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**TRANSMIT/RECEIVE MODULE TECHNOLOGY FOR FUTURE
SPACEBORNE AND AIRBORNE RADAR SYSTEMS**

by

Jeffrey William Smuk, B.E.Eng.

**A thesis submitted to the Faculty of Graduate Studies and Research in partial
fulfillment of the requirements for the degree of**

Master of Electrical Engineering

Department of Electronics

Carleton University

Ottawa, Ontario

April 2, 1987

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ISBN 0-315-39429-3

The undersigned recommend to the Faculty of Graduate Studies and Research the
acceptance of the thesis:

"TRANSMIT/RECEIVE MODULE TECHNOLOGY FOR FUTURE SPACEBORNE AND
AIRBORNE RADAR SYSTEMS"

submitted by Jeffrey William Smuk in partial fulfillment of the requirements
for the degree of Master of Engineering.

Thesis Supervisor

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ABSTRACT

This thesis explores transmit/receive module technology that may be applied in future spaceborne and airborne radar systems. The feasibility of a space-based radar is discussed and the transmit/receive module, a key component of such a system, is studied. Microwave circuits which may be required for a transmit/receive module are developed using both Monolithic Microwave Integrated Circuit (MMIC) and Microwave Integrated Circuit (MIC) technology. MMIC technology is used to form C-band actively matched amplifiers while MIC technology is used to investigate one of the most promising new microwave devices, the High Electron Mobility Transistor (HEMT).

ACKNOWLEDGEMENTS

I wish to thank my supervisor Professor J.S. Wight for his encouragement and support.

I also would like to express my sincere appreciation to the management of Canadian Astronautics Limited (CAL) and the Communications Research Center (CRC) where this research work was carried out. Thanks are especially due to Mr Ian Christison of CAL and Dr Malcolm Stubbs of CRC for their guidance and support. I would also like to thank Mrs Elizabeth Bala, Mr Steven Scribner and Mr Julian Noad, all of CRC, for their technical assistance.

I would also like to thank the Natural Sciences and Engineering Research Council and CAL for their financial assistance.

Finally, I would like to thank my parents for their inspiration, constant encouragement and tolerance.

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CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION

RADAR (RADio Detection And Ranging) uses the reflection of a transmitted pulse of electromagnetic energy to determine the presence of a target. By analysing this reflection, the target's velocity, position and structure may be determined. The ability of a radar to determine this information has led to its pervasiveness. Today, radars mounted on ships, aircraft, and land platforms are used for target surveillance, tracking and identification as well as land mapping. Recent advances in the aerospace field coupled with advances in microwave circuit technology may make orbiting satellites the newest surveillance radar platform.

With their high altitudes, these space-based radars could provide surveillance over large regions of the earth. Also, low flying aircraft that may be masked by the earth's features from a ground-based radar may be detected. These characteristics have made space-based radars attractive and research is being performed to develop such a system. This thesis, which includes a hypothetical space-based radar system analysis as well as experimental work aimed at developing key technology needed for such a project, is a part of the research effort.^[1]

Two main radar topologies, monostatic and bistatic, are possible for use in a space-based radar. In a monostatic radar, the same antenna aperture is used for both transmit and receive, whereas, in a bistatic radar, separate antennas are used for each. Both topologies require the application of the relevant form of the Radar Equation to analyse the performance of the radar. The analysis requires an iterative solution to determine the many variables available in the radar design. The choice of these variables is limited by the radar mechanization.^[2]

A major system driver in the space-based radar analysis is the implementation of the antenna aperture. An active phased array antenna which is amongst the most recent advances in radar hardware is assumed in the radar analyses. Compared with passive reflector or array antennas, active phased arrays allow a radar to direct more of the radar's finite energy at targets rather than into empty space. To construct the phased array, many thousands of individual transceivers are connected to the antenna's

radiating elements. Each transceiver may control the phase and amplitude of the electromagnetic energy passing through itself onto the radiators. Such a transceiver is referred to as a transmit/receive (T/R) module.^[3]

The first part of this thesis is composed of two studies dealing with space-based radar. The first study is a radar system analysis based on a spaceborne surveillance radar used to cover Northern Canada. A picture of such a system is shown in Figure 1.1. Here, an orbiting radar is shown illuminating a portion of its available surveillance zone. The second study concentrates on a critical part of said radar, the transmit/receive module.

Figure 1.1 SPACE-BASED RADAR STRUCTURE

The second part of the thesis is experimental work based on developing some of the technology used in T/R modules. This includes two sets of experiments. The first set uses Monolithic Microwave Integrated Circuit (MMIC) technology while the second set makes use of Microwave Integrated Circuit (MIC) technology to construct test circuits.

MMIC technology used to construct future microwave circuits appears promising. In a MMIC, all the circuit elements are placed on a common Gallium Arsenide (GaAs) substrate measuring only a few millimeters on a side. A scanning electron micrograph of a section of one of the MMICs constructed in this thesis is shown in Figure 1.2. Note that the circuit elements are labeled where R is a resistor, C is a capacitor, L is an inductor and M is the edge of a metal semiconductor field-effect transistor (MESFET). A circuit module or chip fabricated using MMIC technology should have a low unit cost in large quantities and possess high performance at any frequency in the microwave spectrum. It is widely believed that this technology should make the development of large active array antennas with many thousands of elements feasible.^[4]

Since GaAs chips are small relative to a wavelength, the design of MMICs makes use of lumped element theory in order to match the devices for maximum performance. Also the biasing networks, which require considerable space in MICs, may be positioned on the chip. Two two-stage amplifiers using MMIC technology are developed in this thesis. The circuits developed, operating at C-band, are novel in their use of active matching and biasing circuitry. Active matching allows the substitution of an active device in the place of larger passive matching circuitry. In the lower microwave frequency bands, such an arrangement reduces the cost and complexity of an MMIC. The major difference between the two designs constructed is found in the bias circuits. In one design, a transistor may be biased on and off forming a switch while in the other design, all biasing voltages are derived from a single applied external voltage simplifying its connection into a system such as a T/R module.

Figure 1.2 MMIC Scanning Electron Micrograph

Another promising device for use in microwave circuits is the recently developed High Electron Mobility Transistor (HEMT). These devices, based on the precision vertical doping of a tertiary semiconductor to form a heterojunction, rely on recent advances in semiconductor processing technology. The unique heterojunction structure of a HEMT results in lower impurity scattering which gives them lower noise and higher gain than that of a similar MESFET. At low temperatures, where phonon noise becomes negligible, the HEMT's performance advantage should increase. Such devices may be applied in the low noise amplifiers of a T/R module in order to achieve the maximum performance.[5]

A comparison of HEMT and MESFET devices using discrete devices in hybrid (MIC) circuits is the basis of the second set of experiments. Here, discrete devices are used to construct Ku-band amplifiers since HEMTs are not yet available as part of a monolithic process. Two experiments designed to characterize the noise parameters of the two devices and to observe the effect of temperature on them are performed in this thesis.

1.2 THESIS OBJECTIVES AND ORGANIZATION

This thesis is divided into two parts. The first part is comprised of two studies relating to space-based radar while the second part is comprised of experimental work related to the development of space-based radar technology. The studies were performed in cooperation with Canadian Astronautics Limited whereas the experimental work was performed at the Communications Research Center.

A tree diagram of the organization of the thesis is shown in Figure 1.3.

The thesis begins with an analysis to determine the power requirements of a space-based radar in Chapter 2. Starting with no preconceptions, both a monostatic and a bistatic architecture are analysed. In both cases, the amount of transmitter power required to detect a target with a small radar cross section flying over the northern region of Canada is determined using the appropriate form of the Radar Equation. The process is iterative with the final solutions detailed in the chapter.

Chapter 2 begins with the development of the basic geometrical relationships that are unique to a radar placed on a platform in low earth orbit. Considering the velocity and position effects resulting from the dynamics of a low earth orbit, the most suitable altitude and waveform are chosen for a monostatic radar. Next, the surveillance architecture along with a suitable antenna aperture is chosen. These constraints allow the transmitter power to be determined for a given probability of target detection and false alarm rate assuming the radar is noise limited. This is followed by a bistatic radar analysis using a process similar to that for the monostatic radar. Here, the receiver is located in geostationary orbit while many ground-based transmitters are positioned along a constant latitude arc in Canada's North. The chapter ends with a recommendation of the most suitable architecture.

In the second chapter, the antenna for both the monostatic and bistatic radar architecture is assumed to be an active array which provides the optimum system performance. Chapter 3 shows the many advantages of an active array and discusses the T/R module which is the key component of an active array. The objective of the chapter is to determine the best internal architecture for a T/R module and to outline the most appropriate implementation of the microwave circuitry which forms the module's sub-assemblies. Note that for the bistatic radar, the modules are split into transmit-only modules at the transmitter and receive-only modules at the receiver. Being subsets of the T/R module, these modules will not be discussed.

Figure 1.3 THESIS STRUCTURE

The chapter commences with the theory of array antennas. The advantages of array antennas compared with space fed passive reflector antennas are described. The theory of active arrays is then briefly discussed with the many advantages of such an implementation stressed. Finally, an architecture for a T/R module using microwave technology available as of 1986 is proposed. The module is broken into six sub-assemblies including a phase shifter, low power switch, solid state power amplifier, high power T/R switch, integral antenna, and low noise receiver. The module's novel feature is its division into both MMIC and MIC technology in order to obtain the highest overall performance.

Although the original aim of this thesis was to construct the T/R module proposed in Chapter 3, funding was not available. Instead, experimental work concentrating on some of the key microwave circuits required for a T/R module was undertaken. The following two chapters detail the development of circuits using both MMIC and MIC technology, respectively.

MMIC development work forms the foundation of Chapter 4 in which the design, construction and testing of two 6 GHz amplifier/switch circuits is discussed. The chapter begins with a description of the semiconductor processing steps required to construct a MMIC. This is followed by the theory required to construct an amplifier at microwave frequencies. The design steps required to access a foundry and produce a useful module are then discussed. Next, the computer aided design of two actively matched, two-stage amplifiers is detailed. The test results, including gain, return loss, power transfer characteristics, isolation and noise figure are then listed and compared with those predicted in the original design. The chapter ends in an analysis of the maturity of the technology and its suitability to T/R modules.

Two MIC technology based experiments are described in Chapter 5. The experiments are based on a comparison of HEMT and MESFET devices. The objective of the experiments was to confirm the superiority of the HEMT while carrying out useful device characterization. The first experiment determines the noise parameters of both devices at 12 GHz using a series of different input reflection coefficients. These results allow the devices to be applied in low noise amplifier circuits of many systems including T/R modules. The more research oriented second experiment compares the devices at room temperature and liquid nitrogen temperatures. The noise figure and gain of the devices are determined over a range of frequencies from 8 to 14 GHz. Here, the dominance of the HEMT at cold temperatures was to be observed.

Chapter 5 begins with the theory of distributed element matching, HEMT device theory, and transistor noise theory. Next, the design of the experiments to determine the noise parameters using a selection of different input reflection coefficients and to temperature test the two devices is detailed. The experimental results including the gain and noise figure measurements at the two temperatures as well as the empirically determined noise parameters are then summarized. This is followed by a discussion of the suitability of the devices in the low noise amplifier of a T/R module.

Final conclusions and recommendations of future work are discussed in Chapter 6.

CHAPTER 2

SPACE-BASED RADAR - A RADAR ANALYSIS

2.1 INTRODUCTION

Today, radar is the primary wide-area surveillance sensor for aircraft detection. Since electromagnetic energy in the microwave band is usually used, radars are a line-of-sight system. Any target that avoids being illuminated will remain undetected. Hence, low flying aircraft may mask themselves with the earth's terrain in order to avoid detection by ground-based radar. To supplement this system, many expensive airborne look-down radars may be used. Another possibility is a radar mounted on a space-based platform. Such a radar is referred to as a space-based radar (SBR) and it will be analysed in this chapter.

Advances in spacecraft systems may allow space-based radar to supplement future surveillance systems. From space, the radar could cover a large region and aircraft flying nap-of-the-earth would be detectable over a large portion of the surveillance region. Also, non-geostationary orbits would allow the radar to pass over most of the earth's surface. Such a surveillance system could find uses in air traffic control and territorial defence. It should be noted that the capability to detect these types of moving targets via spaceborne radar has yet to be demonstrated by any nation.^A Hence, a space-based radar would be a revolutionary step in present surveillance technology.

This chapter is comprised of a study of a hypothetical space-based radar in which the amount of power required to detect a small radar cross section target flying in Canada's North is calculated. Both monostatic and bistatic radar architectures are considered. Initially, the velocity and position effects resulting from a radar in low earth orbit are determined. This allows the most suitable altitude and waveform for a monostatic radar to be chosen and considering this information, a surveillance architecture is proposed and the required amount of transmit power is determined. A similar method is used to determine the power requirements of a bistatic radar. The chapter concludes with a comparison of the two radars and a recommendation.

^A Although the Soviet Union has an orbiting surveillance radar, it is limited to detecting aircraft carrier size targets in the relatively low clutter background of the open ocean.^[6]

2.2 LOW EARTH ORBIT RADAR RELATIONSHIPS

The first step in the radar analysis is to determine the effect of an orbiting platform on the radar parameters. Using basic geometry, the following section derives many relationships that will be required for future radar calculations. These relationships include the influence of altitude and grazing angle on the doppler bandwidth and slant range that will dictate the mechanization for a radar placed in a circular earth orbit.

A scale drawing of a platform at an altitude of 1,100 km with a beam directed at a 3° grazing angle with the earth is shown in Figure 2.1. Also shown are the symbols, described below, which will be used in the following sections to calculate the velocity and position effects of a radar in low earth orbit.

Alt	altitude
α	squint angle
γ	depression angle
R	slant range
S	ground coverage distance
Θ_g	grazing angle
Θ_{BW}	antenna beamwidth (Θ_{BW0} at $\alpha = 0$)
V_r	velocity of radar platform
V_{rg}	ground track velocity of radar platform
V_t	target velocity
X	earth's subscribed angle

2.2.1 VELOCITY EFFECTS on an ORBITTING RADAR

In the following analysis, the velocity and the resulting doppler bandwidth of an orbiting radar will be calculated over a range of altitudes. Later in Section 2.3, the doppler characteristics will be required in order to determine the best radar waveform for a spaceborne radar in low earth orbit. A qualitative discussion of the effect of antenna orientation on the resulting spectra will also be discussed. This will be used to set grazing angle limits allowing the radar's coverage to be determined. The radar coverage will be required in order to determine the optimum altitude and surveillance structure for a SBR.

Figure 2.1 PHYSICAL MODEL for an 1,100 km ALTITUDE ORBIT

The velocity of a body in an earth orbit is given by equation 2.1. Here, R_e is the radius of the earth (6378 km), G_c is the gravitational constant ($6.7E-11 \text{ Nm}^2/\text{kg}^2$) and M_e is the mass of the earth ($6E24 \text{ kg}$). The equation was applied to determine the velocity of an orbiting platform over a 500 to 2,500 km range of altitudes. This region is of interest due to the finite lifting capability of available launch vehicles and the minimum acceptable orbit lifetime that a large satellite requires.

$$\begin{aligned} r &= R_e + \text{Altitude} \\ v &= \sqrt{G_c \cdot M_e / r} \end{aligned} \quad (2.1)$$

The velocity of the platform over the altitude range is listed in Table 2.1 and is plotted in Figure 2.2. Over this range, the platform velocity is approximately 7 km/s (14,000 knots). The velocity is relatively insensitive to altitude with only a 12% reduction in velocity obtained by increasing the altitude from 500 km to 2,500 km. Hence, a spaceborne platform will travel at least an order of magnitude faster than most aircraft. This is quite different from airborne radar platforms which seldom travel faster than the target .

Before proceeding, the Doppler effect will be discussed.^[7] The Doppler effect is an apparent shift in frequency resulting from either the source of oscillation or the observer being in motion. For a radar separated from a target by a changing distance R , the total phase ϕ of the electromagnetic wave is $4\pi R/\lambda$ radians. Since $w_\phi = \partial\phi/\partial t = (4\pi/\lambda) \cdot \partial R/\partial t = 4\pi \dot{V}_r / \lambda$, the Doppler frequency shift f_d will be described by equation 2.2 where V_r is the relative velocity between the target and radar.

$$f_d = 2 V_r / \lambda \quad (2.2)$$

The resultant spectrum of a continuous wave radar with its antenna boresight aligned along the radar velocity track and a closing target is shown in Figure 2.3. The main beam clutter is shown centered about F_{dr} , the doppler offset created by the radar platform's velocity. Its spread is due to the main beam illuminating a region comprised of different relative velocities due to the change in both depression angle and azimuth angle within the main beam's footprint. The target, assumed to be a point scatterer creating a single spectral line, is shown at F_{dt} , the doppler created by the relative velocity between the target and radar platform.

Table 2.1 SPACECRAFT VELOCITY vs ALTITUDE

ALTITUDE (km)	VELOCITY (m/s)
500	7593.55
550	7566.1
600	7538.95
650	7512.08
700	7485.5
750	7459.2
800	7433.17
850	7407.42
900	7381.93
950	7356.7
1000	7331.73
1050	7307.02
1100	7282.55
1150	7258.32
1200	7234.34
1250	7210.59
1300	7187.07
1350	7163.78
1400	7140.72
1450	7117.88
1500	7095.26
1550	7072.85
1600	7050.65
1650	7028.66
1700	7006.87
1750	6985.29
1800	6963.9
1850	6942.71
1900	6921.71
1950	6900.9
2000	6880.28
2050	6859.84
2100	6839.58
2150	6819.5
2200	6799.59
2250	6779.86
2300	6760.3
2350	6740.91
2400	6721.69
2450	6702.62
2500	6683.72

Figure 2.2 VELOCITY (m/s) vs ALTITUDE (km)

The altitude line clutter shown in Figure 2.3 is created by the antenna side lobes illuminating the ground directly beneath the spacecraft resulting in specular reflections. This clutter is of large magnitude in airborne applications creating blind ranges. Here, targets at the same range as the altitude of the radar are lost. In a spaceborne application, this effect should decrease due to the increased altitude/slant range ratio and the convex curvature of the earth. The side lobe clutter is due to the antenna side lobe illumination of the earth at all velocities in the range $\pm V_r$ and a uniform side lobe distribution is shown in the figure for simplicity.

Figure 2.3 BASEBAND RADAR SPECTRUM

For a SBR, the maximum doppler shift occurs when the antenna boresight is directed along the spacecraft's velocity track and the target is closing at its maximum speed. The maximum doppler shift between the satellite and a 1,200 knot target flying tangential to the earth's surface was calculated using the relationships of equation 2.3. Here, V_r is the velocity of the satellite, V_t is the target velocity, Θ_g is the grazing angle and α is the angle of the target off antenna boresight which is assumed perpendicular to the earth.

$$f_d = 2 \cdot V_{\text{relative}} / \text{wavelength}$$

$$V_{\text{relative}} = V_r \cdot \sin(\alpha) + V_t \cdot \cos(\Theta_g) \quad (2.3)$$

Using Equation 2.3 for a wavelength of 0.2m, assuming an L-band radar with a 1.5 GHz center frequency, the doppler over a wide range of altitudes and grazing angles of 3, 10, 20 and 50 degrees was calculated. The results are listed in Table 2.2 and plotted in Figure 2.4. It is shown that for small grazing angles, the doppler shift is fairly constant since α is relatively constant. Overall, there is a more noticeable effect of altitude on doppler than on spacecraft velocity since a decrease in squint angle, α , with increasing altitude, lowers the spacecraft's relative velocity component. Here, a 30% doppler reduction occurs when the altitude is increased from 500 to 2,500 km at 3° grazing. Note that at 1,100 km and low grazing angles, the doppler of the platform alone is a large 62 kHz. Hence, side lobe clutter will be spread over ± 62 kHz at baseband. These high doppler spreads are unique to a low earth orbit radar platform.

Most look-down radars use doppler processing in which the different doppler shifts of moving targets and stationary ground targets may be separated in the frequency domain. Usually, the difficulty of separating a target from the clutter is considerable. For a SBR operating in the look-down mode, the problem of clutter removal is complex and only a qualitative analysis will be discussed.

The side lobes and main lobe of an orbiting antenna are shown in Figure 2.5. Note that the clutter spectrum will follow the antenna's gain pattern since side lobes offset in either depression angle or azimuth angle will see a different relative velocity ($\partial L/\partial t$).

Figure 2.5 ANTENNA SIDELobe DISTRIBUTION

Table 2.2 DOPPLER (Hz) vs ALTITUDE (km) and GRAZING ANGLE

ALTITUDE	3°	10°	20°	50°
500	76819	75756	72286	49446
550	76060	75007	71571	48957
600	75313	74271	70868	48477
650	74580	73548	70179	48005
700	73860	72838	69501	47541
750	73153	72140	68835	47086
800	72457	71455	68181	46638
850	71774	70781	67538	46199
900	71103	70119	66906	45767
950	70443	69468	66285	45342
1000	69794	68828	65675	44924
1050	69156	68199	65074	44513
1100	68528	67580	64484	44110
1150	67912	66972	63904	43712
1200	67305	66373	63333	43322
1250	66708	65785	62771	42938
1300	66121	65206	62218	42560
1350	65543	64636	61675	42188
1400	64975	64075	61140	41822
1450	64415	63524	60614	41462
1500	63865	62981	60096	41108
1550	63323	62447	59586	40759
1600	62790	61921	59084	40416
1650	62265	61403	58590	40078
1700	61748	60893	58103	39745
1750	61239	60391	57624	39417
1800	60737	59897	57153	39095
1850	60244	59410	56688	38777
1900	59758	58931	56231	38464
1950	59279	58458	55780	38156
2000	58807	57993	55336	37852
2050	58342	57535	54899	37553
2100	57884	57083	54468	37258
2150	57433	56638	54043	36968
2200	56988	56200	53625	36681
2250	56550	55767	53213	36399
2300	56118	55341	52806	36121
2350	55692	54922	52406	35847
2400	55273	54508	52011	35577
2450	54859	54100	51621	35311
2500	54451	53697	51237	35048

Figure 2.4 DOPPLER (Hz) vs ALTITUDE (km) and GRAZING ANGLE

The azimuth and depression angles of the antenna beam with respect to the spacecraft's velocity vector will also influence the baseband spectrum. Figure 2.6 illustrates the effect of various depression angles on the baseband spectra for a spacecraft at an altitude of 1,100 km. Note that the depression angle for a 3° grazing angle is 31.6° due to the curvature of the earth. Initially, at the minimum depression angle, a target assumed to be closing head-on is shown clear of both the narrow main beam clutter and wide side lobe clutter. The target's detection is limited by noise.

As the depression angle increases, the main beam clutter spectra moves down in frequency and broadens due to the effects of the $\cos \psi$ relative velocity profile. Also, the target moves closer to the main beam clutter spectra and may become immersed in the side lobe clutter. The target's detection is now limited in its detection by interference comprised of both noise and clutter. The largest interference occurs when the antenna is pointed directly below the spacecraft. Here, the target is at D.C. and immersed in both main beam clutter and one level of side lobe clutter making target detection by doppler processing impossible. For a satellite in an 1,100 km orbit with a grazing angle of 50°, the doppler drops to approximately two-thirds of the low grazing angle value. Above this angle, doppler discrimination will become less effective in separating the target from the clutter. The corresponding velocity is referred to as the minimum detectable velocity. Hence, the upper limit on grazing angle, referred to as nadir, is chosen to be 50° for the following work.

Similar spectral behavior occurs when the antenna is pointed in azimuth away from the direction of platform motion. The head-on condition, $\Theta_{az} = 0^\circ$, corresponds to $\psi = 31.6^\circ$ (3° grazing), $\Theta_{az} = 50^\circ$ corresponds to the $\psi = 57.1^\circ$ nadir spectra and $\Theta_{az} = 90^\circ$ corresponds to the $\psi = 90^\circ$ case. Therefore, both the azimuth and depression angles of the antenna will effect the resultant spectra complicating the design of a SBR. Note that the $\Theta_{az} = \psi = 90^\circ$ cases will create a blind region in a spaceborne radar's coverage area where doppler processing will not detect the target. The extent of the blind area, referred to as the nadir hole, will depend on the main beam clutter spread and the minimum detectable velocity selected.

Figure 2.6 SPECTRA vs DEPRESSION and AZIMUTH ANGLE

Using the 3° grazing angle and 50° nadir limit, the ground coverage of a SBR will now be calculated. The azimuth angle limit of $\pm 50^\circ$ will be neglected in order to simplify the computation. Equation 2.4 describes the relations used to compute the size of the annular coverage zone using the symbols corresponding to Figure 2.1. Both the average radius (R_{ave}) and width (W) of the annulus were calculated over a range of altitudes from 500 to 10,500 km. The results are listed in Table 2.3. Note that for a change in altitude from 1,000 to 10,000 km, the coverage annulus only increases about 2.5 times in radius and less than two times in width.

$$\sin(\alpha)/R_e = \sin(\Theta_g + \pi/2)/(R_e + \text{altitude})$$

$$\pi/2 + \Theta_g + \alpha + X = \pi$$

$$S = X \cdot R_e \quad (2.4)$$

Substituting 3° and 50° for Θ_g (grazing and nadir), we find:

$$R_{ave} = (S_{3^\circ} + S_{50^\circ})/2 \quad \text{and} \quad W = S_{3^\circ} - S_{50^\circ}$$

The ground distances covered for both the 3° grazing and 50° nadir cases are plotted in Figure 2.7 over the same altitude range. The difference between the two curves indicates the thickness of the annulus of coverage. Below approximately 2,000 km the coverage rapidly increases whereas above 4,000 km, the coverage slowly increases. Clearly, increasing the altitude to extend the coverage is more effective at lower altitudes.

Plots of the coverage zones for orbits of 500, 1,100 and 2,000 km altitude with a 3° grazing angle and a 50° nadir limit are superimposed on the earth in Figure 2.8. These plots are based on viewing the earth from infinity since the coverage zones were projected onto a flat surface. The effect of diminishing returns in coverage area is clearly evident in rising from 1,100 to 2,000 km where the increase in the coverage zone due to extending the 3° grazing length is being offset by a growing nadir hole.

Table 2.3 COVERAGE RADIUS (km) and COVERAGE WIDTH (km) vs ALTITUDE (km) at 3° GRAZING

ALTITUDE	AVE. RADIUS	WIDTH
500	1257	1754
1000	1867	2345
1500	2320	2717
2000	2683	2984
2500	2986	3190
3000	3244	3355
3500	3469	3492
4000	3666	3607
4500	3841	3706
5000	3998	3793
5500	4140	3869
6000	4268	3937
6500	4385	3998
7000	4492	4053
7500	4591	4103
8000	4683	4149
8500	4767	4190
9000	4846	4229
9500	4919	4265
10000	4988	4298
10500	5052	4328

Figure 2.7 COVERAGE LENGTH (km) at GRAZING and NADIR ANGLES vs ALTITUDE (km)

Figure 2.8 EARTH COVERAGE AS SEEN FROM INFINITY
(3° GRAZING and 50° NADIR)

2.2.2 POSITION EFFECTS on an ORBITTING RADAR

In this section, the unique effects resulting from the position of an orbiting radar platform will be considered. It will be seen that although a satellite may view a large region of the earth, the altitude that makes this possible creates a large slant range that complicates the radar design. These complications come from the large path loss and echo delay as well as the dependence of path loss, squint loss and ground swath on grazing angle. Again, these effects will be determined over a wide range of altitudes. The results will be used later in the monostatic radar calculations of Section 2.4.

Using the relationships of equation 2.5, the slant range, echo delay and path loss were calculated. The path loss is shown to be proportional to R^4 since electromagnetic energy disperses in all directions at both the radar and target. Hence, for a noise limited radar, a slant range increase will increase the required transmit power by this amount to the fourth power. Note that the path loss is calculated relative to that at an altitude of 1,100 km. The echo delay is determined by dividing the round trip distance by the speed of light, c (3E8 m/s).

$$R/\sin X = R_0/\sin \alpha$$

$$\text{Path loss} = 40 \log(R) - \text{Loss at 1,100 km altitude} \quad (2.5)$$

$$\text{Echo delay} = 2R/c$$

The slant range, echo delay and path loss over a 500 to 2,500 km altitude range for 3° grazing are listed in Table 2.4 while path loss vs altitude is plotted in Figure 2.9 and echo delay vs altitude is shown in Figure 2.10. Note that increasing the altitude from 1,100 to 2,000 km requires that the transmitted power be quadrupled. This power increase has to be weighed against the increased coverage that higher altitudes provide when determining the optimum altitude for a SBR. Also, the echo delay is seen to increase linearly from 15 to 39 milliseconds over the 500 to 2,500 km altitude range. This long delay is another unique property of a spaceborne platform due to the long slant ranges involved.

