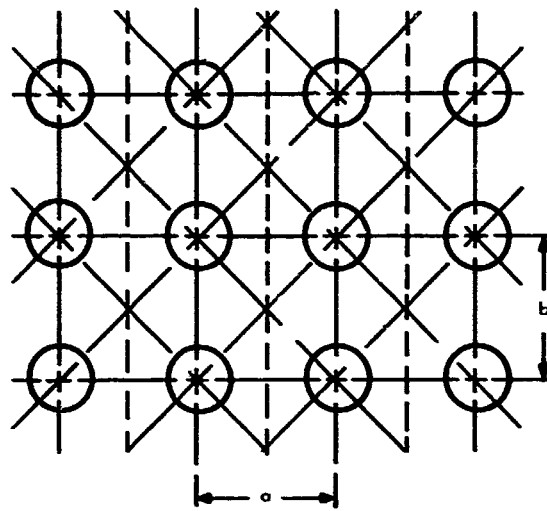
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Figure A-14. Linear Antenna Element Located in a Waveguide



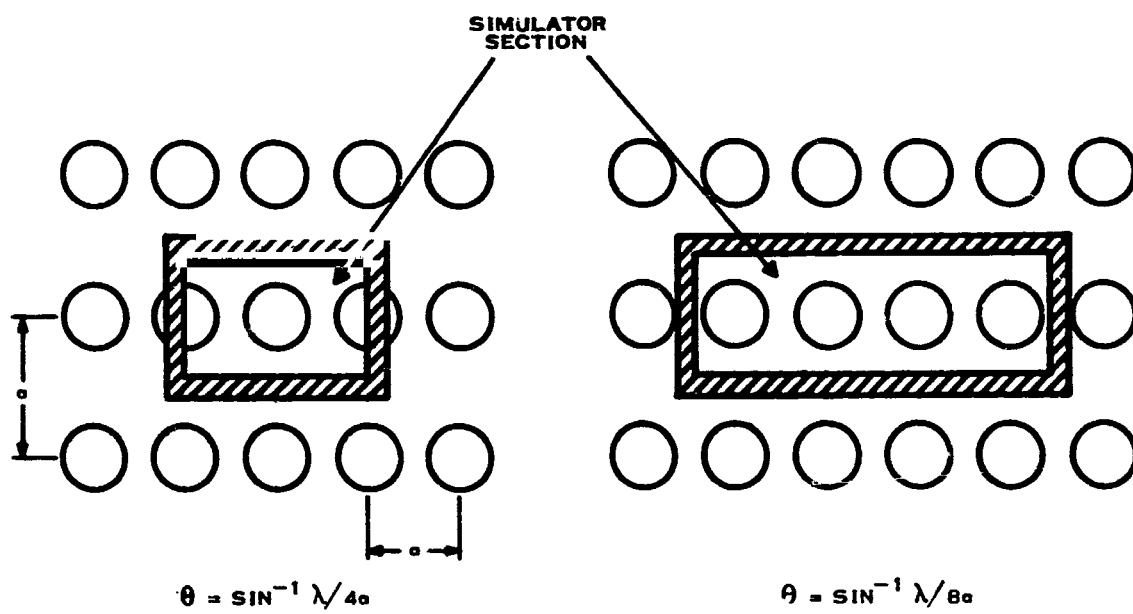
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Figure A-15. Geometrical Definition of Waveguide



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Figure A-16. Symmetry Planes for a Circular Element



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Figure A-17. H-Plane Simulators for Linear Polarization

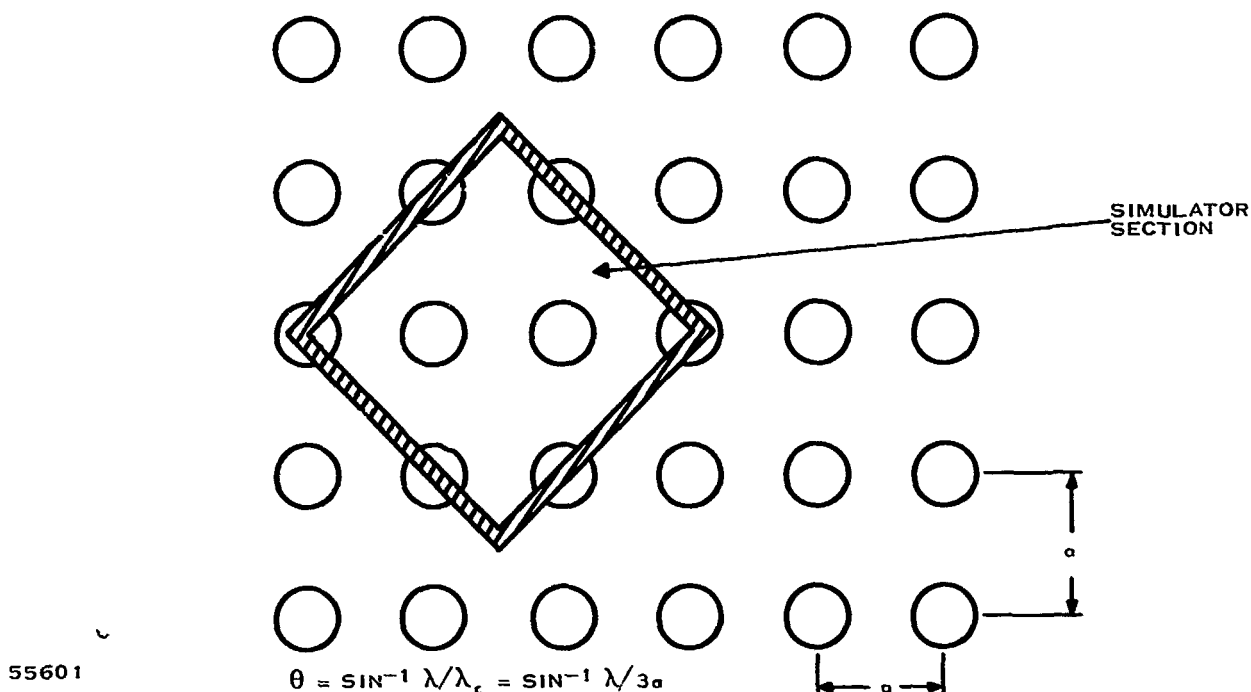


Figure A-18. E-Plane Scan Simulator

For the E-plane scan, the linear phase front must exist across the simulator elements in the E-plane. This may be realized by exciting a TM mode in a simulator section. As an example, consider the square simulator section shown in Figure A-18, constructed by using the diagonal symmetry planes. By exciting the TM_{11} mode, the waves propagate diagonally down the guide. Hence, the simulator represents an E-plane cardinal-scan condition. Also, note the same simulator can be used for the same H-plane cardinal-scan angle by exciting the TE_{11} mode in the waveguide section.

An array of elements capable of supporting only one linear polarization has fewer symmetry planes than an array of circular elements. In this case, the diagonal-symmetry planes no longer exist as indicated by Figure A-19. Thus, the simulator of a linear array has more limitations with regard to the number of scan planes which can be simulated. However, for the scan planes which can be represented, the simulator principles can be applied in the same way as the circular element.

Impedance measurements of an array element can be made quite easily. The array simulator consists basically of four parts.

1. Mode exciter
2. Waveguide transition
3. Simulator waveguide
4. Terminated elements.

By constructing a slotted line in the simulator waveguide section, accurate impedance measurements can be obtained looking into the array. Based upon these measurements, an equivalent circuit can be derived from which the impedance looking out of the array can be obtained.⁸

In summary, the array simulator is a useful tool for designing an array element. The significant advantages are that the technique allows the array designer to:

- Make active impedance versus scan measurement of an element in an infinite array
- Measure the bandwidth of a radiating element at given scan angle
- Measure the axial ratio of a circularly polarized element
- Design and match elements at a certain scan angle.

Although the simulator has several excellent features, there are several disadvantages that should be noted. These disadvantages are as follows:

- Only a finite number of scan angles may be simulated
- It is difficult to construct low scan-angle simulators since many elements must be used; in fact, it is impossible to simulate a broadside scan condition
- For linear elements only a limited number of scan planes can be simulated.

These disadvantages are unfortunate, but even so, the simulator is obviously an important experimental tool in the design of phased arrays.

III. ARRAY GRID CONSIDERATIONS

The appearance of grating lobes (secondary maximums) in visible space limits the usable scan volume of an array. In general, this scan volume is governed by the array spacing and grid configuration of the elements. The use of element grid configurations other than rectangular presents interesting possibilities for minimizing the number of elements in a filled array. For a conical-scan region, it is possible to reduce the number of elements required to maintain a given aperture size by using an equilateral triangle grid.

To pictorially represent the occurrence of grating lobes, considering the beam motion in the complex T-plane ($\sin \theta$ plane) as described by Von Aulock⁹⁵ is desirable. The position of the beam is given by

$$T = \sin \theta e^{i\phi} \quad (A-14)$$

or

$$T = \cos \theta_x + i \cos \theta_y. \quad (A-15)$$

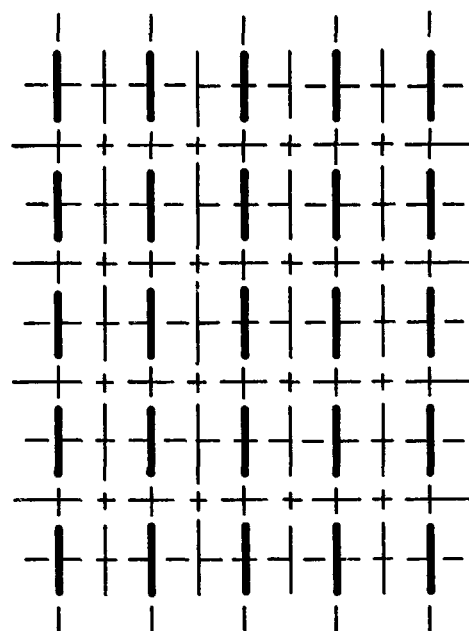
Thus, visible space is confined to the unit circle about the origin. A rectangular array of radiators has an infinite number of secondary maximums spaced (λ/D_x) on the $(\cos \theta_x)$ axis and (λ/D_y) on the $(\cos \theta_y)$ axis. These maximums move with the primary maximum as the beam is scanned and grating lobes will appear in the array pattern if they are allowed to move into visible space. As an example of this, consider an infinite array with $(\lambda/2 < D < \lambda)$ square spacing. The maximum scan along the $(\cos \theta_x)$ and $(\cos \theta_y)$ axes without the occurrence of grating lobes is

$$\sin \theta_m = \lambda/d - 1. \quad (A-16)$$

At this angle, a secondary maximum moves into the unit circle. Figure A-20 illustrates this fact and also shows that a larger scan angle is allowable in the diagonal plane without the appearance of secondary maximums. Hence, we would expect maximum impedance variations with scan along the two principal axes which was pointed out previously in Paragraph II. C. 2.

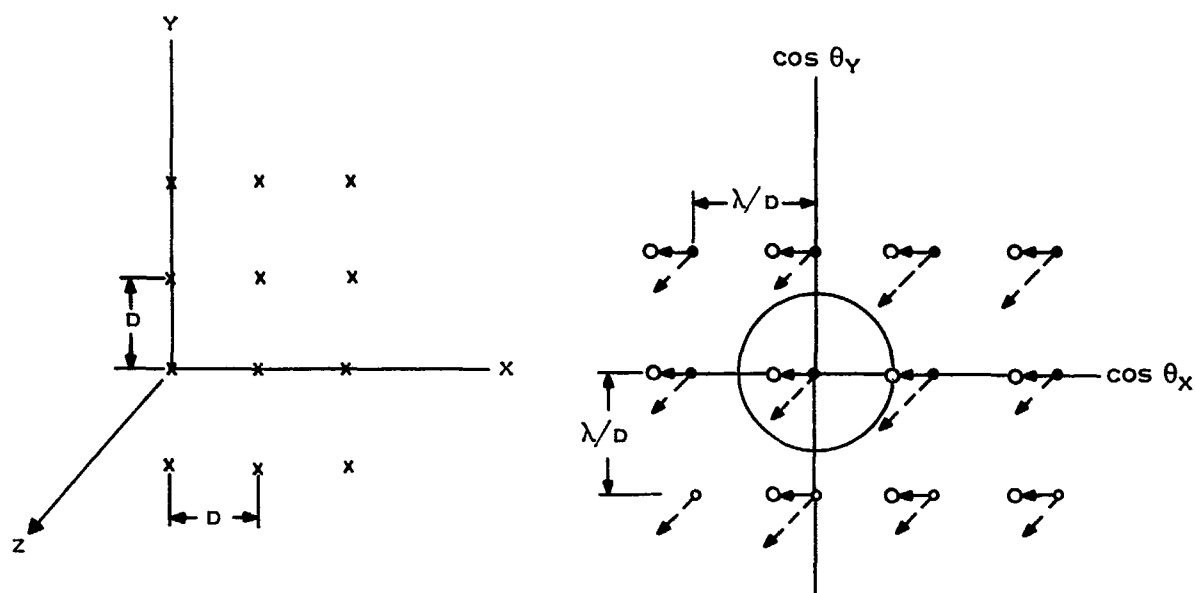
For a triangular grid arrangement the plot of the secondary maximums is somewhat different from that of the rectangular grid. The easiest way to approach the triangular grid problem is to consider the array as a superposition of two identical rectangular arrays.⁹⁵ Considering only one rectangular array, the "sin θ " plane pattern can be plotted as previously described. The triangular array can be constructed by interposing the second array halfway between the elements of the first in both directions. This causes the elimination of alternate grating lobes on the "sin θ " plane as illustrated by Figure A-21.

Two types of elements grids, the square and the equilateral triangular, are shown in Figure A-22, with dimensions for comparable maximum scan. In this case, the grating lobe will not appear for scan angles out to 90 degrees. The area of one element in the square grid with $(\lambda/2)$ spacing is $0.25\lambda^2$. An element in the triangular grid occupies a larger area and as a result, this grid will require 13.5 percent fewer elements than the comparable square grid array.⁹⁶ The scan impedance of two-element orientations for the triangular grid are illustrated in Figure A-23. These cases are comparable to the square array shown in Figure A-22. The shapes of the impedance curves obtained for the two grid types are quite different. However, for a



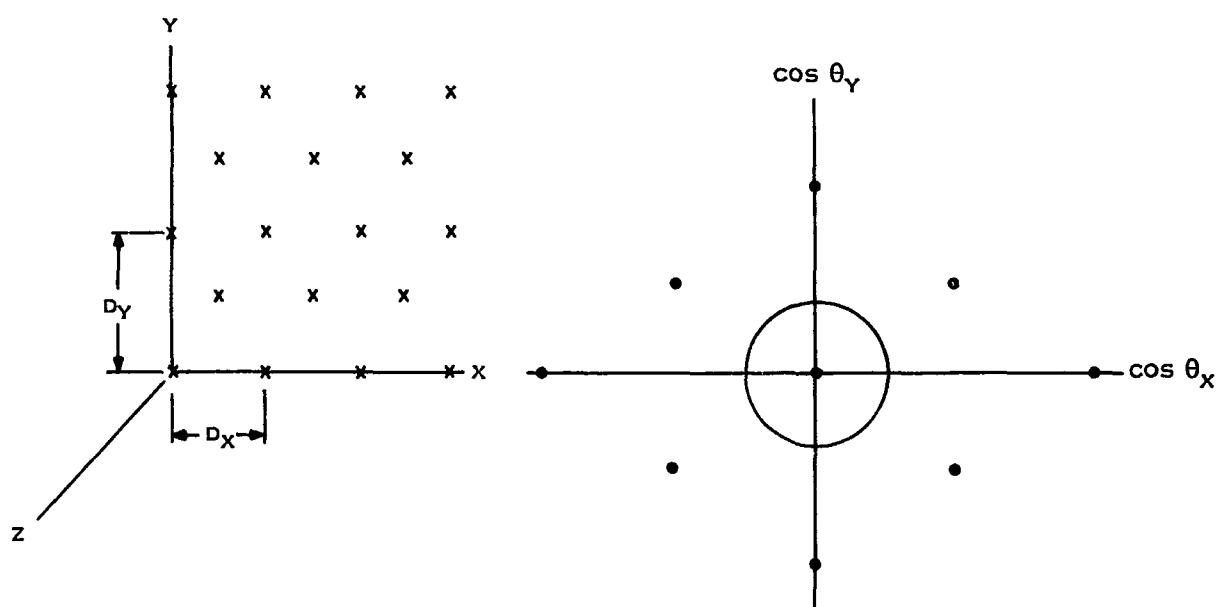
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Figure A-19. Symmetry Planes for a Linear Element



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Figure A-20. Plane Pattern for Square Grid Array



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Figure A-21. Sin θ Plane Pattern for Triangular Grid Array

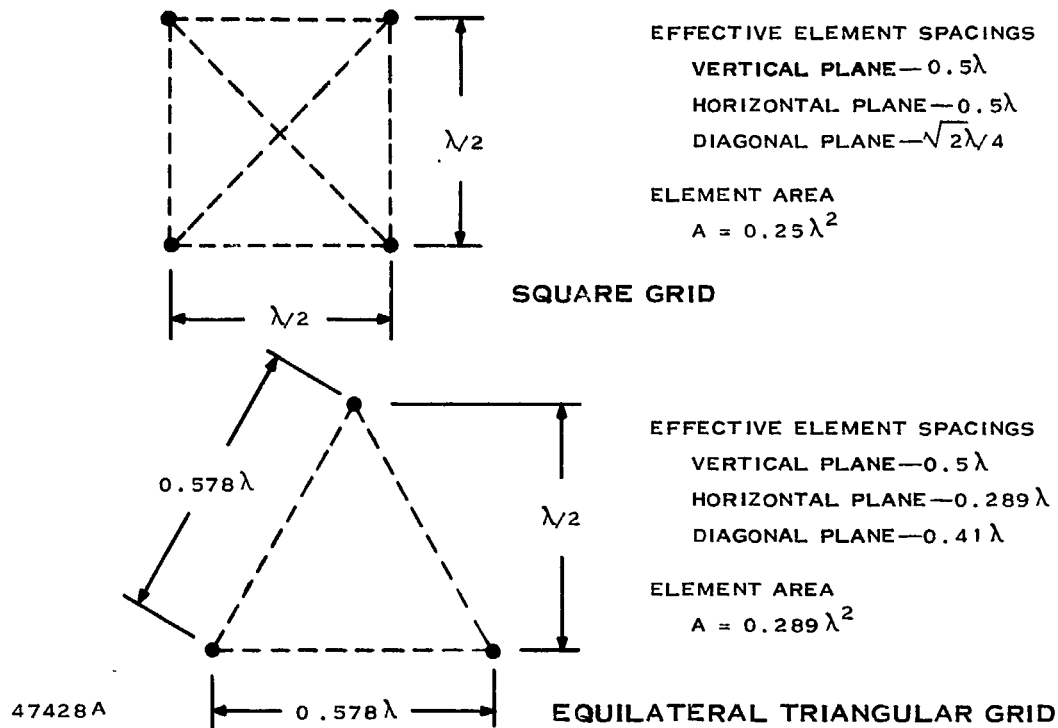


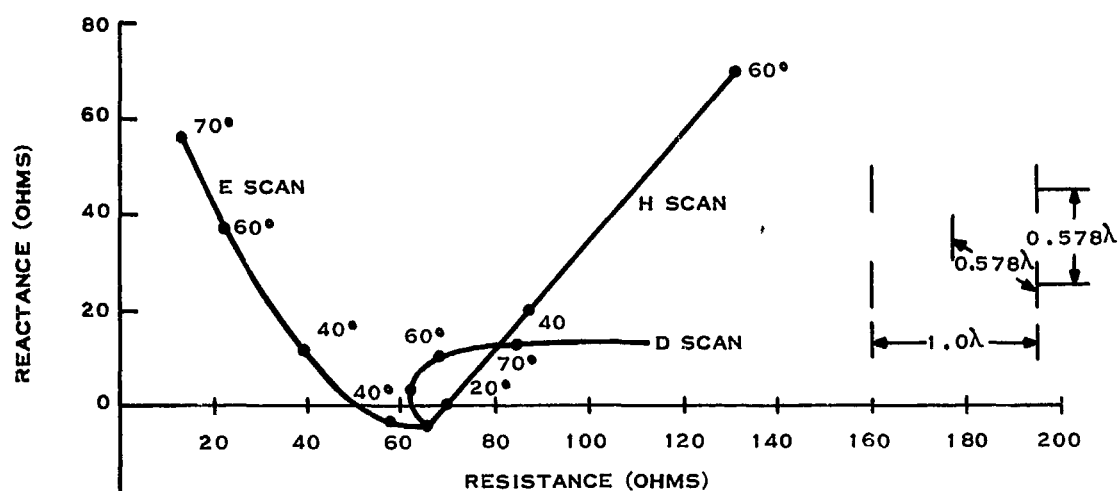
Figure A-22. Array Element Grids

broadside match both grids exhibit similar VSWR with scan. Thus, the most significant difference between the two grids is the reduction in the number of elements required for a given aperture obtained by using the triangular grid. Unfortunately, this reduction is obtained at the expense of an increase in the number of element rows in the array (in other words, by 1.73 times) which increases the complexity of the scan computer.

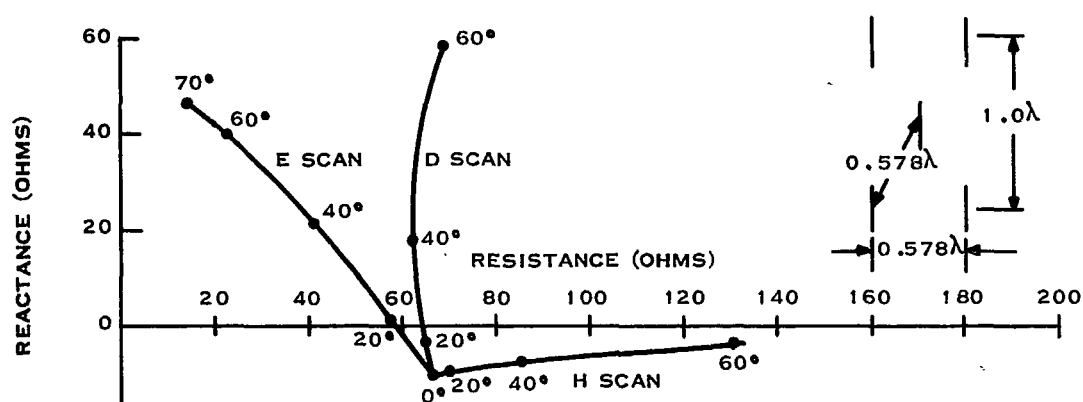
IV. FAN BEAM GENERATION

Beam broadening in a phased array refers to the ability to change the aperture distribution such that the beamwidth broadens or the beamshape is modified. A multipurpose phased array will have several beamshape requirements necessary for successful achievement of the mission. In most multimode arrays the pencil beam will be the beamshape in greatest demand. However, for some search modes a pencil beam that is broadened in either one or two dimensions would offer superior scan coverage.

Fan-beam formation or beam broadening can be accomplished by defocusing the array. The phase of the phase shifters can be adjusted to produce a parabolically shaped phase runoff across the array as shown in Figure A-24.



(A) HORIZONTAL TRIANGULAR GRID



(B) VERTICAL TRIANGULAR GRID

54621

Figure A-23. Impedance Scan for an Equilateral Triangular Grid
Infinite Array of Dipoles in Free Space

The phase change required at each radiating element for the example case is obtained by the following expression:

$$\delta = P \frac{d^2}{D} \quad (A-17)$$

where

δ = phase adjustment of element

P = total phase runoff across the array

d = distance from the element to the center of the array

D = distance from the end of the array to the center.

43925

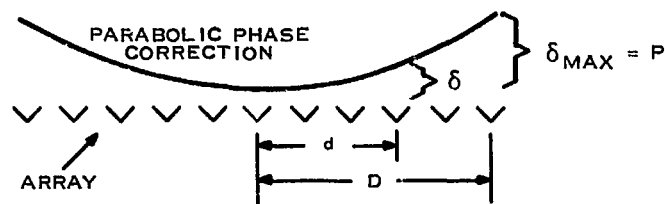
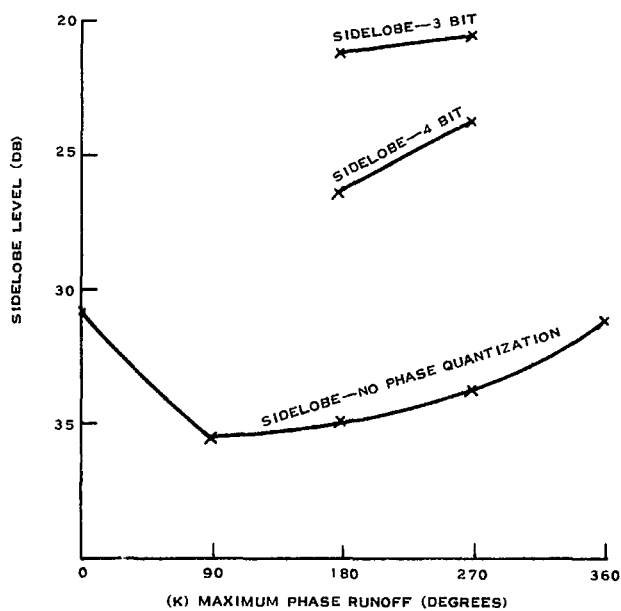


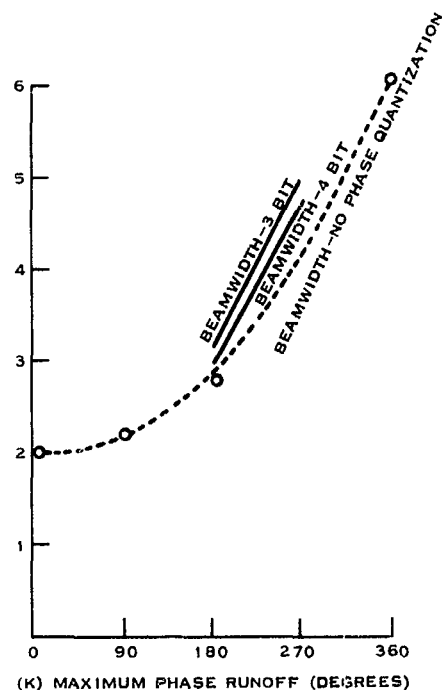
Figure A-24. Beam Broadening Configuration

This technique can be used to broaden the beam while maintaining a good beam shape. To determine the sidelobe level of the array as a function of beamwidth, the far-field pattern of two linear arrays was computed for various beam broadening factors. Figures A-25 and A-26 show amplitude tapers producing a 30-dB modified $(\sin x)/x$ pattern, and Figures A-27 and A-28 show 35-dB Taylor aperture illuminations. The solid lines show the half-power beamwidth and the sidelobe variation as the maximum phase runoff is varied from 0 to 360 degrees. The dashed lines show the effects of 3-bit and 4-bit quantization. A note of explanation is required as to the variation of the sidelobe level of the $(\sin x)/x$ pattern as the phase runoff is increased to 90 degrees. The decrease in sidelobe level occurs because the first sidelobe is



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Figure A-25. Sin x/x Amplitude Taper with Beam Broadening (Sidelobe Characteristics)



44956

Figure A-26. Sin x/x Amplitude Taper with Beam Broadening (Beamwidth Characteristics)

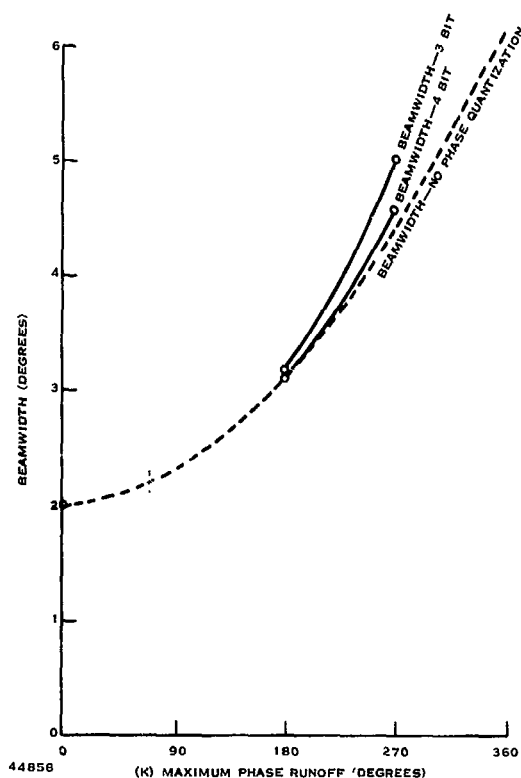


Figure A-27. 35-dB Taylor Amplitude Taper with Beam Broadening (Sidelobe Characteristics)

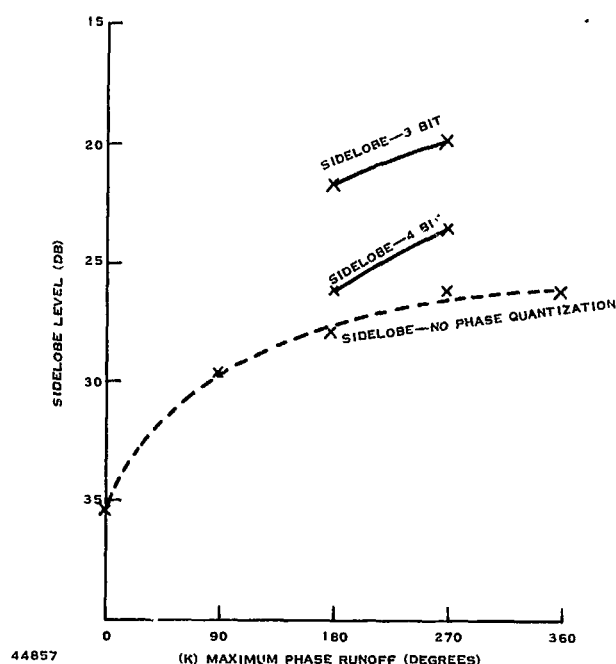


Figure A-28. 35-dB Taylor Amplitude Taper with Beam Broadening (Beamwidth Characteristics)

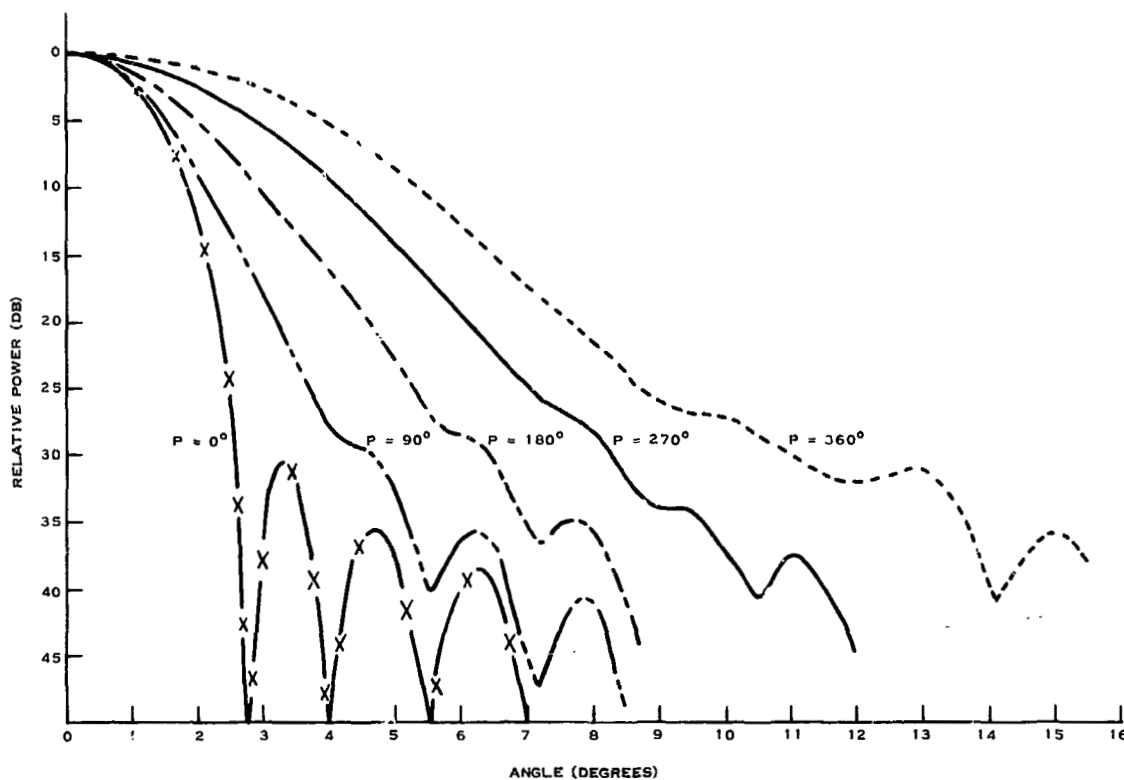
absorbed in the main beam, leaving the second sidelobe the determining factor. This sidelobe was less than 35-dB down with no phase runoff. The shape of the far-field patterns is shown in Figure A-29 for the modified $(\sin x)/x$ illumination.

V. PHASED-ARRAY RADIATING ELEMENTS

The design of the radiating element for phased-array applications is important for polarization requirements. Multipurpose phased arrays require various combinations of polarization for maximum mission effectiveness. In general, both linear and circular polarization may be required. If dual receivers are used, orthogonal polarizations may be simultaneously detected for optimum target enhancement.

A. Linear Polarized Elements

Three types of elements are generally used as radiating elements in a linearly polarized array. These elements are the dipole, slot, and open-ended waveguide.



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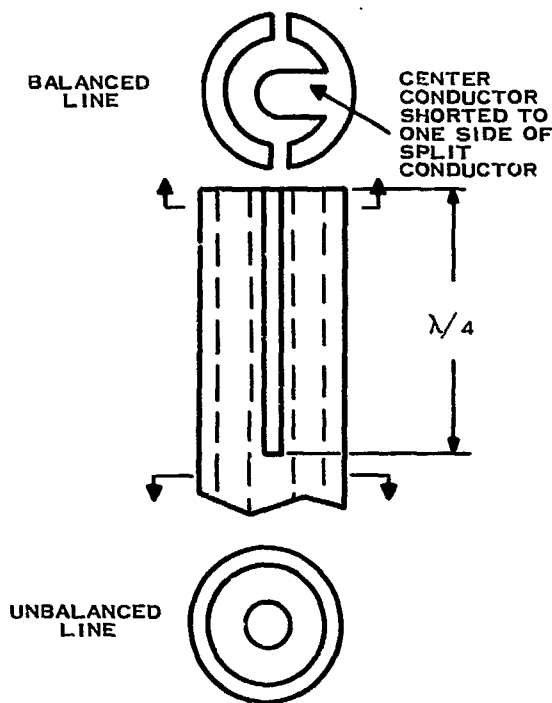
Figure A-29. Far-Field Patterns for Modified Sin x/x Distribution with Beam Broadening

1. Dipole

The dipole element is the simplest element from the standpoint of construction and analysis. There is a large amount of analytical data available for dipole arrays.^{98,99} The dipole array may be easily analyzed as discussed in Section I. The construction of a dipole for frequencies through 9.0 GHz remains relatively simple; but as frequencies increase above 9.0 GHz, the dipole is much less attractive compared to a waveguide radiator because fabrication becomes very difficult.

The dipole element also has the advantage of light weight as compared to open-ended waveguide. When compared to a well-designed slot radiator, there is very little, if any, weight advantage for the dipole.

Several methods are available to excite the dipole. The dipole is a balanced structure requiring a balanced feed. A balanced line can be used to feed the dipole; but the transmission line will radiate for large mismatches, thus requiring the balanced line to be shielded. A more desirable feed would be the split-tube-fed dipole.¹⁰⁰ The split-tube transmission line (balun) shown in Figure A-30 transforms the balanced line at the dipole to an unbalanced transmission line. There is a 4:1 impedance transformation from the unbalanced line to the balanced line feed; also, a reactance is contributed by the balun structure. Riblet¹⁰¹ defined the equivalent circuit of the balun and verified it experimentally.



54622

Figure A-30. Split-Tube Balun

The split-tube balun structure has excellent strength and can be used to support the dipole. This configuration was used for the MERA radiating element as shown in Figure A-31 with a sectional view of the dipole shown in Figure A-32. Figure A-33 illustrates an S-band dipole of similar design.

One of the major disadvantages of the dipole is that it cannot be easily flush mounted. A radome is generally required to cover the dipole array which presents all the radome problems normally encountered plus the array performance versus scan problems.