Table 2.4 SLANT RANGE (km), ECHO DELAY (msec) and PATH LOSS (dBr) vs ALTITUDE (km)

ALTITUDE	SLANT RANGE	ECHO DELAY	PATH LOSS
500	2262	15.08	-7.99
550	2391	15.94	-7.02
600	2516	16.77	-6.14
650	2636	17.57	-5.33
700	2753	18.35	-4.57
750	2866	19.1	-3.87
800	2976	19.84	-3.22
850	3083	20.55	-2.61
900	3187	21.24	-2.03
950	3290	21.93	-1.48
1000	3390	22.6	-.96
1050	3488	23.25	-.46
1100	3584	23.89	0
1150	3679	24.52	.45
1200	3772	25.14	.89
1250	3863	25.75	1.3
1300	3953	26.35	1.7
1350	4042	26.94	2.09
1400	4130	27.53	2.46
1450	4217	28.11	2.83
1500	4302	28.68	3.17
1550	4386	29.24	3.51
1600	4470	29.8	3.84
1650	4553	30.35	4.16
1700	4634	30.89	4.46
1750	4715	31.43	4.76
1800	4795	31.96	5.06
1850	4875	32.5	5.34
1900	4953	33.02	5.62
1950	5031	33.54	5.89
2000	5108	34.05	6.16
2050	5185	34.56	6.41
2100	5261	35.07	6.67
2150	5337	35.58	6.92
2200	5412	36.08	7.16
2250	5486	36.57	7.4
2300	5560	37.06	7.63
2350	5633	37.55	7.85
2400	5706	38.04	8.08
2450	5778	38.52	8.3
2500	5850	39	8.51

Figure 2.9 PATH LOSS (dBr) vs ALTITUDE (km) **Figure 2.10 ECHO DELAY (ms) vs ALTITUDE (km)**

The effect of changing the grazing angle on both path loss and squint loss will now be discussed. Squint loss is an additional loss created by scanning the beam of a planar phased array antenna away from boresight. The planar array antenna, to be discussed in detail in Chapter 3, is the probable choice for the track-while-scan and sequential scanning requirements of a SBR.^[8] A picture of this type of antenna structure scanning a pencil beam is shown in Figure 2.11.

Figure 2.11 ANTENNA STRUCTURE

The squint loss is calculated as the two-way loss in effective area of the antenna aperture which is proportional to $(1/\cos^2(\alpha))$.^[9] Table 2.5 lists path loss and squint loss vs grazing angle for a satellite at an altitude of 1,100 km. The results are also plotted in Figure 2.12. Note that the path loss shown is relative to the loss at 3° grazing. Regrettably, at the lower grazing angles where path loss is a maximum, squint loss is also a maximum.

Also note that the dynamic range required for scanning from 3° (grazing) to 50° (nadir) is 21 dB. Clearly the probability of target detection would increase closer to the nadir hole. However, the main beam clutter will increase in both magnitude and spectral spread in this region which will raise the minimum detectable velocity of the radar regardless of the signal to noise improvement.

Table 2.5 PATH LOSS (dBr) and SQUINT LOSS (dB) vs GRAZING ANGLE (degrees) at 1,100 km

GRAZING ANGLE	PATH LOSS	SQUINT LOSS
0	1.49	5.64
5	-2.97	5.55
10	-3.37	5.3
15	-5.65	4.93
20	-7.77	4.46
25	-9.71	3.95
30	-11.47	3.42
35	-13.04	2.9
40	-14.43	2.41
45	-15.65	1.96
50	-16.72	1.55
55	-17.65	1.18
60	-18.43	.87
65	-19.08	.6
70	-19.6	.38

Figure 2.12 PATH LOSS (dBr) and SQUINT LOSS (dB) vs GRAZING ANGLE (degrees)

The amount of ground coverage provided by the 3 dB beamwidth will also be required for the power calculations of Section 2.4 in order to determine how a given surveillance area will be scanned. The ground distance or swath covered by the 3 dB beamwidth at different altitudes and grazing angles for both a 0.55° beamwidth and a 1.1° beamwidth will now be calculated using the expressions listed in equation 2.6 as a pseudo-coded program. The program accounts for the upper section of the beam missing the earth at low grazing angles as well as the broadening of the main beam as $1/\cos(\alpha)$ as the planar phased array antenna is scanned away from boresight. [9]

With the symbols defined earlier in Figure 2.1:

$$\Theta_{BW} = \Theta_{BW0} / \cos(\alpha); \quad \alpha_1 = \alpha - \Theta_{BW}/2; \quad \alpha_2 = \alpha + \Theta_{BW}/2$$

$$\sin(\Theta_1 + \pi/2) / (R_e + \text{Altitude}) = \sin(\alpha_1) / R_e; \text{ and solve for } \Theta_1$$

$$\text{therefore } X_1 = \pi/2 - \Theta_1 - \alpha_1$$

$$\text{Let } A / (R_e + \text{Altitude}) = \sin(\alpha_2) / R_e; \text{ and solve for } A.$$

If A is greater than 1, the beam misses the earth. (2.6)

$$\text{therefore let } B = \text{MIN}(A, 1) \text{ and } \Theta_2 = \arccos(B)$$

$$\text{therefore } X_2 = \pi/2 - \Theta_2 - \alpha_2;$$

$$\text{and the arc lengths are: } S_1 = R_e * X_1; \quad S_2 = R_e * X_2$$

$$\text{giving a swath of } S_2 - S_1.$$

Table 2.6 lists and Figure 2.13 plots the swath lengths over the 500 to 2,500 km altitude range for a 0.55° beamwidth. Note that at an altitude of 1,100 km, the swath increases by a factor of 30 from nadir (21km) to the outer perimeter (641 km) which is set by the 3° grazing angle.

Table 2.7 lists and Figure 2.14 plots the swaths over the same altitude range for a larger 1.1° beamwidth. Unexpectedly, the 10° grazing swath exceeds that at 3° for altitudes over 2,200 km. The merging of the 3° and 10° grazing angle swaths at high altitude is caused by the fact that at the lower grazing angles, most of the upper half of the beam misses the earth. At an altitude of 1,100 km, the swath decreases by a factor of 19 from 788 km to 41 km at nadir. Note that the swath is not simply double the 0.55° beamwidth's swath due to the non-linear relationship involved. Considering that the radar will be in orbit, a complex antenna scanning pattern will be required in order to provide uniform coverage of a fixed surveillance area.

Table 2.6 SWATH LENGTH (km) vs ALTITUDE (km) and GRAZING ANGLE (degrees) for a 0.55° BEAMWIDTH

ALTITUDE	3 deg	10 deg	20 deg	50 deg
500	572.82	235.48	68.53	9.96
550	581.17	243.6	72.69	10.89
600	588.7	251.48	76.62	11.81
650	595.58	258.16	80.35	12.72
700	601.92	264.78	83.91	13.62
750	607.81	271.04	87.32	14.51
800	613.33	277.01	90.58	15.4
850	618.51	282.71	93.72	16.27
900	623.42	288.19	96.75	17.14
950	628.09	293.48	99.69	18.01
1000	632.54	298.59	102.52	18.86
1050	636.8	303.54	105.28	19.71
1100	640.9	308.36	107.96	20.55
1150	644.84	313.06	110.57	21.39
1200	648.66	317.65	113.12	22.22
1250	652.35	322.14	115.61	23.04
1300	655.93	326.53	118.05	23.86
1350	659.42	330.85	120.43	24.68
1400	662.81	335.09	122.77	25.49
1450	666.12	339.27	125.07	26.29
1500	669.35	343.38	127.33	27.09
1550	672.51	347.44	129.54	27.89
1600	675.61	351.44	131.73	28.68
1650	678.64	355.4	133.88	29.46
1700	681.62	359.31	136	30.25
1750	684.55	363.18	138.1	31.03
1800	687.42	367.01	140.16	31.8
1850	690.25	370.8	142.2	32.57
1900	693.04	374.57	144.22	33.34
1950	695.79	378.3	146.21	34.11
2000	698.5	382	148.19	34.87
2050	701.17	385.68	150.14	35.63
2100	703.81	389.34	152.07	36.39
2150	706.41	392.97	153.99	37.14
2200	708.99	396.57	155.89	37.89
2250	711.53	400.16	157.77	38.64
2300	714.05	403.73	159.64	39.38
2350	716.54	407.28	161.49	40.12
2400	719	410.82	163.33	40.86
2450	721.45	414.34	165.15	41.6
2500	723.86	417.85	166.96	42.34

Figure 2.13 SWATH (km) vs ALTITUDE (km) and GRAZING ANGLE

Table 2.7 SWATH LENGTH (km) vs ALTITUDE (km) and GRAZING ANGLE (degrees) for a 1.1° BEAMWIDTH

ALTITUDE	3 deg	10 deg	20 deg	50 deg
500	666.81	508.22	137.68	19.93
550	682.08	526.14	140.04	21.79
600	695.79	542.82	153.94	23.63
650	708.23	558.51	161.45	25.45
700	719.64	573.35	168.6	27.25
750	730.18	587.49	175.45	29.04
800	739.99	601.05	182.01	30.8
850	749.16	614.09	188.33	32.56
900	757.4	626.7	194.43	34.3
950	765.97	638.94	200.32	36.02
1000	773.72	650.86	206.04	37.73
1050	781.11	662.49	211.58	39.43
1100	788.18	673.89	216.98	41.11
1150	794.96	685.07	222.24	42.79
1200	801.48	696.07	227.37	44.45
1250	807.76	706.9	232.38	46.1
1300	813.84	717.6	237.28	47.74
1350	819.72	728.18	242.09	49.37
1400	825.42	738.66	246.8	50.98
1450	830.96	749.05	251.43	52.59
1500	836.36	759.37	255.98	54.19
1550	841.62	769.62	260.45	55.79
1600	846.75	779.83	264.85	57.37
1650	851.76	790	269.19	58.94
1700	856.67	800.14	273.47	60.51
1750	861.47	810.26	277.69	62.07
1800	866.18	820.38	281.86	63.62
1850	870.8	830.49	285.98	65.17
1900	875.34	840.61	290.05	66.7
1950	879.9	850.75	294.07	68.23
2000	884.18	860.92	298.05	69.76
2050	888.5	871.11	302	71.28
2100	892.75	881.35	305.9	72.79
2150	896.94	891.64	309.77	74.3
2200	901.07	901.98	313.61	75.8
2250	905.14	912.38	317.41	77.29
2300	909.17	922.86	321.18	78.78
2350	913.14	933.42	324.93	80.27
2400	917.06	944.07	328.64	81.75
2450	920.94	954.83	332.33	83.22
2500	924.77	965.69	336	84.69

Figure 2.14 SWATH (km) vs ALTITUDE (km) and GRAZING ANGLE

2.2.3 CONCLUSIONS

The effects of altitude on doppler, coverage, path loss, echo delay and beam swath have been discussed in this section. Note that many differences exist between a SBR in low earth orbit and a modern airborne pulse doppler surveillance radar. Table 2.8 summarizes these differences for a 3° grazing angle SBR at an altitude of 1,100 km and an AWACS system. Compared to the airborne radar, the SBR must see nine times as far, has over twenty times the clutter smearing, and may cover an area over fifty times as large using under 17% of the available power. Clearly, a monostatic SBR placed in low earth orbit will require major advances in present technology.

TABLE 2.8 SBR vs AIRBORNE RADAR PARAMETERS

	SBR	AWACS ^[10]	Ratio
Slant Range (km)	3,580	400	9
Platform Velocity (km/s)	7.3	0.3	24
Platform Vel/Target Vel	11	0.5	22
Coverage Area (10^6 km^2)	8.7	0.5	57
Antenna Beam	spot	fan	---
Power Available (kVA)	100 max	600+	< 0.17

Both the effects caused by the unique velocity and position of a SBR as well as other system constraints including the capacity of available launch vehicles, the position of the radiation belts that encircle the earth and the minimum safe altitude for a large satellite to remain in orbit indefinitely must be considered in order to select the optimum altitude for a spaceborne radar. Combining these considerations, an altitude of 1,100 km was selected for the waveform analysis of Section 2.3 and the monostatic analysis of Section 2.4.

2.3 PULSE REPETITION FREQUENCY SELECTION

Before proceeding with the power calculation for the monostatic architecture, the design of the radar waveform must be considered. In this section, basic radar waveform theory will be discussed. This will allow the optimum pulse repetition frequency (PRF) for a monostatic radar in a low earth orbit to be selected. Here, the chosen PRF will provide a simplified method of separating a moving target from the stationary ground clutter.

A pulsed signal, $f(t)$, of period T shown in Figure 2.15(a) and described by equation 2.7 is the most common waveform found in radar systems. We call $1/T$ the pulse repetition frequency (PRF) of the signal. Applying Fourier analysis, the waveform is expressed in the frequency domain by equation 2.8 where f_0 is the PRF. This spectrum is plotted in Figure 2.15(b). Note that the spectra is repeated every PRF and is weighted by the Fourier transform of the pulse. [11]

$$\begin{aligned} f(t) &= A ; & -t_p/2 \leq t \leq t_p/2 \\ &= 0 ; & \text{elsewhere} \end{aligned} \quad (2.7)$$

$$f(t) = \frac{At_p}{T} + \frac{2At_p}{T} \sum_{n=1}^{\infty} \text{sinc}(n\pi f_0 t_p) \cdot \cos(2n\pi f_0 t) \quad (2.8)$$

FIGURE 2.15 (a) TIME WAVEFORM of a PULSE

b) PULSE SPECTRUM

If this signal modulates a microwave carrier at frequency f_c such that $f(t) = A \cos(2\pi f_c t)$, the whole spectrum will be translated in frequency becoming centered about f_c . Similarly, when such a signal is demodulated, the baseband signal will look like Figure 2.15(b). Note that individual echos will be repeated at $1/T$ or every PRF in the frequency domain. This leads to ambiguous velocity and range limits that will be described next.

Doppler processing is normally used to separate the echo of a moving target from that of stationary clutter and transmitter noise.^[2] Here, a moving target's doppler will shift its echo spectra in frequency relative to that shown in Figure 2.15(b). Provided that such a shift does not exceed half the PRF, the target's unambiguous velocity may be determined. Using equation 2.2, the maximum unambiguous relative velocity may be expressed as $\lambda/4T$.

Since a pulse of electromagnetic energy travels at the speed of light (c), in time T , a pulse may return from a target placed at a maximum distance of $cT/2$. If a target is placed further than $cT/2$ where T is the pulse repetition interval ($1/\text{PRF}$), its echo will return after the next pulse is transmitted. If it arrives $\hat{c}$ seconds later, it is impossible to differentiate this echo from that of a target placed at a distance of $c\hat{c}/2$. The target is said to be outside the unambiguous range of the radar. Hence, the unambiguous range is $cT/2$ for a radar with a single PRF.

Note the tradeoff in radar waveform design. If the PRF is increased, the target's true velocity may be obtained while the range information is lost. This tradeoff will now be made in choosing the PRF for a monostatic SBR. The following is based on assuming a radar platform in a circular 1,100 km altitude orbit.

The first waveform structure to be discussed is referred to as Low pulse repetition frequency (LPRF). LPRF radars determine the unambiguous range and ambiguous velocity of a target. LPRF radars usually have a PRF of under 1 kHz and are used widely for air traffic control. Since the velocity information is ambiguous due to the LPRF, it is usually derived by tracking a target over a fixed time interval or by PRF staggering. The clutter experienced by these radars is either ground clutter centered around DC or slowly moving clutter caused by weather or angel returns that seldom exceeds more than a sixth of the doppler band. MTI processing is usually employed to filter out the narrow clutter band leaving the doppler passband.^[12]

In a SBR in low earth orbit, the platform's rapid motion will cause side lobe clutter to extend over $\pm F_{dr}$ as shown in Figure 2.3. If a LPRF waveform design were used, this clutter will foldover in the frequency domain ($2 \cdot F_{dr} / \text{PRF}$) times as illustrated in Figure 2.16. Note that the side lobe clutter (SLC) level from many separate ground locations will add since the main beam clutter (MBC) spectra is repeated every PRF over the target's doppler range. The lower the PRF, the greater the number of SLC echos of different frequencies corresponding to targets at different Θ_{az} and Ψ will be present throughout the passband. Due to the rapid motion of the platform, the clutter will vary rapidly with time. This dynamic clutter will effectively raise the noise floor by a substantial amount degrading the ability to separate moving and stationary targets using doppler processing.

Figure 2.16 LPRF APPLIED to a SBR

An analysis of the SLC level for SBR has yet to be done. However, in airborne look-down radars, the SLC problem has been so severe that a high PRF (HPRF) system is normally used for the detection of moving targets over land. Due to the more severe requirements of a SBR with respect to airborne radar as summarized in Section 2.2.3, a LPRF system would be saturated with clutter and would fail to detect either range or velocity information. [13] [14]

The second waveform structure to be considered is referred to as High PRF (HPRF). Usually, the PRF is set high enough to determine the target's true relative velocity. This occurs when the PRF is over twice the SLC frequency extent plus the maximum target doppler as shown in Figure 2.17. Note that if the antenna is directed along the velocity track and the target is closing, or if the antenna is directed aft and the target opening, then the target will appear in a clutter free region and detection will be limited only by noise. For other antenna orientations, the target will be in at least one level of SLC.

In a SBR, the PRF required to form a true HPRF system will exceed 140 kHz. In order to cover the entire line-of-sight annulus, the antenna will have to be scanned in both azimuth and elevation making the target compete with one level of SLC most of the time. The situation shown in Figure 2.17, where the target is clear of clutter interference, will seldom occur.

Figure 2.17 HPRF APPLIED to a SBR

The question must be asked: what have we given up by using a HPRF waveform? Clearly the HPRF will create ambiguous range information. A PRF of 140 kHz will yield an unambiguous range of only 1.1 km. However a multi-PRF ranging system may be used to extend the unambiguous range up to 50 times this value. Note that this technique creates blind velocities which may be treated as a reduction in the signal to noise ratio when applying the radar equation. For a SBR, where the PRFs have to be around twenty times higher and the slant range is around nine times longer than for the airborne case, it is evident that this system would still result in ambiguous range information or an unacceptable system loss. Also, multi-PRF systems are unable to deal with multiple targets in a single scan region. Hence, in return for both the SLC and MBC being fixed in frequency and easier to reject, the true range of the target will be lost and a larger dynamic range processor will be required to handle the higher main beam clutter level.

Another important difference between an airborne and a platform in low earth orbit is the proportion of the clutter free band to the total band (PRF). In the airborne case, the target is usually travelling faster than the platform so the clutter free region is over 50% of the band. With a SBR, the target's maximum speed is under 5% of the platform's speed giving a noise limited region of under 5% of the total band. However, the unique property of the SBR's high velocity can be used to further reduce SLC interference. Since the target may only exist in the $\pm F_{dr}$ band around the MBC spectra as shown in Figure 2.17, the frequency bands external to this may be removed by filtering. This will reduce the SLC to only F_{dr}/F_{dr} of the level that would be calculated assuming total band, all aspect, SLC interference. Another way to view this is that only the close-in antenna side lobes would need to be reduced since far-away side lobes will create SLC of doppler outside the target's doppler band. This characteristic is unique to a low earth orbit SBR.

Another possible waveform is a compromise between LPRF and HPRF and is referred to as medium PRF. As stated earlier, a HPRF system trades unambiguous velocity and a higher MBC level over a narrow band for quasi-static SLC over the entire band. However, in airborne radars it has been found that for a closing target in tail-on aspect, the detection range for a HPRF system is severely degraded. This has led to the development of medium PRF systems. Here, the PRF is decreased giving both ambiguous range and velocity measurements. Although the maximum range in some directions is degraded, the overall range in all azimuth directions is more or less constant.^[15]

In the latest pulse doppler radars, multimode systems are employed. Here, a different PRF or waveform structure is used depending on the type of clutter, antenna orientation or type of target being detected. In a SBR, it would probably be of advantage to use a medium PRF waveform in the sidelooking orientations to decrease the minimum detectable velocity and a HPRF waveform when scanning head-on. In conclusion, a multimode design, with its inherent complexity, is the probable waveform architecture for a SBR in low earth orbit.

2.4 MONOSTATIC POWER and CLUTTER CALCULATIONS

In this section, the power required to detect a small target using an L-band monostatic radar placed in an 1,100 km circular orbit will be calculated. In order to perform the calculation, the radar equation will be applied to a proposed surveillance scenario for a given target. The analysis will assume interference limited to only thermal noise. Although the effect of clutter interference will be neglected in the calculation, the level of clutter will be approximated and the problems that it will create as well as some possible solutions will be discussed. Finally, some of the advantages and disadvantages of the monostatic architecture will be described.

2.4.1 THE RADAR EQUATION^[7]

In order to calculate the power requirements of a SBR, the radar equation will be applied. The monostatic radar equation will now be derived. Note that a monostatic radar uses a common antenna aperture to both transmit and receive.

An antenna of gain G transmitting power P_t , will create a power density of $P_t G / 4\pi R^2$ at a distance R along the axis of the main beam of the radar's antenna. Targets are characterized by their radar cross section (RCS) which is given the symbol σ . The RCS is a measure of the amount of power reflected back to the radar by the target and the RCS values quoted later in this section are an all-aspect average value at L-band. Averages are used since the probability of a radar detecting a target is a statistical quantity. Hence, for a receiving antenna of effective area A_e , the power received from a target of RCS σ is given by equation 2.9.

$$P_r = (P_t G / 4\pi R^2) (\sigma / 4\pi R^2) A_e. \quad (2.9)$$

Equation 2.9 may be rearranged to determine the maximum range of a radar capable of detecting an echo signal, P_r , with a minimum power level S_{min} as shown in equation 2.10. Note that the transmit power varies as the range to the fourth power.

$$R_{max}^4 = P_t G \sigma A_e (4\pi)^{-2} S_{min}^{-1} \quad (2.10)$$

Since the noise figure, F_n , of a receiver is equal to the ratio of the input signal to noise ratio $(\text{SNR})_{in}$ divided by the output signal to noise ratio $(\text{SNR})_{out}$ and assuming thermal input noise $N_i = K T_o B_n$ where K is Boltzmann's constant, T_o is the effective system temperature and B_n is the bandwidth, we find that S_{min} may be expressed as $K T_o B_n F_n (\text{SNR}_{out})_{min}$. Substituting this into equation 2.10, we may obtain equation 2.11 which is a form of the radar equation. From statistics, $(\text{SNR})_{out}$ is selected to be the minimum value required to provide a desired probability of detection and probability of false alarm.

$$P_t = (4\pi)^2 R_{max}^4 K T_o B_n F_n (\text{SNR}_{out})_{min} / G A_e \sigma \quad (2.11)$$

The above equation is based on the return of a single pulse. In reality, the transmitter's power requirement may be reduced by integrating the returns from several pulses. In a coherent system, the pulse's signal voltage vectors will add whereas thermal noise will add on a power basis. Hence, for n pulses $(\text{SNR})_{integrated(I)} = n * (\text{SNR})_{1 \text{ pulse}}$. For incoherent integration, the SNR improvement may be as low as $\sqrt{n}$. The more useful form of the radar equation incorporating integration gain is shown below in equation 2.12 where SNR_n is the per pulse SNR required to form SNR_I from n pulses.

$$P_t = (4\pi)^2 R_{max}^4 L_s (K T_o B_n F_n) (\text{SNR}_n) / G A_e \sigma \quad (2.12)$$

Note that an additional term L_s has been added to equation 2.12. This term represents system losses which will occur in any physical implementation. Possible system losses are listed below.

1. Atmospheric propagation loss due to precipitation attenuation and polarization rotation.
2. Antenna mismatch between the feed network and the radiator(s).
3. Eclipsing of the receive window by the finite transmit time.
4. Doppler filter straddle in a coherent processor where a signal is spread over more than one velocity bin during the processing interval.

5. Antenna losses include a swath loss due to the main beam's gain rolling off near the beam's edges, squint loss, as previously discussed in Section 2.2.2, and a weighting loss resulting from weighting the antenna aperture in order to reduce the side lobe levels. Table 2.9 illustrates the tradeoff of antenna efficiency (main lobe loss) versus side lobe reduction. The main lobe losses are referenced to a uniformly weighted aperture. This weighting loss will be incorporated in the swath loss in the following calculations.

Table 2.9 WEIGHTING EFFECTS^[14]

Weighting	First Side Lobe Level (dBr)	Main Lobe Loss (dB)
uniform	-13.5	0
triangular	-27	1.2
$\cos^2$	-31.5	1.75
hamming	-42.8	1.74
hanning	-31.6	1.76

6. Field degradation due to system aging, etc.

2.4.2 MONOSTATIC ANALYSIS

The following analysis is based on the detection of small radar cross section airborne targets. Three different types of targets with their RCS and maximum speed listed in Table 2.10 may be detected by an SBR. Slow moving, large radar cross section (RCS) Type A targets would be encountered over the sea while high speed, medium RCS Type B targets and medium speed, low RCS Type C targets may operate over the entire globe. Airborne experience has shown that the detection of the small Type C targets is the most challenging. Hence, the following power calculations are based on the detection of Type C targets.

Table 2.10 TARGET CHARACTERISTICS

	RCS (m^2)	Maximum Velocity (knots)
Type A	20,000	30
Type B	5	1,200
Type C	0.1	500

The surveillance structure for the detection of these targets will now be defined. A fixed surveillance fence located along the line $170^\circ W - 10^\circ E$ longitude and spanning ± 70 degrees latitude across the North Pole is used as the surveillance region in this analysis. This fence, shown in Figure 2.18 below, is 4,450 km long and 530 km wide and is comprised of many individual cells corresponding to the antenna's spot beam coverage. From Figure 2.7, a satellite in an 1,100 km orbit with a 3° grazing angle coverage limit can cover an annulus of 6,370 km outer ground distance diameter. Hence, the 4,450 km long fence could be completely covered by one satellite assuming that its ground track passed close to the middle of the fence's length. Since the speed of the Type B target is highest, it will determine the re-scan time of a stationary surveillance fence. However, the small RCS of the Type C target will determine the required transmitter power.

Figure 2.18 North American FENCE STRUCTURE

In these calculations, it is assumed that a gravity stabilized, square, phased array antenna structure of Figure 2.11 is utilized. Although a mechanically pointed antenna would reduce the squint loss, severe mechanical steering problems would be created due to the complex scanning pattern required. Also, mechanical scanning would probably create an inertia/stability problem since the antenna would be a large percentage of the total mass of the spacecraft.

In choosing the antenna weighting, the following tradeoff must be considered. In addition to the main lobe gain loss, weighting will also widen the main lobe. When the side lobes are reduced, decreasing the minimum detectable velocity, the broader main beam clutter will tend to increase the minimum detectable velocity. Clearly, a lower minimum detectable velocity will need to be traded for prime power. A compromise solution between a highly weighted, high main beam loss configuration and an unweighted, low main beam loss with higher minimum detectable velocity but lower power requirements resulted in a $\cos^2 x$ amplitude weighting being chosen. This weighting produces side lobes under 31 dB, a main beam gain loss of 1.8 dB, and a 3 dB beamwidth, in degrees, of $83.2 \lambda / (\text{aperture length})$. A 30m square aperture generating a 0.55° wide pencil beam at L-band (1.5 GHz) was chosen for the following analysis. It is believed that this would be the largest possible antenna for a spaceborne platform.

The 530 km width of the fence is set by the antenna's 0.5 dB swath length at 3° grazing. Note that if the 3 dB swath calculated in Table 2.6 is used, a 6 dB loss occurs at the cell edges. Hence a 0.275° 0.5 dB beamwidth is used resulting in a 1 dB two-way spot beam periphery loss. This gives an additional average loss of 0.6 dB across the main beam for a total swath loss of 2.4 dB. The azimuth extent, assuming a 3 dB beamwidth coverage, is simply the azimuth beamwidth times the slant range which is 34 km for this antenna.

For this surveillance structure, a Type B target would take 530 km / 2,200 kPH or 870 sec to penetrate the fence in range and only 34 km / 2,200 kPH or 56 sec in azimuth. Hence, after a maximum of 870 sec, the entire fence must be re-scanned and a maximum of 56 sec is available to dwell on each spot beam's coverage area which will be referred to as a cell. The fact that there are 4,450 km / 34 km or 130 cells further limits the cell dwell time to 870 sec / 130 or 6.7 sec assuming a worst case analysis based on all the cells being located along the periphery of the coverage zone.

If the cell is adjacent to the nadir hole (50° grazing angle), the slant range will decrease from 3,580 km at grazing to 1,370 km giving a SNR improvement of 16.7 dB due to the lower path loss. Also, the antenna's effective area will increase due to the reduced squint loss resulting in a 4 dB gain. Hence, an overall SNR gain of 20.7 dB should occur at nadir. However, considering that the nadir's cell will be reduced to 13 km in azimuth extent and 15 km in swath length, many more cells will need to be scanned in the same overall re-scan time. If all the cells were in the nadir region, the earlier fence would require $4,450 \times 530 / (13 \times 15)$ or 12,000 cells reducing the cell dwell time to $870 \text{ sec} / 12,000$ or 0.07 sec. Assuming perfect integration, this would reduce the SNR by $6.7 \text{ sec} / 0.07 \text{ sec}$ or 19.7 dB. Hence, the net SNR gain is only 1 dB at nadir.

In reality, the fence will be comprised of cells ranging in size from 13 X 15 km at nadir to 34 X 530 km at a 3° grazing angle. Some surveillance cells inside the fence as seen from the satellite passing over the fence are shown in Figure 2.19. Clearly, this will result in a complex antenna pointing system if the fence is to be uniformly illuminated.

Figure 2.19 CELL SIZE vs POSITION

Note that a Type C target travels at approximately 42% the speed of a Type B target. Hence, the Type C will be present for 2.4 scans of the fence and a single scan probability of detection of 0.8 will lead to a cumulative probability of detection of 97.9%.