2. Slot

The slot radiator is useful for flush-mounted applications. Mutual coupling data calculated for a dipole array in free space is analogous to data for a slot array by Babinet's principle. The E- and H-planes must be interchanged and the impedance is given by

$$Z_s = \frac{\eta^2}{4Z_D} \quad (\text{A-18})$$

where

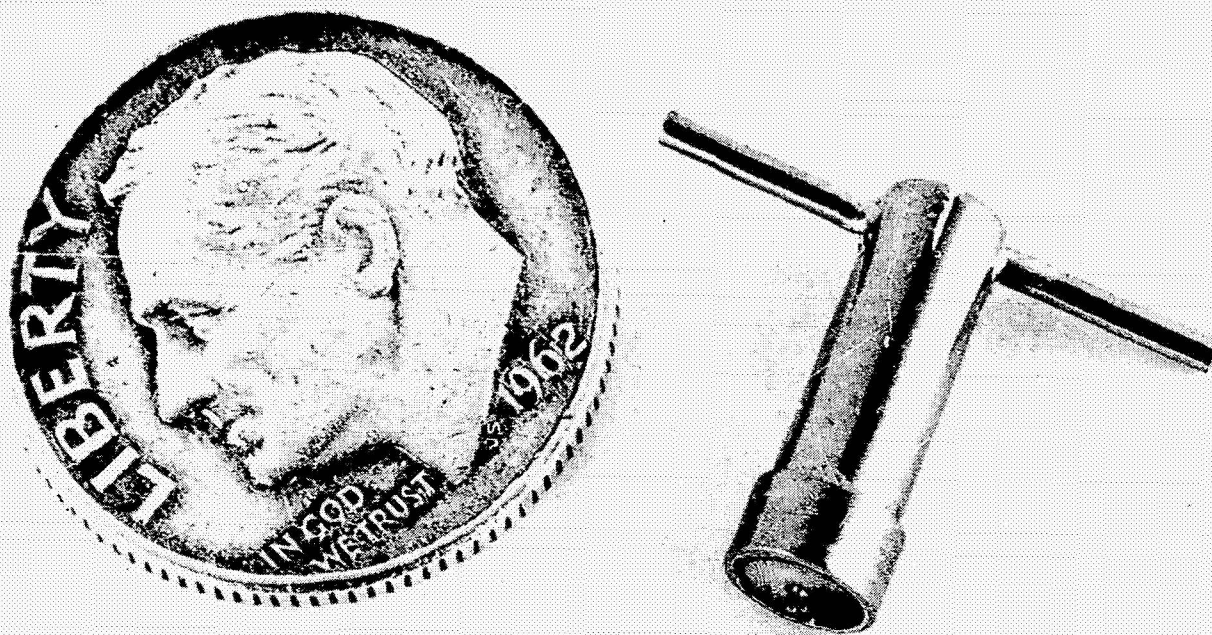
Z_s = slot impedance

Z_D = dipole impedance

η = intrinsic impedance of free space.

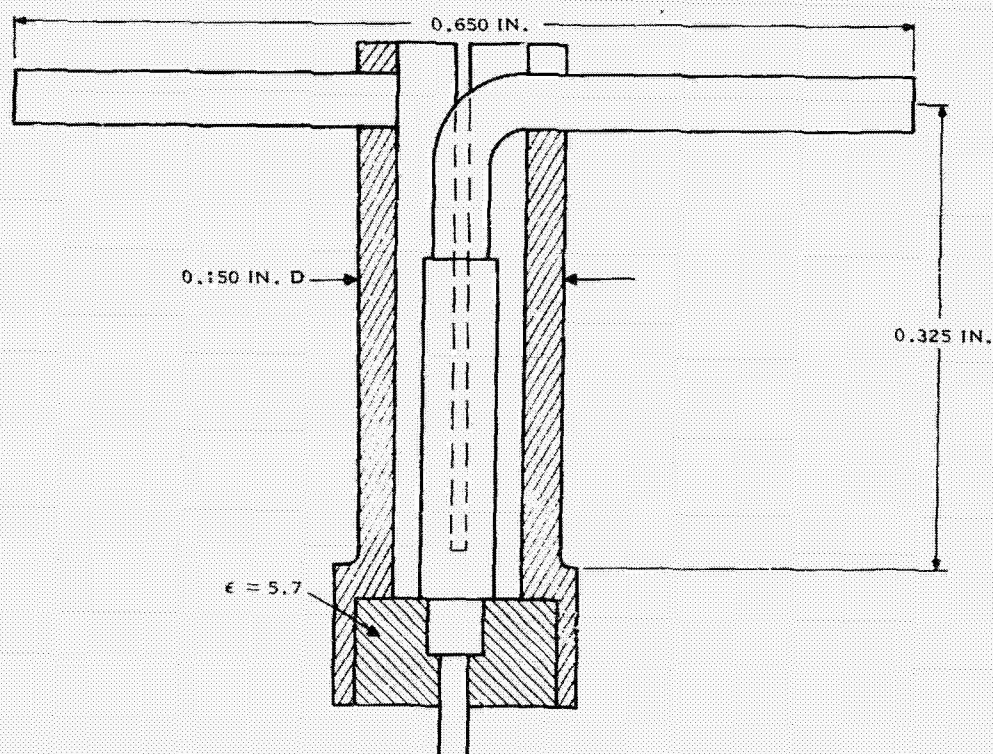
A disadvantage of the slot is the complexity of the feed. Either a cavity which backs the slot must be excited or the slot must be excited directly. Three techniques for exciting the slot radiator are shown in Figure A-34. The stripline-excited slot is feasible over the same frequency range as the dipole antenna, and the cavity-backed slot with the cavity excitation may be used at frequencies above the upper limit of the dipole.

The impedance variation versus scan of the slot antenna is somewhat worse than the dipole mounted $\lambda/4$ above a ground plane. A comparison is shown in Figure A-35. This difference in variation may limit the maximum scan of a slot array to a greater extent than a comparable dipole array.



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Figure A-31. MERA Dipole



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Figure A-32. Sectional View of MERA Dipole

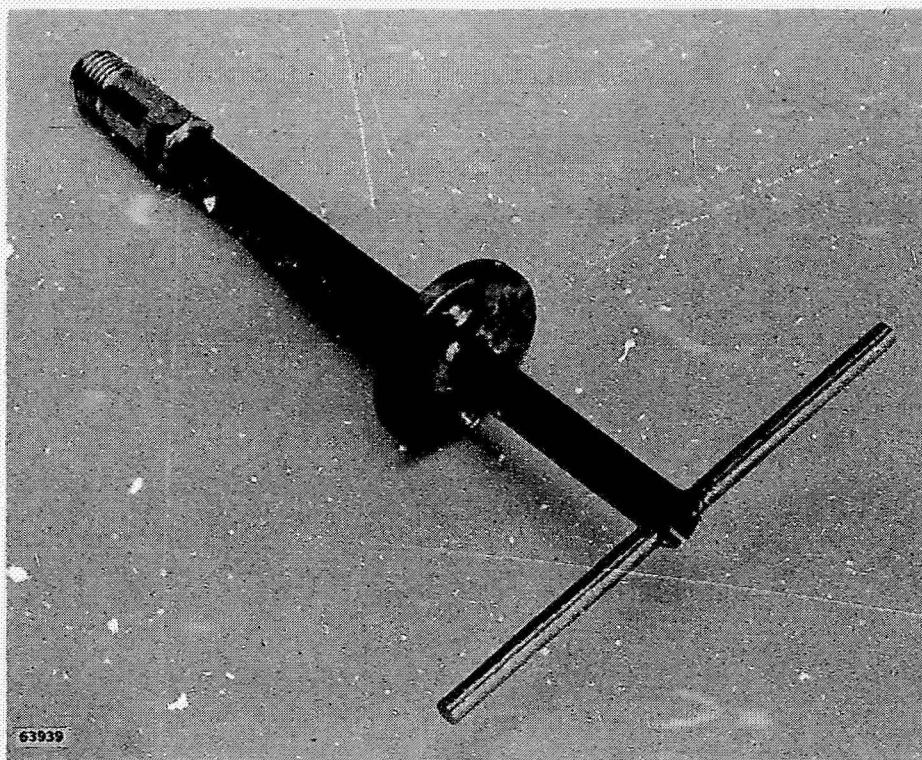


Figure A-33. S-Band Dipole

3. Open-Ended Waveguide

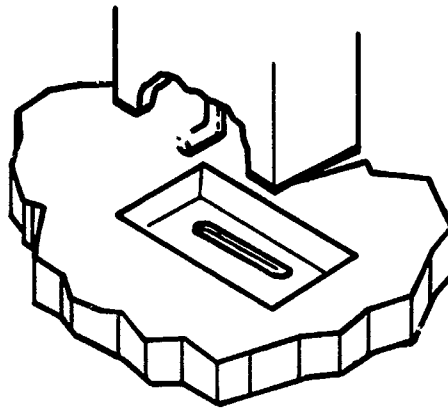
The open-ended-waveguide radiator may be considered as a wide slot. Diamond ⁹⁵ has shown good correlation between data measured on an open-ended-waveguide array and data calculated for a dipole array in free space; therefore, the slot coupling data was used in the design of open-ended-waveguide arrays.

An unloaded rectangular waveguide may be used in a linearly polarized array by stacking the elements in a brickyard fashion as shown in Figure A-36. For the element spacing shown in the figure, no grating lobe will appear for scanning in the E- or H-planes, but a grating lobe will appear for a 60-degree scan in the diagonal plane. The waveguide must be fed from the rear because of the tight packing.

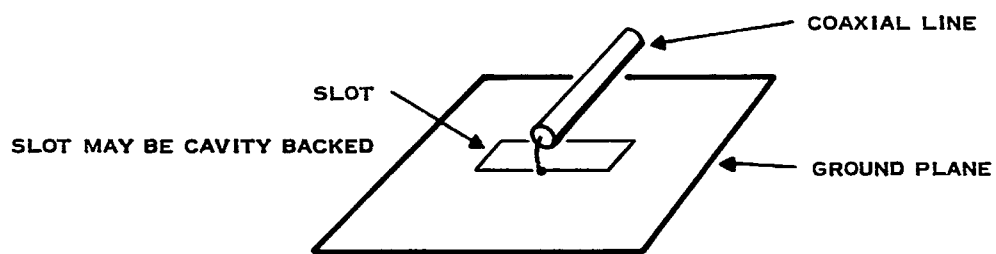
This element is also useful for flush mounting, and by filling each aperture with the proper dielectric, the radome can be eliminated.

B. Circularly Polarized Elements

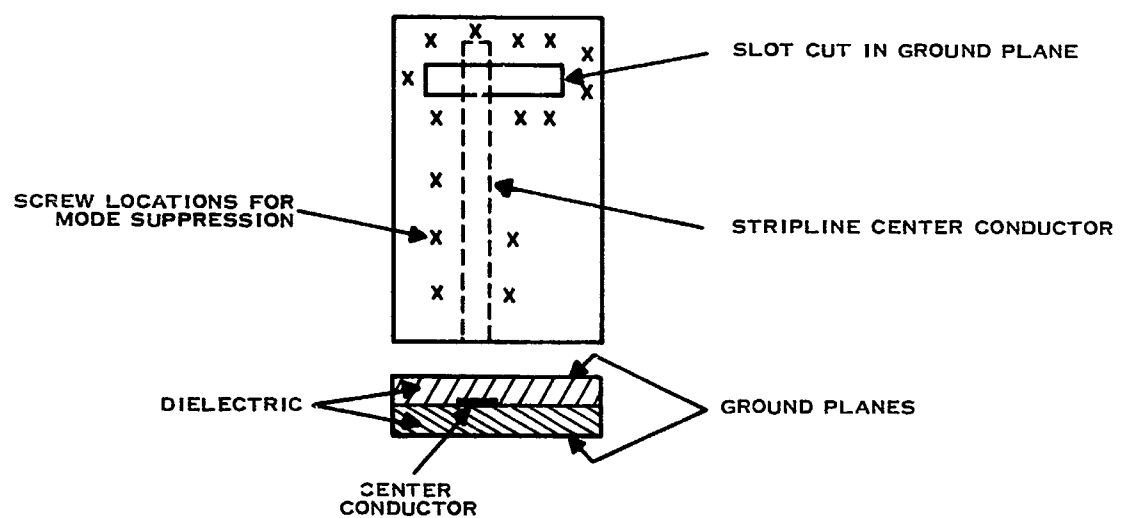
Five types of elements are considered for phased arrays in which only circular polarization is required. The elements discussed are the Archimedean spiral, helix, crossed-dipole, dipole-slot, and waveguide. Linear polarization using these elements can be achieved with proper addition



A. CAVITY EXCITED SLOT



B. DIRECTLY EXCITED WITH COAX



C. STRIPLINE EXCITED SLOT

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Figure A-34. Types of Slot Excitation

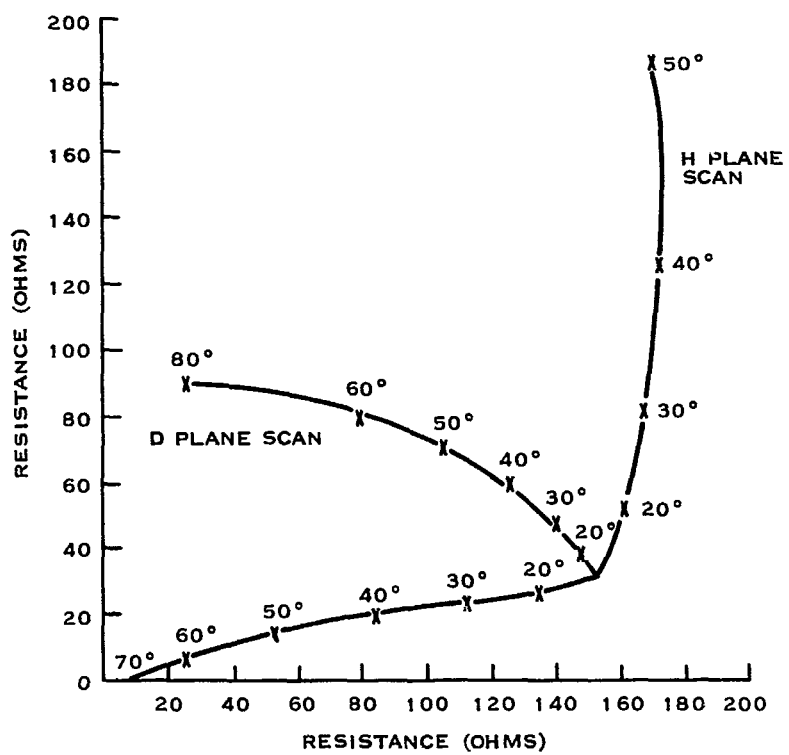
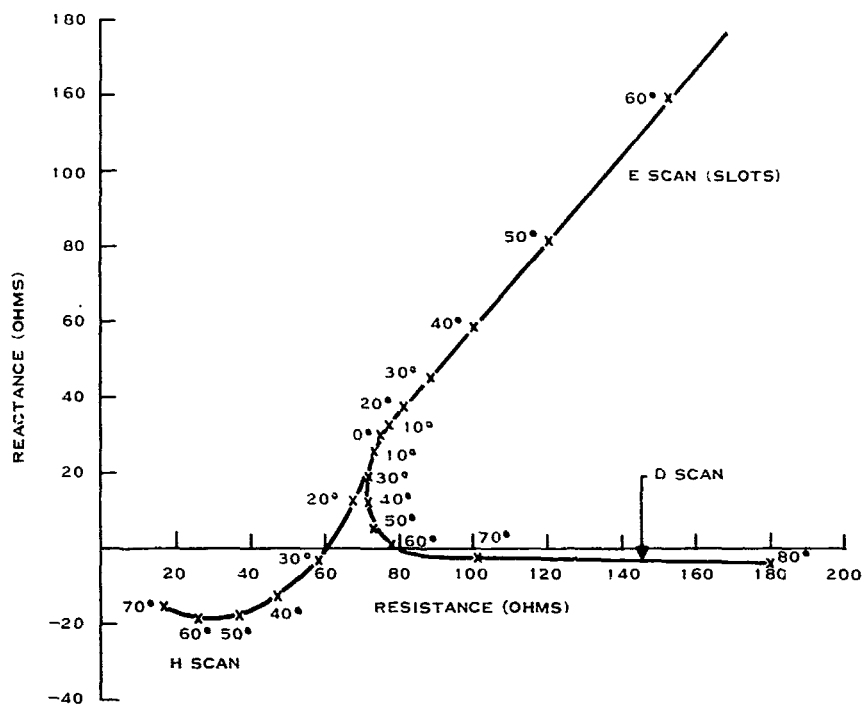
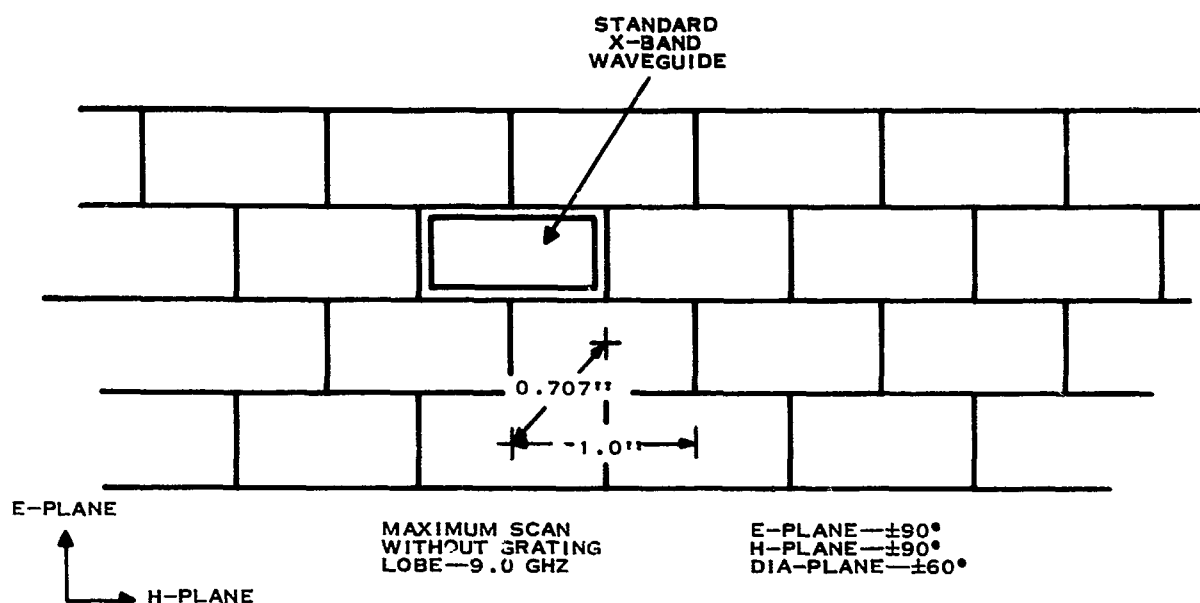


Figure A-35. Resistance Versus Reactance for Infinite Arrays of Dipoles and Slots



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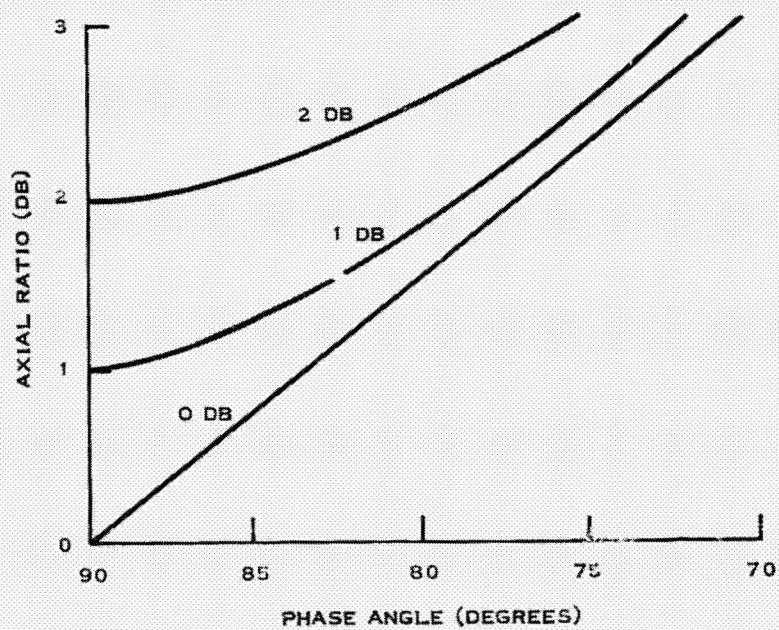
Figure A-36. Open-Ended-Waveguide Array

of the orthogonal circular polarizations. Because of mutual coupling, both the amplitude and phase of the orthogonal components generating the circular polarization will vary. This variation of amplitude and phase will cause a degradation of the axial ratio. The variation of axial ratio as a function of phase and amplitude is shown in Figure A-37.

1. Spiral

A log spiral or an Archimedean spiral can be used in the array application for circular polarization. In general, X-band is the highest frequency presently recommended for use of a spiral radiator. Figure A-38 illustrates an X-band Archimedean spiral. The maximum diameter is 0.5 inch which is small enough that two spirals could be placed in the space for an X-band rf building block.

The small ring around the outside diameter of the spiral is 50-ohms-per-square resistive material to absorb currents which were not radiated in the Mode 1 region. To achieve this small size, some gain was sacrificed (greater than 5 dB). For a $\lambda/2$ spaced array of spirals, some loading would still be required but the amount of gain reduction would not be as severe. If a load ring were not used, the currents which were not radiated would be reflected from the ends of the spiral and would be radiated in the Mode 1 region but in the orthogonal sense of polarization. This design



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Figure A-37. Ratio of Orthogonal Components

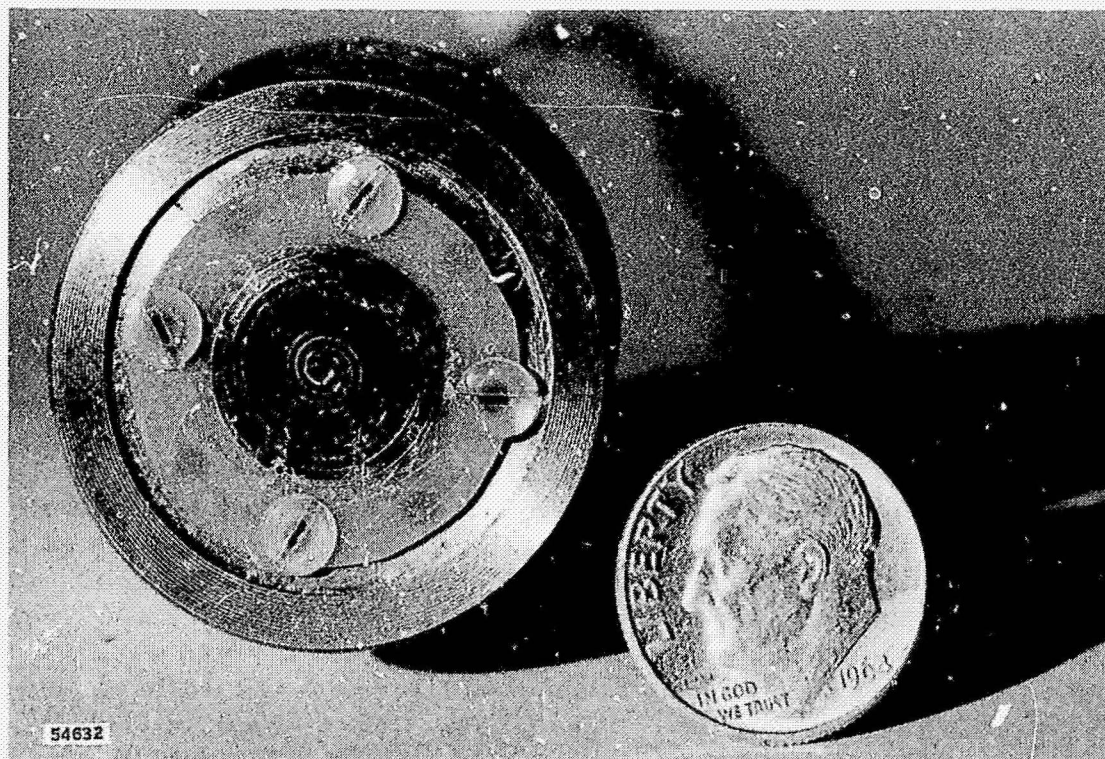


Figure A-38. X-Band Spiral Antenna

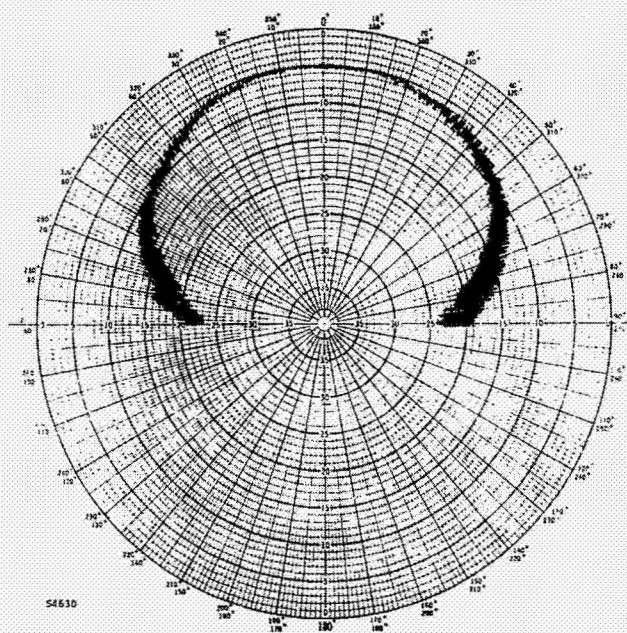
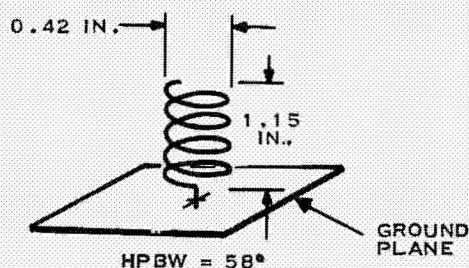


Figure A-39. Pattern of X-Band Spiral

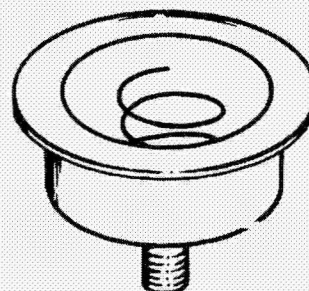
2. Helix

The helix ¹⁰² antenna can also be used for a circularly polarized array. The helix was only considered in passing because it would be similar to the spiral antenna. The approximate dimensions for an axial mode helix are shown in Figure A-40a. Flush mounting is possible by using the helix to excite a cavity as shown in Figure A-40b.

The generation of linear polarization is the same as for a spiral doublet and will not be discussed.



(A) AXIAL MODE HELIX



(B) CAVITY MOUNTED HELIX

Figure A-40. Helix Array Element

would give a higher axial ratio since the orthogonal radiated polarization would add to give an elliptically polarized wave. The pattern of the X-band Archimedean spiral is shown in Figure A-39. The axial ratio is superimposed on the far-field pattern by rotating a linearly polarized source while taking the pattern.

If linear polarization were also required, a spiral doublet (two orthogonally polarized spirals) would be needed. The two senses of circular polarization with equal amplitudes can be added at the proper phase to give any linear polarization. However, for the case where linear polarization is the primary polarization for a radar, an element which is normally linearly polarized is the better choice for improved cross polarization characteristics.

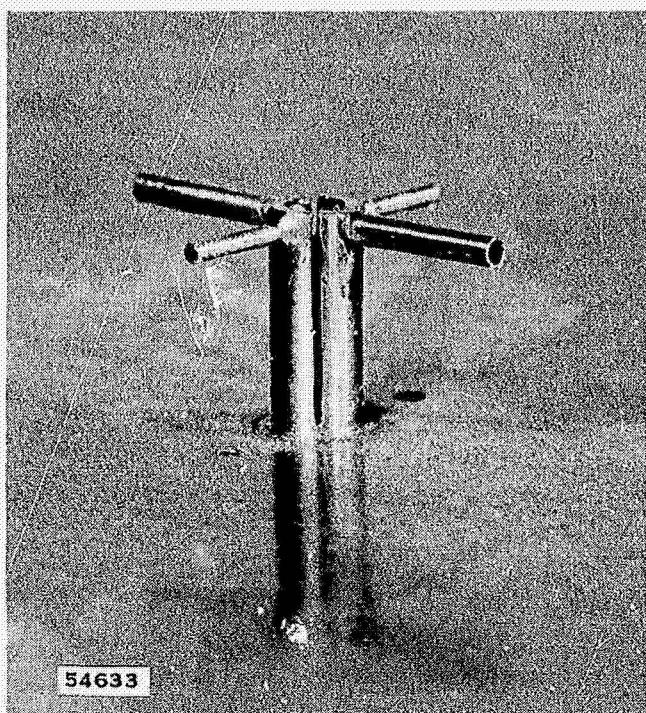


Figure A-41. Crossed-Dipole Antenna

3. Crossed Dipole

A crossed-dipole antenna is shown in Figure A-41. This crossed-dipole design is discussed by Bolster.¹⁰³ The feed is a simple slit-tube balun as used for a dipole, but in this case extra elements are added. To achieve circular polarization, orthogonal linear elements are added in phase quadrature. The dipoles are fed with equal amplitudes by keeping the two dipole L/D ratios equal. Phase quadrature is achieved by variation of the dipole lengths until there is a 90-degree difference in the two dipole impedances.

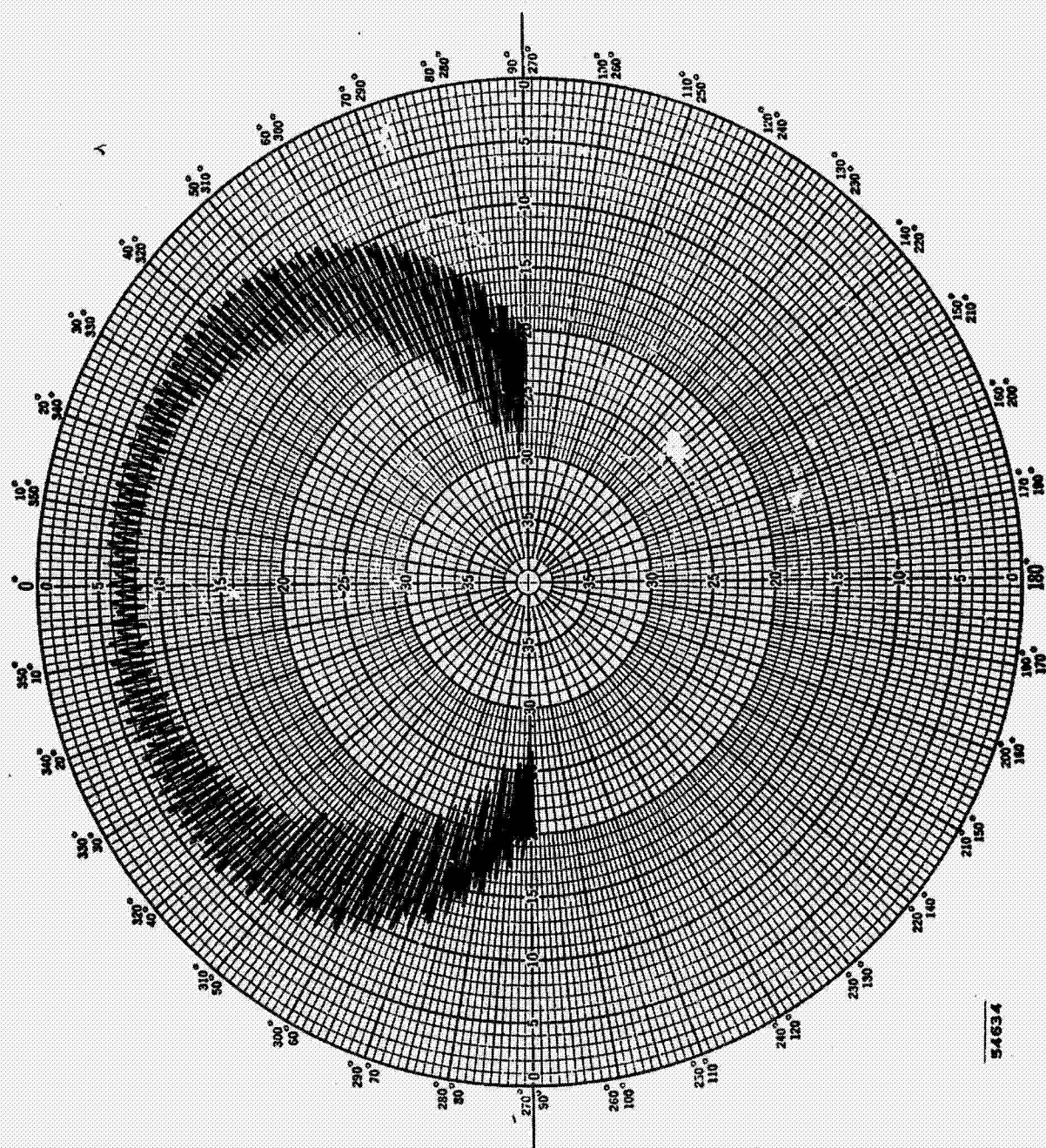
As seen in Figure A-42, this element is not as useful for circularly polarized phased arrays because the off-axis axial ratio is very poor. The axial ratio of Figure A-42 could be improved, but generally this element is not as useful for the 45-degree scan angles expected in a phased array.

4. Slot-Dipole

The slot-dipole is a dipole mounted in a cavity as shown in Figure A-43. The cavity acts as a parasitic element to give circular polarization. The operation of this type of element is explained by Wilkinson.¹⁰⁴ A pattern of the slot-dipole is shown in Figure A-44.

5. Waveguide

The last element to be discussed for circular polarization only is the open-ended waveguide with polarizer. To achieve circular polarization, a symmetrical waveguide element (circular or square) is fed to set up the dominant mode (TE_{10} for square waveguide or TE_{11} for circular waveguide) and a dielectric is used to give phase quadrature. In the case of a circular waveguide, the polarizer is placed at 45 degrees to the polarization so that one component of the E vector is parallel to the polarizer and the other component is perpendicular as shown in Figure A-45. The component parallel to the polarizer will be slowed because of the change in velocity of propagation by the dielectric. Phase quadrature required for circular polarization is achieved by adjusting the length of the polarizer to give the 90-degree phase difference between the two components.



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Figure A-42. Crossed-Dipole Pattern

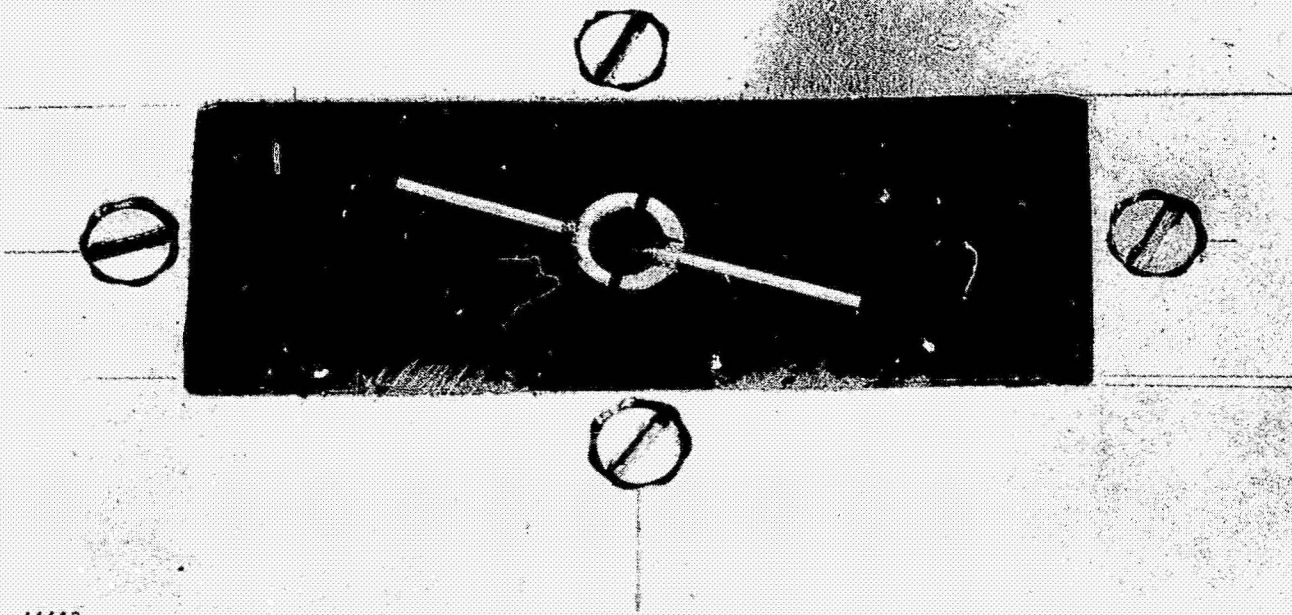


Figure A-43. Slot-Dipole Element

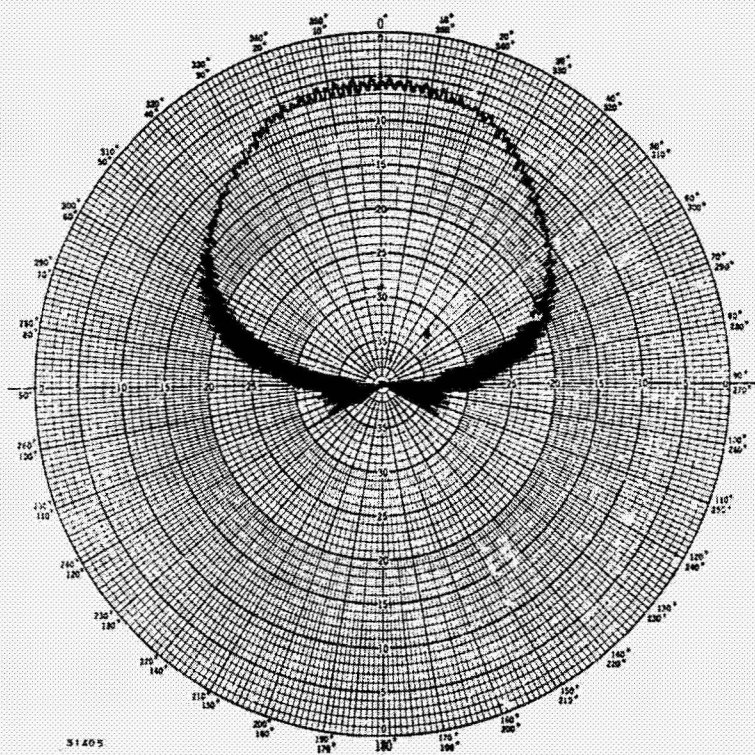
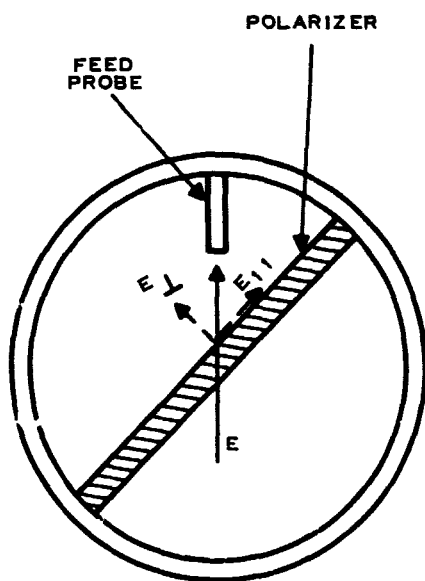


Figure A-44. Slot-Dipole Pattern



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Figure A-45. Circular Waveguide with Polarizer

are excited with a balanced line achieved with ring hybrids. An illustration of the dipoles with processing required for multiple polarization is shown in Figure A-46. The 3-dB branch line coupler feeding the ring hybrids is used to give the required phase quadrature. The available polarizations are shown in Figure A-46.