The spacecraft's 7.3 km/s velocity will further complicate the fence scanning problem. In 870 seconds, the radar will travel 6,350 km in space corresponding to 5,420 km on the ground as shown in Figure 2.20. This vast movement during a single scan is significant since the radar will not be able to re-scan the fence in order to take advantage of the presence of the Type C over the 2.4 scans. At best, the fence could be re-scanned $6,370 \text{ km} / 5420 \text{ km}$ or 1.2 times. Also, when side-looking, the satellite will travel in azimuth $T_{\text{cell dwell}} \cdot V_r$. At 7.3 km/s, a nadir element only 13 km wide will be travelled through in 1.8 sec. Hence a lower integration time or cell dwell time must be used.

Figure 2.20 SATELLITE MOVEMENT DURING A SINGLE FENCE SCAN

In the following calculation, the maximum integration time will be set at one second per cell requiring the entire fence to be scanned in a maximum of 130 seconds. Before the satellite moves out of the fence's range, it will re-scan the fence at least $6,370 \text{ km} / [7.3 \text{ km/s} \cdot (6,380 \text{ km} / 7,480 \text{ km}) \cdot 130 \text{ sec}]$ or 7.9 times. During the $1.18 \cdot 870 \text{ sec}$ or 1,023 sec of fence scan time, a Type C target will penetrate only $250 \text{ m/s} \cdot 1,023 \text{ s}$ or 256 km and a Type B target will penetrate 614 km. Hence, the Type B will be present for 6.8 scans from one satellite and assuming a constellation of satellites, the Type C will be present for 7.9 scans from the first satellite, 7.9 scans from the next satellite and 0.6 scans from the following satellite for a total of 16 scans. The large number of scans as well as the ability to view the same cell from sixteen different azimuth/depression angles will make target RCS reduction extremely difficult causing the cumulative detection probability to rise.

The amount of power required to detect a Type C target will now be calculated. The following calculation assumes the worst case target being the Type C with a 0.1 m^2 RCS. Although Section 2.3 found a multimode waveform to be optimum, the following calculation is based on a HPRF design. However, it applies equally well for other waveform designs which would only differ in their MBC to SLC ratios since this analysis assumes the best case of being noise limited. No calculation of the increased probability of false alarm due to clutter is considered.

The many variables required to apply equation 2.12, the monostatic radar equation, will now be considered.

A PRF of 140 kHz was chosen which includes a safety margin over the $2 \times 68.5 \text{ kHz}$ requirement for minimizing SLC. Assuming a $1/4$ second coherency time where perfect integration occurs, a processing gain of $140,000/4$ or 45.5 dB is possible. Over the one second dwell time, four $1/4$ second coherency periods are assumed to give a $\sqrt{N}$ integration gain of 3 dB. Hence a total processing gain of 48.5 dB is assumed.

The worst case range (R) is the slant range at 3° grazing which was found in Section 2.2 to be 3,584 km.

The effective area (A_e) of the 30m X 30m antenna is reduced by squint to 900×0.66 or 595 m^2 resulting in an antenna gain (G) of $4\pi A_e / \text{wavelength}^2$ or 52.7 dB

The Type C target is Swerling 2 with an assumed RCS (σ) of 0.1 m^2 .

From Skolnik, for a probability of detection of 0.5 per scan and a probability of false alarm of 10^{-7} , the SNR correction is 1 dB for a Swerling 2 target and the required $\text{SNR}_{\text{integrated}}$ is 12 dB resulting in a $\text{SNR}_{\text{integrated}}$ of 13 dB being required. Hence, a cumulative probability of detection for a Type C target of $1 - (0.5)^{16}$ or 0.99998 should be realized. [3]

The HPRF waveform used is shown in Figure 2.21. It has a noise bandwidth (B_n) of $1/(\text{pulse length } t_p)$ or 280 kHz giving a false alarm time of 35.7 seconds. So, for every 36 cells scanned, the radar will need to re-scan one cell in order to confirm that a target is present or that a false alarm occurred. Hence, $KT_0 B_n F_n$ equals $1.741 \text{E-}15$ Watts assuming F_n is 1.5 and T_0 is 290°K .

Figure 2.21 TIME WAVEFORM

The system losses are listed below in Table 2.11. Note that the overall system loss is 16 dB. This loss is mainly determined by the antenna architecture. An active array antenna which will be described in Chapter 3 is assumed in this table.

Table 2.11 SYSTEM LOSSES in dB

Propagation loss	1
Antenna mismatch	2
Eclipsing	1 (assumes a pulse burst waveform which will be discussed later in this section)
Doppler filter straddle	1
Squint loss	5.6
Swath loss	2.4
<u>Field degradation</u>	<u>3</u>
Total Loss (L_s)	16

Substituting these variables into equation 2.12, one obtains:

$$P_t = (4\pi)^2 (1.741E-15) (10^{1.3}) (3.584,000^4) (10^{1.6}) (10^{-5.27}) (595^{-1}) (0.1^{-1})$$

$$= 3.25 \text{ GW for a single pulse}$$

For a processing gain of 48.5 dB, P_t is $3.25E9 / 10^{4.85}$ or 45.94 kW. For a 50% duty cycle waveform, the average RF power requirement becomes 22.97 kW. Assuming a 33% power conversion efficiency results in an average power generating requirement of 69 kW to provide a probability of detection of 99.9%, probability of false alarm of 10^{-7} and false alarm time of 35 seconds for a Type C target in a noise limited environment. This is a substantial amount of power. For comparison, most commercial satellite transponders produce under 10 W of power at C-band.

The qualitative effect of ground clutter will now be discussed. A rough calculation of the expected magnitude of the clutter will be attempted and the problems that this creates along with some possible solutions will be detailed.

Since a SBR is operating in a look-down mode, the target's echo will be immersed in a background of surface reflections referred to as clutter. Clutter is characterized by a radar cross section density σ^0 . The magnitude of the clutter echo will be $\sigma^0 \cdot A$ where A is the area on earth intersected by the pulse transmitted within the swath of the main beam.

Over land, the worst clutter environment, σ^0 is a function of the type of surface terrain, grazing angle, frequency, pulse length and polarization. Table 2.12 summarizes the σ^0 values for an L-band radar at 8° grazing. The V and H columns represent vertical and horizontal polarization.

TABLE 2.12 SURFACE RCS vs POLARIZATION and TERRAIN [3]

Arizona terrain	σ^0 (dB)	
	V	H
desert	-35	-30
hills	-30	-25
mountains	-25	-24

Although vertical polarization seems to reduce the clutter, we are unable to use this σ^0 as a function of polarization property to reduce the magnitude of the clutter since the radar will view the clutter over a range of polarizations due to the dynamics of the problem. That is, the antenna will be fixed in orientation due to control system constraints and as the radar changes its scan angle, the polarization of the beam striking the ground will also change as shown in Figure 2.22. Considering the lack of σ^0 values for the Arctic a value of -20 dB will be used in the following calculation. This value is considered conservative.

Figure 2.22 POLARIZATION ROTATION

The worst case clutter occurs for 3° grazing. Due to the HPRF, no range gating is possible and the clutter competing with the target is the entire ground swath or cell area. Hence, the MBC/RCS_{target} is simply the area of the cell multiplied by σ^0 or $(530,000\text{m})(34,000\text{m}) \cdot (0.01) / (0.1\text{m}^2)$ giving a clutter to signal ratio of 93 dB maximum. To prevent saturation of the receiver by the clutter, over 16 bits of digitization will be required.

To reduce the high MBC and eclipsing loss of the 50% duty cycle waveform shown in Figure 2.21, a pulse burst waveform may be employed.^[16] This technique takes advantage of the long echo delay of from 9 msec to 24 msec for a SBR at an altitude of 1,100 km. For 3° grazing, we may send $23.9\text{ms} / 3.57\mu\text{s}$ or 6,700 pulses and then receive the echos as shown in Figure 2.23. A guard time interval of 3.9 msec is required to account for the time delay through the 530 km swath.

Due to the guard time and long receive window pulse compression techniques may be employed. Pulse compression is a technique in which matched filtering is used to synthesize a very short pulse. This results in better range resolution reducing the clutter level. For a 100:1 pulse compression ratio, the MBC/RCS_{target} ratio would be reduced by 20 dB to a more manageable level of 73 dB.

Figure 2.23 PULSE BURST WAVEFORM

In addition to the large value of stationary clutter, the problem of main lobe clutter smear must be considered in a doppler based detection system. The target's minimum detectable velocity will be determined by the large clutter return being smeared in frequency due to platform motion. Below this velocity, the radar will be unable to separate the target and clutter. The target will pass undetected. A qualitative discussion of the problem and some solutions for the specific case of a SBR follow.

CASE 1 BROADSIDE

The doppler distribution of a side-looking antenna is shown in Figure 2.24. Note that the doppler follows a sine curve about the origin. Hence, small changes in azimuth angle will create large changes in doppler.

Figure 2.24 MAIN BEAM DOPPLER SMEAR

At an altitude of 1,100km, a 0.55° boresight beamwidth will grow to 1.05 degrees at 3° grazing. The velocity spread in the 6 dB two-way region of the main beam will be $7.3 \text{ km/s} \times \sin(1.05/2)$ or 67 m/s.

To be more accurate, the antenna's first null width should be considered due to the high level of the MBC. For a $\text{SINC}^2(x)$ beam pattern, this will occur at approximately three times the 3 dB beamwidth giving a velocity spread of 201 m/s. [9] Considering the maximum velocity of the Type B is 651 m/s and that of the Type C is 250 m/s, MBC will cover approximately $201/651$ or 31% of the doppler band. Any Type C targets further than 36 degrees away from head-on aspect will be lost in the MBC.

CASE 2 HEAD-ON

The doppler distribution of a forward looking antenna is shown in Figure 2.25. Note that the main beam doppler follows a cosine curve about the origin. Hence, the small angle of the main beam will create little doppler spread at 3° grazing.

Figure 2.25 HEAD-ON SMEAR

The worst case will occur near the nadir hole where we are operating on the 57° region of the cosine curve. Fortunately, the main beam spread at nadir is smaller due to the reduced squint angle. A 0.55° boresight beamwidth is 0.66° wide at nadir resulting in the clutter's velocity spread being $7.3\text{km/s} \cdot (\cos(57+0.66/2) - \cos(57))$ or 35.3 m/s. Using the first null beamwidth, this doppler spread corresponds to $106/651$ or 16% of the doppler band. Type C targets would be detectable until the angle between V_r and V_t exceeded 65°.

Clearly, MBC raises the minimum detectable velocity and this must be reduced in order to have a reasonable probability of detecting Type C targets. Some of the possible solutions to this problem are described next.

1. Assuming that the MBC is symmetric about DC at baseband, the upper and lower sidebands may be subtracted to yield the target which will only appear in one of the sidebands. The doppler processing in which complementary frequency bin values are subtracted is relatively simple to implement. However, the question of whether tundra clutter will yield a symmetric MBC spectra is yet to be answered and the dynamic range of the radar must be large enough to handle the clutter to signal ratio of over 70 dB.

2. The second solution is the enlargement of the physical antenna. As the beamwidth decreases, so will the clutter spread allowing the minimum detectable velocity to decrease. There is a direct relationship between increasing the antenna aperture and decreasing the minimum detectable velocity. Practical problems of fabrication and pointing stability will limit this solution.

3. Possible real-time synthetic aperture techniques should be available by the early 1990's. Today, synthetic aperture radar processing runs at one six-hundredth real-time. However, intensive work is presently under way to use synthetic aperture radars for rear echelon real-time surveillance of ground vehicles and by the 1990's, processor speeds should increase approximately 250 times. Hence, the possibility of using a synthetic aperture radar for a future SBR exists.^{[3] [17]}

4. A displaced phase center antenna is the last solution to be discussed. Here the phase center of the antenna is shifted after every pulse by rapidly re-weighting the antenna aperture to synthesize a quasi-static platform.^[18] The amount of shift of the phase center required for the SBR is $V_r \cdot PRF^{-1}$ or 5 cm maximum. Displaced phase center techniques have been applied to some low speed airborne platforms.^[19] However, implementing a displaced phase center antenna on a SBR is complicated by the significant echo delay since upon reception, the correct sequence of echos must be combined to form short pulse sequences where the antenna was effectively stationary. Hence, the sequences must be tagged to determine their beginning and end which will complicate the mechanization. Also the physical aperture is not used fully for each pulse transmission resulting in another antenna loss.

2.4.3 ADVANTAGES and DISADVANTAGES

The advantages and disadvantages of a monostatic SBR will now be discussed.

ADVANTAGES

Considering the 6,367 km footprint for a satellite at an altitude of 1,100 km and the 40,074 km circumference of the earth, a single satellite would spend under 16% of its total orbital period being able to scan the single Northern fence. During 84% of its orbit, the power generated would be wasted. Clearly, the SBR in a low earth orbit would be more intelligently used to gather information from around the globe by laying down at least six other fences. This global coverage capability of a SBR in a low earth orbit is its main advantage.

DISADVANTAGES

A SBR could be jammed using simple noise jamming due to its low earth orbit of constant altitude. The usual burn through at a distance below $R_{\max}$ due to the R^4 radar power relationship and the R^2 jammer power relationship will not occur due to the fixed orbit radius. Hence, a monostatic SBR could be blinded at a proportionately lower cost by using ground-based jammers.

The overall system is extremely complex. Both antenna pointing and clutter rejection processing will require a spaceborne computer far more advanced than anything yet built.

Another disadvantage is the high transmit power required due to the two-way path loss for a monostatic configuration. Also a low earth orbit will create frequent solar eclipsing of the spacecraft. No spaceborne, high power generator meeting these requirements has been demonstrated in the West.^[20]

A satellite will be utilized for under a sixth of its orbital period if only a Northern fence were to be covered. Also many satellites, approximately 20, will be required due to the limited coverage that a low earth orbit provides.

The communications with the satellite will be difficult as a result of the low earth orbit. Many ground stations with moderately complex tracking systems will be required.

2.5 BISTATIC POWER CALCULATION

In the following section, the amount of power required to detect an airborne target using a bistatic radar will be calculated. Before proceeding with the calculation, a bistatic architecture will be proposed, the bistatic radar equation will be derived and the surveillance scenario developed. The results of the calculation will be compared to those of the monostatic radar allowing the advantages and disadvantages of this architecture to be discussed.

2.5.1 INTRODUCTION^[2]

The bistatic radar architecture uses physically separate receive and transmit antennas. A general ground-based bistatic radar is shown in Figure 2.26.

Figure 2.26 A GENERAL BISTATIC CONFIGURATION^[3]

The physical separation between the transmitter and receiver give the bistatic configuration an inherent isolation between the transmitted pulse and the reflected echo. This allows both the transmitter and receiver to be operational simultaneously. Hence, a high duty cycle waveform such as continuous wave may be used.

Again, doppler processing is usually used to detect moving targets. With reference to Figure 2.26, the target's spectra will be doppler shifted by an amount proportional to $\partial L / \partial t$ where $L = L_t + L_r$. Thus, a moving target may be separated in the frequency domain from the direct signal and any of the echos from stationary clutter.

The bistatic radar equation will now be derived. Note that since it is a bistatic configuration, the antennas may have different gains as shown in Figure 2.26. That is, the transmit antenna has gain G_t and the receive antenna has an effective area of A_{eRx} . Hence, the power density received at the target is $G_t P_t / (4\pi L_t^2)$.

The amount of energy reflected by the target in the direction of the receiver is proportional to the bistatic RCS σ_b . Hence, the amount of power received from the target is given by equation 2.13.

$$P_r = [G_t P_t / (4\pi L_t^2)] [\sigma_b] [A_{eRx} / (4\pi L_r^2)] \quad (2.13)$$

Using a method similar to that of the monostatic equation we may derive one version of the bistatic radar equation as shown in equation 2.14. The new terms were defined earlier in Section 2.4. Note that the bistatic radar equation is similar to the link equation for a passive repeater found in communication links.

$$P_t = (4\pi)^2 (KTB_n F_n) (SNR)_t L_t^2 L_r^2 L_{total} / G_t A_{eRx} \sigma_b \quad (2.14)$$

Note that for the L_{total} term, the propagation loss along L_t and L_r may be different. Also, the bistatic RCS is not necessarily the monostatic RCS. If the enclosed angle between the transmitter and receiver beams (θ) is in excess of 165 degrees, a target's RCS is enhanced. For a 180° enclosed angle, forward scatter occurs and the RCS follows equation 2.15 where A is the target area intercepted by the radar beam.^[21]

$$\sigma_f = 4\pi A^2 / \lambda^2 \quad (2.15)$$

It should be noted that even if the target absorbs all the incident microwave power, σ_f will still exist as experimental work has confirmed.^[22] The RCS enhancement of a bistatic configuration should render possible stealth techniques impotent.^[23] For a Type C target of minimum dimension 1m X 1m (nose-on aspect) and a carrier frequency of 1.5 GHz, the enhanced RCS is 63 m² which is a 28 dB gain over the 0.1m² monostatic RCS used in Section 2.4. Note the definite advantage of using higher carrier frequencies if the target is in the forward scatter zone.

2.5.2 BISTATIC ANALYSIS

For any bistatic radar system, energy is lost if the receive and transmit antenna beams are not exactly aligned. If a dual-satellite bistatic system were placed in low earth orbit, beam misalignment would require that approximately 10 dB more power be required than was used in the monostatic case.^[8] This power is clearly not available. Hence, the only possible bistatic implementation requires that one of the path lengths be shortened to compensate for this extra loss. A possible bistatic surveillance system that would satisfy this requirement is shown in Figure 2.27. In this configuration many closely spaced transmitters are placed on the earth and a receiving satellite is placed in geostationary orbit.

Figure 2.27 BISTATIC SYSTEM.

Note that if the distance from the ground station to the fence, L_1 , is kept small, the path loss $L_r^2 \cdot L_t^2$ will be lower than that of the R_{sat}^4 loss for the monostatic radar in low earth orbit. Hence, the power requirement of this configuration will be smaller and more importantly, the space segment is passive resulting in a much lower spaceborne power requirement. Conventional solar arrays should be capable of generating the required power.

The largest advantage of this configuration is the removal of the complex signal processing that the monostatic radar required to remove the clutter interference. The main lobe and side lobe clutter problem which limited the minimum detectable velocity and increased the probability of false alarm in the monostatic radar has been removed due to the fixed nature of the bistatic configuration. Here, a geostationary receiver and

ground transmitter will be at the same position with respect to each other removing the velocity effects discussed earlier for a monostatic radar. Also, the spaceborne antenna subsystem is greatly simplified since no complex pointing system is required.

Other advantages of this configuration are the simplification of the communications with the geostationary satellite and the waveform design. Considering all these advantages, this proposed system will be used in the following analysis.

The targets being detected are the same as for the monostatic case. Again, a Type C target will be the hardest to detect while the Type B target will determine the required re-scan time of the surveillance area.

The surveillance structure will now be described. From basic geometry, the path distance from a geostationary satellite in the equatorial plane at an altitude of 35,000 km to a position on the earth at a given latitude may be derived. Table 2.13 lists the latitude covered and ground to satellite path distance over a range of grazing angles.

Table 2.13 PATH LENGTH vs LATITUDE

Grazing Angle (deg)	Latitude (deg)	Path Length (km)
3	78.2	40,763
5	76.2	40,532
10	71.3	40,031
15	66.5	39,480

Note that we are limited to a grazing angle of 15° due to the enhanced RCS phenomenon which becomes insignificant for β under 165° . Regretfully, the latitudes for good forward enhancement correspond to the small grazing angles which are unsupportable on the ground in the Northern hemisphere due to the Arctic Ocean. Hence, the analysis that follows is based on a 15° grazing angle and a 3,190 km long fence along the Arctic Circle which is mainly on the Canadian mainland.

The structure of the ground station transmitters is shown in Figure 2.28. Regretfully, the target may enter under the ground station's line of sight if the radar is located at low altitude. By raising the ground radars onto 100 m towers or similar high ground, the line of sight coverage is extended to 43 km (based on an effective earth radius of $4/3$ normal) allowing each station to cover a 50 km section of the fence. Eighty ground stations, separated every 40 km, will be required to complete the fence which is a minimum of 3,200 km wide.

Figure 2.28 GROUND STATION COVERAGE

A 1,200 knot Type B target will take a maximum of 6.6 km / 651 m/s or 10.1 seconds to penetrate the fence. Assuming a $\cos^2 x$ weighting, a 2.44 m high by 2 m wide ground station antenna will provide an 8.3° (Θ_{az}) by 6.8° (Θ_{el}) beamwidth at 1.5 GHz. This will cover a sector 50 km long by 3 km in altitude as shown in Figure 2.28. Ten scans of one second duration would be required for the antenna to scan the 62.4° arc of coverage allowing the individual beams to overlap at the 2 dB points. This will give a transmitter antenna gain of 30 dB.

The receiver's antenna aperture is limited by the requirement to view the entire 3,190 km long fence as shown in Figure 2.29. Assuming a uniform weighting, a 1,180 m by 2.2 m antenna will produce 3 dB beamwidths of $\Theta_{el} = 0.0086^\circ$ and $\Theta_{az} = 4.6^\circ$ which will cover an area approximately three times the fence's width. This will ease the pointing accuracy and lower the swath loss to 1 dB for the antenna.

Figure 2.29 SATELLITE RECEIVER COVERAGE

The bistatic radar equation will now be applied to this proposed surveillance structure in order to determine the amount of transmitter power required. The bistatic array will be a CW system using doppler processing to filter targets from the clutter centered at DC in the baseband spectra. A continuous wave (CW) waveform was selected as the transmitted waveform since it will lay down the maximum possible energy on a target for any given peak power limitation. As shown in Figure 2.30, the maximum target doppler is proportional to $\partial L/\partial t$ where $L = L_t + L_r$.

Figure 2.30 DOPPLER TOPOLOGY

For a Type C target with a velocity of 651 m/s, $\partial L_r/\partial t$ is 651 m/s (maximum). $\partial L_r/\partial t$ is $-651 \text{ m/s} \cdot \cos 15^\circ$ or -629 m/s . Hence $\partial L/\partial t$ is 22 m/s resulting in a maximum doppler shift of 220 Hz. The doppler band will be only 220 Hz wide. For a one second dwell time, a perfect doppler processor would provide 1Hz resolution or noise bandwidth.

System losses similar to those used in the earlier monostatic calculation will also exist in the bistatic system. However, no squint loss or main beam broadening effects will occur at the receive antenna over the narrow 5° arc that the fence length occupies removing the previous 5.6 dB squint loss. Also, the swath loss is reduced to 1.5 dB and the eclipsing loss vanishes since a CW signal is used while the propagation loss will triple due to the increased atmospheric attenuation along L_t . This results in a total system loss of 11 dB.

The Type C's RCS is assumed to be enhanced by 10 dB due to both a forward scatter enhancement and a tail-on aspect illumination. Note that this is lower than the optimum enhancement that forward scatter should produce.

Using the above assumptions, the variables required for equation 2.14 are listed in Table 2.14.

Table 2.14 BISTATIC EQUATION VARIABLES

$$A_{eRx} = 2600 \text{ m}^2 \quad (1,180 \text{ m} \times 2.2 \text{ m})$$

$$G_{Tx} = 30 \text{ dB} \quad (2 \text{ m} \times 2.44 \text{ m})$$

$$kT_o B_n F_n = 6E-21 \text{ W} \quad (B = 1 \text{ Hz, } T_o = 290^\circ, \text{ and } F_n = 2)$$

$$\sigma_b = 1 \text{ m}^2 \quad (10 \text{ dB enhancement})$$

$$L_{\text{total}} = 11 \text{ dB}$$

Due to the thin fence width, the Type C target will be scanned a minimum of twice during its penetration. Using the tables in Skolnik, a SNR_i of 17.5 dB producing a single scan detection probability of 0.8 and a probability of false alarm rate of $1E-6$ will be chosen. [7] This will result in a cumulative probability of detection of 96% for the Type C target. Substituting this value and the variables of Table 2.14 into equation 2.14, we find:

$$P_i = (4\pi)^2 (6E-21) (10^{1.75}) (41,300^2) (39,480,000^2) (10^{1.1}) / (10^3) (2600) (1) \\ = 685 \text{ W} = P_{\text{ave}}$$

Assuming a 33% power conversion efficiency, approximately 2 kW of power will need to be generated.

The bistatic configuration is seen to have a substantially reduced power requirement for each of its transmitters compared to that of the monostatic radar. More importantly, the high power section will be ground-based. Only a small transmitter at each earth station will be required. Note however, that eighty of the transmitters would be required to construct the fence.

2.5.3 ADVANTAGES and DISADVANTAGES

The advantages and disadvantages of the bistatic radar will now be summarized.

ADVANTAGES

The main advantage is due to the removal of the problems associated with a high speed platform in low earth orbit. There is no need for complex antenna pointing and doppler processing subsystems. Hence, although the satellite will be a complex doppler processor, this is seen as an evolutionary, rather than revolutionary, extension of existing satellite technology.

The average power is lower due to the smaller $L_t^2 \cdot L_R^2$ path loss involved and the high power segment is ground-based allowing existing solar power systems to power the satellite receiver. The lower infrared signature of the satellite will help to decrease its vulnerability. Also, using a geostationary fixed satellite will reduce the number of satellites to one or two lowering the cost of the system.

Communications to the ground transmitters may be accomplished through the geostationary radar platform reducing the overall system cost. Communications with a geostationary satellite is an established area and the link may share the satellite's radar aperture. Note that this communications link is vital to verify that the ground transmitters are still operational.

Solid state modules, with their high reliability, are more appropriate for CW signals than for low duty cycle waveforms. Solid state microwave electronics required for a SBR will be discussed at length in Chapter 3.

DISADVANTAGES

The satellite's large antenna is over one kilometer long. This is over fifty times the size of any antenna ever built for space. The fixed location makes the satellite vulnerable to electronic and physical neutralization. Also, targets flying tangential to the fence may not be detected.

Many ground transmitters, approximately 80, of moderate complexity would be required. They may be destroyed without advance warning of the target's presence due to the forward scatter RCS enhancement method employed. However, a short range monostatic radar at each site would solve this problem.

2.6 SUMMARY and RECOMMENDATIONS

The major difference of the two radar architectures are summarized below in Table 2.15.

Table 2.15 ARCHITECTURE DIFFERENCES

	MONOSTATIC	BISTATIC
Power Generation Requirement	69 kW*	2 kW**
Number of Satellites	~20	1 or 2
Complex Subsystems	Antenna (pointing and construction) Doppler Processing Power Generation	Antenna (construction)
Coverage	Global	Continental

Of the two architectures studied, the geostationary bistatic structure looks most promising. Canada is fortunate in that its geographic location would allow such a system to be possible. Compared to the monostatic constellation, the bistatic radar should provide a higher performance detection system at a lower cost.

Although the bistatic configuration looks like a viable solution to provide a surveillance fence, it is interesting to note that if 80 independent ground-based monostatic radars spaced every 40 km are used instead, the average power required to detect a 0.1 m² target is of the order of 100 W. It would seem that a ground-based northern fence comprised of many closely spaced radar stations would be superior to any form of space-based system. However, such a system would require a satellite communications link and bistatic receivers placed on the required communication satellites could supplement this system by providing a counter to stealth targets.

* onboard the spacecraft

** at each of eighty ground transmitters

CHAPTER 3

TRANSMIT/RECEIVE MODULES for ACTIVE ARRAY

ANTENNAS

3.1 INTRODUCTION

The majority of airborne radars presently use high power amplifiers based on electron tube technology. These devices, including magnetrons, klystrons and travelling wave tubes, require high voltage power supplies resulting in low system reliability. Mean time between failure rates of ten hours are common in many high performance radars mainly because of the high power amplifier.^[24] Today, intensive research is underway to develop active array antennas to solve this reliability problem while producing more capable systems. Such active arrays may find a use in spaceborne radar as well as airborne radar.

It is widely believed that high performance radar systems of the future will be comprised of active arrays in which the high power amplifier is distributed amongst many transmit/receive(T/R) modules which are placed across the antenna aperture. Each T/R module will incorporate a solid state power amplifier, a low noise amplifier and a phase shifter. Hence, in addition to the high reliability of a solid state power amplifier, the active array will possess other advantages which will be detailed in this chapter.

With reference to Chapter 2, no matter which topology is chosen for a future radar surveillance system, T/R module technology will play a key role whether it be in a monostatic radar or broken into separate transmit-only and receive-only modules for a bistatic radar. This chapter will discuss the theory of active arrays and some applications as well as a proposed architecture for a T/R module using today's technology. However, before this is discussed, basic array antenna theory required to understand active arrays will be summarized.

3.2 ANTENNA ARRAY THEORY

3.2.1 PASSIVE ARRAY THEORY^[9]

In this section, the basic theory of array antennas will be discussed. First, the far-field radiation pattern of a uniformly weighted array will be developed. This will be followed by a description of how array antennas may be electronically aimed or squinted. It will be shown that wide squint angles will result in unwanted grating lobes which produce an interferometer pattern if the antenna is improperly designed. The ability to weight the aperture in order to synthesize almost any beam pattern will also be discussed. Finally, the ability to form multiple beams from the same aperture will be described.

From electromagnetic theory, the vector sum of the electric (E) fields of the individual elements of an array will produce the total electric field. The total E field of an array is given by equation 3.1. Usually, the individual radiating elements are non-directional resulting in a far-field radiation pattern that is dependent on the position of the elements and their excitation which is described by the array factor.

$$E(\text{total}) = E(\text{single element of the array}) * (\text{array factor}) \quad (3.1)$$

The array factor of the uniformly excited N-element linear array of Figure 3.1 will now be derived to illustrate equation 3.1.