The axial ratio of a crossed dipole is affected by the phase between the dipoles and the difference in power level between the dipoles. With correct phasing, the axial ratio varies as shown in Figure A-47.¹⁰⁵ This variation is caused by the amplitude difference of the E- and H-plane pattern illustrated in Figure A-48.

Figure A-49 illustrates an X-band crossed dipole built by Texas Instruments. Ring hybrids were constructed to feed the X-band crossed dipole and the phase quadrature was achieved with a length of line. The processing for the crossed dipole can be constructed on microstrip but must be placed inside the rf module. A branch line coupler, two ring hybrids, and the switching will be required for a polarization diverse system. The isolated element pattern with axial ratio superimposed is shown in Figure A-50. Radiators with separated phase centers as shown in Figure A-51 are useful only if the spacing "S" is on the order of the array element spacing required for other array considerations. If standard dipoles are used as shown in Figure A-51 then ($S \approx \lambda$) which will cause severe axial-ratio degradation from the peak of the beam. The separation of phase centers causes a phase shift between polarization as the array is scanned. However, if loaded slots are used (or a

C. Multiple Polarization

Array elements useful for multiple polarization must be able to support orthogonal linear polarizations in most cases. For instance, a crossed dipole can be used to support orthogonal polarizations. Circular or square waveguide can also be used for orthogonal polarizations at the higher frequencies, where a waveguide radiator is required for fabrication considerations. Groups of orthogonal linear polarized elements, such as slots, can also be used if the group phase center spacing meets other array requirements.

1. Crossed Dipole

A crossed dipole can be used to generate the orthogonal polarizations required for circular polarization. For best wide scan angle characteristics orthogonal $\lambda/2$ dipoles with the same phase center can be used. The dipoles

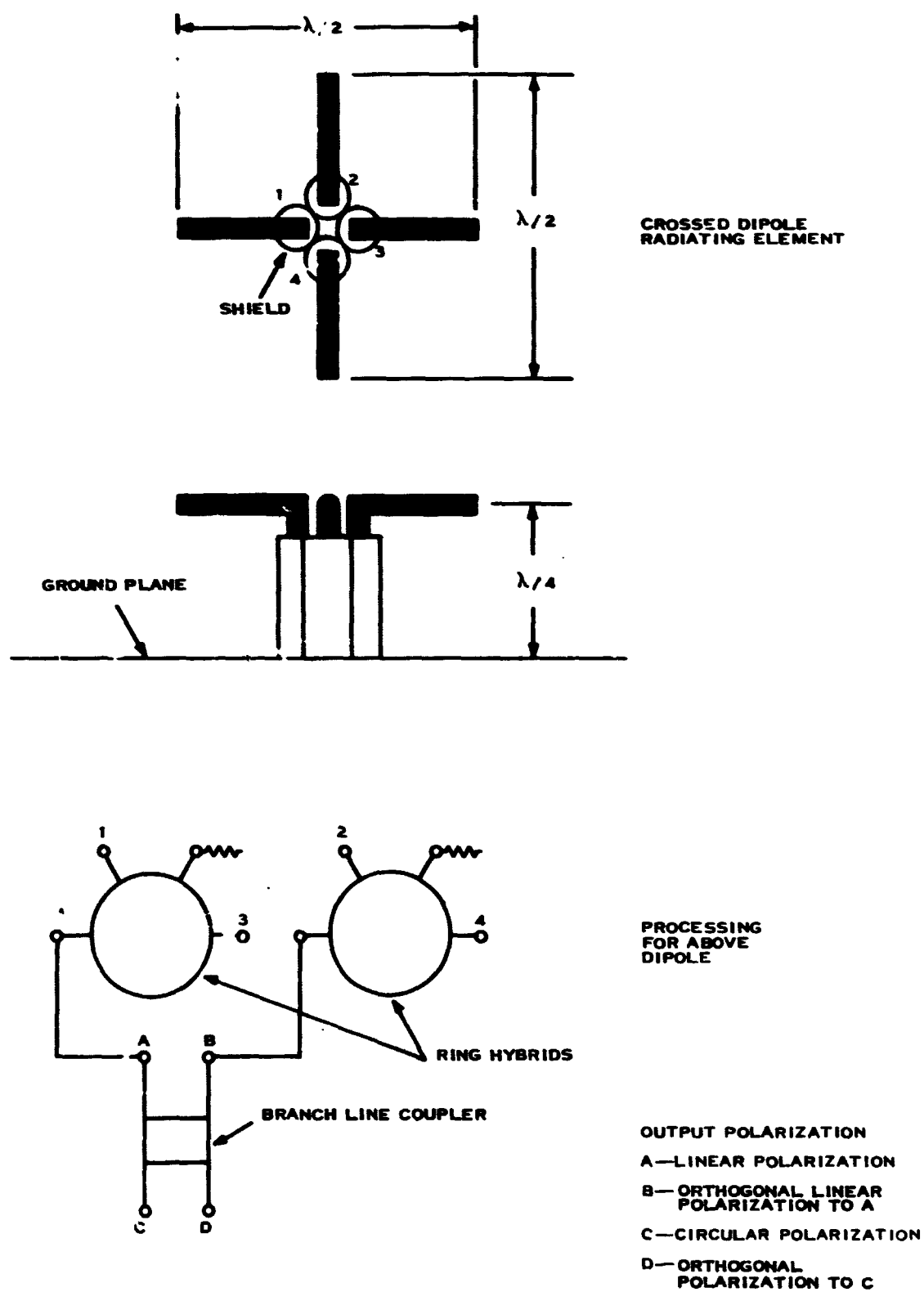


Figure A-46. Crossed-Dipole Configuration

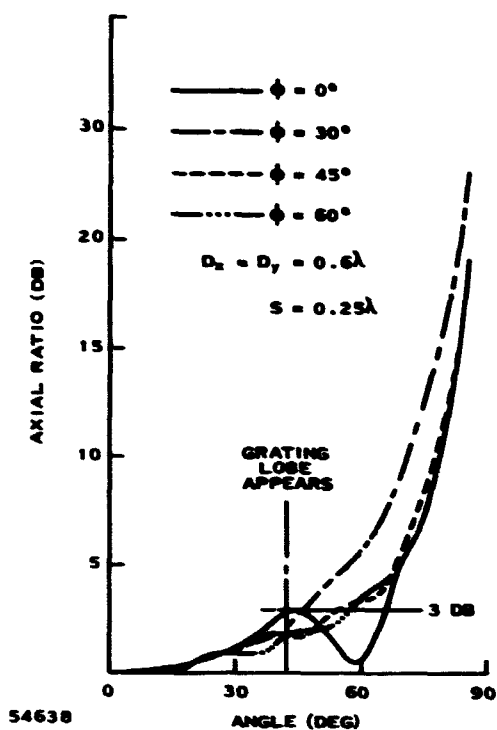


Figure A-47. Axial Ratio of Central Element in a Planar Array of Crossed-Dipole Antennas

similar reduced-size radiator) then the spacing may be reduced so that ($S \doteq \lambda/2$) and the polarization characteristics of the array will meet the scanning requirements.

2. Waveguide Elements

The decision to use a waveguide element will depend heavily upon the choice of frequency. In general, waveguide elements should be considered for X-band and K_u -band applications and crossed dipoles or slots for L-, S- and C-bands. Since the waveguide elements should be symmetrical in order to propagate orthogonal polarizations, the waveguide shape will generally be square or circular.

Waveguide elements for closely spaced arrays ($\lambda/2$ spacing) must be loaded in order to propagate. Both dielectric and ridge loading are used in the design of radiating elements. Table A-I shows the advantages of each method of element loading. One

of the most important considerations is the weight of the radiator. Figure A-52 illustrates the importance of weight in a spaceborne phased array using X-band element weights as an example. X-band represents a general cross-over point in that dual-mode waveguide for lower frequencies is generally much heavier than corresponding crossed dipoles or crossed-slot radiators. At K_u -band the dipole elements would be impractical from a construction viewpoint and the loaded-waveguide elements would be lighter in weight than shown in Figure A-52. For instance, the weight of a single K_u -band double-ridge-waveguide radiator constructed by Texas Instruments is 0.03 oz.

To obtain dual or circular polarization from the ridge-waveguide radiator, four ridges are used so that orthogonal linear polarizations can be propagated in the guide. A rear-probe feed is required because the array elements are so tightly packed that a sidewall feed is not feasible. The probes excite the TE_{10} and TE_{01} modes in the square waveguide. Figure A-53 shows several engineering prototypes of quad-ridge waveguide and illustrates one method of using a rear-fed probe to excite the E field in the waveguide (for a single mode). Shown (from left to right) are quad-ridge elements for K_u -band, X-band and C-band.

Table A-I. Loading of Waveguide Elements for Phased-Array Radars

	Dielectric Loading	Ridge Loading
Element Weight	Can be excessive for large arrays for some dielectrics	Hollow ridges may be used for weight reduction
Waveguide Type	Best suited for circular waveguide	May be used for either circular or rectangular waveguide
Impedance Matching	Excessive loading creates a larger discontinuity which must be matched	Ridge size and spacing can be varied for matching purposes
Phase-Shifter Types	Well suited for matching into ferrite phase shifters or diode phase shifters in a waveguide configuration	Best for matching into diode or ferrite phase shifters in a microstrip configuration
Surface Array Modes (causing "holes" in patterns)	Will tend to propagate a surface wave mode in some cases	Not as likely to support surface modes

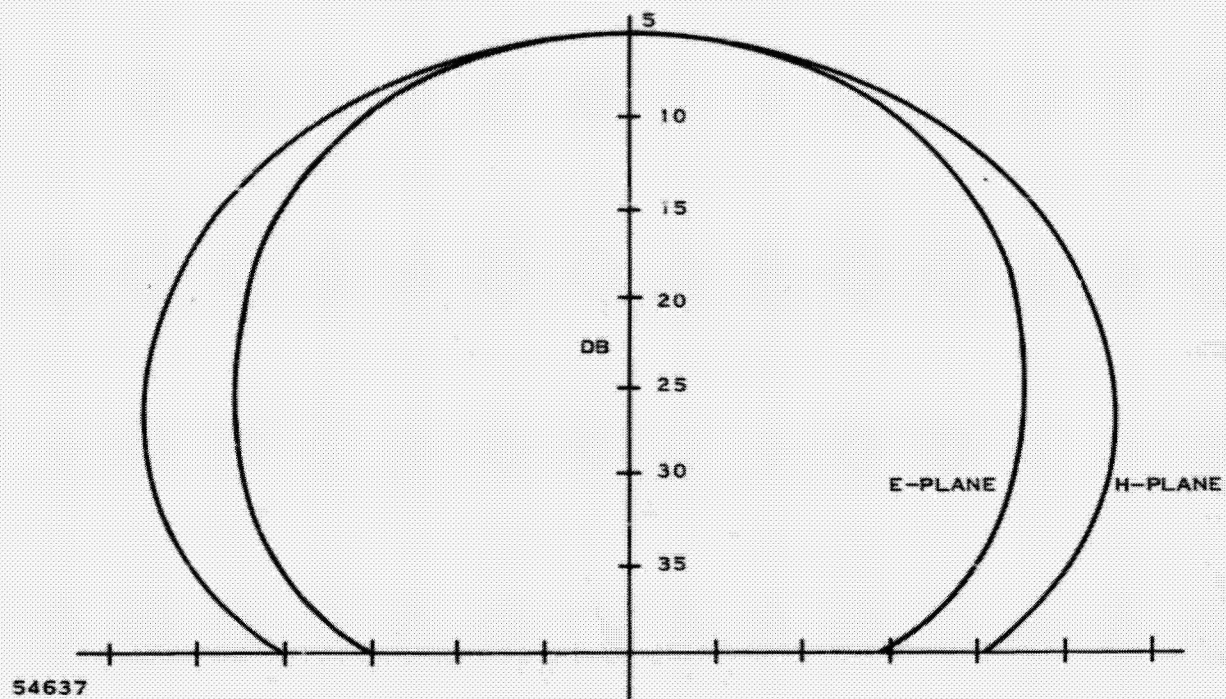


Figure A-48. Dipole Pattern, Dipole Element



Figure A-49. λ -Band Crossed-Dipole Antenna

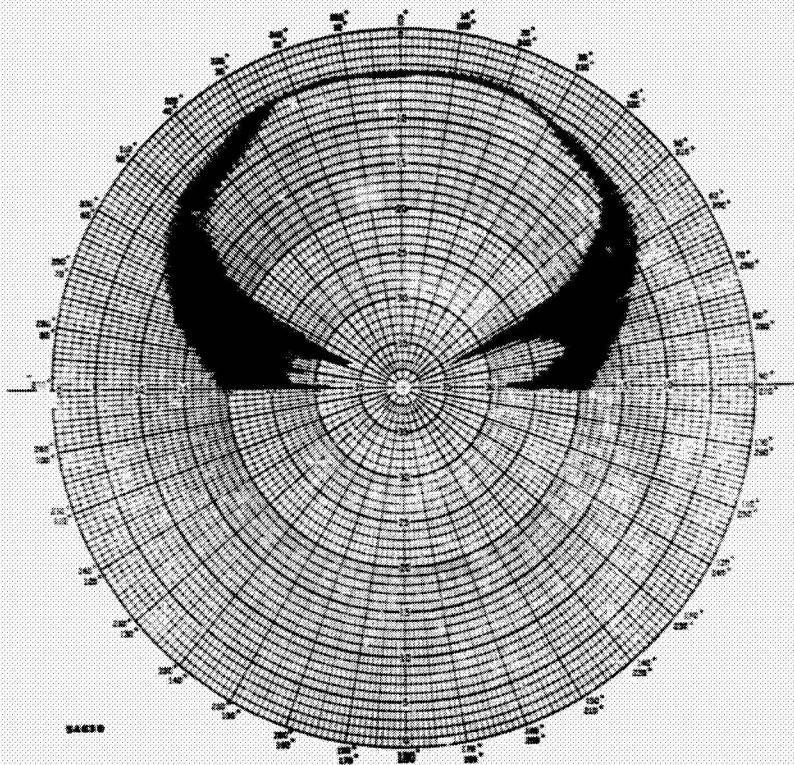
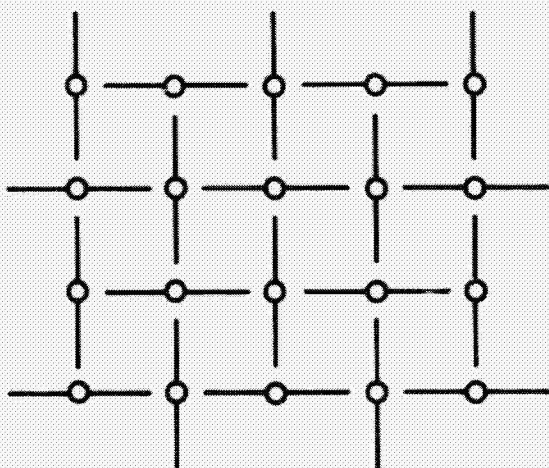


Figure A-50. Crossed-Dipole Pattern

To achieve circular polarization, the probes are phased in quadrature as shown in Figure A-54. The circuit for linear plus orthogonal circular polarization requires a one-bit phase shifter as indicated in the diagram.

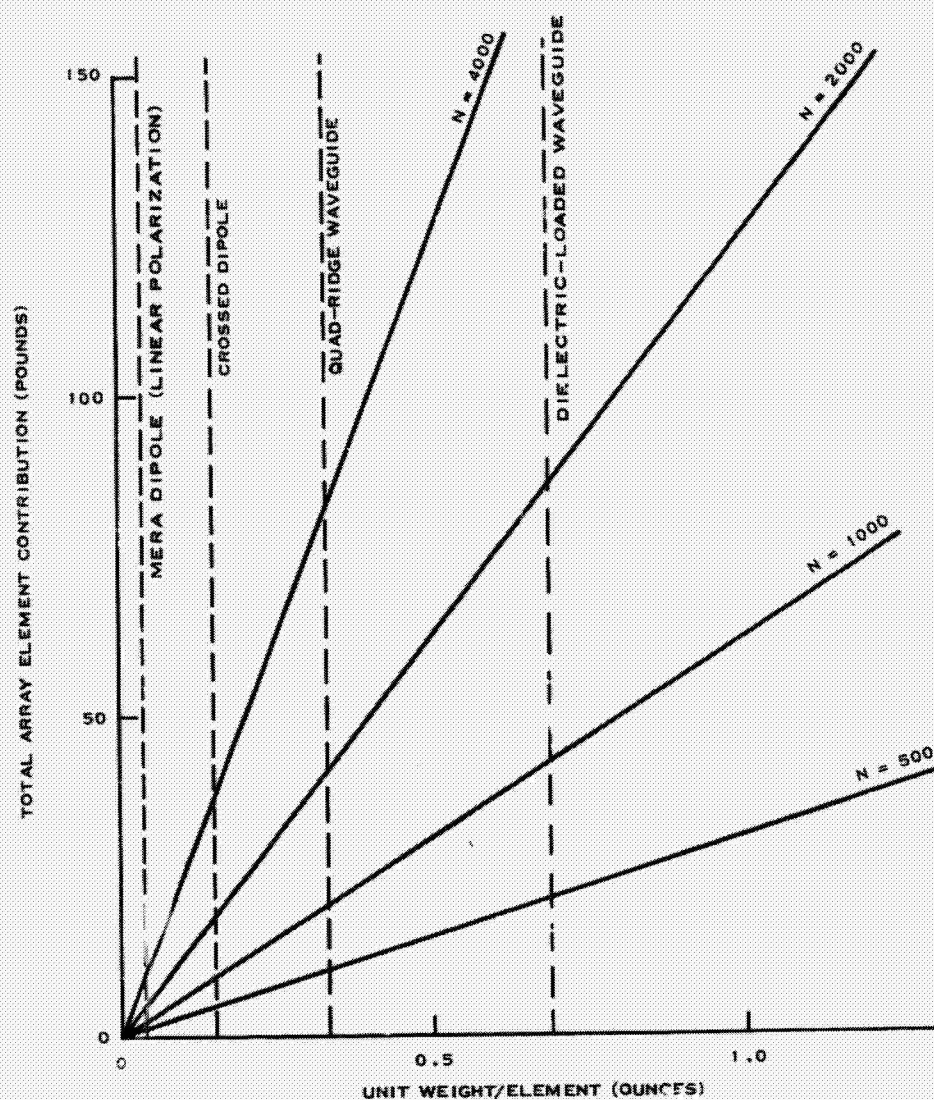


The linear polarization is an "in-phase" combination of the TE_{10} and TE_{01} modes, and the resulting polarization will be a 45-degree slant polarization with respect to the orientation of the waveguide.