Figure 3.1 FAR-FIELD GEOMETRY of an N-ELEMENT LINEAR ARRAY^[9]

Since each element has the same amplitude and a phase shift of β occurs between elements, summing the E fields at infinity results in the array factor as expressed by equation 3.2. Here $Y = k d \sin \Theta + \beta$, k is the propagation constant $2\pi/\lambda$, Θ is the angle off boresight and d is the element spacing.

$$\text{Array Factor} = 1 + e^{j(k d \sin \Theta + \beta)} + e^{j2(k d \sin \Theta + \beta)} + \dots + e^{j(N-1)(k d \sin \Theta + \beta)}$$

$$= \sum_{n=1}^N e^{j(n-1)Y} \quad (3.2)$$

Using the geometric series formula of equation 3.3 and the mathematical identity $e^{jx} - e^{-jx} = 2j \sin x$, we may express the array factor as shown in equation 3.4 which is plotted in Figure 3.2. Note that the radiation is maximum at $Y = 0^\circ, 360^\circ, 720^\circ$, etc. and provided that $Y < \pi$ radians, only one maxima will occur resulting in one main beam in the physical world.

$$\sum_{i=0}^{N-1} x^i = (x^N - 1) / (x - 1) \quad (3.3)$$

$$\text{Array Factor} = (e^{j[(N-1)/2]Y}) \cdot \sin(NY/2) / \sin(Y/2) \quad (3.4)$$

Figure 3.2 ARRAY FACTOR MAGNITUDE - $\sin(NY/2) / \sin(Y/2)$ ^[3]

For a linear phase taper with $\beta = -k d \sin \theta_0$, the beam will be squinted to an angle θ_0 off boresight as illustrated in Figure 3.3. The angle is referred to as the squint angle and is equal to $\arcsin(n \lambda/d)$ where n is an integer.

The array factor amplitude distribution $[\sin(NY/2)/\sin(Y/2)]$ will be the far-field radiation pattern of the array provided that the elemental radiation patterns exist over this range of Y . Note that this distribution may be approximated by $N \text{sinc}(NY/2)$ for small Y . Hence, for the region near $Y = 0^\circ$, the beam is identical to that of a uniformly weighted aperture of similar size and a good approximation to the main beam characteristics of any array are provided by taking the Fourier transform of the spatial amplitude distribution of a similarly weighted continuous aperture.

For a given d , as the number of elements increases, the aperture increases and the main beam narrows increasing the directivity of the antenna. This corresponds to larger N in Figure 3.2. Also note the periodicity of the radiation pattern in Θ -space. This repetition of the main beam every $Y/2 = \pi$ radians is unique to array antennas and will now be discussed in more detail.

Note that $[\sin(NY/2)/\sin(Y/2)]$ has a maximum at $Y/2 = 0, \pi, 2\pi, \text{etc.}$ However, $Y/2 = (k d \sin \theta + \beta)/2$ and assuming $\beta = 0$, we obtain maximum radiation given by equation 3.5 resulting in the squint angles of equation 3.6.

$$\begin{aligned} 0 &= k d \sin \theta / 2 \\ \pi &= k d \sin \theta / 2 \\ 2\pi &= k d \sin \theta / 2 \\ &\text{etc.} \end{aligned} \quad (3.5)$$

$$\begin{aligned}
 \Theta_1 &= 0^\circ \\
 \Theta_2 &= \arcsin(2\pi/kd) = \arcsin(\lambda/d) \\
 \Theta_3 &= \arcsin(4\pi/kd) = \arcsin(2\lambda/d) \\
 &\text{etc.}
 \end{aligned} \tag{3.6}$$

If the elements are spaced too far apart, large d , there will be a maxima before $\Theta_2 = 90^\circ$ (imaginary space), and an interferometer pattern will result. This pattern is usually not required for radars. Hence, to obtain a single main beam along the antenna boresight in real space, a spacing of $d < \lambda$ is used.

If the antenna is squinted off boresight, multiple maxima will occur even if $d < \lambda$. These extra maxima are called grating lobes. For example, if $\beta \neq 0$ and $k d \sin \Theta = -\beta$ at the maxima, the squint angle (Θ_1) is $\arcsin(-\beta/kd)$. For a squint angle of 90° (end-fire operation), $\beta = -kd$ and the first grating lobe will appear at $\gamma/2 = \pi$ corresponding to $\Theta_2 = \arcsin((2\pi - \beta)/kd) = \arcsin((2\pi + kd)/kd)$. For the grating lobe to be confined to imaginary space, $d < \lambda/2$. Hence, for array elements spaced less than one-half wavelength apart, the main beam may be scanned $\pm 90^\circ$ without the appearance of grating lobes.

For the 2-D arrays considered in Chapter 2 for the monostatic radar, where $\Theta_{\text{squint}} = \pm 57^\circ$, an element spacing of 0.6λ will produce a single main beam. The 30m square aperture antenna required for the monostatic SBR would require $(30m / (0.2m \cdot 0.6))^2$ or 62,500 elements at L-band. Clearly it will be important to minimize the unit cost of the T/R modules.

An array antenna's aperture may be amplitude weighted by varying the intensity of individual elements or by varying the spatial density of identical elements. Using the former method, for an array of $2M$ elements with symmetrical amplitude excitation about the origin, the array factor (AF) is expressed in equation 3.7.

$$\begin{aligned}
 (AF)_{2M} &= a_1 e^{j(kd/2)\sin\Theta} + a_2 e^{j(3kd/2)\sin\Theta} + \dots + a_M e^{j((2M-1)kd/2)\sin\Theta} \\
 &+ a_1 e^{-j(kd/2)\sin\Theta} + a_2 e^{-j(3kd/2)\sin\Theta} + \dots + a_M e^{-j((2M-1)kd/2)\sin\Theta} \tag{3.7}
 \end{aligned}$$

Since the element excitations a_n are independent, equation 3.7 can not be reduced to the simple form obtained earlier. The most compact form of the equation becomes equation 3.8.

$$(AF)_{2M} = \sum_{n=1}^M a_n \cos(((2N-1)/2)kdsin\Theta) \quad (3.8)$$

This leads to the topic of antenna pattern synthesis since different coefficient structures may be applied to form antennas with different main lobe to side lobe proportions as was previously shown in Table 2.9 of Chapter 2. The ability to control the amplitude and phase of each element in an array gives the designer many more degrees of freedom compared to reflector antennas where aperture weighting is fixed by a rigid space feed. Also, a non-periodic spacing of array elements may be utilized to provide an additional degree of freedom. This has resulted in the highest performance (lowest side lobe) antennas used in radar being arrays.

Array antennas may produce multiple beams by the addition of feed networks. By splitting the same elements' power and summing a combination of many elements into different feed networks with different elemental phase and amplitude weights, many differently shaped beams may be simultaneously scanned in different directions using a single antenna aperture. An example of this is the Butler feed matrix of Figure 3.4 where four similar beams are formed using hybrid power splitters. The beams will be at the angles given in equation 3.9.

Figure 3.4 MULTIBEAM ANTENNA FEED^[5]

$$\begin{aligned}
 \Theta_1 &= \arcsin((3\pi/4)kd) \\
 \Theta_2 &= \arcsin((\pi/4)kd) \\
 \Theta_3 &= \arcsin((-\pi/4)kd) \\
 \Theta_4 &= \arcsin((-\pi/4)kd)
 \end{aligned}
 \tag{3.9}$$

Note that this method of using passive power splitters is usually limited to a few beams due to the loss of power through the feed network. For P beams, each will have $1/P^{\text{th}}$ the power of a single beam.

3.2.2 ACTIVE ARRAY THEORY

The theory and application of active array antennas is a large discipline. References are available on the manufacture of the T/R modules and many volumes of information are available on the optimum use of active arrays at a systems level. This section will briefly discuss the theory while the advantages and disadvantages of an active array antenna will be stressed. The internal architecture of the modules will be discussed later in Section 3.3.

Figure 3.5 shows the simplified block diagram of a transmit/receive (T/R) module. Included in each module is a phase shifter, amplifiers, and T/R switches. The electronic phase shifter allows the phase taper (β) across the array to be rapidly changed and the amplifiers consist of a power amplifier for transmit and a low noise amplifier (LNA) for receive. The T/R switch selects the appropriate amplifier chain.

Figure 3.5 TRANSMIT/RECEIVE MODULE BLOCK DIAGRAM^[8]

If a T/R module is placed at each radiating element, an active array is formed. Figure 3.6 illustrates such an array.

Figure 3.6 ACTIVE ARRAY ANTENNA

The ability of the T/R modules to quickly change the phase contour over the array provides the antenna with inertialess scan. The beam may be re-aimed in milliseconds compared with the seconds of a mechanically gimbaled antenna allowing the radar to operate in a Track-While-Scan mode and apply Sequential Detection.

In Track-While-Scan, a surveillance sector is scanned serially. If a target is detected, the radar may return to the target's old position at a later time, update its track and illuminate the target in the process. It should be stressed that this update may occur at any time(s) during the sector scan. Unlike a mechanical antenna, this update does not sacrifice time to slew and stabilize the aperture. Adaptive techniques allow the radar to return to targets depending on their previous behaviour allowing the radar to make the best possible use of its surveillance time.

Sequential detection is a similar method. If a possible target is found, the radar re-scans that location as many times as necessary to obtain the required probability of detection and probability of false alarm. This allows the radar's finite energy to be directed at targets and not simply spread uniformly into free space. Compared to a fixed scan pattern, in most scenarios, sequential detection halves the required transmitter power. [27]

The freedom to amplitude weight every element in the array by using programmable gain amplifiers in each T/R module allows any aperture weighting to be selected. This may allow a dynamic antenna pattern where high side lobe/gain weightings are used in look-up modes whereas low side lobe/gain weightings are used in look-down modes in order to minimize ground clutter or jammers.

The power amplifier at each element allows the transmit power to be distributed among many small amplifiers. Solid state amplifiers may replace the expensive, low reliability, high voltage tubes. Also, if a few of the many thousands of T/R modules fail, the radar will continue to operate. The antenna/transmitter will degrade gracefully.

The LNAs will set the system noise figure. Plumbing losses from the array face to the exciter/receiver are no longer important since the signal is at an intermediate power level. Also, the elements' amplified output may be split to many beam forming networks without the loss of SNR as described earlier. Antennas looking in over ten directions at once are possible providing a far greater surveillance capability. One form of multiple beams is useful in monopulse tracking. Here, a planar array is segmented into four sub-arrays with their outputs summed and subtracted to perform single-pulse tracking. In addition, the low power level at each LNA, compared with that at the output of a passive antenna's feed network, gives the radar greater resistance to jamming and electromagnetic pulse damage.

The only disadvantages of active arrays are cost and complexity. For an N element array, the cost of the array at least $N \times$ (T/R module cost) and the requirement to control the phase and gain of N modules in real-time in an airborne/spaceborne environment will be quite challenging.

3.2.3 APPLICATIONS^[28]

The many advantages of active array antennas over conventional rotating reflector antennas has led to many applications in the surveillance radar field. Over the last decade, radars have begun to employ active apertures. These include the fully active PAVE PAWS, a land-based radar with UHF T/R modules, AEGIS, a destroyer-based system and the B1-B aircraft's offensive avionics which use digital phase shifters.

Today, much research is underway to develop fully active arrays for ship-based radar in the U.K. and for airborne radar in the U.S.A.. The following sections will concentrate on T/R module technology required for a spaceborne active array. T/R modules with their small size, light weight, high reliability and radiation hardness will be a key part of any SBR.

3.3 T/R MODULE ARCHITECTURE

3.3.1 INTRODUCTION

In this section, a T/R module architecture, using microwave technology available as of 1986, is proposed. Both Monolithic Microwave Integrated Circuits (MMIC) and Microwave Integrated Circuits (MIC) technology will be used in the required microwave circuits in order to obtain the highest performance.

The block diagram of the proposed module is shown on the next page in Figure 3.7. The receive half of the module, without the T/R switches, could be used on the satellite receiver of the geostationary bistatic SBR receiver and the transmit half could form the ground-based transmitter array modules. This module could also be applied to the short range ground-based chain of monostatic radars which would complement any SBR.

Included on the diagram are typical noise figures, gains and one dB compression power levels of the module. A 5W saturated output power per T/R module was chosen for the 62,500 element L-band array required for the monostatic SBR of Chapter 2. The radar will operate in the linear regime of the power amplifier allowing complex waveforms to be transmitted for ECCM or clutter rejection purposes.

The proposed module is composed of the following sub-assemblies:

- Four Bit Phase Shifter
- Low Power Switch
- Solid State Power Amplifier
- High power Transmit/Receive Switch
- Integral Antenna
- Low Noise Receiver

Note that the solid state power amplifier and the low noise amplifier include some of the same circuit elements. The output section of the low noise amplifier and the input section of the solid state power amplifier are identical.

A description of T/R module operation follows. A common low power transmit/receive signal is fed to the module from the exciter/receiver unit (not shown). On transmit, the signal is fed through a 4-bit phase shifter and solid-state power amplifier to an integral antenna. On receive, the return signals are directed to the receiver through the high-power T/R switch, the Low Noise Amplifier (LNA) and the same phase shifter.

Figure 3.7 TRANSCEIVER BLOCK DIAGRAM

(3) SOLID STATE POWER AMPLIFIER

A pictorial view of the T/R module is shown below in Figure 3.8. Note that the small size and simple interfaces will allow the T/R module to be used in applications other than for a SBR. Opportunities for the module exist in beam-hopping communications systems, expendable decoys and drones and conformal phased array antennas.

Figure 3.8 T/R MODULE PICTORIAL

Note the division of the module into hybrid and monolithic circuits as shown in Figure 3.7. The reason that the module is split into these two technologies will be discussed next.

3.3.2 MIC and MMIC GaAs TECHNOLOGY

The module's microwave circuits will utilize Gallium Arsenide (GaAs) FET technology in order to provide the performance required in a space/military application. The physical differences of GaAs with respect to Silicon lead to many performance advantages. GaAs dominates high speed devices due to its low-loss, semi-insulating substrate and its low electron effective mass which creates high mobility. Also, its larger bandgap allows higher ambient temperature operation compared with silicon. Operating temperature ranges are -200 to 200°C and -55°C to 125°C respectively. The partially ionic bonding of a binary compound give GaAs superior radiation resistance. Figure 3.9 (a), (b), (c), show its superior performance under neutron and gamma radiation.^[29] For low earth orbit satellites or electronic circuit operation near nuclear fuel sources, GaAs technology is the clear choice. With this in mind, the module will utilize only GaAs-based MMIC and MIC technology for the microwave circuits.

Hybrid (MIC) circuits make use of well established manufacturing techniques. The circuits are laid out creating rubilith masks which are used to etch or sputter copper or gold transmission lines on a low-loss substrate. The transmission lines are used to interconnect and match discrete active/passive devices. Since the lines are on the order of a wavelength, they may form distributed high Q elements for matching. In space applications, alumina is used as the substrate due to its inherent stiffness and high thermal stability compared with plastic laminates. Discrete devices are wire bonded to the alumina substrate's microstrip transmission lines to form hybrid circuits. Such a circuit is shown in Figure 3.10.^[30] The reader is referred to Chapter 5 for an in depth discussion of MIC design.

Figure 3.9 Radiation Characteristics of GaAs and Silicon

Figure 3.10 4 GHz MIC Amplifier and Equivalent Monolithic Circuit

A typical MMIC chip is shown in Figure 3.11.^[31] All the circuit elements are on a single millimeter-sized GaAs substrate. A photolithographic masking/etching technique similar to that used in Silicon I.C. production is used to define the circuit. The elements include resistors, capacitors, inductors, field-effect transistors (FETs) and diodes. Nine steps are required to form these elements for the process used by Raytheon as shown in Figure 3.12.^[32] Recently, commercial foundrys have become available allowing microwave circuit designers from small organizations to access large scale production facilities for a fee. Foundry services will be described in Chapter 4 along with a detailed description of the MMIC design process.

Figure 3.11 A Typical MMIC Chip

Figure 3.12 A Typical MMIC Fabrication Process

Figure 3.10 contrasts the size of a MIC and MMIC implemented amplifier. The small black rectangle is the MMIC equivalent. The MMIC's millimeter-sized package allows the circuit elements to be treated as lumped elements. Hence, MMIC circuits produced at the same time from a single wafer, are phase and gain matched. This is a major advantage of MMICs in T/R module design since the thousands of T/R modules of an active array must be phase and gain matched in order to be useful. Adjustments of hybrid modules to provide similar matching would be labour intensive and prohibitively expensive.

Comparing the two technologies, both have advantages and disadvantages. Discrete MIC devices possess the highest performance since the substrate for all the MMIC elements is common. Devices that require custom vertical doping such as varactor diodes and High Electron Mobility Transistors (HEMTs) can not be integrated at the present time. Also, GaAs is a poor thermal conductor making MMICs power limited compared to MIC circuits which may use microstrip power combiners to step-up the power levels. Also, hybrid technology is proven while MMIC circuits still suffer from processing inconsistencies and unrefined modeling software. On the other hand, MMICs are small, light, and cheap in large quantities making them ideal for the large arrays. Hence, the T/R module proposed next uses a combination of both technologies to form a Quasi-MMIC circuit where the best of both technologies is adopted. Hybrid MICs will be used for the low noise and high power sections as shown in Figure 3.7 while all the other circuitry will be MMIC.

MMIC LNAs give inferior noise performance compared to hybrid MICs using discrete FETs due to the higher input matching network losses of MMICs caused by their lumped elements' lower Qs. MMIC Qs of 20 compare poorly to microstrip distributed elements' Qs of 200. These lossy elements limit MMIC noise figures to 3 dB at present whereas discrete HEMT devices, which have unequalled noise figures, are now available. Using these, noise figures of 1 dB for the LNA are possible. Note that only the first stage of the LNA will be hybrid. Once intermediate power levels are achieved, MMICs will be used for further amplification.

Power stages in MMICs are presently limited to under 0.5 W/mm of gate periphery. Also, the R.F. output power is limited to under 2W due to thermal limits. High power MMICs require the substrates be thinned from the normal 250 μ m to 50 to 100 μ m. This creates higher transmission line losses and, hence, gain is sacrificed for power.

Also, the extra processing required to thin the substrate and the long gate periphery of power FETs decreases yield. Therefore, for the power amplifier stages, a hybrid amplifier will be used.

The proposed layout for the transceiver module is shown in Figure 3.13. Note that all GaAs chips are attached to an alumina motherboard and interconnected using conventional hybrid bonding techniques. Although the power amplifier and first stage of the LNA are discrete, they will not dramatically increase the size of the module due to the requirement for control and bias circuitry as well as an antenna which are fixed in size.

Figure 3.13 Physical Layout

3.3.3 SUB-ASSEMBLY DESIGN

The remainder of this section will describe the six individual units (sub-assemblies) of the transceiver as broken down in Figure 3.7. Possible circuit topologies for the units will now be discussed

1. Phase Shifter

Several types of phase shifting techniques have been used in MMICs. The most common methods are listed below:

- i) Switched Filter
- ii) Switched Line
- iii) Reflection
- iv) Vector Modulator

Switched filter phase shifters make use of the differential phase shift between high and low pass filters. For example, at high frequencies, a RC circuit produces a -90° phase shift whereas a CR series circuit gives a 0° phase shift. Therefore, switching between these filters will form a 90° phase shifter. This example has unacceptable differential amplitude. However, by operating at the -3 dB frequency of both circuits where each have a $|45^\circ|$ phase shift, a 90° low differential amplitude phase shifter may be realised.

In actual circuits, ' π ' and ' τ ' type filters are used. Figure 3.14 shows a three element ' τ ' high pass/low pass filter producing a 45° phase shifter.^[33] This design incorporates the 'off' capacitance of the FET into the filter resulting in broadband operation. Although compact, this topology has high insertion loss and requires the parasitic capacitance of the FETs to be known and reproducible.

Figure 3.14 45° Switched Filter Phase Shifter

The switched line phase shifter switches series transmission lines of different physical length in and out of the signal path. Hence, a set of phase shifts corresponding to the different time delays is formed.

Although simple in theory, the fixed delay lines require a large GaAs area. For example, a 4-bit, X-band phase shifter developed at Raytheon required a chip 6.4 mm x 7.9 mm. Since the yields of chips larger than 3 mm square are low resulting in higher processing costs, an alternative approach where the required delay line selections are 'off chip' and the transmission line selection switches are integrated has been constructed. This is shown in Figure 3.15^[4] where the squares are the input/output bonding pads used to connect the MMIC switches to microstrip delay lines on an alumina substrate. The disadvantages of this design are the limited bandwidth and large size.

Figure 3.15 4-Bit 360° Switched Line Phase Shifter

A variation of the switched line phase shifter is the reflection-type phase shifter. The reflection-type phase shifter makes use of a 3 dB coupler feeding two transmission lines which have their terminations switched between open and short circuits as shown in Figure 3.16.^[34]

Figure 3.16 90° Reflection Type Phase Shifter

Although a savings of area is realized by the two-way path in reflection phase shifters, the phase shift path requires a large area. For example, a C-band 6-bit phase shifter developed by ITT required a 9.4 mm x 4.2 mm chip. Again, the bandwidth is limited.

The vector modulator phase shifter is an active configuration. Although complex, it is more tolerant to processing than the previous passive phase shifters which require the switch parasitics to be quantized.

A block diagram of a vector modulator is shown in Figure 3.17. Here, the signal is broken into inphase (I) and quadrature (Q) components which are amplitude weighted to produce a unity magnitude vector at any angle from 0 to 90°. By adding additional 0/180° and 0/90° splitters to the vector modulator, a complete 0 to 360° phase shifter is formed. This is shown in block form in Figure 3.18(a) and its schematic is shown in Figure 3.18(b).^[35] GE built this 5-bit C-band phase shifter using segmented-dual-gate MESFETs to amplitude weight the I and Q channels. The chip is only 2.5 mm square and the resulting phase shifter is based on device ratios resulting in high yields independent of processing variations. Bandwidth is limited to an octave by the phase splitters.

Figure 3.17 Vector Modulator Block Diagram

Figure 3.18 Active Phase Shifter a) Block Diagram b) Schematic

Of the topologies possible, the vector modulator looks most promising. Its small size is not dependent on the number of bits of resolution required. Resolution may be easily increased from four bits if an application requires higher accuracy. However, this phase shifter is unilateral. In order to be used in the transceiver, a DPDT switch will be required to replace the SPDT low power switch in Figure 3.7. This will allow the connections to the phase shifter to be reversed during receiver operation.

2

MICRO

2. Low Power Switch

A single D-MESFET with passive matching may be used as a SPDT switch at low power levels. Figure 3.19 shows the equivalent circuit of a FET.^[36] With no gate bias, the drain-to-source resistance is determined by the channel's bulk resistance. The device is 'on'. With a gate bias greater than pinchoff, the channel is pinched off and the drain to source impedance is limited by the drain-to-source parasitic capacitance. The device is 'off'. At L-band, the on-to-off impedance ratio may exceed 5,000 allowing a single FET to form a useful switch at low power levels.

Figure 3.19 A GaAs FET Switch

Compared with PIN switches, GaAs FETs are faster, consume less power and are easily integrateable. The high speed properties of GaAs allow nanosecond switching times. The reverse biased gate limits bias current to negligible levels for either switch state. Also, the small size of FETs and their three terminals allow complex switching networks to be formed on MMICs.

The most common switch topologies are shown in Figure 3.20.^[37] A series or shunt configuration is selected depending on whether high isolation or low insertion loss is the primary requirement. Since the dual-gate switch is not bilateral, it is of limited use for transceiver circuits. The matched switch is a tradeoff of greater circuit complexity and higher insertion loss for increased bandwidth.

Figure 3.20 Various FET Switch Configurations

The circuit in Figure 3.21 is a high isolation, large bandwidth DPDT switch.^[36] This topology would allow a vector modulator to become bilateral, as the module requires, while providing excellent isolation.

Figure 3.21 High Isolation DPDT Switch Schematic

3. Solid State Power Amplifier

As previously stated, the power amplifier was broken into MMIC and hybrid sections. This was due to the limited power handling capability of MMICs compared with discrete FETs. Figure 3.22 clearly shows how discrete FETs (solid line) outperform MMICs (symbols) at C-band.^[38] Also note that the MMICs that operate near the discrete FET line are very large (over 30 mm²) and are impractical to develop using commercial foundrys. A more subtle effect limiting MMICs is the non-linear behaviour of the bulk GaAs bias resistors. For high power linear operation, thin film NiCr bias resistors are required. These require another processing step which no foundry offers at present.

Figure 3.22 MMIC Output Power vs Frequency

The proposed configuration shown in Figure 3.7 makes use of the LNA's final stage MMIC for the power amplifier pre-driver. This level of operation is possible for foundry produced MMICs. However, if non-linear effects are encountered, several MMICs may be combined using hybrid couplers on the alumina motherboard. This is represented by the thick dotted lines on Figure 3.7.

Four possible MMIC amplifier topologies are travelling wave, feedback, lossy reactively matched and actively matched. They will now be discussed.

A travelling wave or distributed amplifier is shown in Figure 3.23.^[39] In this circuit, the FET's parasitic capacitances are absorbed by the gate and drain lines to form pseudo transmission lines. This topology, the simplest, gives superior bandwidth (lowest gain ripple) and the highest power capability due to the wide separation of heat producing power FETs. However, gain is traded for bandwidth requiring many stages to be cascaded. For radar applications, where a small number of narrowband, high gain stages are preferable, this topology becomes inadequate.

Figure 3.23 Travelling Wave Amplifier Schematic

MMIC amplifiers make use of shunt feedback to attain wideband, low noise and low VSWR operation. Figure 3.24 shows the schematic of a feedback amplifier.^[37] Here, a drain-to-gate inductor is used to compensate for the drain-to-gate parasitic capacitance which limits the high frequency roll off. Hence, low gain, wideband feedback amplifiers are possible. A possible configuration for the medium power MMIC stages could make use of many simple feedback amplifiers connected with narrow band hybrid filters in order to minimize noise and provide interstage isolation.

Figure 3.24 Two-Stage S-Band Feedback Amplifier

The lossy reactive matched topology makes use of the most common method of solid state amplifier design. Figure 3.25 shows a three stage X-band amplifier.^[37] Here, shunt and series reactive components provide the input/output and interstage matching. This topology yields the highest gain per stage, low noise and narrow bandwidth. However, processing tolerances can shift the centre frequency of operation a few hundred MHz at C-band as will be seen in Chapter 4.

Figure 3.25 3-Stage Reactively Matched Amplifier

Active matching techniques may also be used in MMIC amplifiers. Here, active GaAs FETs replace passive L-C matching networks. Figure 3.26 shows a common gate / common source / source follower amplifier.^[37] The common gate and source follower stages provide a low VSWR match to 50Ω while the common source stage provides most of the gain. This technique requires the least area for the input/output matching circuitry at the lower microwave frequencies.

Figure 3.26 Actively Matched Amplifier

The higher power stages make use of hybrid technology. The two-stage design with single stage gains of 9 dB shown in Figure 3.7 is possible using 'off the shelf' components. In order to improve the efficiency of the power amplifier, different classes of operation may be used. Class A amplifiers are limited to an efficiency of 40% while class AB or class B topologies promise higher efficiency. An X-band push-pull amplifier is shown in Figure 3.27.^[40] Although complex, it has a high efficiency and is able to drive balanced loads such as a dipole antenna directly.

Figure 3.27 4-Stage Push-Pull Power Amplifier

4. High Power Transmit/Receive Switch

As shown in Figure 3.7, this switch must handle a high RF power level from the power amplifier to the antenna with a low insertion loss. The T/R switch cannot be monolithic following the same reasoning behind the power amplifier final stages being hybrid.

Figure 3.28 shows a high power SPDT T/R switch topology.^[41] When transmitting, the receiver is protected by turning on the lower FET creating a short circuit. This short is transformed through a $\lambda/4$ transmission line of ' Z_2 ' characteristic impedance to become an open circuit at the antenna port. ' Z_2 ' is chosen in order to reduce the short circuit current and hence the size of the lower power FET. ' Z_1 ' is chosen to allow maximum power flow in the transmit path without turning the upper FET on. A shunt inductor may be used to resonate out the parasitic source-to-drain capacitance in order to increase the isolation.

It should be noted that a passive circulator may remove the need for a switch depending on the antenna's VSWR and the capability of the limiter to protect the LNA.

Figure 3.28 High Power Switch

5. Antenna

The module includes an integral antenna. Including a matched antenna in the module will achieve the highest overall performance. For use in phased arrays requiring a large scan angle, the antenna's pattern should be as non-directional as possible. Many antenna topologies are available. Both horn antennas milled out of the enclosure and microstrip antennas formed on an extension of the alumina motherboard are possible.

The microstrip dipole antenna shown in Figure 3.29 looks promising. This is a bow-tie antenna constructed from metallizing one side of an alumina substrate that may be an extension of the alumina motherboard with the bottom ground plane etched away. The radiation pattern from this antenna is shown in Figure 3.30 (a) and (b).^[42]

Figure 3.29 Bow-Tie Antenna

Figure 3.30 Bow-Tie Radiation Pattern a) E-Plane b) H-Plane

6. Low Noise Receiver

The first stage of the receive chain will probably be a limiter. Shunt GaAs diodes which turn on at high power levels are a common limiter configuration. Figure 3.31 shows the simple schematic of a commercial limiter module and its equivalent circuit.^[43] The possible removal of the limiter allowing the LNA to saturate and thereby limit is an option. In either case, the effect of saturation on LNA recovery time and possible accelerated failure should be considered in the limiter design.