Figure A-55 illustrates a K_u -band module for a front-fed K_u -band phased array which uses dual-ridge waveguide. Ridge size was chosen so that additional ridges could be added at a later time for a multipolarization module.

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Figure A-51. Orthogonal-Dipole Array with Separated Phase Centers



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Figure A-52. X-Band Element Weights

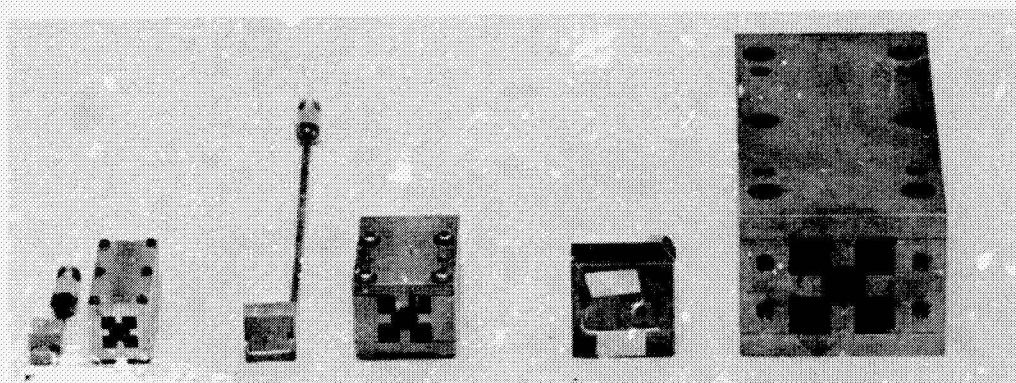
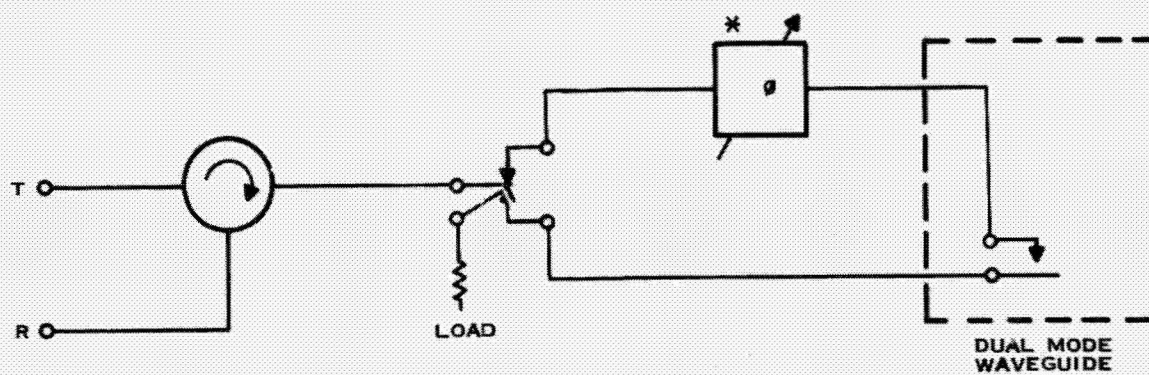


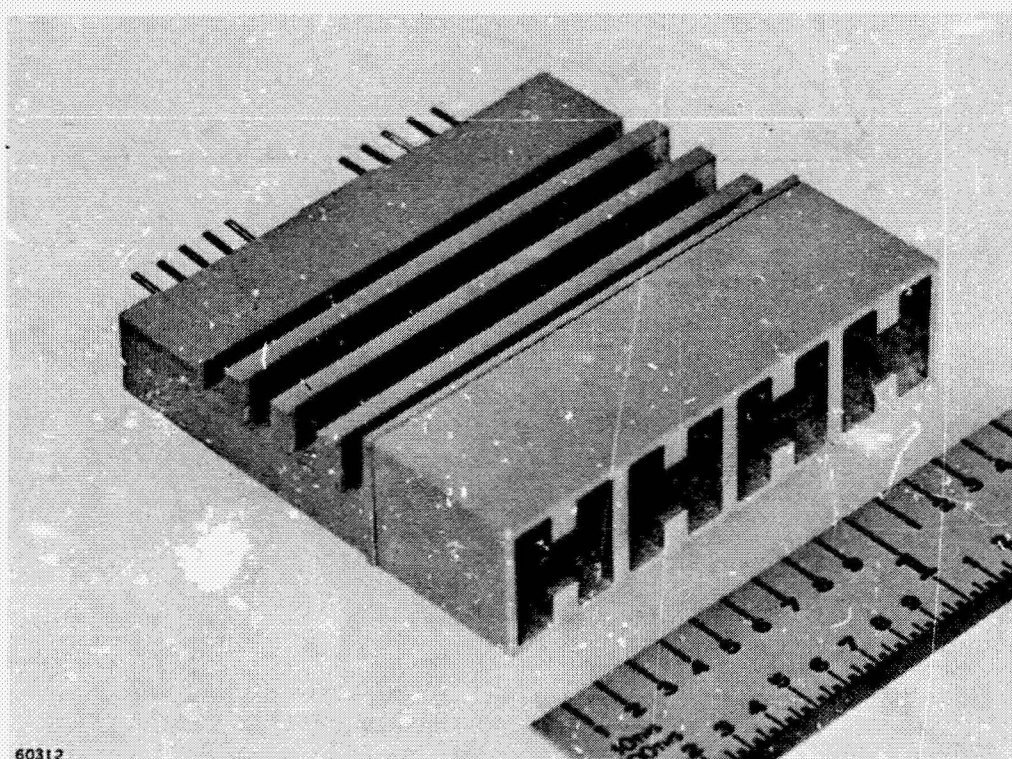
Figure A-53. Quad-Ridge Waveguide Elements



MODE	POLARIZATION	* 1-BIT PHASE SHIFTER PHASE
1	LINEAR	0°
2	CIRCULAR	90°

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Figure A-54. Multipolarization Switching Network



60312

Figure A-55. Space-Fed Array Phase-Shift Module

APPENDIX B

MATCHED-FILTER THEORY AND PULSE COMPRESSION

I. INTRODUCTION

Because the Combined Acquisition and Tracking (CAT) Radar requirements do not indicate the need for an extremely high pulse compression ratio, at least for this first look at the problem, it was decided not to devote a major part of the main body of the study report to the subject of pulse compression. Rather, the topic will be discussed from a general standpoint in this appendix section. In this way it is felt that the treatment will be of more general usefulness to someone using the results of this report when more definitive conditions have been established for the CAT radar program.

Quite possibly, in the near future there will be a new component of general usefulness to the whole area of matched filter applications. Work is underway at Texas Instruments on this component called a surface-wave device ¹⁰⁶ (SWD). This device can be classified in a gross sense as a tapped delay line. What distinguishes the SWD from other tapped delay lines is the promise of flexible tap locations, small size, higher SNR, and many desirable attributes attendant to the batch fabrication method.

The remainder of this appendix section will consist of a discussion of radar and communication system trends, matched-filter theory, and a summary of filter requirements and state of the art.

II. TREND IN RADAR AND COMMUNICATION SYSTEMS

One of the goals most consistently emphasized in the development of modern radar and communication systems has been the maximizing of the information transfer rate from transmitter to receiver under the bandwidth, signal power, and equipment constraints of particular systems or applications. In the case of radar systems much research and development effort has been directed toward means of increasing the maximum range capability of peak-power limited systems (without compromising on range resolution) and of improving their target-background (subclutter) visibility. In communication systems, the research and development effort has been focused mainly on efficient methods of coding and decoding data such that the minimum energy per received bit is required to transfer data from the transmitter to receiver at some maximum acceptable bit error rate.

In both radar and communication systems, the greater performance efficiency being sought is very often achievable by means of specially designed signals or codes for transmission and specially designed processors or matched filters for detection (or decoding) at the receiver. Generally, the implementation of the matched-filter signal generator at the transmitter is

much simpler than the counterpart detector implementation at the receiver. Although there is much work to be done on the design of special matched-filter signals for various applications, the accomplishments so far in this area appear to have out-distanced the work of finding or developing better and simpler means of implementing the matched filter at the receiving end.

The two subsections that follow present information on the mathematical theory of matched filters and on various radar system requirements and state-of-the-art in the implementation of matched-filter radar systems. Although the emphasis in these discussions is on radar applications, linear matched filters have significant defense applications in antijam and low-detectability communication systems. Thus, the successful development of a batch-fabricated, relatively inexpensive linear matched filter will have a major impact on both radar and communication systems technology.

III. MATCHED-FILTER THEORY

A. General

A matched filter is the optimum linear filter for signals embedded in white Gaussian noise in the sense that the maximum signal-to-noise ratio is obtained at the output. The theory and concept of the matched filter was developed by D. O. North in 1943.¹⁰⁷

While signal-to-noise ratio considerations led to the development of the matched-filter concept, more recent investigation in the field of detection and estimation theory have shown that a matched filter is an integral part of the optimum receiver.¹⁰⁸

In addition to the theoretical justification mentioned above, the use of matched filters in location and communication systems had additional desirable attributes from a system design standpoint. One of these is an inherent immunity to interference. Since the response to an unmatched signal produces a much reduced output compared to the response to the matched signal of equal energy, without exact information about the structure of the transmitted signal, an enemy intercept receiver has difficulty in detecting the transmission, or an enemy jammer must resort to inefficient barrage noise jamming techniques.

Another advantage for radar system design is that the signal energy can be increased (thus achieving greater detection range) by using a signal of long duration. A matched filter then compresses the received signal into a high-amplitude pulse of short duration which permits a range resolution factor determined by the duration of the compressed pulse rather than by the duration of the transmitted pulse as in ordinary pulse radar.

Alternately the signal bandwidth can be increased without reducing the signal energy. This capability is important in detecting targets in clutter

since subclutter visibility is directly proportional to signal bandwidth, but sufficient signal energy is still required to prevent masking by the white noise background.

B. Matched-Filter Transfer Function

A matched filter can be described in terms of its transfer function,¹⁰⁹ which must be the complex conjugate of the input signal spectrum with a constant time delay. This relationship is written as follows:

$$H(w) = S^*(w) e^{-swt_0} \quad (B-1)$$

where

$H(w)$ = required filter transfer function
 $S(w)$ = input signal spectrum
 $S^*(w)$ = complex conjugate of $S(w)$
 t_0 = signal duration.

A matched filter may also be described in terms of its impulse response,¹⁰⁹ which must be the time inverse of the input signal. The impulse response is given by:

$$h(t) = s(t_0 - t) \quad (B-2)$$

where

$h(t)$ = filter impulse response
 $s(t)$ = input signal
 t_0 = signal duration.

Either one of these equations can be obtained from the other since they are a Fourier transform pair.

C. Matched-Filter Characteristics

The filter parameters of principal interest from a system design standpoint are the signal bandwidth, the signal duration, and the product of the two (or time-bandwidth product). Signal bandwidth is important for a radar application since it determines the range resolution obtained. Signal duration is of interest since it determines the signal energy and therefore the maximum detection range for fixed peak power and a given target size. In addition, the frequency resolution obtained is equal to the reciprocal of the signal duration when the radar is required to measure target velocity. The improvement in signal-to-noise ratio is a function of the time-bandwidth product,

$$(S/N)_{out} = 2BT(S/N)_{in} \quad (B-3)$$

where

$(S/N)_{out}$ = signal-to-noise ratio at filter output
 $(S/N)_{in}$ = signal-to-noise ratio at filter input
B = signal bandwidth
T = signal duration.

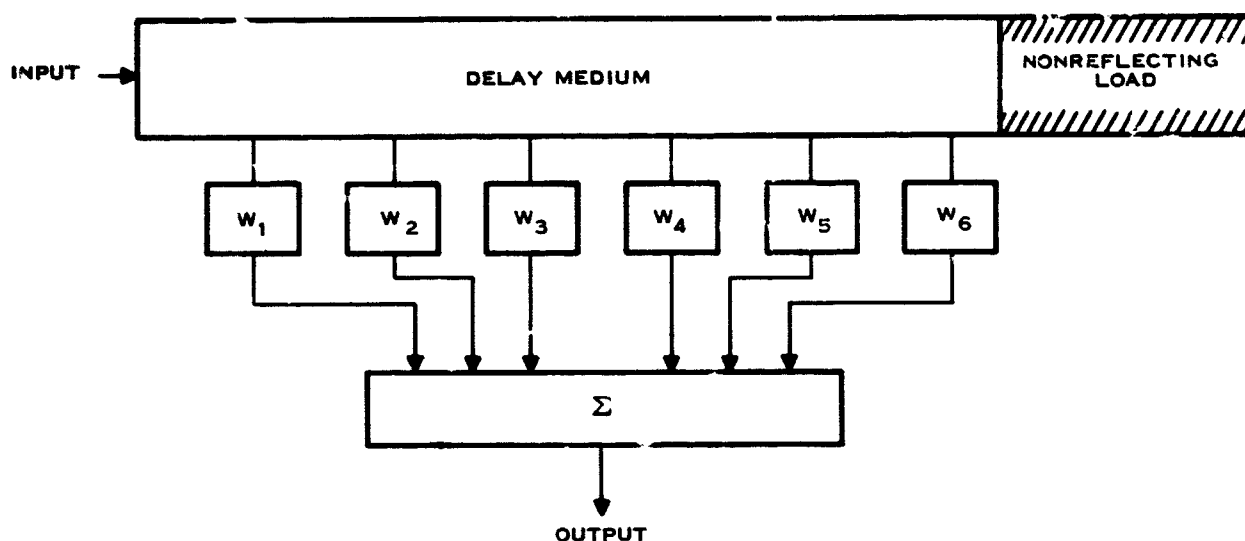
The time-bandwidth product is also a measure of the immunity of the signal-to-noise or interference. Shannon's sampling theorem states that $2BT$ numbers are required to uniquely specify a signal of duration T and bandwidth B . This leads to the concept of a signal as a point in multidimensional space with each of the $2BT$ numbers being a coordinate of this point. The likelihood that a random signal will have a set of coordinates that match the filter becomes smaller as the number of coordinates ($2BT$) is increased.

The magnitude of these parameters will vary with the particular application. For radar application the bandwidth will usually lie between 0.5 and 50 MHz and the time-bandwidth products between 50 and 1000 MHz.

D. Realization of Matched Filters

A variety of techniques has been used in the construction of matched filters, as will be shown in the next section where a review and tabulation of matched-filter requirements and state of the art are presented. In brief, the principal difficulty in their realization is the selection of a suitable delay medium; consequently, matched filters are often classified in terms of the type of delay structure employed. Those that have been used in the past include tapped, nondispersive delay lines, dispersive delay lines, and cascaded all-pass filters. Cascaded all-pass filters provide a method for the synthesis of a dispersive delay line using lumped-constant components. Each of the first two types may be further divided in terms of the delay media used in the lines.

The surface-wave device previously discussed can be classified as a tapped-delay line. It should be a versatile device since any band-limited filter can be replaced by a tapped-delay-line filter¹¹⁰ such as that shown in Figure B-1. The theory of such filters is based on Shannon's sampling theorem which states that any signal limited in both time and frequency can be recovered from its sampled values taken at intervals no greater than one-half the reciprocal of the signal bandwidth. Therefore, the delay line is tapped at intervals of $1/B$, and each tap is weighted by the sample values of the desired filter impulse response in inverse order.¹¹¹ Hence, any matched filter can be constructed by using proper weights on the taps. If the signal has a bandpass spectrum the taps can be spaced further apart but the weights will be complex, that is, amplitude as well as phase weighting.



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Figure B-1. Tapped-Delay-Line Filter

IV. REVIEW OF REQUIREMENTS AND STATE-OF-THE-ART

The demands made by modern radar missions upon system parameters have required the design of increasingly complex transmit waveforms. Waveform design considerations include ambiguity response, resolution, parameter estimation accuracy, and state-of-the-art matched-filter implementation. At present, theoretical waveform design capability has far exceeded technological capability for matched-filter implementation. It is the filling of this "hardware gap" that the surface-wave device exhibits promising potential. In Table B-1, some typical radar applications, along with required waveform and matched-filter characteristics, are presented.

These requirements depend on the application and the type of source used. Antiballistic missile systems need, even with high-power transmitters, wide bandwidths and long pulses to provide detection, tracking accuracy and discrimination on the small cross-sections of incoming missiles. Long range air surveillance radars are not presently power limited using high-power transmitters but transmit low-resolution (1 to 10 μ sec) pulses. These long pulses cannot penetrate severe weather conditions and the addition of pulse compression can greatly enhance performance. The matched filter must be inexpensive, though, or retrofitting cannot be afforded. New solid-state phased array surveillance radars must transmit very long pulses (100 to 300 μ sec) to achieve long-range capability. Here pulse compression is mandatory to maintain good range resolution.

Table B-1. Waveform and Matched-Filter Requirements of Typical Radar Systems

APPLICATIONS		TYPICAL WAVEFORM CHARACTERISTICS			Comments on Matched-Filter Requirements
		Bandwidth	Time Duration	Time Sidelobes	
Antiballistic Missile System	Search Mode	20 kHz	500 μ sec	-40 dB	Insertion loss should be low to avoid the use of a large number of post-filter power amplifiers. Phase linearity must be tightly controlled across the pass-band to prevent waveform distortion. The system is ground based; therefore, weight is not critical.
	Acquisition Mode	100 kHz	50 μ sec		
	Track Mode	20 MHz	4 μ sec		
	Discriminate Mode	20 to 500 MHz	1 μ sec		
Aircraft Surveillance Radar	Long-Range Search and/or Track Conventional W/O PC W/ PC Solid-State	0.1 to 1.0 MHz	1 to 10 μ sec	30 to 35 dB	Weight is an important factor due to the transportable nature of the radar. Phase and amplitude linearity must be tightly controlled. Insertion loss should be < 40 dB.
		1 to 5 MHz	100 to 300 μ sec		
		5 MHz	$\approx$ 1 μ sec		
Airport Surveillance Radar				-40 dB	Phase linearity must be maintained within a few percent to prevent sidelobe addition in the post-filter MTI circuitry. Weight is not a critical factor.
Combat Surveillance Radar		5 MHz	0.5 μ sec	-40 dB	The radar is man-transportable; therefore, size and weight are extremely critical.
Satellite Detection		1 MHz	1000 μ sec	-35 dB	The large delay required implies high insertion loss. This loss should be minimized.
Surveillance in Sea Clutter		500 to 1000 MHz	1 μ sec	-40 dB	Extremely low self-clutter must be maintained along both the time and doppler axes.

Widely varying requirements placed on signal (matched-filter) design are illustrated in Table B-I by the contrast between satellite detection and sea surveillance systems. While the BT products required are approximately the same (1000), the bandwidth ratio between the two is approximately 1000. Techniques for processing a long pulse with relatively low bandwidth are radically different from those used in processing a short pulse with high bandwidth. Another important consideration is the ratio of target velocity (doppler frequency spectrum) to signal bandwidth (resolution). Targets having high velocity often require a parallel channel matched filter (doppler filter bank). Existing techniques for implementing filters of this kind require the use of a number of separately packaged filters, each tuned to a different center frequency. The resulting matched filter is usually large, heavy, and inefficient. Multiple filters fabricated in a single device would greatly enhance the state-of-the-art in this area.

The requirements for combat surveillance (man-transportable radar) places heavy emphasis on equipment being lightweight and small in size. Pulse compression is used to increase subclutter visibility (by providing good range resolution), and to permit the use of low power, high reliable and lightweight solid-state sources with no sacrifice in performance. For this application, the matched filter must be very small, lightweight, inexpensive and very rugged.

In short, the main practical considerations which are of concern in various matched-filter implementations are size, weight, bandwidth, storage capacity, phase and amplitude control in passband, dynamic range, temperature stability, reliability, and cost.

There is at present no general technique which even comes close to meeting most of these practical requirements. In general, existing implementation techniques for pulse compression matched filters include a wide variety of delay media and devices:

1. Media which have intrinsic frequency dispersion characteristics
2. Tapped-linear-delay-line networks with external phase and amplitude weighting to obtain dispersive characteristics
3. Shift-register digital correlators
4. Microwave devices utilizing such characteristics as media geometry and guided waves.

Primary restrictions of some of the above techniques are:

1. Inaccessability of delay media for external connection (tapping)
2. Bandwidth limitation due to such factors as restricted passband phase control, lack of real-time processing ability, state-of-the-art analog circuitry, digital conversion limitations, etc.

3. Delay time limitations due to physical size
4. Weight
5. Temperature instability
6. High signal insertion loss.

Each of the existing techniques is constrained by one or more of the above limitations which results, in many cases, in a failure to achieve the desired radar parameter estimation and detection capability.

With respect to applications, matched-filter requirements fall into one of three broad categories:

1. Short-pulse/high-bandwidth techniques
2. Medium time-bandwidth techniques
3. Long-pulse/low-bandwidth techniques.

These categories, along with present state-of-the-art capability in filter implementation, are presented in Tables B-III through B-IV which are generally self-explanatory.

Study of the tabulated information points out a number of regions (with respect to the BT product) where present-day technology is deficient or inefficient with respect to weight, size, power loss, information storage, etc. The most pressing requirements include:

Pulse lengths on the order of milliseconds and bandwidths > 5 MHz

Bandwidths > 25 MHz for medium dispersion times

Matched filters of extremely small size and weight

Small, wideband linear delay lines which can be externally tapped at selected delay intervals

Contiguous doppler filter banks

Matched filters for medium bandwidths (5 MHz) and relatively long pulses (> 200 μ sec)

Generation of sophisticated phase and frequency codes.

The surface-wave device appears to lend itself readily to meeting the requirements in several of the above areas and perhaps extending the state-of-the-art.

Table B-II. Short Pulse/High Bandwidth Techniques

Technique	State-of-the-Art	Comments and Applications
Meander line techniques at microwave frequencies	A meander line dispersive network has been designed with a bandwidth on the order of 750 MHz and duration of $\approx 0.8 \mu\text{sec}$. Considerable amplitude ripple occurs across the pass-band and post-filter compensation will be required. Phase linearity and insertion loss (40-70 dB) are a strong function of frequency and tight control is difficult. Technique has been demonstrated for 400 MHz bandwidth but $\sim 0.1 \mu\text{sec}$ dispersion time.	This device is lightweight and relatively reproducible. Considerable post-filter amplification and linearity compensation will be required in practical systems. Possible applications include ABM and submarine detection systems.
Synthesis of all-pass filters using stripline circuits at microwave frequencies	Technique has been demonstrated for 400 MHz bandwidth but $\sim 0.1 \mu\text{sec}$ dispersion time.	Similar to meander line techniques, and the comments above apply. An additional problem of reflections, which add in phase when stripline circuits are cascaded to get longer dispersion time, is incurred.
Optical correlators using ultrasonic light modulators	Bandwidth is presently limited to $< 100 \text{ MHz}$ by available ultrasonic light modulators. Transducers should be available in the near future which will make 1000 MHz bandwidths feasible. Insertion loss is $\approx 30 \text{ dB}$. Dynamic range is limited due to device characteristics.	Any FM function, including linear FM, can be generated and compressed using this technique. Sidelobe weighting can be accomplished with optical filters that are much easier to implement than electronic filters. Transmitted waveforms can easily be changed to correct for distortion in other system components. Possible applications include ABM, air surveillance, and surveillance in sea clutter.
Wedge shaped dispersion filter	Technique has been demonstrated for bandwidths up to 40 MHz for duration $< 5 \mu\text{sec}$. Components necessary for bandwidths to 500 MHz have been demonstrated in other applications. Insertion loss is 50 dB.	Possible to synthesize a wide range of FM functions including linear FM. Input signal power is limited to $+40 \text{ dBm}$.

Table B-III. Medium Time Bandwidth Techniques

Technique	State-of-the-Art	Comments and Applications
Metallic strip dispersive delay lines, linear FM.	For long dispersion times (200-500 μ sec) the highest bandwidth obtainable is <1 MHz. ⁸ If pulse duration <25 μ sec are acceptable, bandwidths to 10 MHz can be obtained. Insertion loss is 40 dB. Very linear phase and amplitude characteristics have been obtained.	These devices are reproducible, relatively small and light, and can be used for most radar applications having acceptable BT requirements.
Synthesis of all pass filters using lumped LC circuits.	Technique has been demonstrated with a 30-MHz bandwidth and delays on the order of 1 μ sec.	These units are extremely large and heavy. Both sidelobe weighting and phase equalization are required at the filter output due to phase and amplitude nonlinearities. Application is restricted to large ground based systems.
Frequency and phase-coded techniques using tapped-delay line as matched-filter receiver. Phase and amplitude weights are required on each tap.	The resolution, transmitted-pulse energy, and sidelobe requirements of an advanced airborne system have been demonstrated in pulse compression system employing variations of this technique.	Bandwidth is limited by the coded element pulsewidth. Sidelobes weighting of coded waveforms is possible. Extremely low sidelobes as a function of BT product and size can be obtained using polyphase codes.

Table B-IV. Long-Pulse/Low-Bandwidth Techniques

Technique	State-of-the-Art	Comments and Applications
Parallel Channel dispersive line network. Dispersion time per channel is T/N . Linear delay lines and amplitude weights are required in each channel.	A feasibility model has been implemented with a dispersion time of 1000 μ sec and a bandwidth of 1 MHz. ¹⁰ Results obtained were within brass-board tolerances. Insertion loss is on the order of 40 dB per channel.	The total dispersion matrix is extremely bulky. Many practical problems exist. These include phase, amplitude, and delay matching between channels. Possible applications include large ground based systems such as satellite detection and ABM systems.
Optical techniques using film as the delay media.	At low bandwidth ($B < 0.5$ MHz) dispersion times on the order of milliseconds can be obtained.	This technique is eliminated from most practical radar applications due to nonreal-time operation.
Digital correlation techniques with possible use of the direct or modified Cooley-Tukey Fast Fourier Transform.	Laboratory models of a direct digital correlator have been implemented for BT on the order of 250. ¹¹ A performance evaluation has not been completed at this time. Cooley-Tukey processes for large BT products have not been implemented at this time but theoretically BT products on the order of 10 ⁵ could be processed. Insertion loss is < 3 dB. Dynamic range constraints are imposed by the number of quantization levels.	(BT) ² operations are required in direct digital correlation. Cooley-Tukey processing would reduce the number of operations required to (BT/2) log BT. A modified form of the Cooley-Tukey algorithm could reduce the number of operations by another 50 percent for certain forms of signal modulation. The advent of Large Scale Integration could make digital filtering particularly attractive where lightweight, compact units are required. Bandwidth limitations are imposed by sampling and A/D conversion constraints (present state-of-the-art is approximately 25 MHz, 8-bit).

APPENDIX C

EXPERIMENTAL PHASED-ARRAY ANALYSIS

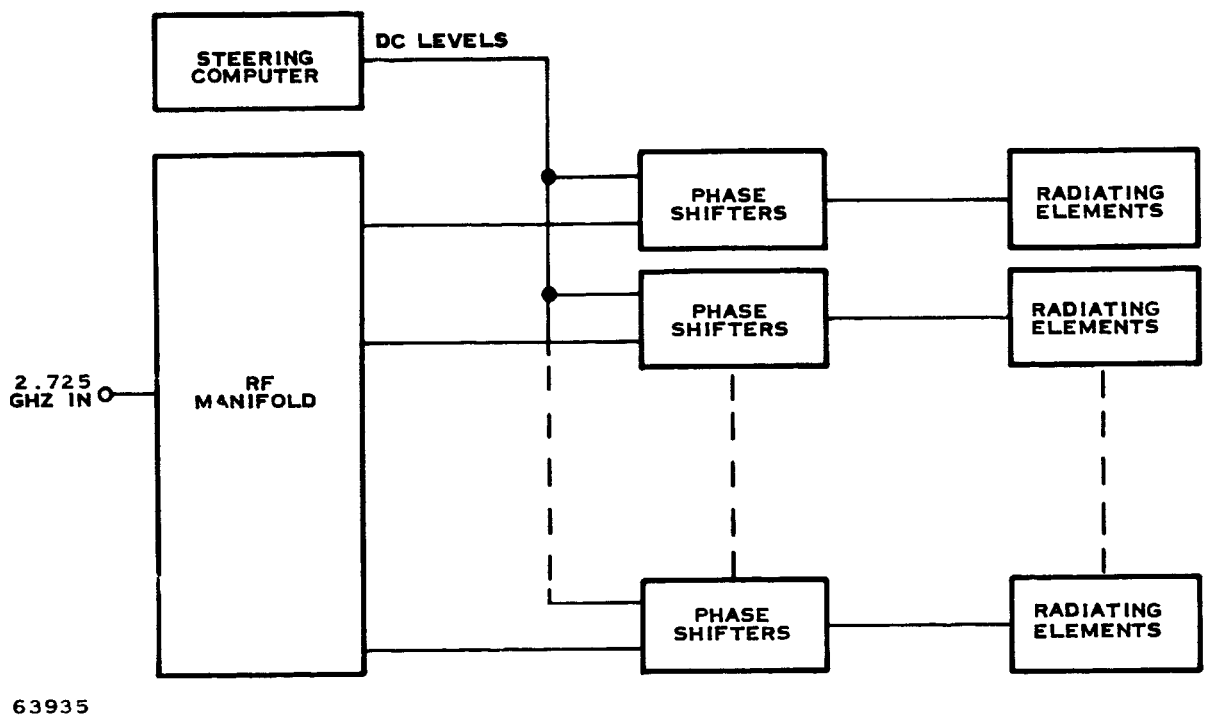
An experimental program has been undertaken at Texas Instruments to study the array radiation characteristics of various electronic steerable S-band phased arrays. The initial phase of the program was concerned with a 16-element linear array with emphasis being placed on beam steering accuracy. The antenna was an H-plane array of dipoles with a radiating element spacing of 0.534 wavelengths. A block diagram of the array is shown in Figure C-1. S-band diode phase shifters were used to provide the required phasings at the radiating elements with the phase-shifter control inputs being provided by a steering computer. A description of the array follows along with some experimental results.

The 4-bit, 360 degree, PIN diode phase shifter is a key component in the array. The phase-shift network makes use of thin-film circuitry and is of the diode-loaded transmission-line type. The circuitry consists of an rf wafer containing the PIN switching diodes and rf circuit elements and a control wafer with digital integrated circuits and other components. The network is of the constant-time-delay variety; that is, time delay through the network is independent of frequency.

To simplify the steering computer, a portion of the logic has been included in the phase-shift network package. This includes memory elements, network driver circuitry and a series on/off switch. Figure C-2 is a block diagram of the phase shifter. The diode switch allows the rf line of the network to be opened, making possible the automatic removal of an element when thinning analysis is required. Performance characteristics of the phase shifter are outlined in Table C-I.

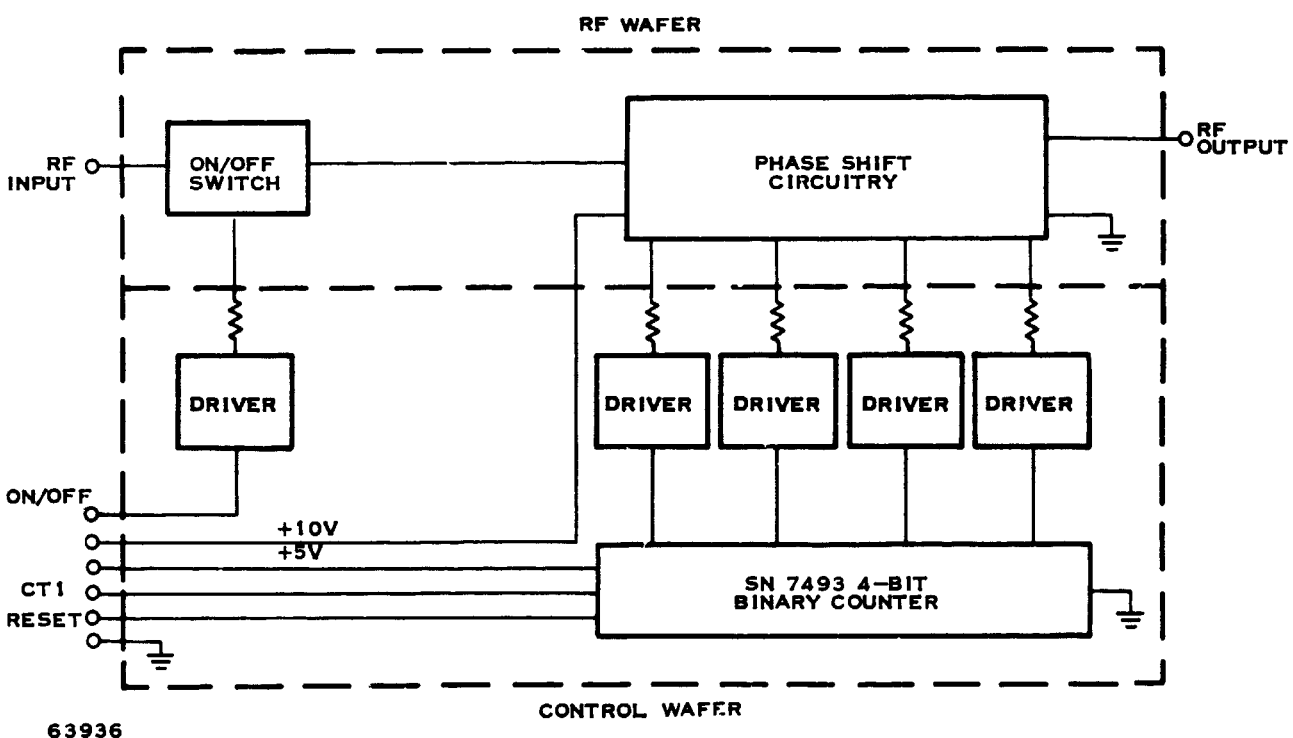
TABLE C-I. PERFORMANCE SPECIFICATIONS

PARAMETER	SPECIFICATION
Frequency Range	2.725 GHz; 100 MHz bandwidth
Number of Bits	Four
Operating Power	5mW maximum CW
Phase Deviation from nominal	±15 degrees rms
VSWR	1.5:1 maximum
Insertion loss	3.0 dB maximum
Unbalance in Insertion Loss	1.0 dB difference between minimum and maximum loss at one frequency between phase settings.



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Figure C-1. Block Diagram of Array



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Figure C-2. Phase-Shift Module Schematic Diagram

The steering computer provides dc inputs to the phase shifters to control the phase settings. The steering computer operates upon a digital gradient value to calculate the phase value needed on each phase shifter. The digital gradient value represents the desired difference in phase value from element to element along the linear array, and is represented with an accuracy which exceeds that which can be obtained with the 4 bit phase shifters. The gradient is successively added to a register to form phase values, the most significant of which are then transmitted to each phase shifter.

The steering computer has two modes of operation, manual and automatic scanning. The manual mode allows the beam to be pointed in one of 128 possible positions. The phase shifters are properly phased for this beam position and remain that way until changed by the selection of another position. This gives the opportunity to take antenna patterns of the steered beam in the conventional manner. In the automatic scanning mode the computer steps through each beam position in order, with the scan rate being a variable. This provides a dynamic pattern taking capability.

The radiating element in the array is a slotted balun half-wavelength dipole. This type dipole was chosen because of the ease of construction and the ease with which the balun-impedance transformer can be modified to provide a match to different array configurations.

The manifold feed contains four levels of two-way stripline power division resulting in 16 output ports. The two-way dividers are of the resistor loaded-reactive type. The resistor placed between the two output lines provides isolation of 20 to 30 dB between output ports without materially contributing to losses. This compact feed is easily constructed and provides equal amplitude and phase at the output along with low insertion losses and matched input and output ports. Figure C-3 is a photograph of a 4-way divider, a 16-way divider, a balun-fed dipole and the phase-shift package.

The 16 element array has been tested in both the manual and automatic scanning modes, and Figures C-4 and C-5 are array patterns recorded in the manual mode of operation. With an element of 0.534 wavelength, a grating lobe should begin to appear at a scan angle of about 50 degrees. As can be seen in Figure C-5 the grating lobe does appear at a scan angle of about 70 degrees. The grating lobe has reached equal amplitude with the main beam. At greater scan angles, the grating lobe becomes the principal beam. As can be seen, the accuracy of pointing the main beam was quite good.

Figure C-6 shows element patterns recorded in the array environment and shows the effects of variation of mutual coupling on the edge elements. Element 1 was an element on the edge with elements 8 and 9 being center elements. Only three elements from the array edge, the element pattern is not appreciably affected by edge effects. The pattern of element 3 is almost identical with that of element 9. The apparent ripple at the center of the element pattern is presently unexplained.