Figure 3.31 Limiter Schematic and Equivalent Circuit

The first stage of the LNA will be a hybrid circuit utilizing discrete FETs. As already stated in Section 3.3.2, HEMTs are now available and they seem to outperform normal low-noise MESFETs. LNA circuits using HEMTs will be discussed in Chapter 5.

The intermediate power amplifier to follow the hybrid LNA stage will be the same used in the power amplifier pre-driver as was described earlier. By using the same amplifier chip in both transmit and receive chains, the MMIC design will be focused on fewer circuits and production costs will decrease.

3.4 CONCLUSION

In this chapter, the active array antenna, one of the more recent advances in radar hardware, was described. Although the cost and complexity of an active array has limited its use in current systems, new advances in Monolithic Microwave Integrated Circuits and Very High Speed Integrated Circuits technology, respectively, seem capable of solving these problems.

The development of the T/R modules described in section 3.3 would allow the development of active array radars with excellent performance. Time and financial limitations prevented an entire T/R module from being constructed in this thesis. However, key parts of the module formed the basis for the experimental work discussed in the following chapters. Chapter 4 will concentrate on an actively matched C-band intermediate power amplifier/switch using MMIC technology while Chapter 5 will compare HEMT and MESFET devices for use in LNAs using MIC technology.

CHAPTER 4

MONOLITHIC MICROWAVE INTEGRATED CIRCUIT

DEVELOPMENT

4.1 INTRODUCTION

This chapter is the first of two describing experimental work in which the design and testing of two MMIC chips will be detailed. One design is an amplifier/switch and the other is a similar amplifier requiring only a single bias voltage. Both are two-stage designs using a combination of active and reactive matching techniques. These circuits could find a use in the intermediate-power amplifier of the T/R module discussed in Chapter 3.

The Adams-Russell foundry, located in Boston, Mass., manufactured the circuits while the Communications Research Center provided the facilities required to design and test the microwave circuits. Since they provided access to the foundry, the author was restricted to designs operating in the communications section of C-band. However, the results from this experimental work are relevant to an L-band T/R module design.

This chapter begins with a discussion of GaAs processing technology which is required to fully exploit MMIC technology. This will be followed by the theory required to design a microwave amplifier. A brief discussion of the MMIC design process will precede the actual design of two circuits developed in this chapter. Finally, the results of testing these devices and comparing them with theory will be detailed.

4.2 MMIC PROCESSING TECHNOLOGY

4.2.1 SURFACE GaAs PROCESSING^[44]

GaAs has lagged behind the development of Silicon for integrated circuits due mainly to the lack of a stable native oxide. Although more expensive processing is required to solve this problem, GaAs is now used for the high performance/speed devices in digital and analog circuits. This section will describe the processing required to produce analog MMICs.

In the last decade, semi-insulating bulk GaAs has become available in commercial quantities. The Liquid-Encapsulated Czochralski method is used to produce cylinders of bulk GaAs which are sliced into thin wafers two or three inches in diameter. These wafers form the substrate onto which circuits are defined using photolithography.

Due to its larger bandgap, pure GaAs has a resistivity approximately 2,000 times higher than for pure Silicon.^[45] By adding a trap producing dopant such as Chromium, the GaAs becomes semi-insulating (S.I.). This property minimizes leakage currents through the substrate to ground allowing low-loss microwave circuits to be constructed.

MMIC planar processing is based on the vertical doping of a S.I. GaAs substrate and the formation of both Schottky and ohmic contacts at the interface. Many steps are required to process a MMIC circuit. The steps used by the Adams-Russell foundry are shown in Figure 4.1.^[44] These steps will now be explained.

The S.I. wafer forms the starting material onto which active layers are deposited. Epitaxial deposition or ion implantation is used to inject elemental Si into the GaAs lattice. The Si atoms replace Ga to form an n-type, conducting layer within the insulating substrate^[46] which is referred to as the active layer. In order to form low-loss, ohmic (non-rectifying) contacts on the circuit at a later time, areas are heavily doped to produce N^+ layers. These steps are illustrated in Figure 4.1 as step 1 (Alignment) and step 2 (Contact).

To pattern the contact areas and any other areas, the resist/liftoff method is used. A metallization definition step is illustrated in Figure 4.2.^[44] Here, an organic resist is deposited uniformly over the wafer and a mask, a metal pattern on a low thermal expansion glass, is placed above the wafer. Any holes in the mask will transmit

Figure 4.1 WAFER FABRICATION [44]

Figure 4.2 RESIST/LIFTOFF PROCESSING [44]

ultra-violet light to expose the resist which is removed with a selective etchant. Following this, the metal layer is deposited through the windows in the resist onto the bare GaAs. The remaining (masking) resist is then etched away with a different solvent. The metal on top of the resist is "lifted off" at the same time leaving the required metal pattern. For the N^+ contact step, the metal would be replaced with a beam of Si^+ ions which penetrate the uncovered GaAs surface to further dope the active layer.

In order to separate circuit elements from each other, the active layer is etched away as shown in step 3 (Isolation) of Figure 4.1. All elements requiring an active layer are formed on the remaining "high ground" which is referred to as a "mesa".

Step 4 (Ohmic) shows the formation of ohmic contacts. Here, ohmic metal, a Ni:Ge:Au:Ni alloy, is deposited over the N^+ doped region and alloyed at a high temperature. The metal/GaAs interface becomes smeared creating a low resistance connection.

Step 5 (Gate) shows the formation of a rectifying contact. Gate metal, layers of Ti:Pt:Au:Pt, is deposited on top of the active GaAs layer forming a Ti:GaAs Schottky diode which is useful for the gate of a MESFET or the anode of a diode. The gold layer creates a low-loss transmission line and the Pt buffer layer prevents the gold from diffusing into the substrate at elevated operating temperatures. The gate metal may also form the first level of interconnect lines to other elements and the bottom plate of capacitors.

Step 6 (Passivation) shows the passivation layer covering the bare GaAs. This is composed of the polymer polyimide which forms a low dielectric layer ($\epsilon_r = 3.5$). This cap prevents the As from diffusing out of the active layer to be replaced with impurities which would slowly degrade the device over time while also protecting the circuit from physical damage. Polyimide is used to minimize the stray surface capacitances inside the elements.

Step 7 (Nitride) is the formation of the dielectric used for the capacitors. Si_3N_4 is used to form an $\epsilon_r = 6.8$ dielectric with a high voltage breakdown resistance.

Step 8 (Bridge) is the formation of thick resist patterns to support the next metal layer. After this metal layer is deposited on top, the resist is etched away leaving an airbridge. Hence, the resist allows the vertical patterning of interconnect lines which are useful for low capacitance crossovers found in inductors, MESFETs and interconnect lines.

The Step 9 (Metal), is the deposition of the second metallization layer. This gold layer is patterned to interconnect the various circuit elements. It also forms the source and drain contact pads and the capacitor top plates.

The final step (not shown) is the thinning of the substrate to a thickness of 250 μ m and the metalizing of its underside. This places the ground plane the correct distance from the surface to form the required microstrip transmission lines.

The resulting circuit schematic is shown at the bottom of Figure 4.1 as a FET and a capacitor in series.

4.2.2 CIRCUIT ELEMENTS

The circuit elements produced using MMIC processing include resistors, capacitors, inductors, MESFETs and diodes. These elements will now be discussed in depth.

1. Resistors may be formed by bounding a region of n or n⁺ GaAs with ohmic contacts. The higher doping is used for the lower resistance values. Values of resistance from 33 Ω to 10 k Ω were used in the following designs. A scanning electron micrograph (SEM) of a 368 Ω resistor used in Design #2 is shown in Figure 4.3. Note the ohmic contacts at the ends of the resistor which are rough due to alloying.

Figure 4.3 A 368 Ω RESISTOR

Resistor dimensions determine their value. Since the resistivity of the GaAs is expressed in ohms per square (R_{Ω}), the resistance is simply the number of squares (the length divided by the width) multiplied by R_{Ω} plus twice the contact resistance.

Bulk GaAs is only ohmic for low electric fields. Its electric field - velocity profile is shown in Figure 4.4. At high electric fields, above 200 V/mm, the electron velocity saturates and the resistance no longer obeys the earlier equation. A correction factor is applied in the design of short bias resistors. However, for high RF power applications, bulk resistors create non-linear effects which are unacceptable in most power amplifiers. Thin film NiCr resistors requiring an additional processing step are substituted in these cases.^[37]

Figure 4.4 ELECTRON VELOCITY vs ELECTRIC FIELD^[37]

The Adams-Russell foundry provides a cell library of standard resistance values. All resistors are bulk GaAs and the standard cell shapes are tailored to form any required value. The equivalent circuit model used in the following designs is a simple series resistor which neglects the substrate leakage currents.

2. Capacitors, as shown in Figure 4.1, are of metal-insulator-metal construction. These are simply parallel plate capacitors where gate metal forms the lower plate, Si_3N_4 forms the dielectric and the metal layer forms the upper plate. The nitride dielectric layer is of uniform thickness so the capacitance values are linearly related to the surface area of the plates. Figure 4.5 shows a SEM of two 0.4 pF tuning capacitors in series which are used in Design #2. Note the airbridges which are used to jump over the dielectric's edges in order to minimize the possibility of a short with the lower plate.^[47] A piece of dust is visible on the left capacitor.

Figure 4.5 0.42 pF CAPACITORS

Again, the Adams-Russell foundry provides a cell library of standard capacitor values. At C-band, these capacitors are limited to 20 pF before they resonate. However, finite real estate in the following designs limited the values to 5 pF for the bias bypass capacitors. Device Qs of 20 and 30 were originally used for the large bypass capacitors and the small tuning capacitors, respectively.

3. Inductors are constructed from high impedance transmission lines. As they are made longer, their inductance increases until they resonate at a length of $\lambda/4$. Above $\lambda/4$, they appear capacitive. In order to fit large inductors on the chip, they are wound up into octagonal shapes as shown in the SEM of Figure 4.6. Transmission lines of width $6\text{ }\mu\text{m}$ (thin) and $10\text{ }\mu\text{m}$ (thick) are available. The wider lines are used where high bias currents flow. This figure shows a 4 nH , six-turn, thin choke. Note how the center is fed with a gate metal line and the spirals use airbridges to jump over it.

Figure 4.6 SIX-TURN THIN OCTOGONAL INDUCTOR

Adams-Russell provides a cell library of n -turn (n is an integer) inductors. The closest oversized inductor is chosen and coils are peeled back to form the required value. The equivalent circuit of a spiral inductor is shown in Figure 4.7. This circuit accounts for the parasitic capacitance to ground, adjacent winding capacitance and transmission line resistance. Above resonance, the coil will behave like a capacitor which limits the maximum size of the thin inductors to approximately six turns at C-band.^[48]

Figure 4.7 SPIRAL INDUCTOR EQUIVALENT MODEL

For a FET structure, Grove has found that the drain current follows equation 4.1. Here V_i is the drain-to-source voltage across the gate region, V_{Bi} is the Schottky barrier height, and $g_0 = q\mu N_D W A / L_g$ and $V_{po} = qN_D A^2 / 2\epsilon_r \epsilon_0$ where W is the gate width, $\epsilon_r \epsilon_0$ is the substrate permittivity, μ is the electron mobility, N_D is the active layer doping density and A is the active layer depth.[50]

$$I_{ds} = g_0 [V_i - 2[(V_i + V_{Bi} - V_g)^{3/2} - (V_{Bi} - V_g)^{3/2}] / 3V_{po}^{1/2}] \quad (4.1)$$

The small signal gain of the FET is dependent on the transconductance, g_m , which is described by equation 4.2.

$$g_m = \partial I_{ds} / \partial V_g = g_0 [(V_i + V_{Bi} - V_g)^{1/2} - (V_{Bi} - V_g)^{1/2}] / V_{po}^{1/2} \quad (4.2)$$

Overlaid on Figure 4.8 are the internal model elements of a FET. This leads to the equivalent model for a FET shown in the schematic of Figure 4.10 (a). The values of these elements for the Adams-Russell FET with a gate 300 μm wide and 0.7 μm long operated at $I_{dss} / 2$ and $I_{dss} / 5$ with a 4 V drain bias are listed in Figure 4.10 (b).[44]

Figure 4.10 MESFET MODEL

a) Schematic

b) A-R 300 μm FET Element Values

	0.2 I_{dss}	0.5 I_{dss}	Units
Rg	0.7	0.7	ohms
Ri	13.3	10.0	ohms
Rs	2.3	2.3	ohms
Rd	4.4	4.4	ohms
Rds	533	533	ohms
Cgs	0.306	0.383	pF
Cdg	0.028	0.021	pF
Cds	0.061	0.069	pF
gm	0.024	0.028	S
tau	4	4	psec
Lg	0.050	0.050	nH
Ls	0.030	0.030	nH
Ld	0.035	0.035	nH

With reference to the small signal model of Figure 4.10, the cutoff frequency (f_T) at which the current I_g is equal to I_{ds} is $g_m / 2\pi C_{gs}$ where C_{gs} is the

gate-to-source capacitance. The device is usually operated well below f_T . Note from equation 4.2 that if the gate length, L_g , is decreased, C_{gs} will decrease while the transconductance and f_T will increase extending the useful frequency regime of the FET. For the 300 μm Adams-Russell FET, f_T was calculated to be 12 GHz which limits amplifier designs to below X-band. Other foundries offer 0.25 μm FETs which allow MMICs to operate up to 20 GHz. Soon, HEMT processing should extend the frequency regime of MMICs performing relatively complex functions to 30 GHz.

A SEM of a 300 μm Adams-Russell MESFET is shown in Figure 4.11. This figure shows a compact interdigitated structure where three source contacts and two drain contacts are separated by four 75 μm wide X 0.7 μm long gates. The source pads jump over the gate bus using airbridges. Note the rough surface on the source and drain which is caused by alloying the ohmic contacts. The polyimide passivation layer covering the channels is also visible.

Figure 4.11 A 300 μm MESFET

Adams-Russell provides a cell library of several FETs including 'T' and interdigitated structures with gate widths from 150 μm to 1200 μm . The equivalent models for the interdigitated structures are also supplied. The FETs used in the following designs were both 300 μm interdigitated structures.

4.3 MICROWAVE AMPLIFIER DESIGN THEORY

A brief review of microwave amplifier design theory required for the following MMIC designs follows. It is strongly based on reference 51.

4.3.1 S-PARAMETER BASICS^[51]

At microwave frequencies, circuits are characterized by scattering (s) - parameters. Low frequency H, Y and Z parameters require open and short circuits in order for currents and voltages to be measured. At high frequencies, opens and shorts are hard to define, voltmeters and ammeters are unavailable and some active devices will oscillate when terminated in these loads. Hence, s-parameters which utilize transmission lines of characteristic impedance Z_0 , usually 50 Ω , to terminate the ports of the device under test, are used.

For any two-port network with its input and output terminated in characteristic impedance Z_0 , the magnitude and phase of the incident and reflected voltage waves may be measured. These measurements may be converted to a set of s-parameters which characterize the two-port. Figure 4.12 shows the voltage waves.

Figure 4.12 INCIDENT and REFLECTED VOLTAGE WAVEFORMS

The voltage waves may be normalized according to equation 4.3 as shown in Figure 4.13. Note that $|a|^2$ and $|b|^2$ are power units. Hence, the magnitudes of a_1 and b_1 are easily determined using power measurements.

$$a_1 = E_{i1}/\sqrt{Z_0}, \quad a_2 = E_{i2}/\sqrt{Z_0}, \quad b_1 = E_{r1}/\sqrt{Z_0}, \quad b_2 = E_{r2}/\sqrt{Z_0} \quad (4.3)$$

Figure 4.13 NORMALIZED VOLTAGE WAVES

Any two-port may be characterized by its s-parameters which are defined by equation 4.4. These s-parameters are found by selectively terminating the input and output transmission lines in Z_0 and measuring the complex quantities a_i and b_i . For example, if the output transmission line is terminated in Z_0 , no reflection will occur at the load, i.e. $a_2 = 0$, and $s_{11} = b_1 / a_1$.

$$\begin{aligned} b_1 &= s_{11}a_1 + s_{12}a_2 \\ b_2 &= s_{21}a_1 + s_{22}a_2 \end{aligned} \quad (4.4)$$

Today, vector network analysers are available which determine s-parameters for frequencies up to 26.5 GHz (see Section 4.6.2). Here, voltage waves are injected into the ports of a device and the reflected and transmitted waves are monitored. Both the magnitude ratio and phase differences of the voltage waves are determined using power meters and phase detectors, respectively. A built-in processor computes the four s-parameters automatically. Only circuits which become unstable when "seeing" Z_0 cannot have their s-parameters determined directly.

4.3.2 AMPLIFIER DESIGN^[51]

Once the s-parameters of a two-port are known, the power gain of the device may be determined. Figure 4.13 may be re-written as a flow graph as shown in Figure 4.14. Note that Γ_l and Γ_s are the load and source reflection coefficients defined by equation 4.5. Applying Mason's gain rule, we may derive equation 4.6, and since the transducer power gain G_T is defined as the power delivered to the load divided by the power available from the source it is expressed by equation 4.7.

$$\Gamma_l = (Z_l - Z_0) / (Z_l + Z_0); \quad \Gamma_s = (Z_s - Z_0) / (Z_s + Z_0) \quad (4.5)$$

$$b_2 / b_s = s_{21} / (1 - s_{11}\Gamma_s - s_{22}\Gamma_l - s_{21}s_{12}\Gamma_l\Gamma_s + s_{11}\Gamma_s s_{22}\Gamma_l) \quad (4.6)$$

$$G_T = [|s_{21}|^2 (1 - |\Gamma_s|^2)(1 - |\Gamma_l|^2)] / [(1 - s_{11}\Gamma_s)(1 - s_{22}\Gamma_l) - s_{21}s_{12}\Gamma_l\Gamma_s]^2 \quad (4.7)$$

Figure 4.14 TWO-PORT FLOW GRAPH

In the following designs, the first step toward maximizing the gain is to let $\Gamma_s = s_{11}^*$ and $\Gamma_l = s_{22}^*$, where $*$ is the complex conjugate, for each amplifier stage. This follows from the equation for G_T assuming unilateral devices, ie. $s_{12} = 0$. The Super Compact microwave computer aided design program which considers $s_{12} \neq 0$ is then applied. This program also accounts for the complex interactions of many stages and the element parasitics which would overwhelm a manual design.

Stability is also an important consideration in amplifier design. Assuming that $\Gamma_s = s_{11}^*$ and $\Gamma_l = s_{22}^*$ we may define the K_S -factor by equation 4.8. For $K_S > 1$, the amplifier will be stable provided Γ_l and Γ_s are less than one (passive terminations).

$$K_S = (1 + |s_{11}s_{22} - s_{12}s_{21}|^2 - |s_{11}|^2 - |s_{22}|^2) / 2|s_{12}s_{21}| \quad (4.8)$$

Stability circles, the $K_S = 1$ contours for different frequencies, are plotted on a Smith chart using the relations $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$ which are defined in equation 4.9.

$$\begin{aligned} \Gamma_{in} &= b_1 / a_1 = s_{11} + s_{21}s_{12}\Gamma_l / (1 - s_{22}\Gamma_l) \\ \Gamma_{out} &= b_2 / a_2 = s_{22} + s_{21}s_{12}\Gamma_s / (1 - s_{11}\Gamma_s) \end{aligned} \quad (4.9)$$

Equating the first part of equation 4.9 to one, we find the load stability circles of radius and center shown in equation 4.10. Similar equations may be derived for the source. The designer must keep the actual load and source impedances a safe distance away from the unstable regions. In practice, Super Compact is used to compute and plot these circles.

$$\begin{aligned} \text{radius} &= |s_{12}s_{21} / (|s_{22}|^2 - |\Delta|^2)| \\ \text{center} &= (s_{22} - \Delta s_{11}^*)^* / (|s_{22}|^2 - |\Delta|^2) ; \quad \Delta = s_{11}s_{22} - s_{12}s_{21} \end{aligned} \quad (4.10)$$

4.3.3 SUPER COMPACT REVIEW^[52]

Some of the Super Compact programs used to design the MMIC amplifiers are placed in Appendix I for reference. To understand the structure of the programs, the circuit "FETCG" drawn in Figure 4.15 and listed in Appendix I-1 will be discussed.

Figure 4.15 COMMON GATE FET

The Super Compact code listed in Appendix I-1 is broken into sections. Each section is terminated by an "END" statement. The first block begins with "NOI". All element nodes are numbered in agreement with Figure 4.15. The two-port is called "FETCG" with its input at node 5 and output at node 2. The missing inductor's node 4 is grounded by default. The next block "FREQ" is the frequency range over which the circuit will be analysed. Here, a 2-18 GHz sweep with one GHz steps is defined. The block "OUT" selects the output option. Here, the two-port's s-parameters will be listed. The last section "DATA" lists the s-parameters and noise parameters of a 300 μm FET in the common source configuration.

In order to model the resistors, inductors, capacitors and FETs previously discussed, the respective Super Compact element models "RES", "SPI", "TFC" and "FET" will be used. Microstrip transmission lines are also simulated in Super Compact and they are defined by the "TRL" element and the "MS" data block.

The most difficult MMIC element to model accurately is the inductor. Super Compact version 1.8 models both spiral and rectangular inductor geometries. From previous work at the Communications Research Center, it was found that the octogonal Adams-Russell inductors were closely modeled for frequencies under 10 GHz using the spiral model. One by-product of the following work is a verification of the validity of this Super Compact inductor model at C-band.

4.4 MMIC DESIGN STEPS

Recently, commercial GaAs MMIC foundry service has become available. Table 4.1 shows some of the foundries available as of 1986. It includes their technical capabilities and fee. The fee is an estimate based on producing ten samples of ten different designs with an individual chip area of 1.5 mm^2 . The design steps required to access a foundry and produce a useful module will now be discussed.

Table 4.1 FOUNDRY CAPABILITY

	Adams-Russell	Avantek	Gould	Harris	Triquint	Ma-Com
Max. Chip Size (mm^2)	16	16	64*	16	9	16
Metal Thickness (μm)	1.5	1.5	2	3	1.5	2
Min. Gate Length (μm)	0.7	0.5	0.5	0.5	0.5	0.25
Cost per Iteration (US\$) §	40K	40K	65K	65K	50K	65K
Comments		No device models supplied by foundry. HEMTs under development	HEMTs within a year			

* maximum chip size defined by reticle size and not yield.

§ as of January, 1986

The steps required to design a MMIC are illustrated in Figure 4.16. Unlike MIC circuits, the size of MMICs prevent tweaking of the final circuit. Also, the turn-around time using a foundry is four months instead of a few weeks. Hence, considerable effort is expended at the design stage in order to ensure a satisfactory end product.

Figure 4.16 MMIC Design Steps^[53]

The first step in the design process is to determine the specifications of the MMIC. In this work, a C-band communications amplifier was designed.

To implement these specifications, microwave amplifier design techniques are applied. Using a foundry, the choice of active devices is limited. However, the devices have been previously characterized at the foundry saving the designer considerable time and effort. Once the devices are known, a matching topology is chosen using the manual design techniques where unilateral FETs, lossless capacitors and inductors, and zero length interconnect transmission lines are assumed.

Super Compact is then used to optimized the basic circuit. During optimization, different weights are applied to the gain, return loss, noise or stability response in order to obtain the required result. The designer must realize the conflicting nature of these requirements. For a few elements, few degrees of freedom are available which limits the control over the final circuit characteristics.

The optimized inductor and capacitor values are then converted into equivalent n-turn spirals and fixed area, thin-film capacitors using the foundry's design rules. Circuit layout may now proceed. Standard cell elements and bonding pads are drawn to scale on graph paper and cut apart. The paper elements are positioned inside a boundary representing the maximum die size to form the appropriate circuit.

The length of the interconnections and bonding pads must be carefully considered during the "paper" layout. Since the interconnecting lines behave as small series inductors, their length, especially on the FET's source lines to ground, is usually minimized. The placement of bonding pads requires that the designer be aware of the jig bonding arrangement of the finished die since the die is usually wire bonded into a hybrid circuit for testing. Ground pads are placed close to the jig's ground plane and the RF input/output pads are placed close to their respective input/output hybrid microstrip lines in order to minimize the influence of the wire bonds.

The "paper" design is then laid out on a CAD workstation. This is a high resolution display for which a scale of 100 pixels per μm is possible. Foundry supplied standard cells can be moved around to form a circuit similar to the paper design. However, foundry design rules limit element orientation and spacing. A plot of the circuit is then produced.

This plot is measured and the layout is transferred back into an enhanced Super Compact model where the interconnect transmission lines and bonding wires are considered. This circuit is then re-analysed. If it performs satisfactorily, a CIF-coded tape including all the mask layers required to form the MMIC is produced. If this layout does not meet the specifications, it must be re-optimized. However, clearer physical limits are now available for the circuit elements, particularly the interconnect lines, for use in the optimization. The layout is then modified to agree with the new design. If this is not possible, a compromise layout is made and the earlier process is repeated until a satisfactory final simulation results.

The CIF tape representing the mask layers is sent to the foundry where the code is used on an automated pattern generator to form the mask set. After four months, sample chips arrive and they are visually inspected under an optical microscope. If acceptable, DC and RF testing occurs.

4.5 DESIGN of a C-BAND AMPLIFIER/SWITCH

In this section, the design of two MMICs referred to as Design #1 and Design #2 will be described. The first design will be very detailed and Super Compact listings of the design at various stages of completion are placed in Appendix I for reference. The second design, being similar, will be described briefly.

4.5.1 ACTIVE/REACTIVE MATCHING and INTRODUCTION

The low frequency small-signal model of a MESFET is shown in Figure 4.17 along with its simplified device equation. The FET is shown configured in the common gate mode.

Figure 4.17 Low Frequency MESFET Model

Note that at low frequencies, the input impedance (Z_{in}) is simply the inverse of the device transconductance. That is: $Z_{in} = V_{in} / I_{in} = V_{gs} / I_{ds} = 1 / g_m$ and for a g_m of 0.020 S, Z_{in} is 50 Ω . In a 50 Ω system, the FET's input will appear matched. This is called an active match. In the following designs, the Adams-Russell 300 μm gate width FET is chosen since it has the closest available transconductance with a value of 0.024 S at $I_{dss} / 5$. Note that bias adjustments may be used to change I_{dss} and g_m to obtain a better match. This FET, in a common gate configuration, provides a wide-band input match for both amplifier designs.

To provide interstage and output matching, reactive matching is used. A shunt inductor placed before the second (gain) stage alters the phase of its reflection coefficient and a shunt-L, series-C high pass filter is added at the output to counter the $1/f^2$ rolloff of the MESFET and to match the amplifier to the transmission line over a narrow band. The simplified schematic of both amplifiers is shown in Figure 4.18.

Figure 4.18 AMPLIFIER SCHEMATIC

Both amplifiers are designed to operate in the 5.9 to 6.4 GHz communications band. Although the active input match conserved space, its main purpose is to determine the useful bandwidth of such an arrangement using today's foundry technology. The high Q interstage and output matches test the inductor models and magnify any processing variations. Also, the common gate arrangement allows the first stage to be biased "on" and "off" to form an RF switch in Design #1 and in Design #2, the single bias tests the precision of the resistive elements.

4.5.2 DESIGN #1

A discussion of the first MMIC design follows. This design has individual control over the V_{DS} and V_G bias lines of the two FETs. A discussion of the design of the active input matching circuit will be followed by the design of the interstage and output matching circuits.

Appendix I-1 lists the results of the 300 μm FET operated at $0.2 I_{DSS}$ in the common gate configuration. $|s_{11}|$ of under 0.1 up to 6 GHz and $|s_{21}|$ of approximately unity provides a useful 20 dB input return loss and 0 dB insertion loss matching circuit. Above 6 GHz, the effect of the intra-element parasitic capacitance alters the $I_{DS} = g_m V_G$ equation and V_G and I_{DS} are no longer in phase. This creates an imaginary impedance component causing a mismatch. Note that at 6 GHz, $s_{22} = 0.96e^{-j13.7}$.

The s-parameters for a 300 μm FET biased at $I_{DSS} / 2$ in the common source configuration are shown in the Super Compact output of Appendix I-2. At 6 GHz, $s_{11} = 0.68e^{-j75.2}$ and $s_{22} = 0.78e^{-j18.5}$.

To provide the interstage match for maximum power gain, the second stage impedance (Z_{in}) must be brought close to s_{22}^* of stage 1. This is partially achieved with a single shunt inductor which alters the second stage input impedance to Z_{in}' as shown on the Smith chart of Figure 4.19. Note that we require $B = 0.84 = 50 / \omega L$ at 6 GHz or $L = 1.6 \text{ nH}$ to form such a susceptance. From the Adams-Russell design manual, this corresponds to a 4.5 turn wide spiral inductor. A wide inductor (width = 10 μm , spacing = 6 μm) is chosen to provide the drain bias for the first FET.