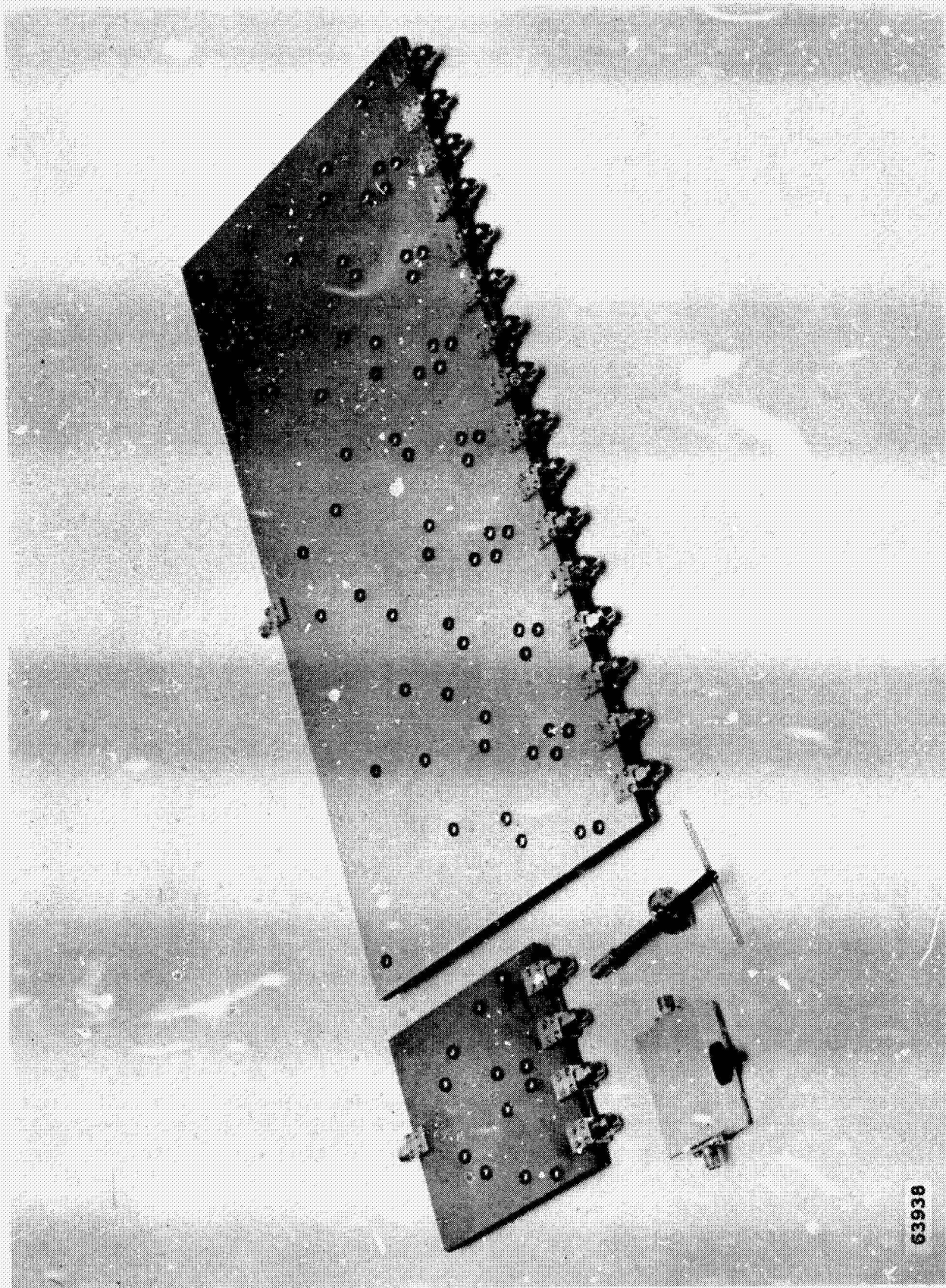
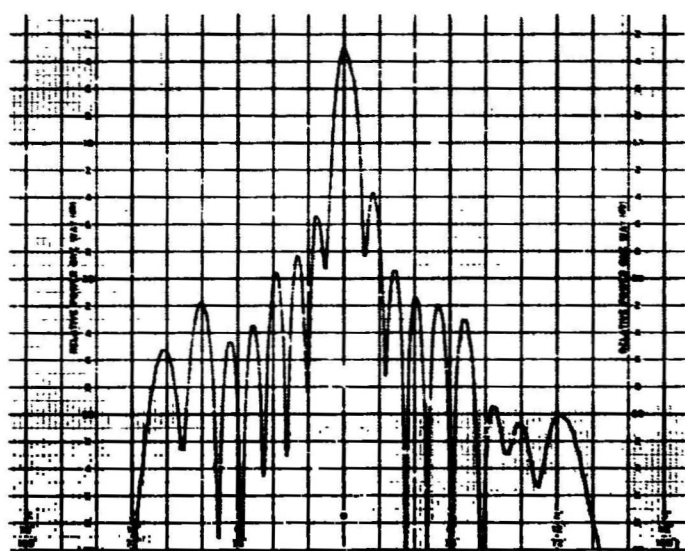
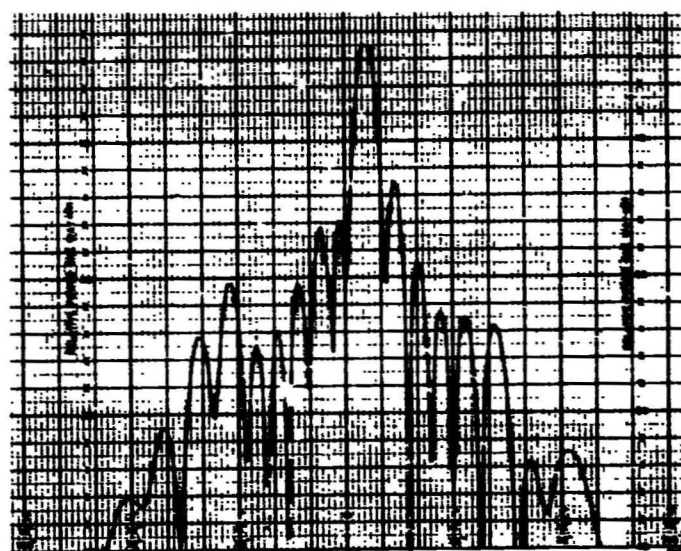


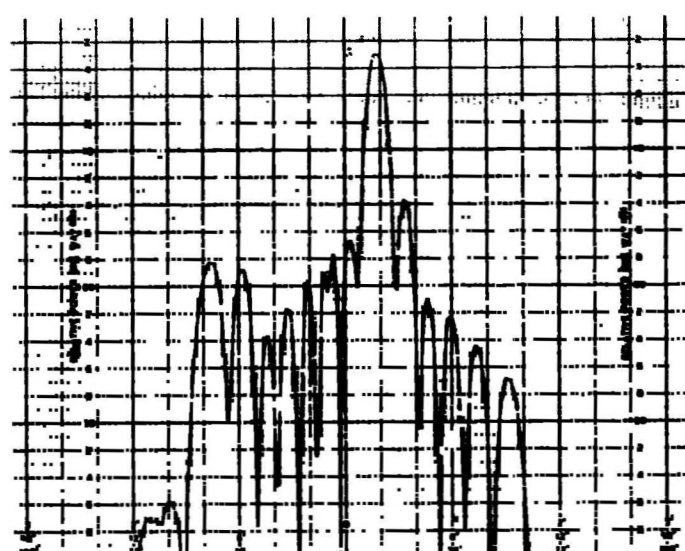
Figure C-3. Phase-Shift Package, Balun-Fed Dipole and Power Dividers at S-Band



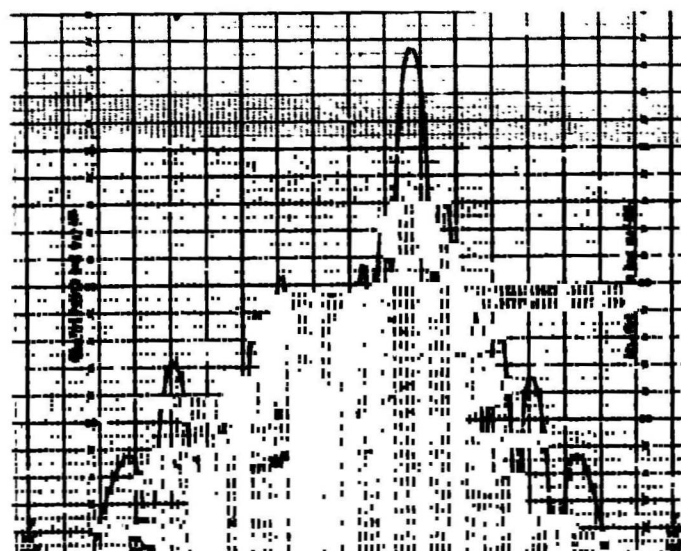
(a) $\theta = 0$



(b) $\theta = 6.7$

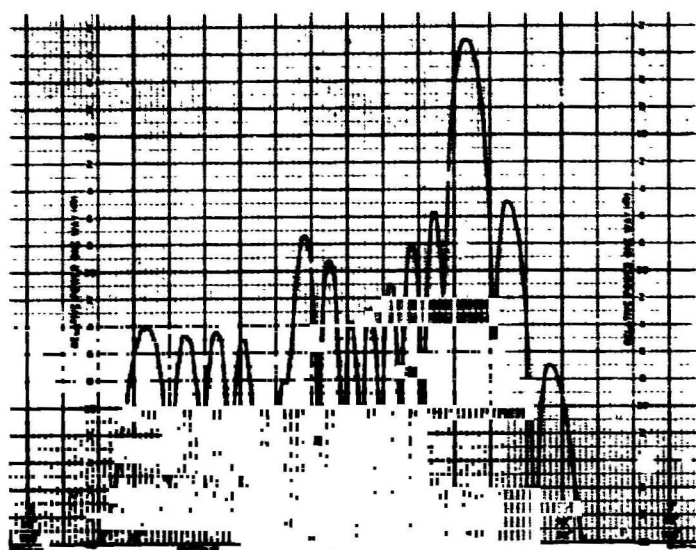


(c) $\theta = 10.1$

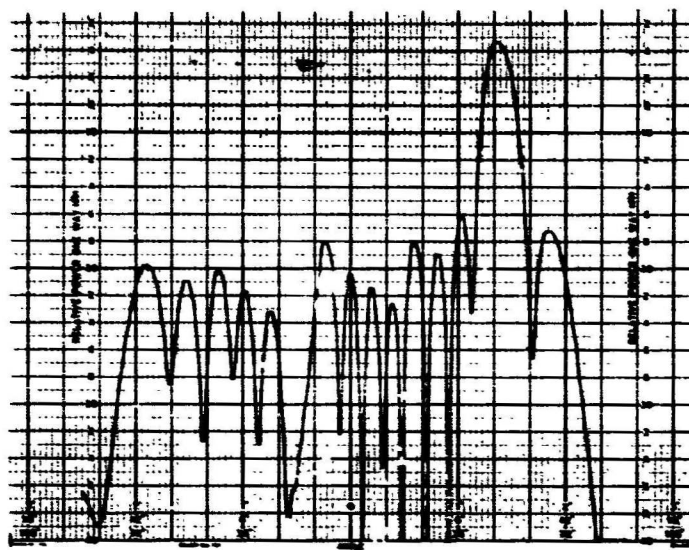


(d) $\theta = 20.6$

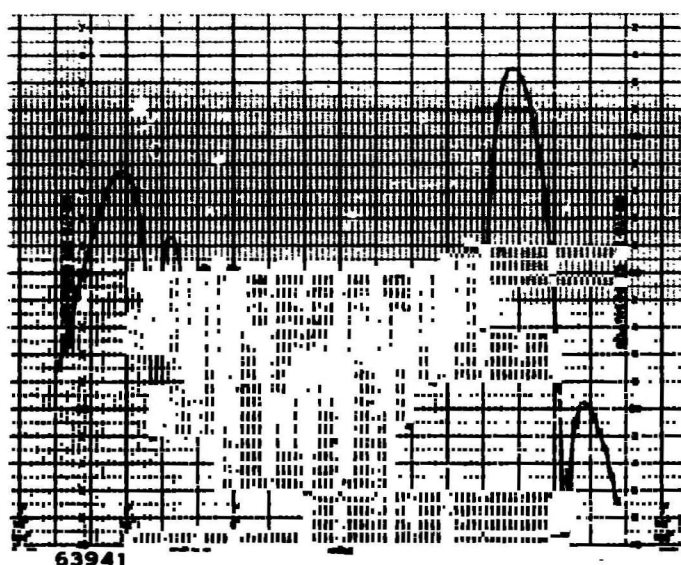
Figure C-4. Measured Patterns of 16-element linear array (a) 0 degree scan (b) 6.7 degree scan (c) 10.1 degree scan (d) 20.6 degree scan



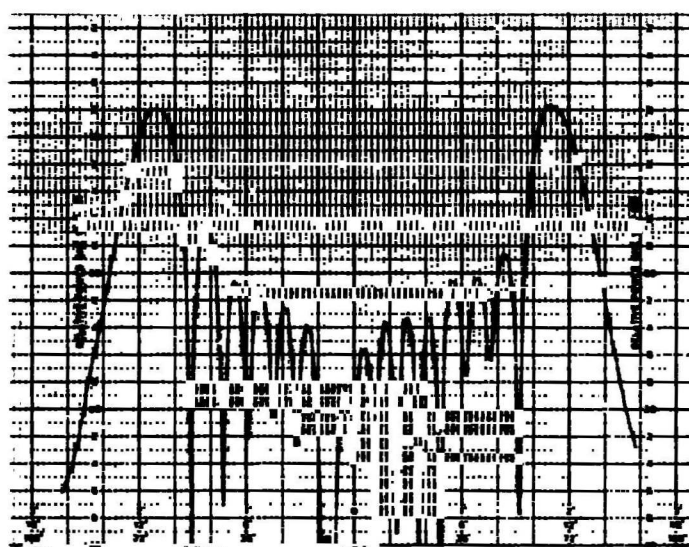
(A) $\theta = 40.1$



(B) $\theta = 49.5$

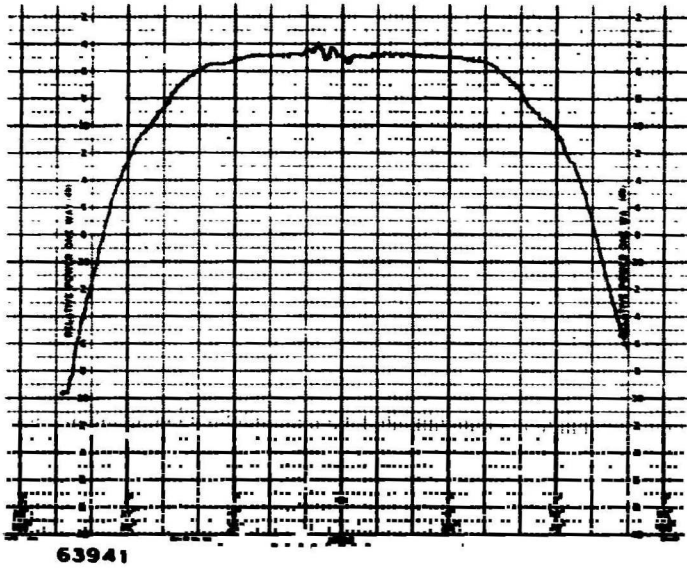


(C) $\theta = 59.7$

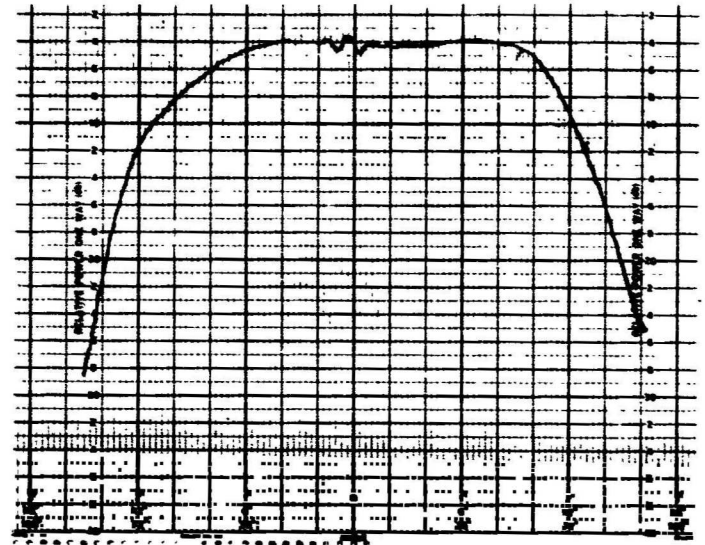


(D) $\theta = 69.5$

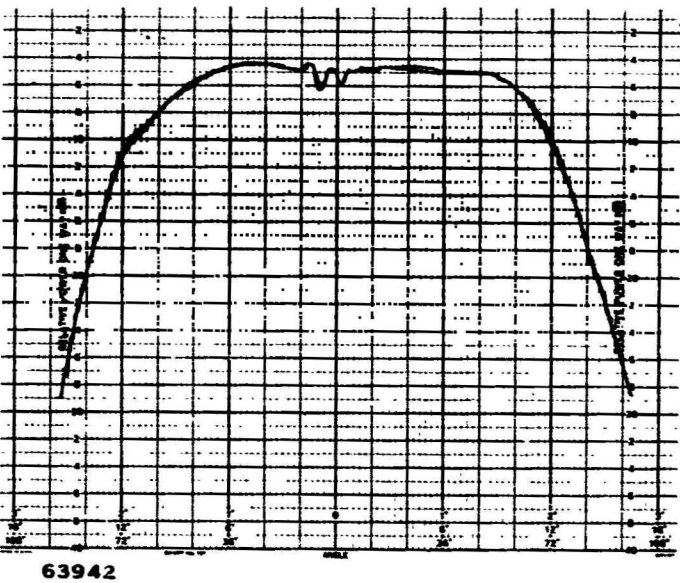
Figure C-5. Measured Patterns of 16-element linear array (a) 40.1 degree scan (b) 49.5 degree scan (c) 59.7 degree scan (d) 69.5 degree scan



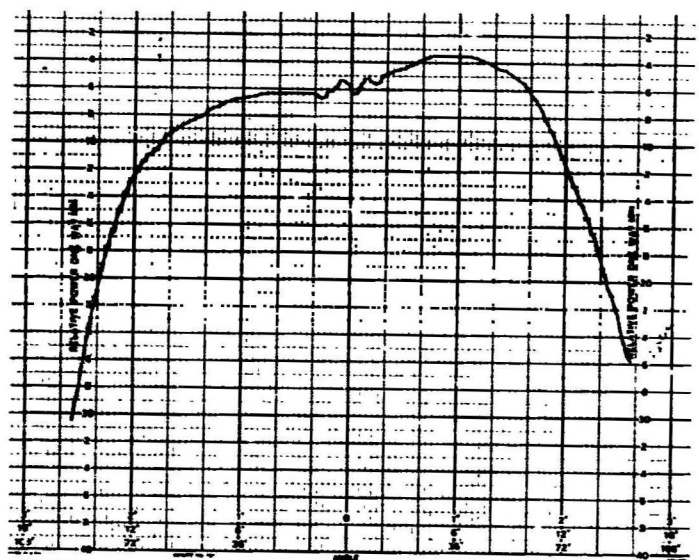
ELEMENT NO. 9



ELEMENT NO. 1



ELEMENT NO. 3



ELEMENT NO. 2

Figure C-6. Element Patterns-Elements 9, 3, 1, 2

Plans have been made to analyze larger arrays, planar arrays, and conformal arrays, and to explain analytically the element pattern ripple. A cylindrical ground plane has been fabricated which will permit the analysis of a 64-element ring array with an element spacing slightly greater than 0.5 wavelengths. This ring array subtends an arc of about 150 degrees on the ground plane.

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