Also shown in Figure 4.19 are the first stage stability circles at 4, 6 and 8 GHz. The load that the second stage presents to the first stage after the shunt tuning inductor is close to the 6 GHz stability circle. Note that providing a perfect match where the second stage Z_{in}' equals the first stage s_{22}^* would be unstable. This is normal for common gate configurations making them widely used as oscillators. Although there was a possibility that process variations could merge the location of Z_{in}' into the unstable region, this was felt to be a reasonable risk for a research project. In a commercial

Figure 4.20 OUTPUT MATCH

NAME	TITLE	DWG NO
SMITH CHART FORM ZY 81-N	ANALOG INSTRUMENTS COMPANY NEW PROVIDENCE N.J. 07974	DATE

NORMALIZED IMPEDANCE AND ADMITTANCE COORDINATES

The bias circuits are then added to the basic Super Compact file. The first stage gate is biased through a four-turn thin inductor to provide isolation and to allow the gate to be switched quickly and the second stage gate is biased through a 5 k Ω resistor with a shunt 5 pF bypass capacitor. The sources and drains can not be isolated with large resistors due to the substantial bias current. Instead, the first FET's drain is biased through the interstage matching inductor with a 5 pF decoupling capacitor to ground and the source is biased to ground "off chip" using a bias tee. The second FET had its source connected to ground and the output matching inductor was used to bias the drain. A 5 pF blocking capacitor isolated the two stages.

The effect of the bias circuitry on the RF performance is negligible. Only the interstage inductor changed after optimization to become 4.0 turns while the gain dropped to 14.5 dB and the input return loss became 18 dB at 6 GHz.

A sensitivity analysis was performed on the basic design. The results are shown in Figure 4.22 (a) for a 4, 4.5 and 5 turn interstage matching inductor and in Figure 4.22 (b) for overall capacitance variations of $\pm 20\%$. Clearly the designs will be very sensitive to process variations and inductor modeling errors. This high sensitivity will allow both the process control and modeling accuracy to be observed.

The circuit layout on graph paper followed. The approximate lengths of the interconnecting transmission lines were added to the Super Compact file. Re-optimizing for a flat gain profile over 5.9 to 6.4 GHz produced a circuit with a gain of 13 dB (-0.7 dB to -1.0 dB over this band). Only the capacitor's value changed becoming 0.14 pF.

The circuit was then laid out on an Avera CAD workstation. The layout was plotted to scale. All transmission line lengths were determined from the plot and the Super Compact model was modified to reflect any changes. Two more layout/optimization iterations were required before a satisfactory design was obtained.

Figure 4.22 CIRCUIT SENSITIVITY TO:

a) INDUCTANCE VARIATIONS

SUPER-COMPACT 24-OCT-86 16:28:27 File: TSTMOD.DAT Circuit: FETCG

b) CAPACITANCE VARIATIONS

SUPER-COMPACT 12-JUN-86 16:34:33 File: FINMOD.DAT Circuit: FETCG

The schematic of the final circuit is shown in Figure 4.23. The rectangles represent microstrip transmission lines with their dimensions noted. The Super Compact model and table of results are listed in Appendix I-5,6. Note that the tuning elements have become 4.0 turns, 4.25 turns and 0.21 pF. The small 0.21 pF output tuning capacitor was constructed of two 0.42 pF capacitors in series (recall Figure 4.5) due to the limited dimensional tolerances of liftoff processing. Although an interdigitated capacitor could be used, the lack of standard cells precluded its use.

Figure 4.23 FINAL DESIGN #1 SCHEMATIC

Gain and return loss plots are shown in Figure 4.24 and 4.25 respectively. Note that s_{21} is the device gain, s_{11} is the input return loss and s_{22} is the output return loss. The lack of many tuning elements limited the number of degrees of freedom available to shape the response. After considerable optimization effort, it was realized that a tradeoff between gain flatness and return loss would have to be made. The final design concentrated on minimizing the input return loss resulting in a value below 20 dB over the 5.9 to 6.4 GHz band while the output return loss and gain flatness were sacrificed.

Figure 4.24 Design #1 GAIN RESPONSE (THEORETICAL)

Figure 4.25 Design #1 RETURN LOSS RESPONSE (THEORETICAL)

4.5.3 DESIGN #2

The second design is similar in function to the first. However, it has a single 15V bias supply for the entire amplifier. Internal resistor networks set the bias levels for the FETs so the ability to RF switch the first stage is lost. This design tested the DC processing tolerances of the foundry.

The design method is similar to that for the first design and will not be repeated. However, the design of the bias circuitry will be discussed. The schematic diagram of the bias circuits is shown in Figure 4.26.

Figure 4.26 DESIGN #2 BIAS TOPOLOGY

The calculation of the resistor networks will now be detailed. The DC characteristics of the 300 μm FET, computed from the design manual, are listed below.

$$I_{\text{dss}} = 73.5 \text{ mA}$$

$$V_{\text{ds}} = 4 \text{ V}$$

$$V_{\text{gs}} = -1.2 \text{ V for } I_{\text{dss}} / 2$$

$$V_{\text{gs}} = -2.1 \text{ V for } I_{\text{dss}} / 5$$

For the first stage, $I_{D1} = 14.7$ mA and $I_{D1}(R_1 + R_2 + R_{L1} + R_{L2}) = 15 - 4$ V where R_{L1} and R_{L2} are the inductors' resistance. Computing R_{L1} and R_{L2} from the design manual metal resistivities, we find $R_1 = R_2 = 368 \Omega$. To bias the FET we require G_1 at $5.5 - 2.1 = 3.4$ V. For good isolation, R_3 is assigned the largest standard value of $10 \text{ k}\Omega$ making $R_4 = 2.93 \text{ k}\Omega$.

For the second stage, $I_{D2} = 36.75$ mA and $I_{D2}(R_5 + R_6 + R_{L3}) = 15 - 4$ V. To bias the gate with the gate grounded requires that $V_{S2} = 1.2$ V. Hence, R_6 is 32.7Ω and R_5 is 237Ω accounting for the inductor resistance R_{L3} and the saturation of the bulk resistors.

Note that chokes $L1$ and $L2$ are added to Design #2 in order to isolate the first stage. Both are chosen to be as large as possible; six-turn spirals. Due to the small bias current for the first stage, thin spirals are used. It is interesting to note that the order of the R-L elements effects the response. This is due to the large parasitic shunt capacitance to ground of the spiral inductors. To further increase the isolation, shunt capacitors, designed to resonate at 6 GHz, are placed across the chokes. The addition of these capacitors increased the gain by 2.5 dB to 10.3 dB.

The schematic of the final design is shown in Figure 4.27 and the gain and return losses are plotted in Figure 4.28 and 4.29, respectively. Compared with Design #1, approximately 2.5 dB of gain has been lost through the first stage bias chokes while the return losses are similar.

Figure 4.27 FINAL DESIGN #2 SCHEMATIC

Figure 4.28 Design #2 GAIN RESPONSE (THEORETICAL)

Figure 4.29 Design #2 RETURN LOSS RESPONSE (THEORETICAL)

The layout of both designs is shown in Figure 4.30. Design #1 is located at the top while Design #2 is at the bottom. Inputs are on the right and left, respectively. This plot shows four of the nine mask layers required to construct the chip which is only 1.5 mm X 1.0 mm.

Figure 4.30 DESIGN #1 and #2 FINAL LAYOUT

4.6 EXPERIMENTAL RESULTS

In this section, measurements of the two designs will be summarized and compared with the expected results.

4.6.1 TEST PROCEDURE and APPARATUS

In order to RF test the MMICs, they are mounted on a holder referred to as a jig. This jig provides the transition from the chip's bonding pads to the more familiar SMA connectors. A pictorial view of the test jig developed by the Communications Research Center is shown in Figure 4.31 (a) and a photo of Design #2's jig is shown in Figure 4.31 (b).

Figure 4.31 MMIC Test Jig a) Pictorial

b) Jig for Design #2

The MMICs are attached to the jigs with silver epoxy and they are thermocompression bonded into the circuit with 0.7 mil diameter gold wire. Optical pictures of the wire bonding required for each design is shown in Figures 4.32 and 4.33.

Figure 4.32 DESIGN #1 BONDING PICTURE

Figure 4.33 DESIGN #2 BONDING PICTURE

To determine the amplifier's s-parameters, an 8409 automated vector network analyser (ANA) is required. A block diagram of the setup is shown in Figure 4.34.[54]

Figure 4.34 8409 VECTOR ANA EXPERIMENTAL SETUP

An ANA is similar to the network analyser described earlier in that phase detectors and power detectors are used to measure the incident and reflected voltage waves. However, by automating a network analyser, the accuracy of the measurements is improved and more data may be taken. The 8409 is calibrated with a series of open, short, matched load, and through terminations and the reflections from port 1 and transmission characteristics are stored in memory. With the device connected, these values are subtracted from the measured results to provide error cancellation. Since only one port of this ANA is active, the device must be reversed in order to measure s_{22} and s_{12} . Hence, a bias tee is required on both the input and output of the setup as shown.

Also, the 8970A automated noise figure meter was used for a noise figure measurement (see Chapter 5) and a 435B power meter was used for linearity measurements. Both experiments require a bias arrangement similar to Figure 4.34.

Two batches of chips were supplied by the Adams-Russell foundry. The first batch had a poor active layer implant reducing the device gains. The second batch was plagued with poor airbridge quality. A SEM of an inductor with a missing airbridge is shown in Figure 4.35. All the output tuning capacitors of Design #1 were similarly damaged due to over-etching. Another SEM showing the amount of undercutting on this 0.42 pF capacitor is shown in Figure 4.36.

Figure 4.35 TYPICAL BATCH #2 INDUCTOR

Figure 4.36 TYPICAL BATCH #2 CAPACITOR

Table 4.2 summarizes the process variations where C_n is the capacitance per area and R_{sl} is the active layer resistance. These values compare the mean of sixteen process control monitor measurements for each batch (wafer) with the original foundry data. Overall, the normalized standard deviations (standard deviation/mean) for the g_m , R_{sl} and C_n were 3%, 12%, and 3%, respectively. These variations were incorporated into the initial FET models producing new sets of s-parameters which will be used to compare the measured results. In the model, the transconductance and implant resistors were scaled in proportion to Table 4.2. For example, the 300 μm 0.5 I_{dss} FET's transconductance was reduced from 0.028 to 0.026 S. Similar changes to R_{ds} , R_i , R_g , R_d and R_s were also made. Also, the capacitance variation was incorporated by a change in the CN statement.

TABLE 4.2 PROCESS VARIATIONS^[55]

	g_m	R_{sl}	C_n
BATCH #1	-5%	+15%	+10%
BATCH #2	-1%	+17%	-5%

In addition to the processing deviations, a design error was made. The inductor's center feed wire was included in the SPI model. It should have been treated as an additional separate transmission line. This error caused the actual inductance to be underestimated during the simulations.

In the following sections, the design simulations are re-run taking both the process variations and the design error into account. These simulations are then compared with the experimental results.

4.6.3 DESIGN #1 RESULTS and a COMPARISON with SIMULATION

Two sets of experimental data were obtained for the first design. All the devices were from the first batch of chips since all the samples from the second batch of Design #1 were rendered useless due to the output capacitor's airbridges being over-etched (see Figure 4.36). Although input return loss could be improved by varying V_{g1} , all the following data is for the designed bias conditions of $V_{d1} = V_{d2} = 4$ V, $V_{g1} = -2.1$ V and $V_{g2} = -1.2$ V.

It should be noted that from batch #1, another two chips were tested and failed to operate. One had a shorted capacitor and the other had a FET which could not be pinched off. Both problems were due to poor processing.

Plots of the gain and insertion loss of the first chip, referred to as chip #1js, are shown in Figure 4.37 and 4.38, respectively. Note s_{21} is the device gain, s_{11} is the input return loss and s_{22} is the output return loss. Similar plots for the second chip, chip #1-2, are shown in Figure 4.39 and 4.40.

Figure 4.37 CHIP #1js GAIN RESPONSE

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Figure 4.38 CHIP#1js RETURN LOSS RESPONSES

Figure 4.39 CHIP #1-2 GAIN RESPONSE

Figure 4.40 CHIP#1-2 RETURN LOSS RESPONSES

These results will now be compared with the simulations. Figure 4.41 consists of four gain plots. They include simulations of the original design and the design with foundry variations and the inductor feed error accounted for as well as the two experimental results #1-2. The center frequency of the modified simulation shifted down from the original 5.9 GHz to 5.2 GHz while the peak gain remained at 13 dB. The experimental results had their center frequencies at 4.6 GHz and 4.9 GHz with associated gains of 14.4 dB and 9.7 dB.

In the modified simulation, approximately one-third of the frequency shift was due to the design error which increased the inductor values. Note that the modified simulation only shifted the results half way towards the experimental results. To match the center frequencies, further simulations required that the interstage matching inductor be increased by a half turn. The plot of this simulation and the experimental results is shown in Figure 4.42.

Figure 4.41 DESIGN #1 GAIN RESPONSE COMPARISON

Figure 4.42 DESIGN #1 MODIFIED GAIN RESPONSE COMPARISON

Similarly, Figure 4.43 compares three return loss plots for both the input and output of the device. Again, the experimental curves are shifted lower in frequency with respect to both simulations. Note that only one experimental result is plotted in order to allow easier interpretation.

The measured input return loss, the parameter that the design was optimized for, is better than either simulation predicted with the return loss exceeding 20 dB from 2 to 6 GHz. Also, the return loss of over 10 dB up to 9 GHz would be useful for many applications. Hence, active matching using commercial foundry technology does provide a broadband input match for the C-band amplifier.

The experimental output return loss showed a broader match than expected. Instead of being a single high Q notch, it appears to be the sum of two closely spaced notches. This second notch could not be simulated and may be due to device parasitics which are not accounted for in the models.

The experimental and predicted return loss responses plotted in Figure 4.44 include the half turn increase to the interstage inductor that was required to match the earlier gain responses. Although not identical, the two plots are much closer in frequency. Again, the output return loss simulation does not exhibit the experimental double resonance.

Figure 4.43 DESIGN #1 RESPONSE COMPARISON

a) INPUT RETURN LOSS

SUPER-COMPACT

4-NOV-86

17:23:00

File: FIN.DAT

Circuit: FETCG

b) OUTPUT RETURN LOSS

SUPER-COMPACT

4-NOV-86

17:23:00

File: FIN.DAT

Circuit: FETCG

Figure 4.44 DESIGN #1 MODIFIED RESPONSE COMPARISON**a) INPUT RETURN LOSS**

SUPER-COMPACT 26-NOV-86 17:45:54 File: FMODECO.DAT Circuit: FETCG

b) OUTPUT RETURN LOSS

SUPER-COMPACT 26-NOV-86 17:45:54 File: FMODECO.DAT Circuit: FETCG

A lack of internal bonding pads prevented the chip from being subdivided to see if the individual FETs performed as modeled. This will be done in the next section for Design #2.

Linearity measurements were also performed on the devices. Figure 4.45 is a plot of output power (P_{out}) vs input power (P_{in}) for device #1-2. The average one dB compression power level for this design was 11.5 dBm and the saturated output power level was 14.1 dBm. The results from the second device are similar.

Figure 4.45 DESIGN #1 P_{out} vs P_{in}

An additional measurement of the RF switching characteristic was obtained by varying the bias level on the first stage gate. Gain decreased from 13 dB to -9 dB for a drop in gate voltage from 0 V to -5 V. The 22 dB dynamic range is limited by the parasitic C_{ds} element which limits the "off" isolation.^[56] Further investigations into the switching speed and noise figure recovery time were omitted due to time considerations.

4.6.4 DESIGN #2: RESULTS and a COMPARISON with SIMULATION

Three sets of experimental results were obtained on the second single-bias design. One device was from the first batch and the other two were from the second batch of chips. All data is for a single +15 V bias applied to the amplifiers. It should be noted that a lower bias resulted in gains approximately 2 dB higher.

Figure 4.46 and 4.47 show a superimposed plot of the three results for the gain and return loss, respectively. It should be noted that another chip from batch #1 failed to operate due to a shorted capacitor and a chip from batch #2 oscillated.

A gain response comparison of the experimental results and the simulations is shown in Figure 4.48. Only one experimental result is plotted to simplify the figure. The center frequency of the simulated results shifted from 6.1 to 5.7 GHz when the processing variations and design error were included while the gain remained approximately constant at 10.5 dB. The experimental result is centered at 5.6 GHz with a large 17.4 dB of gain. The other experimental results from batch #2 produced even larger gains of 18.8 and 19.7 dB at the same frequency. By adding one-quarter turn to the matching inductors and a fifth of a turn to the chokes, the center frequency of the simulation was lowered to 5.6 GHz with a gain of 11.5 dB. The experimental gain results are still far larger than the simulations predict.

Figure 4.49 shows a similar set of curves for the input and output return loss. The experimental results of both return losses are quite different from those simulated. Another simulation with the inductors lengthened, as above, did not bring the curves substantially closer. Note that on the input, there seems to be a sharp resonance at the center frequency of the gain response and the input return loss decreases from 20 dB to 10 dB after this resonance. This is far above the predicted value. Similarly for the output return loss, a resonance occurs at the center frequency.

Figure 4.47 THREE DESIGN #2 RETURN LOSS RESPONSES

Figure 4.48 THEORETICAL AND EXPERIMENTAL GAIN RESPONSES

SUPER-compact 4-NOV-86 16:42:50 File: BIAS.DAT Circuit: FETCG

Figure 4.49 THEORETICAL AND EXPERIMENTAL RESPONSES

a) INPUT RETURN LOSS

SUPER-compact 4-NOV-86 17:01:22 File: BIAS.DAT Circuit: FETCG

b) OUTPUT RETURN LOSS

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SUPER-COMPACT

4-NOV-86

17:01:22

File: BIAS.DAT

Circuit: FETCB

In order to discover the cause of the return loss resonance and high gain, the chip was subdivided and the first stage was characterized alone. The first stage is isolated from the second stage since the interconnecting transmission line was scraped off with a probe. A wide transmission line leading to the gate of the second FET allowed a wire bond to be connected to the output of the first stage. The schematic of the circuit is shown in Figure 4.50.

Gain measurements of this circuit were relatively flat with a maximum value of 0 dB from 5.6 to 9 GHz. However, de-embedding the circuit from the alumina jig and connectors and calculating the maximum available gain (MAG) provided a sharp resonance at 5.7 GHz with a possible gain of 11.2 dB. This indicates that the first stage is capable of providing over 10 dB of gain if matched appropriately. Clearly some form of matching is occurring and it is believed that it is caused by interactions with the choke/capacitance resonant circuits, the only RF difference of this design with Design #1. To confirm this, simulations were performed and the results will be discussed next.

Figure 4.50 FIRST STAGE SCHEMATIC

Super Compact runs of the first stage with the chokes lengthened, as described earlier, produced a similar MAG response. Figure 4.51 compares this simulation with the experimental data. Again, the simulation predicts lower gain but its shape and frequency are similar. If one considers that Design #2's gain is approximately 6 dB higher than Design #1, it becomes apparent that the first stage is providing this gain increment.

When the chip was biased from the single +15 V supply, the bias current was 41 mA instead of the designed 50.5 mA. It is believed that this may have changed the FETs' s-parameters enough to cause the first stage to be matched at the choke/capacitor's resonant frequency which will provide a large amount of additional gain. This would correspond to the original design where the interstage matching inductor would bring Z_{in} close to the first stage's load stability circle.

Note that the Q of the small capacitors placed across the chokes was increased from 30 to 100 to produce the theoretical MAG curve with a similar resonant structure. Hence, the loss in the circuit elements including the inductors, capacitors and FETs may be lower than originally assumed.

Figure 4.51 EXPERIMENTAL and THEORETICAL FIRST STAGE MAG

Linearity measurements were also performed on the three amplifiers. Figure 4.52 is a plot of output power (P_{out}) vs input power (P_{in}) for device #2-2. The average one dB compression power level was 10.5 dBm and the average saturated output power level was 13.5 dBm. This is slightly lower than Design #1 and is probably due to the lower bias current.

Figure 4.52 DESIGN #2 P_{out} vs P_{in}

Noise figure (NF) data was obtained for the batch #1 device. A plot and table of results is shown in Figure 4.53. A poor minimum noise figure of 7 dB was obtained. The poor noise performance was expected since an actively matched configuration suffers from the drain-to-source channel resistance at the input. Clearly active matching would be inappropriate for a LNA. However, this topology could be used at intermediate or high power levels.

Figure 4.53 DESIGN #2 NF and GAIN vs FREQUENCY

Table 4.3 summarizes the results for both designs. The simulated results take the process variations and design error into account.

TABLE 4.3 MMIC RESULTS

PARAMETER	DESIGN#1		DESIGN#2	
	Simulated	Experiment ^a	Simulated	Experiment ^b
center frequency(f_0)	5.2 GHz	4.8 GHz	5.9 GHz	5.5 GHz
max gain (dB)	13	11.9	10.6	18.6
3dB bandwidth	1.25 GHz	0.8 GHz	1.5 GHz	0.6 GHz
S11 @ f_0 (dB)	-20	-21	-24	-10
S22 @ f_0 (dB)	-9	-16	-14	-11
minimum S11(dB) ^c	-15	-19.5	-18	-9
P _{1dBc} (dBm)	-----	11.5	-----	10.5
minimum NF (dB)	-----	-----	-----	7 ^d
Switch Isolation(dB)	-----	22 ^d	-----	-----

Notes:
^a average of two results
^b average of three results
^c in the 3dB bandwidth
^d a single measurement

In both designs, for the center frequencies of the simulations to agree with experiment, the inductors required lengthening. In Design #1, half a turn was added to the interstage inductor and in Design #2, one-quarter turn was added to the tuning inductors and one-fifth of a turn was added to the chokes. Hence, the Super Compact inductor models are underestimating the actual inductance.

Note that the Super Compact inductor models do not consider the surface coupling between different inductors and other elements.^[57] To minimize this effect, earlier Communications Research Center designs separated the inductors by at least their radius from adjacent elements. Figure 4.30 clearly shows that the close spacing of both designs allows many possible coupling paths to exist. The earlier analysis that indicated that the inductor models are only off by a fraction of a turn seems to indicate that this coupling is a second order effect. Hence, for low Q circuits, a spacing similar to Figure 4.30 should be sufficient to neglect surface coupling.

4.7 CONCLUSION

It is unlikely that a narrow band two-stage design could be produced in a single design iteration using today's commercially available design tools. In addition to the inductor models needing refinement, the FET models may also need modifications. The two-stage designs of this thesis magnified these errors since a small error in the s-parameters of either FET was compounded at the interstage match to produce the large overall errors that were observed in the results. Hence, a few iterations with empirically based corrections would probably be required to produce narrow band MMICs.

Overall, the MMICs did achieve their main goal of a reasonable input match. Considering that the foundry was having early production problems and the immature state of the modeling software, the amplifiers operated amazingly well. The MMIC occupied approximately one-tenth the area of an equivalent hybrid circuit and the single bias design required only two critical bonding wires.

For an L-band design, the active matching topology looks even more promising compared to reactive matching which would require large reactive elements and a large chip. In a T/R module, the die would be placed directly adjacent to each other producing a large ripple bandwidth due to mismatching that would allow a low return loss to be traded for other requirements.

In conclusion, the small size, low weight and low volume cost of MMICs should make them widely used in many systems, including the T/R modules of future radar systems.

CHAPTER 5

MICROWAVE INTEGRATED CIRCUIT DEVELOPMENT

5.1 INTRODUCTION

This chapter is based on a comparison of a discrete High Electron Mobility Transistor (HEMT) and a Metal Semiconductor Field Effect Transistor (MESFET) which may be used in Low Noise Amplifiers (LNAs). In the following work, the noise parameters of both devices will be determined at 12 GHz. Also, the devices will be characterized at room temperature and liquid nitrogen temperatures over a wider band. Circuits using Microwave Integrated Circuit (MIC) techniques will utilize the discrete devices.

An ideal opportunity to compare the HEMT and MESFET discrete devices appeared in November of 1985 when Gould became the first manufacturer to offer a HEMT in production quantities.^[58] This device was based on the surface geometry of an existing MESFET which allowed a meaningful comparison of the two devices to be accomplished in the following experiments. Photographs of the two devices used in the following work are shown in Figure 5.1. Figure 5.1(a) is a scanning electron micrograph (SEM) of the H503 HEMT while an optical picture of the 0503 MESFET is shown in Figure 5.1(b). The layouts are almost identical with both devices having a 280 μm wide channel and the π gate feed arrangement. The only differences are the lack of a passivation layer on the HEMT and its longer gate length. The H503 has a 0.5 μm gate length versus the 0.3 μm gate length of the 0503.

Figure 5.1 DEVICE PHOTOGRAPHS a) H503 HEMT

b) 0503 MESFET

The purpose of this chapter is to observe the behavior of these two commercially available devices in the areas of high gain and low noise. Earlier research using laboratory devices has shown the superiority of the HEMT in these areas. It is believed that future high performance LNAs will use HEMT devices. Such LNAs could be applied in the T/R modules of the active arrays discussed in Chapter 2.

5.2 THEORY

Before the design of the circuits to compare the HEMT and MESFET can be discussed, the theory of distributed element matching, HEMT operation and transistor amplifier noise will be summarized.

5.2.1 DISTRIBUTED ELEMENT MATCHING

The following circuits make use of hybrid or Microwave Integrated Circuit (MIC) technology. These hybrid circuits use microstrip transmission lines on an alumina substrate to bias and match the active discrete transistors. In the following designs, the length of the microstrip transmission lines used to interconnect the circuit elements are of the order of a wavelength requiring that distributed transmission line theory be applied in the design and analysis of the circuits.

In the following designs, the discrete devices are placed on a jig that is approximately 25 mm square. At 12 GHz, the guide wavelength on alumina is 9.4 mm so the transmission lines are large relative to a wavelength and distributed transmission line matching techniques must be employed to analyse the circuit. This differs from the earlier MMIC work of Chapter 4 where all "on-chip" interconnecting transmission lines were under $\lambda/10$ in length allowing lumped element matching techniques to be employed for a large part of the design.

Transmission line theory may be used to derive the expression, shown in equation 5.1, describing the impedance at any position on the lossless transmission line shown in Figure 5.2.^[59]

$$Z_{in}(z) = (Z_L + jZ_0 \tan(kz)) / (Z_0 + jZ_L \tan(kz)) \quad (5.1)$$

Here k is the propagation constant ($2\pi/\lambda_0$). Note that if the load Z_L is an open circuit and $\lambda/4 < l < \lambda/2$, the impedance will be $Z_{in} = -j\cot(kl)$ and the load will appear as a shunt inductor. Also note that if $l = \lambda/4$, $Z_{in} = 1/Z_L$ and the transmission line will form an impedance inverter. Both of these techniques will be used in the following designs. Also, the microwave amplifier design techniques discussed previously in Section 4.3.2 will be applied.

Figure 5.2 LOSSLESS TRANSMISSION LINE

In order to realize the transmission lines, microstrip technology is used in the following MIC designs. Here, gold and tungsten thin films are sputtered onto an alumina substrate. The substrate is masked to define the required metal patterns using a photolithographic technique similar to that used to construct the MMICs of Chapter 4. In contrast to the nine MMIC masks, only a single mask was required in the following MIC designs.

The structure of a microstrip line is shown in Figure 5.3. The gold layer forms a low-loss conductor on top of the tungsten layer which is required to form a strong mechanical bond with the substrate. Note that the underside of the substrate is completely metallized to form the ground plane for the microstrip transmission lines. In the following designs, the metallization thickness (t) is $4\ \mu\text{m}$ and the alumina substrate thickness (d) is $0.025''$.

Figure 5.3 MICROSTRIP STRUCTURE

The impedance Z_0 of the transmission lines depends on the aspect ratio w/h as identified in Figure 5.3. Note that since some of the electromagnetic fields pass through an abrupt dielectric discontinuity at the air/alumina interface, the mode of propagation is not pure TEM. It is referred to as quasi-TEM and requires that empirical expressions be used to analyse the impedance of the structure. For $w/h > 1$ and $t \ll h$, the impedance follows the expressions of equation 5.2.^[60] Overall, the wider the line, the lower its impedance. Note that since the electromagnetic fields are not fully confined to the substrate, they fringe into the air lowering the effective permittivity ϵ_{eff} . For microstrip lines of realizeable impedance values, ϵ_{eff} is approximately 7 on alumina ($\epsilon_r = 10$). The Super Compact program which uses similar empirical expressions is used to determine the appropriate line widths in the following work.

$$Z_0 = (120\pi/\sqrt{\epsilon_{eff}}) \{w/h + 1.393 + 0.667(w/h + 1.444)\} \quad (5.2)$$

$$\epsilon_{eff} = (\epsilon_r + 1)/2 + ((\epsilon_r - 1)/2)(-1 + 10h/w)^{-0.56}$$

In addition to the transmission lines, discrete resistors, capacitors, and transistors are used in the following designs. Thin-film, laser-trimmed resistors and high Q $\text{Si}_3\text{N}_4/\text{SiO}_2$ dielectric capacitors are used. These components are attached with silver epoxy for the noise experiment and a stronger, low temperature Au/Ge solder for the environmental experiment. Again, thermo-compression bonding is used to connect the devices into the circuits.

Transmission line theory will now be used to explain the bias isolation circuits used in the following designs. These circuits must provide DC bias to the transistors while having a negligible effect at microwave frequencies. Figure 5.4 (a) shows the basic layout of a bias circuit. Here, an open circuited low impedance $\lambda/4$ transmission line is connected to a $\lambda/4$ high impedance line at point A which joins the $50\ \Omega$ main transmission line which we are trying to bias at point B. Hence, the open circuited low impedance stub is transformed to a short circuit at A which is then transformed to an open circuit at B creating a reflection coefficient near unity at B. This allows the DC bias to flow while preventing RF leakage at a frequency where the stubs appear to be of $\lambda/4$ electrical length. An equivalent lumped element circuit is shown in Figure 5.4 (b). The wide stub appears as a capacitor to ground while the narrow high impedance line may be modeled as a choke. The bias line is connected at A for minimal RF loading.

Figure 5.4 BIAS CIRCUIT TOPOLOGY

a) MICROSTRIP LAYOUT

b) EQUIVALENT CIRCUIT

5.2.2 HEMT STRUCTURE and OPERATION^{[61] [62]}

In order to operate the discrete HEMT and MESFET devices properly, their operating characteristics must be known. The structure and operation of the HEMT will now be discussed. Refer to Section 4.2.2 for a description of the MESFET.

As previously stated, the layouts of the HEMT and MESFET are similar. The major difference between the two devices is found in the active layer doping profile. In a HEMT, the active layer is modulation doped in the vertical dimension to create a "super lattice" which gives the HEMT its superior performance. Only recently has Molecular Beam Epitaxy processing technology allowed the precise control of the vertical doping profile required to form HEMTs.

The vertical structure of a HEMT is shown in Figure 5.5. Notice that the active layer is a sandwich of a doped tertiary compound (AlGaAs), a thin layer of pure (undoped) AlGaAs and a pure binary compound (GaAs) substrate. At the AlGaAs/GaAs interface, a heterojunction forms. Due to the higher bandgap of AlGaAs, the electrons in this doped layer migrate through the thin spacer of pure AlGaAs into the pure GaAs forming a two-dimensional electron gas. In the undoped GaAs, these mobile carriers are free of Coulomb scattering effects. That is, they are unhindered in their movement by collisions with donor ions which would decrease their average velocity. Hence, HEMTs have a higher electron mobility resulting in superior microwave performance.

Figure 5.5 VERTICAL HEMT STRUCTURE[58]

With reference to Figure 5.5 and 4.8, the HEMT may be described by equation 5.3.

$$I_{ds} = g_m[V_g - V_{off} - E_c L_g] \quad (5.3)$$

Here, the transconductance $g_m = g_{m0}/(1 + g_{m0}R_s)$ where g_{m0} is a parameter based on the dimensions of the device ($g_{m0} = \epsilon_r \epsilon_0 v_s W/d_2$) and v_s is the electron's saturation velocity, E_c is the critical electric field where the electron velocity saturates and d_2 is the thickness of the AlGaAs layer. Also $V_{off} = \phi_m - \Delta E_{cond} - V_{p2}$ where ϕ_m is the Schottky barrier height, ΔE_{cond} is the difference in the conduction band levels at the AlGaAs/GaAs interface and V_{p2} is the voltage required to pinch off the AlGaAs channel ($qN_D d_2^2 / 2\epsilon_r \epsilon_0$). Equation 5.3 only holds for operation away from pinchoff. More comprehensive models are being developed to describe the HEMT over its complete operating regime.[63]

The current - voltage characteristic for the H503 is shown in Figure 5.6. Note the linear $I_d - V_g$ dependence that equation 5.3 predicts. This is caused by the fact that the HEMT does not operate like a MESFET. It is similar to a metal oxide semiconductor (MOS) structure operated in accumulation. However, the number of electrons at the heterojunction is directly modulated by the gate bias since the depletion layer width remains constant at d_2 until the device is pinched off. This differs from the MESFET and

MOSFET behaviour since d_2 is fixed by the superlattice doping profile. As previously stated, the transition region between the linear and pinched off mode of operation is not yet completely understood.

Figure 5.6 H503 I-V CURVE^[58]

The main advantage of the HEMT structure compared to the MESFET is its ability to place the heterojunction close to the gate contact by decreasing d_2 during production. As d_2 is decreased, the influence of the gate potential over the two-dimensional electron gas increases. Hence, the transconductance increases which increases the f_T improving the microwave performance of the device. This structure bypasses the transconductance limitation of MESFETs. To improve a MESFET's performance, the active layer's doping concentration is increased which shrinks the channel depth bringing the channel closer to the gate. Again, this increases the influence of the gate on the channel current to increase the transconductance. However, a maximum doping concentration of 10^{18} cm^{-3} limits this technique and the transconductance of a MESFET.

The same equivalent small-signal model is used for the HEMT and MESFET in the following designs. The schematic of the equivalent model is shown in Figure 5.7 (a). Compared to the MMIC's model shown earlier in Figure 4.10, additional parasitic elements have been added due to the higher frequency response of these discrete devices. Figure 5.7 (b) lists the element values determined by Gould for both devices. Comparing the values, the transconductance of the HEMT is approximately 10% higher which should result in a larger forward gain. Also, the HEMT's resistive parasitics R_g , R_i , R_s and R_d are substantially lower due to the added conductivity of the AlGaAs layer outside of the gate region which should improve its noise performance.

Figure 5.7 a) SMALL SIGNAL MODEL [64]

b) TABLE of MODEL VALUES^[64]

element	0503	H503	units
gm	0.040	0.043	mho
Cgs	0.0092	0.0082	pF
Rg	2.21	1.34	Ω
Ggs	1.1E-12	1.1E-12	mho
Cdg	0.03	0.028	pF
Cgs	0.37	0.325	pF
Ri	2.86	2.05	Ω
Cdc	0.008	0.004	pF
Rs	0.71	0.6	Ω
Gds	0.002	0.001	mho
Cds	0.09	0.042	pF
Rd	2.43	1.1	ohm
Cds	0.01	0.02	pF
Lg	0.09	0.16	nH
Ld	0.1	0.15	nH
Ls	0.03	0.04	nH

At room temperature the motion of the HEMT's conduction electrons is Brownian. The electrons travel randomly with their velocity mainly limited by collisions with the vibrating lattice and the fact that there is an absence of ionized donors to collide with is of secondary importance. These collisions are referred to as phonon scattering.

In normal operation, the HEMT operates under saturated velocity conditions. When an electron in GaAs obtains excessive kinetic energy, it is transferred into a slower higher potential energy state corresponding to the upper valley of the energy - propagation constant diagram. This corresponds to the high electric field region below

the gate which has the most significant effect on the device's performance. Only in the regions of low electric field near the source and drain is the HEMT's electron mobility enhanced. Hence, the name "high electron mobility transistor" is a misnomer under normal operating conditions. More appropriate names are Two-Dimensional Electron Gas FET (TEGFET) and Modulation Doped FET (MODFET).

The effect of temperature on the two devices will now be considered. The mobility dependence on temperature is shown in Figure 5.8 for both HEMT and MESFET structures. If the HEMT is cooled to 77K, Coulomb scattering will dominate phonon scattering and it will become a true high electron mobility device. The electrons at the heterojunction will travel "ballistically" and reach velocities approaching 10^8 cm/s which is an order of magnitude larger than the saturated velocities at room temperatures. Note that the MESFET's carrier velocity will be limited at low temperatures by the Coulomb scattering caused by its high channel doping resulting in a marginal mobility increase at 77K. It should be stressed that although the mobility of the HEMT should increase by approximately a decade at 77K, the "overshoot effect" will limit the performance gain to a fraction of this value.^[65]

Figure 5.8 ELECTRON MOBILITY vs TEMPERATURE^[63]

Before the circuits for low noise operation are designed, the basic theory of how to match a transistor to obtain the minimum noise figure will be discussed. In general, any linear two-port may have its noise characteristics defined by equation 5.4.^[51]

$$F_n = F_{\min} + (r_n/g_o) [(g_s - g_o)^2 + (b_s - b_o)^2] \quad (5.4)$$

Here, $F_{\min}$ is the minimum noise figure possible which occurs when the source impedance $[g_s, b_s]$ is equal to the optimum noise impedance $[g_o, b_o]$. The parameter r_n is referred to as the equivalent noise resistance. Note that the noise figure is only dependent on the input match to the device.

If equation 5.4 is plotted on a Smith Chart, the constant noise figure contours will form circles. Normally Γ_o is defined as the reflection coefficient at $[g_o, b_o]$ allowing the constant noise figure circles to be defined by equation 5.5.^[51] Here, F_i is the desired noise figure value. Once $F_{\min}$, r_n and Γ_o are known, Super Compact may be used to plot these circles.

$$\begin{aligned} \text{center} &= \Gamma_o / (1 + N_i) \\ \text{radius} &= \sqrt{N_i^2 + N_i (1 - |\Gamma_o|^2) / (1 + N_i)} \\ \text{where } N_i &= (F_i - F_{\min}) / (1 + |\Gamma_o|^2 / 4r_n^2) \end{aligned} \quad (5.5)$$

Γ_o and r_n data for both the 0503 and H503 had yet to be determined by the manufacturer when this work was begun. Hence, one of the objectives of the following experiments is to determine these parameters.

For MESFETs, Fukui has found that equation 5.6 models the device's minimum noise figure correctly. Note that K_f is a fitting factor and in MESFETs with a high quality substrate, K_f has been found to be 2.5.^[66] The same equation also holds for HEMTs with $K_f = 1.5$.^[63] Hence, this equation will be used to determine $F_{\min}$ for both devices.

$$F_{\min} = 1 + 2\pi K_f f C_{gs} \sqrt{(R_g + R_s)/g_m} \quad (5.6)$$

Once $F_{\min}$ and Γ_0 have been determined, r_n may be found from rearranging equation 5.5 into 5.7 as shown below. Note that $\Gamma_s = 0$ is simply a matched (50 Ω) transmission line on the input.

$$r_n = [F_n \Gamma_s = 0 - F_{\min}] [1 + \Gamma_0]^2 / 4 |\Gamma_0|^2 \quad (5.7)$$

The noise parameters of both a HEMT and MESFET may be obtained from existing theory if the device's geometry and doping is known. Brookes has modified Pucel's MESFET noise model for HEMTs with sub-micron gate lengths and obtained reasonable agreement with experiment. However, the calculations are quite involved and require processing data.^[67] Sufficient processing data on both the HEMT and MESFET was unavailable to determine Γ_0 using this theory. Hence, an experimental approach was used to determine Γ_0 for both devices in the following work.

At room temperature, both devices have their noise dominated by thermal noise created by lattice vibrations. However, at cryogenic temperatures, hot-electron noise dominates. Theory predicts that the noise temperature of the device varies linearly with absolute temperature.^[68] To observe the effects of temperature on both devices, another experiment in which both devices were characterized at 290K and 77K was performed. The design of these experiments will now be discussed.

5.3 CIRCUIT DESIGNS to COMPARE MESFET and HEMT TECHNOLOGY

5.3.1 NOISE PARAMETER DETERMINATION

This first MIC experiment was designed to determine the noise parameters of both the HEMT - H503 and the MESFET - 0503 in the Ku communications band. As previously stated, the noise behaviour of these transistors may be described by the parameters $F_{\min}$, r_n and Γ_0 using equation 5.4. Theory will be used to confirm the manufacturer's value of $F_{\min}$ while r_n and Γ_0 will be determined in the following experiment.

Using the small signal model data of Figure 5.7(b) in equation 5.6, $F_{\min}$ was calculated for both devices at 12 GHz. These results along with the $F_{\min}$ values supplied by the manufacturer are listed in Table 5.1. As seen, the results confirm the manufacturer's data which will be applied later in Section 5.4.3 to determine r_n and Γ_0 .

TABLE 5.1 $F_{\min}$ (dB) at 12 GHz

	THEORY	MANUFACTURER
MESFET	2.0	1.9
HEMT	1.1	1.2

The majority of the following work is involved in the experimental determination of the parameter Γ_0 . In this experiment, a selection of source reflection coefficients (Γ_s) is presented to the input of both devices. As Γ_s nears Γ_0 the noise figure will decrease. By measuring the noise figure over a range of Γ_s values, the most likely Γ_0 value will be determined allowing the parameter r_n to be calculated using equation 5.7.

In designing this experiment, the first requirement was to determine the approximate location of Γ_0 . Previous experimental experience in MESFET amplifier design was used to choose an appropriate range of Γ_s values. In MESFET design, it has been found that the magnitude of Γ_0 is approximately 80% of s_{11} and its angle is

normally within the range of s_{11}^* 's angle $\pm 20^\circ$.^[69] Hence, the following experiment was designed to have Γ_s values surround this region.

The manufacturer's supplied s-parameters were used to determine s_{11}^* for both devices. Experimental work characterizing the devices in a 50Ω environment, presented later in Section 5.4.2, verified these s-parameter values. For the HEMT, s_{11} is $0.77e^{-j132.7}$ and for the MESFET, s_{11} is $0.72e^{-j132.5}$ at 12 GHz. It was determined that a single open circuit stub ranging in length from $\lambda/4$ to $\lambda/2$ would provide a series of source reflection coefficients near the required value.

In order to provide a variety of Γ_s values, an array of five open circuit stubs with three or four segments shown in Figure 5.9 is used in this experiment. One at a time, the five stubs are connected onto the input transmission line by wire bonding and their length is varied by connecting up to four of their sections. Stub #1 with four sections is shown connected. Note that this array allows eighteen different Γ_s values to be presented to the input of the device. Since the s_{11} values of the two devices are close, the same array is used for both devices in the following experiment.

FIGURE 5.9 INPUT STUB ARRAY

Super Compact was used to choose the length of the stubs and the distance that the stubs were placed from the input plane of the transistors. The range of Γ_s values that the stubs will create, along with s_{11}^* is shown in Figure 5.10.

In order to minimize the noise contribution of the device's output circuit on the noise figure measurement, the output of both the MESFET and HEMT were matched for maximum gain. Since the design is similar for both devices, only the output circuit of the HEMT will be described.

The initial manual design of the HEMT's output circuit is shown diagrammatically on the Smith chart of Figure 5.11. Note that s_{22} is $0.623e^{-j50.8}$. The output impedance is shifted along a constant VSWR circle by a 0.2λ long 50Ω transmission line to a real impedance and a two-stage Butterworth transformer of impedance 11 and 31Ω matches this to 50Ω . The final optimized design is shown in the schematic of Figure 5.12(a). The MESFET's schematic, similarly derived, is shown in Figure 5.12(b).

Also shown in Figure 5.12 is the gate and drain bias circuitry. This is similar to the bias topology discussed earlier except for the addition of a series 50Ω resistor - ~~10~~ pF capacitor placed at the junction of the low and high impedance $\lambda/4$ transformers. Although transistors with sub-micron gates provide impressive performance at high microwave frequencies, they suffer from instability at lower frequencies due to their high gain. This R-C filter was added to help stabilize the amplifiers at these lower out-of-band frequencies. It is designed to absorb power outside the bandwidth of the bias circuit which encompasses the unstable lower frequencies. This technique was developed previously at the Communications Research Center.^[70]

IMPEDANCE OR ADMITTANCE COORDINATES

FIGURE 5.12 FINAL CIRCUIT SCHEMATICS

a) HEMT

b) MESFET

A Super Compact plot of the insertion loss of the bias circuit attached to a 50 Ω line is shown in Figure 5.13. Over the 11.7 to 12.2 GHz band for which the following experiment was designed, the loss was under 0.25 dB whereas at 6 GHz, the frequency of instability, the loss was 1.6 dB. Note that the two bias circuits used in the actual design doubled the insertion loss so any oscillations were precluded resulting in a stable circuit.

FIGURE 5.13 BIAS CIRCUIT AMPLITUDE RESPONSE

SUPER-COMPACT 27-AUG-86 15:59:19 File: IDAMP.DAT Circuit: A

The predicted gain and return loss of the Ku-band amplifiers are shown in Figure 5.14 for the HEMT and Figure 5.15 for the MESFET. Note that these plots are for a single input stub placed at the position of stub #1. Its length was chosen to maximize the gain in the 11.7-12.2 GHz band requiring a stub 113 mil long for the HEMT and 111 mil long for the MESFET. Over this band, the HEMT should provide 1.5 dB more gain than the MESFET.

FIGURE 5.14 HEMT RESPONSE a) GAIN

b) RETURN LOSS

FIGURE 5.15 MESFET RESPONSE a) GAIN

b) RETURN LOSS

The layout and bonding configurations for the two circuits are similar so only the HEMT's circuit is shown in Figure 5.16. Note that the transistor, mounted on a separate 0.020" thick shim, is clamped between the input and output hybrid circuits which are mounted on separate brass jigs. This allows the input and output circuits to be characterized individually and the same transistor to be used with different matching circuits.

FIGURE 5.16 HEMT CIRCUIT LAYOUT

5.3.2 LOW TEMPERATURE CHARACTERIZATION

In this second experiment, the noise figure and s-parameters of the HEMT and MESFET were determined at ambient temperatures of 77K and 290K. The purpose of this experiment is to confirm the superiority of the HEMT over the MESFET at low temperatures due to its ballistic carrier flow. The hybrid circuit required to characterize the devices simply consists of 50 Ω input and output microstrip transmission lines.

The production of the jig to allow measurements at the temperature of liquid nitrogen (77K) required careful consideration. As the temperature of the jig is lowered, moisture in the air will condense on the circuit. Considering that the dielectric constant of water is roughly 80, this condensed water film will change the impedance of the microstrip lines as well as create appreciable parasitic capacitances along the surface of the transistors. Since the dielectric constant of liquid nitrogen is only slightly higher than air at 1.45, a circuit fully immersed in this coolant will have its microwave behaviour only slightly effected. Hence, the immersion of the circuit into a container filled with liquid nitrogen was selected to provide the required cryogenic environment.^[71]

The mechanical design of the test jig will now be discussed. Severe mechanical stresses are induced in any compound structure cooled over a large temperature range. Experiments involving the immersion of an alumina substrate attached with silver epoxy to a brass holder ended after a single thermal cycle when the alumina de-bonded. Further experiments using Au/Ge solder to bond the substrate to the carrier and sample thermo-compression bond wires withstood the thermal shock. Therefore, the alumina and transistors are attached to the brass carriers with Au/Ge solder in the following characterization experiment.

The final circuit and holder is shown in Figure 5.17. This is the same type of three-piece jig used in the Γ_0 determining experiment shown earlier in Figure 5.16. However, the solder bonding results in a more rugged assembly. The container used to hold the liquid nitrogen was a slab of styrofoam carved with a hot-wire cutter to match the shape of the test jig and its semi-rigid coax lines. A picture of this is shown later in Figure 5.19.

Figure 5.17 TEST JIG

Another complication of this experiment is the need to de-embed the s-parameter measurements taken at the SMA connectors to obtain the s-parameters at the transistor plane. This is accomplished using the Super Compact reciprocal two-port model where the input and output transmission lines and connectors were subtracted mathematically to obtain the required results. To allow this to be accomplished accurately, the sex of the connectors on the test jig were mixed as seen in Figure 5.17. During calibration of the network analyser, an accurate through was provided by simply connecting the mating semi-rigid coax lines. These coax lines, approximately five inches in length, act as a spacer to insulate the network analyser's expensive cables from the possibly harmful cryogenic environment.

5.4 EXPERIMENTAL RESULTS

5.4.1 TEST EQUIPMENT

The following experiments required automated network analysers to determine the s-parameters of the devices. The 8409, discussed previously in Chapter 4, was used in the determination of Γ_0 while the 8510 was required for the cryogenic characterization experiment.

The 8510 is shown in Figure 5.18. It is required for the cryogenic experiment since the jig could not be reliably disconnected and reversed at 77K in order to obtain all the s-parameters using the 8409. The 8510 solved this problem since its two active ports allow all four s-parameters to be measured at once. Also note that since the 8510 has internal biasing whereby the transistor bias is connected through BNC connectors on the mainframe, no external bias tees are required.

Figure 5.18 8510 ANA SETUP^[72]

For the noise measurements required in both experiments, the 8970A automated noise figure meter and a 346B noise source are used. This equipment determines the noise figure of an amplifier using equation 5.8. Here, N_p is the noise at the amplifier's output, N_a is the noise added by the amplifier and KGB_nT_s is the thermal noise at the input amplified by the amplifier's gain G . In addition, K is Boltzmann's constant and B_n is the bandwidth. By switching the temperature, T_s , of the 346B noise source between two noise temperatures T_c and T_h and measuring the corresponding output noise power N_{p_c} and N_{p_h} , N_a may be determined using equation 5.8. The noise figure is then calculated by comparing N_a to the noise created by a 290K resistor.^[73]

$$N_p = N_a + KGB_nT_s \quad (5.8)$$

A picture of the equipment setup for the cryogenic experiment is shown in Figure 5.19. Here, the signal from the wide band 346B noise source flows through a bias tee to the input of the device under test and out another bias tee into a wideband mixer placed at the input to the 8970A. Since the 8970A is limited to measurements below 1.5 GHz, the mixer and the 8530 sweeper acting as a local oscillator heterodyne the noise at the frequency of interest, 12 GHz, into the 8970A's band. No image rejection filter is used resulting in double sideband noise figure measurements.

Figure 5.19 NOISE FIGURE METER APPARATUS

5.4.2 S-PARAMETER DETERMINATION

The first set of results contrasting the simulations based on the manufacturer's supplied device models with their measured s-parameters will now be discussed. The s-parameters of both devices are compared to Super Compact simulations using the models listed in Figure 5.7 (b).

The s-parameter set for the HEMT is shown in Figure 5.20 and 5.21. A magnitude plot of s_{21} is shown in Figure 5.20 for device HEMT-2. The experimental gain is approximately 1 dB below the simulation and there appears to be a resonance around 13 GHz. The input and output reflection coefficients s_{11} and s_{22} as well as the reverse isolation s_{12} of the same device are shown in Figure 5.21. The phase shift of s_{11} exceeds the simulation while the phase shift of s_{22} over the 2 to 22 GHz band is less than expected. These phase deviations are probably due to errors in de-embedding the 12 mm input/output microstrip lines. Note that at 12 GHz, a 1 mm error in the electrical length of the connector models or the transmission input/output lines will correspond to a 75° phase error. Hence, the experimental phase deviations may be accounted for by small errors in characterizing the test jig. Also note that the magnitude of s_{11} is about 20% larger than expected at 12 GHz and that there seems to be a jig resonance above 15 GHz indicated by the closed circles on the Smith chart plots.

Figure 5.20 HEMT GAIN RESPONSE

Figure 5.21 HEMT RETURN LOSS AND REVERSE GAIN RESPONSE

The MESFET's s-parameters are shown in Figure 5.22 and 5.23. The magnitude of s_{21} is very close to the prediction in Figure 5.22. The remaining s-parameters are shown in Figure 5.23 and they appear to be closer to prediction than the HEMTs. However, s_{11} now appears to be about 20% low in magnitude at 12 GHz and a jig resonance is seen at 11 GHz.

Figure 5.22 MESFET GAIN RESPONSE

Figure 5.23 MESFET RETURN LOSS AND REVERSE GAIN RESPONSE

From the s-parameter characterization, it is generally verified that the devices follow the manufacturer's models. However, it is seen that the jigs resonate. This is due to their three-piece construction and this resonance will make the accuracy of the following 12 GHz measurements marginal. Overall, it was confirmed that the earlier design of the two experiments based on the manufacturer's data was valid.

During the characterization, it was also noted that the saturation current (I_{DSS}) and the transconductance (g_m) varied among the three batches of HEMTs purchased. This variation is clearly shown in the histograms of Figure 5.24. In Figure 5.24 (a), the g_m is seen to cover $0.045 \pm 25\%$ mho and the saturation current shown in Figure 5.24 (b) varies wildly. Clearly the HEMTs exhibit poor processing repeatability which is probably due to Gould's limited production experience with this new device. Hence, in the following experiments, devices with DC characteristics close to the design model are chosen.

Figure 5.24 HISTOGRAM of HEMT DC PARAMETERS

a) TRANSCONDUCTANCE

RELATIVE FREQUENCY

b) SATURATION CURRENT

5.4.3 NOISE PARAMETER RESULTS

As stated earlier, the determination of the optimum source reflection coefficient (Γ_0) was the aim of the first experiment. This experiment also allowed the general noise performance of the two devices to be compared. By selectively connecting the eighteen input stub combinations to the devices and measuring the noise figure these goals were achieved.

The first part of this experiment was to determine the actual impedance that the eighteen stubs were presenting. The reflection coefficient of the stub array was determined by connecting it directly to a 50 Ω output microstrip line and measuring the s-parameters. The results, de-embedded to remove the phase and attenuation effects of the output transmission line, are shown in Figure 5.25. Two arrays were measured and their values were averaged in order to form this diagram. Compared with the simulated response shown earlier in Figure 5.10, a large phase rotation of the impedance values has occurred.

Figure 5.25 MEASURED ARRAY IMPEDANCES

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The measured impedance contour rotated away from the design contour an average of 30° in a clockwise direction. This may be explained by taking the parasitic effect of the unconnected stubs that was neglected in the initial design into account. A single unconnected stub #2 added to the original simulation of stub #1 resulted in a 6° clockwise rotation which seemed to verify this theory. For the following work, the measured Γ_s contour will be used in the analysis. Regrettably, this contour is far removed from the s_{11}^* region where Γ_0 is believed to be. Hence, the following results will not be as accurate as originally planned.

Note that stub #4 is missing from Figure 5.25. It is not connected in the following experiment since its impedance is very similar to that of stub #3 and #5. Hence, only fifteen different input impedances are connected using this jig while a $50\ \Omega$ input transmission line is used to measure $\Gamma_{s=0}$.

The noise figure was measured over an 8 to 14 GHz band with the results at 12 GHz shown in Figure 5.26 for the HEMT and Figure 5.27 for the MESFET. Here, the noise figure, in dB, is superimposed on the stub's impedance contour.

The data in these figures was used to determine the most probable Γ_0 location for both devices. A Super Compact program with a variety of assumed Γ_0 values was run to determine the corresponding noise figure circles. Again, Γ_0 values in an array bounding the s_{11}^* region were used. A "least normalized magnitude" algorithm using the error function described by equation 5.9 was then used to determine the best Γ_0 value. Here, the difference between the measured and simulated noise figure, in linear units, is divided by the measured noise figure at every Γ_s value and summed to form the error function. This error function was calculated for sixteen different possible Γ_0 values and the results were superimposed on the Γ_s array. The Γ_0 value with the minimum error function was then used to calculate r_n . The best values for Γ_0 and r_n , determined by this method for both devices, are listed in Table 5.2.

$$E = \sum NF_{\text{measured}}/NF_{\text{simulated}} \quad (5.9)$$

Figure 5.26 HEMT Noise Figure Measurements

Figure 5.27 MESFET Noise Figure Measurements

Table 5.2 NOISE PARAMETERS at 12 GHz

	Γ_0	r_n	$F_{min}(dB)$
MESFET	$.7e^{j130}$	0.24	1.9
HEMT	$.6e^{j120}$	0.36	1.2

5.4.4 CRYOGENIC and ROOM TEMPERATURE CHARACTERIZATION

In this experiment, the noise figure and gain of the HEMT and MESFET were determined at 77K and 290K over a 2 to 22 GHz band. Three attempts were required to characterize the HEMT at 77K due to cracking problems. Also, the 8510 malfunctioned part way through this experiment requiring that the final data be averaged to remove system noise.

The first step in this experiment was to determine the current-voltage transfer characteristics to allow correct biasing of the two devices in the following measurements. The transfer curve for the HEMT is shown in Figure 5.28 (a) while the MESFET is shown in Figure 5.28 (b). Note that for both devices the saturation current increased approximately 50% at the lower temperature. From these curves, the transistors were biased at the drain currents listed in Table 5.3 which corresponds to $I_{DSS}/4$ at both temperatures. This ratio must remain constant in order to allow a valid comparison of the devices' temperature dependence. Also note that the MESFET was biased at a 3.5V V_{DS} while the HEMT was at 3.0V V_{DS} corresponding to the manufacturer's suggested drain voltages.

Table 5.3 BIAS LEVELS

		$I_D(mA)$	$I_{DSS}(mA)$	$g_m(mho)$
MESFET	290K	20	82	0.04
	77K	30	121	0.04
HEMT	290K	12	47	0.04
	77K	18	70	0.05

Figure 5.28 TRANSFER CURVES

a) H503-2 I_{ds} (mA) vs V_{ds} (V) SYMBOLS (290K) LINES (77K)

b) 0503 I_{ds} (mA) VS V_{ds} (V) SYMBOLS (290K) LINES (77K)

Note the large deviations in Figure 5.28 (a) in the saturation current of the HEMT as V_{DS} is varied at 77K. This is due to the device becoming light sensitive at this temperature. The transfer characteristics were manually measured and during the time it took to complete a range of V_{GS} values at a constant V_{DS} value, the ambient light level changed due to varying cloud cover. Since the light level changed faster than V_{DS} was changed, the device's saturation current fluctuated during the experiment. The more light the device was exposed to, the higher the saturation current became.

This phenomenon has been confirmed by other researchers and is explained by the "deep trap" theory in which mobile conduction electrons are removed from the channel by combining with energy states near the middle of the energy band gap created by lattice impurities or imperfections. At low temperatures, the electrons, once captured, remain immobile for long lengths of time since they do not possess much thermal energy. Hence, the HEMT is showing signs of poor processing which may be due to impurities or lattice imperfections.^[68]

When the HEMT was placed in a light tight container, the saturation current decreased to 60 mA while the low frequency transconductance remained at 0.05 mho and the gain of the device dropped uniformly by 3 dB over the 2-22 GHz band. Hence, in the following results, the HEMT was operated in ambient light.

The forward gain of the HEMT and MESFET at both 290K and 77K are shown in Figure 5.29 and 5.30, respectively. Note the large amount of noise on Figure 5.29 due to a malfunctioning 8510. Since this instrument was serviced after this set of measurements, further experimentation was impossible. The results, although imperfect, are shown and only general trends will be considered. Figure 5.30 appears smoother since the results are averaged. Both devices' gain increased when immersed in liquid nitrogen. The HEMT's forward gain improved from 9 to 11 dB at 2 GHz and over the full band, the improvement is approximately constant at the same 2 dB. The MESFET's forward gain improved from 10.2 to 10.3 dB at 2 GHz and over the full band, it increased by 1 dB.

Figure 5.29 HEMT GAIN vs TEMPERATURE

Figure 5.30 MESFET GAIN vs TEMPERATURE

The raw s-parameter data was further averaged and then de-embedded to obtain the transistors' s-parameters at 1 GHz intervals under 10 GHz. Above this frequency, the noise made the results meaningless. The maximum available gain or maximum stable gain was then calculated using Super Compact and the final results, for both temperatures, are shown for the HEMT in Figure 5.31 (a) and for the MESFET in Figure 5.31 (b). It is evident that at the lower temperature the HEMT exhibited a 2 dB gain increase whereas the MESFET showed none.

Figure 5.31 MAXIMUM GAIN vs TEMPERATURE

a) HEMT MSG (dB) or MAG (dB) vs FREQUENCY (GHz)

b) MESFET MSG (dB) or MAG (dB) vs FREQUENCY (GHz)

The noise figure of the devices was also measured at the two temperatures over 8 to 14 GHz. The results are shown in Figure 5.32 (a) for the HEMT and in 5.32 (b) for the MESFET. Both sets of curves appear to be flat implying that over the frequency band, the 50 Ω match may be at the position of the same constant noise contour for the various frequencies in the band. Overall, the noise figure for both devices decreases approximately 1 dB when they are cooled to 77K.

Figure 5.32 NOISE FIGURE vs TEMPERATURE

a) HEMT NF (dB) vs FREQUENCY (GHz) and TEMPERATURE

b) MESFET NF (dB) vs FREQUENCY (GHz) and TEMPERATURE

The results at 12 GHz, using the 8970A for both gain and noise figure measurements, are summarized in Table 5.4. Although the HEMT's performance is better at both temperatures, it did not dominate as expected. This may be explained by considering the structure of the HEMT, specifically the ohmic contacts. The ohmic contacts rely on a large activated donor concentration for their low resistance. At low temperatures, the percentage of donors ionized will decrease due to the lower thermal energy of the lattice and the resistance of the ohmic contacts will increase.^[45] Hence, R_D and R_S in the small signal model will increase which increases the loss at the gate input degrading the overall performance. It was observed that the input reflection coefficient decreased at the lower temperatures confirming this effect. Hence, although the channel may operate ballistically, the complete structure must be considered in determining the overall performance gain.

Table 5.4 HEMT and MESFET TEMPERATURE DEPENDENCE

	HEMT		MESFET	
	290K	77K	290K	77K
NF (dB)	3.9	3.1	4.7	3.1
S21 (dB)	5.9	8.1	3.9	6.0
Gmax (dB)	7.0	10.6	7.2	11.5

The experiment was further complicated by the cracking of the HEMTs when they were immersed in liquid nitrogen. During the thermal tests, two of three devices cracked along the channel after one thermal cycle. Figure 5.33 shows a SEM of a HEMT that worked until after it was removed from the coolant. Cracks in the active layer are clearly visible in the channel running parallel to the gate electrode. When re-tested, this device had a saturation current of only 6 mA compared to its original value of 50 mA. The MESFET, on the other hand, survived over fifteen thermal cycles and is still working. To explain why the HEMT was so fragile, the structural differences between the devices will be considered. As previously stated, the only differences between the devices are the passivation layer and the vertical structure. Hence, either the passivation layer on the MESFET is acting like a thermal barrier to reduce the thermal gradient or lattice imperfections are creating weak regions which crack when stressed.

Figure 5.33 CRACKS in a THERMALLY CYCLED HEMT

5.5 CONCLUSION

The manufacturer's HEMT and MESFET s-parameters were verified experimentally. These values were used to design two experiments.

The first experiment to determine Γ_0 for the devices did not provide the expected accuracy. Γ_s values far from Γ_0 limited the results' usefulness and a modified experiment would be required to obtain better results. Overall, the noise figure of the HEMT was 0.5 dB lower than the MESFET over the 8 to 14 GHz band. Also, using 70% of the DC power of the MESFET, the transconductance of the HEMT was 10% higher.

The second experiment involving liquid nitrogen testing of the devices was disappointing. The expected dominance of the HEMT over the MESFET due to ballistic carrier flow was not observed due probably to the increased resistance of the ohmic contacts. Also, the HEMT became light sensitive when immersed in liquid nitrogen which may reflect processing problems. In the dark at 77K, the HEMT showed no performance gain compared to room temperature.

The DC parameters of the HEMT varied widely over three batches of received devices. Clearly Gould is having trouble in manufacturing the world's first commercially available HEMT.

Overall, the fact that the HEMT with a 0.5 μm gate length outperformed the 0.3 μm MESFET is remarkable. The performance of a 0.3 μm gate length HEMT should be even better making the use of HEMTs in future microwave LNA circuits look promising. For look-up radar applications, a device such as the HEMT which reduces the system noise and at the same time decreases the number of amplification stages will be a definite improvement over existing technology.

It is predicted that monolithic processing will soon incorporate HEMT circuit elements. If this occurs, MMICs using HEMTs will be a viable option for circuits operating above 20 GHz. Also, lower frequency circuits such as LNAs will definitely apply this new technology in many systems, including T/R modules.

CHAPTER 6

CONCLUSION and RECOMMENDATIONS

6.1 SUMMARY

Two studies, one on space-based radar and another on a related topic, the T/R module were undertaken in this thesis. In addition, microwave circuits, which may be applied in T/R modules, were developed.

This thesis is divided into two parts, a study section and an experimental section. The first study is a space-based radar analysis in which the feasibility of mounting a radar on a spaceborne platform in order to detect low flying aircraft was determined. The second study discussed active array antennas. The key component for such an array, the T/R module, was described and an 'optimum' module architecture was proposed. The experimental work concentrated on developing some microwave circuits that are required for the proposed T/R module. Circuits at both C-band and Ku-band were constructed using MMIC technology and MIC technology, respectively. The results of this work will now be summarized.

A space-based radar analysis in which the transmitter power requirements were determined is described in Chapter 2. Both monostatic and bistatic radar architectures were analysed for a hypothetical problem of detecting small radar cross section targets flying through Canada's North. Both analyses began with no preconceptions. The analyses required the best topology be chosen using an iterative technique. The final solutions for both architectures are detailed in this chapter.

Initially, the doppler, coverage and slant range for a radar mounted on a platform in low earth orbit were determined. The velocity of the platform was nominally 7 km/s (14,000 knots) and relatively insensitive to altitude variations. For an L-band radar, this led to large doppler bandwidths of ± 70 kHz. The coverage area, forming an annulus on the ground limited by the wide swath at small grazing angles and the high altitude return at nadir, was found to be a large fraction of the earth. Also, the slant range was found to vary from two to four times the platform's altitude and was strongly dependent on the altitude as expected. These large values of slant range created large path losses as well as echo delays of tens of milliseconds. Compared to a conventional airborne radar

such as AWACS, a radar in low earth orbit was found to have a slant range, clutter smear and a coverage area many times larger.

Due to the limited power available in a spaceborne system, the monostatic analysis was performed at the lowest possible altitude and tried to exploit the unique velocity and slant range effects resulting from using an orbiting platform. Recognizing the additional loss inherent in using separate transmit and receive apertures for the bistatic architecture, the bistatic analysis concentrated on a novel arrangement similar to that of a line-of-sight communications link. Here, ground-based transmitters and a receiver in geosynchronous orbit were used.

The monostatic analysis begins with the development of the chosen radar mechanization in which a high PRF waveform, active planar array antenna and coherent pulse integration topology was used. The high PRF waveform allowed the dynamic clutter interference, distributed over most of the doppler band, to be isolated and removed by standard doppler processing techniques. The active array antenna allowed sequential detection of targets flying through a fixed fence which allowed coherent pulse integration to be fully exploited in order to reduce the transmitter's power requirement.

One unique result of using a high PRF waveform is that only side lobes close to the main beam of the antenna pattern needed to be minimized. Side lobes far from the main beam, even grating lobes, illuminate clutter outside of the target's possible doppler band which allowed them to be removed with doppler processing.

With this mechanization, in order to detect a small radar cross section target in a noise limited environment, the prime power was calculated to be 69 kW. This was a 'best case' calculation where the effect of increased false alarms due to clutter interference was not considered. Only the qualitative effect of clutter interference on increasing the probability of false alarm was discussed. However, it is clear that even without considering this effect in the power calculation, the monostatic architecture resulted in a system that would require quantum jumps in present space technology. Even if the revolutionary advances in spaceborne power generation and signal processing were accomplished, this implementation would still be vulnerable to jamming. The usual burn-through caused by the R^4 radar and the R^2 jammer power relationships would not occur for a radar in low earth orbit due to its fixed altitude.

Next the bistatic analysis was detailed. A proposed topology using ground-based transmitters and a geostationary receiver reduced the two-way path loss to a more manageable one-way, two-hop loss. This static configuration removed the large doppler bandwidth and its associated dynamic clutter interference problem that plagued the monostatic analysis.

The analysis, assuming a 10 dB radar cross section enhancement due to the forward backscatter phenomenon and a thermal noise limited environment, resulted in a power requirement of approximately 2 kW for each of the eighty ground transmitters. Note that the spaceborne receiver, being passive, would have a low power requirement. It is believed the required power would be similar in magnitude to that of a commercial communications satellite.

Of the two space-based radar architectures studied, the bistatic structure appeared to be the preferred system. The novel structure of the bistatic radar fulfills the hypothetical surveillance requirement for Northern Canada while requiring standard ground transmitters and a space receiver that would require an evolutionary improvement of current space technology. The only high risk development for the bistatic radar would be the construction of the receiver's 1 km long antenna in space.

Since both the monostatic and bistatic radars required active array antennas in order to achieve their highest performance, this topic was the subject of the second study in Chapter 3. This chapter discussed passive array antenna theory, active array antenna theory and the T/R module which is the key component of any active array.

Initially, the far-field antenna pattern of an array antenna was derived. It was shown that the main beam pattern is simply the Fourier transform of the spatial excitation of the aperture elements. By linearly grading the phase of the radiation at the array elements the main beam may be squinted to angles off boresight. For an array squinted at large angles, the grating lobe phenomena limits the spacing of the elements to approximately half a wavelength. The ability to weight the aperture by varying the magnitude and phase of the excitation across the array was then discussed. Having many individually controllable elements allows the designer many degrees of freedom in the synthesis of beam patterns. This has led to the dominance of array antennas in high performance systems. A discussion of the use of separate feed networks to allow multiple beams to be formed simultaneously from the same aperture followed. The main limitation of this method, the large power loss resulting from the division of power between the multiple feed networks, is overcome by using active array antennas.

Here, a low noise receiver at each element fixes the noise figure of the system negating the effect of feed losses.

The structure of an active array antenna requiring a T/R module at each of the radiating elements was then discussed. Many of the advantages and applications of the active array are then listed. With respect to the earlier space-based radar studies, inertialess scan allowing sequential detection, multiple beams allowing aperture re-use and the high reliability of a gracefully degrading solid state system are its main advantages.

The rest of the chapter detailed a proposed internal architecture for a T/R module. The proposed module was based on that required for the monostatic L-band radar discussed in Chapter 2. The module was broken into six sub-assemblies including a phase shifter, low power switch, solid state power amplifier, high power T/R switch, integral antenna and low noise receiver. The design is novel in that both MIC and MMIC technologies were used in the module. MIC technology, which possesses the lowest noise and highest power capabilities, was used in the low noise receiver and high power amplifier whereas MMIC technology, which allows complex repeatable circuits to be constructed in small packages, was used in the remaining sub-assemblies.

Common microwave circuits required to implement these sub-assemblies were then discussed. Among the most promising new developments are the segmented-dual-gate MESFET which allows a high resolution vector phase shifter to be constructed on a very small circuit and the HEMT which allows low noise receivers to significantly reduce their noise contribution.

Since funding was not available to construct the proposed T/R module, only some of the key circuits in the T/R module were developed. This experimental work included the construction of MMIC circuits in Chapter 4 and MIC circuits in Chapter 5.

Actively matched, C-band intermediate power level amplifiers were developed using MMIC technology in Chapter 4. The chapter begins with an explanation of MMIC processing and the structure of the circuit elements that are available from a commercial foundry. This is followed by microwave amplifier design theory and a brief discussion of the computer aided design program SuperCompact. The first of two MMIC designs was described in detail allowing the design process to be understood. The second design, being similar, was described briefly.

Both designs made use of active matching on their input port and reactive matching on their output port. The designs differed mainly in their bias circuits. One circuit used individual bonding pads to bias the individual transistors whereas the other used a resistive divider circuit to derive the bias levels from a single +15 V supply. Design #1, using the individual bias connections, allowed the input transistor to be pinched off to form an RF switch with an isolation of 22 dB. Design #2, requiring only a single external bias level, successfully tested the DC processing tolerances of the foundry.

RF testing of at least two working devices of each design was completed. This work was hampered by the poor processing quality control exhibited by the foundry. Both airbridges and active layer implants were at a research level of refinement. The active matching topology performed better than expected providing a low input return loss from 2 to 8 GHz. However, the high Q of the designed circuits magnified any processing or simulation errors causing the original simulation to differ from the experimental results. The simulations were made to agree with experiment by increasing the size of the tuning inductors by a small fraction of a turn. Considering the close spacing between circuit elements in the designs and the lack of surface coupling being considered in the simulation, the experimental results were reasonably close to those expected.

With respect to T/R modules, clearly MMIC technology with its small size, light weight and low volume cost appears promising. However, using the commercial design tools available today, it is unlikely that a successful narrow band circuit could be designed in the first attempt. Such a design would probably require empirically corrected designs based on the information from several processing iterations.

Chapter 5 discussed the development of Ku-band MIC circuits in which similar MESFET and HEMT devices were compared. The chapter begins with a discussion of the theory of distributed element matching, HEMT operation and transistor amplifier noise. Applying this theory, two experiments were conceptualized and implemented. In one experiment, the noise parameters Γ_0 and r_n were determined for both devices while the other less practical experiment investigated the effect of cryogenic temperature on the devices.

Although the first experiment allowed Γ_0 and r_n to be determined, the accuracy of the results was degraded by an incomplete simulation of the microstrip test circuit. In particular, the parasitic coupling of an array of unused input stubs altered the

impedance that was presented to the devices by a significant amount. During this experiment, it was found that the HEMT produced lower noise and higher gain characteristics than the MESFET. Considering that the HEMT's gate length was two-thirds longer than the MESFET's, these results were remarkable. Hence, at room temperature, the superiority of the HEMT that theory predicted was verified.

The second experiment in which the devices were characterized in liquid nitrogen did not produce the expected results. Instead of seeing the dominance of the HEMT due to ballistic carrier flow, the HEMT became light sensitive. This light sensitivity was attributed to the poor structural quality of these early devices. In the dark, the HEMT showed no performance gain at 77K compared to room temperature. When illuminated in ambient indoor light and immersed in liquid nitrogen, the gain of the HEMT jumped two dB higher than at room temperature while the MESFET's gain remained constant. The noise figure of both devices dropped by approximately the same amount, one dB over an 8 to 14 GHz band, when cooled.

These experiments were complicated by the fact that the HEMT was a recently developed device. The DC parameters of several batches of the devices varied considerably and, when thermally cycled, two out of three HEMTs cracked leading the author to believe that the manufacturer was having early production problems. Overall, although the cryogenic experiment was disappointing, the high gain and low noise performance of the HEMT looks promising. Once the manufacturing process is further refined, it is believed that the HEMT will be applied increasingly in the low noise amplifiers of many systems including active array radars.

6.2 FUTURE RESEARCH WORK

In Chapter 2 it was noted that the monostatic radar analysis assumed the best case of being noise interference limited. Considering that only a small portion of the doppler band is free of ground clutter interference, this analysis probably underestimated the probability of false alarm. Hence, the performance of the monostatic radar should be worse than calculated.

To determine the effect of ground clutter interference, a study to calculate the degradation it causes should be attempted. This analysis will be quite involved with the magnitude of the ground clutter varying with grazing angle and terrain type and the ground illumination being a dynamic function of the radar's position over the fence and the element of the fence being scanned.

As noted in the experimental section of the thesis, the T/R module was not constructed. Although many circuits including the vector modulator phase shifter and the power amplifier would provide an opportunity for future work, only experiments based on lessons learned in performing the experimental work of Chapter 4 and 5 will be proposed.

With respect to the MMIC work of Chapter 4, the designs should be modified to account for the simulation's underestimation of the inductors' actual values. The inductors should be reduced in size by the values stated in the chapter to obtain an empirically modified design that should operate in the original frequency band. Such an experiment will verify the ability to empirically compensate for the unrefined modeling tools.

Due to space limitations, only an active input match was constructed. However, a modified design using a third stage to form an active output match should result in a wide band amplifier design that would be more tolerant of processing variations. Also, it would be interesting to redesign the circuits to operate at L-band in order to observe the size reduction that active matching should produce as compared with reactive matching.

Although Γ_0 and r_n were determined for the HEMT and MESFET in Chapter 5, the accuracy of the results is questionable. In order to improve these results, the noise parameter determination experiment should be modified. To begin, a one piece jig with the input/output lines halved in length should eliminate the problem of jig resonance and reduce the input/output losses which reduced the accuracy of the noise figure

measurements. Also, a series of single stubs mounted on separate substrates should be presented to the input of the devices one at a time. This will eliminate the parasitic effect of the unconnected stubs of the stub array that plagued the experiment discussed in the chapter. Since the approximate location of Γ_0 is known, fewer stubs will be required than were used in this thesis.

Overall, solid state microwave circuit technology is advancing rapidly due mainly to semiconductor processing technology developments. It is believed that HEMTs will soon become available as part of monolithic foundry processes allowing complex, low noise and high gain MMICs to operate well above 20 GHz. Similarly, progress in depositing binary and tertiary compound semiconductors on inexpensive and tough silicon substrates should allow a higher level of integration to be achieved. The day that an entire T/R module may be constructed on a single substrate is fast approaching. When this occurs, radar systems using active arrays will become widespread.

APPENDIX-I

I-1

SUPER-COMPACT Version 1.8 + 002 10/17/85 25-MAY-85 10:50:27

SCG.DAT

NOI

TWO 1 2 3 FET300 X16

IND 3 1 L=0.3MM

IND 3 1 L=0.1MM

FETCG: ZFOR 3 2

END

FREQ

STEP 25MHZ 180MHZ 10MHZ

END

OUT

PRI FETCG :

END

DATA

FET300:5

0.971	-24.8	1.994	157.0	0.032	75.8	0.828	-8.3
0.946	-36.5	1.915	144.2	0.044	69.3	0.817	-12.3
0.902	-47.5	1.820	134.0	0.058	63.5	0.804	-15.7
0.842	-57.7	1.717	124.4	0.069	58.5	0.791	-19.0
0.822	-67.0	1.614	118.0	0.077	54.2	0.778	-22.1
0.785	-75.4	1.515	110.0	0.083	50.5	0.764	-25.0
0.750	-83.5	1.422	102.4	0.089	47.5	0.754	-27.8
0.720	-90.8	1.337	95.8	0.093	45.0	0.747	-30.5
0.693	-97.6	1.259	89.3	0.094	42.9	0.739	-33.1
0.476	-103.8	1.189	83.5	0.099	41.2	0.733	-35.7
0.449	-109.4	1.125	78.0	0.101	39.9	0.727	-38.3
0.432	-115.0	1.067	72.8	0.103	39.0	0.722	-40.9
0.617	-120.0	1.015	67.8	0.105	38.2	0.718	-43.5
0.603	-124.8	0.967	63.2	0.106	37.8	0.715	-46.0
0.592	-129.3	0.924	58.7	0.107	37.5	0.712	-48.4
0.583	-133.5	0.884	54.5	0.108	37.4	0.710	-51.2
0.574	-137.5	0.847	50.5	0.109	37.5	0.707	-53.8

NOI RM 50

40KHZ 1.3 .5 103 .25

80KHZ 2.3 .46 144 .21

800000 NOISE DATA

END

CIRCUIT: FETCG

S-MATRIX, ZS = 50.0+J 0.0 ZL = 50.0+J 0.0

Free	S11	S21	S12	S22	S21	
GHz	Mag	Ang	Mag	Ang	Mag	
2.00000	0.041	142.5	1.052	-11.5	0.082	15.1
3.00000	0.049	133.9	1.050	-17.3	0.088	20.7
4.00000	0.060	150.5	1.040	-23.3	0.094	25.1
5.00000	0.094	146.3	1.044	-29.5	0.103	28.8
6.00000	0.112	142.9	1.039	-35.8	0.112	29.9
7.00000	0.132	140.0	1.032	-42.3	0.121	31.0
8.00000	0.133	134.9	1.023	-49.2	0.129	31.5
9.00000	0.182	134.3	1.012	-56.5	0.137	31.3
10.00000	0.212	131.4	0.998	-64.1	0.144	30.7
11.00000	0.244	128.4	0.981	-72.1	0.149	30.3
12.00000	0.280	125.1	0.959	-80.6	0.153	29.6
13.00000	0.319	121.4	0.933	-89.4	0.156	28.9
14.00000	0.357	117.8	0.903	-99.1	0.158	28.0
15.00000	0.399	113.7	0.868	-109.1	0.158	27.3
16.00000	0.440	109.5	0.831	-119.7	0.159	26.7
17.00000	0.480	105.2	0.791	-130.7	0.160	26.1
18.00000	0.518	100.8	0.748	-142.1	0.162	25.8

CIRCUIT: FETCG

Free	K	S11	S21	S12	S22
GHz	dB	Mag	Ang	Mag	Ang
2.00000	0.908	+	11.08		
3.00000	0.817	+	10.77		
4.00000	0.712	+	10.40		
5.00000	0.610	+	10.04		
6.00000	0.510	+	9.67		
7.00000	0.417	+	9.29		
8.00000	0.325	-	8.89		
9.00000	0.243	-	8.48		
10.00000	0.169	-	8.02		
11.00000	0.099	-	7.51		
12.00000	0.024	-	6.96		
13.00000	-0.053	-	6.38		
14.00000	-0.087	-	5.78		
15.00000	-0.119	-	5.17		
16.00000	-0.152	-	4.54		
17.00000	-0.186	-	3.89		
18.00000	-0.223	-	3.23		

SUPER-COMPACT Version 1.3 + 001 10 17 85 1-NOV-85 11:37:07

* 3STAGE, 61055 FET MODEL

FET 1 G=0.028 CGS=.021FF T=4E-12 IDG=.021FF DGS=.069FF SDS=.00138
 + R1=10 RG=7.7 RD=4.4 RS=2.3 LG=.055NH LD=.035NH LS=.23NH
 FET300: 2P3K
 END
 FREQ
 STEP 20MHZ 130MHZ 13MHZ
 END
 OUT
 FRI FET300 S
 END

CIRCUIT: FET300

S-MATRIX, ZS = 50.0+J 0.0 ZL = 50.0+J 0.0

Freq	S11		S21		S12		S22		S21
GHZ	Mag	Ang	Mag	Ang	Mag	Ang	Mag	Ang	dB
2.00000	0.941	-28.8	2.262	151.9	0.024	79.8	0.827	-7.2	7.09
3.00000	0.881	-41.9	2.129	139.2	0.033	76.8	0.816	-10.4	6.56
4.00000	0.813	-54.0	1.980	127.8	0.039	75.9	0.804	-13.3	5.94
5.00000	0.745	-65.0	1.833	117.6	0.044	77.4	0.793	-16.0	5.23
6.00000	0.682	-75.2	1.695	108.5	0.049	81.2	0.783	-18.5	4.58
7.00000	0.626	-84.7	1.571	100.3	0.053	87.0	0.776	-21.0	3.93
8.00000	0.577	-93.7	1.462	93.0	0.060	93.7	0.770	-23.4	3.30
9.00000	0.534	-102.4	1.366	86.4	0.068	100.4	0.765	-25.9	2.71
10.00000	0.498	-110.8	1.281	80.5	0.080	106.2	0.762	-28.5	2.15
11.00000	0.468	-119.1	1.206	75.0	0.095	110.5	0.759	-31.2	1.63
12.00000	0.442	-127.3	1.139	70.0	0.113	113.5	0.757	-34.0	1.13
13.00000	0.422	-135.5	1.079	65.4	0.135	115.3	0.754	-37.0	0.66
14.00000	0.405	-143.6	1.025	61.3	0.159	116.0	0.752	-40.1	0.21
15.00000	0.392	-151.6	0.975	57.5	0.185	115.8	0.749	-43.3	-0.22
16.00000	0.383	-159.6	0.929	54.0	0.214	115.0	0.745	-46.7	-0.64
17.00000	0.376	-167.6	0.886	51.0	0.246	113.6	0.741	-50.3	-1.05
18.00000	0.372	-175.4	0.846	48.2	0.279	111.8	0.735	-54.1	-1.45

CIRCUIT: FET300

FREQ: FET500		S--- MAX GAIN ---S		S--- CONJUGATE MATCH GAMMA ---S				
Freq GHZ	K	SIGN BI	AVAILABLE	STABLE	SOURCE		LOAD	
			dB	dB	Mag	Ang	Mag	Ang
2.00000	0.392	+		19.76				
3.00000	0.593	+		18.12				
4.00000	0.797	+		17.00				
5.00000	0.996	+		16.16				
6.00000	1.172	+	12.92		0.839	87.9	0.891	25.7
7.00000	1.300	+	11.42		0.783	97.0	0.869	26.5
8.00000	1.358	+	10.33		0.753	105.3	0.864	27.8
9.00000	1.342	+	9.51		0.743	112.9	0.867	29.4
10.00000	1.272	+	8.90		0.750	119.9	0.878	31.2
11.00000	1.174	+	8.50		0.779	126.4	0.896	33.2
12.00000	1.071	+	8.39		0.841	132.5	0.928	35.3
13.00000	0.976	+		9.04				
14.00000	0.894	+		8.10				
15.00000	0.828	+		7.21				
16.00000	0.777	+		6.37				
17.00000	0.742	+		5.57				
18.00000	0.719	+		4.82				

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SUPER-COMPACT Version 1.8 + 001 10/17/85 21-OCT-85 14:48.13

* TWO ST

BLK

TWO 1 2 3 FET300 416

SF1 1 N=74.57 W=10UM S=6UM R=10UM SUB

IND 5 1 L=0.5NM

IND 3 1 L=0.5NM

TWO 1 0 5 STAGE

SF1 5 N=757 W=10UM S=6UM R=10UM SUB

CAP 5 7 F=1.24PF

FETCG: 2 FOR 5 7

END

FREQ

STEP 16MHZ 166MHZ 16MHZ

END

OUT

F=1 FETCG 5

END

DATA

SUB: MS N=250UM ER=12.0 MET1=AU 1.0UM

FET300:5

0.971	-24.8	1.994	157.0	0.032	75.8	0.828	-8.3
0.940	-34.5	1.915	144.2	0.044	69.3	0.817	-12.1
0.902	-47.5	1.820	134.0	0.058	63.5	0.804	-15.7
0.862	-57.7	1.717	126.6	0.069	58.5	0.791	-19.0
0.822	-67.0	1.614	118.0	0.077	54.2	0.778	-22.1
0.785	-75.6	1.515	110.0	0.083	50.5	0.764	-25.0
0.750	-83.5	1.422	102.6	0.089	47.5	0.754	-27.8
0.720	-90.8	1.337	95.8	0.093	45.0	0.747	-30.5
0.693	-97.6	1.259	89.5	0.096	42.9	0.739	-33.1
0.670	-103.8	1.189	83.5	0.099	41.2	0.733	-35.7
0.649	-109.6	1.125	78.0	0.101	39.9	0.727	-38.3
0.632	-115.0	1.067	72.8	0.103	39.0	0.722	-40.9
0.617	-120.0	1.015	67.8	0.105	38.2	0.718	-43.5
0.603	-124.8	0.967	63.2	0.106	37.8	0.715	-46.0
0.592	-129.3	0.924	58.7	0.107	37.5	0.712	-48.6
0.583	-133.5	0.884	54.5	0.108	37.4	0.710	-51.2
0.574	-137.5	0.847	50.5	0.109	37.5	0.707	-53.8

NOI RM 50

46MHZ 1.3 .5 103 .25

86MHZ 2.3 .46 146 .21

8000UM NOISE DATA

5 STAGE:5

0.941	-28.8	2.262	151.9	0.024	79.8	0.827	-7.2
0.881	-41.9	2.129	139.2	0.033	76.8	0.816	-10.4
0.813	-54.0	1.980	127.8	0.039	75.9	0.804	-13.3
0.745	-65.0	1.833	117.6	0.044	77.4	0.793	-16.0
0.682	-75.2	1.695	108.5	0.049	81.2	0.783	-18.5
0.624	-84.7	1.571	100.3	0.053	87.0	0.776	-21.0
0.577	-93.7	1.462	93.0	0.060	93.7	0.770	-23.4
0.534	-102.4	1.366	86.4	0.068	100.4	0.765	-25.9
0.498	-110.8	1.281	80.5	0.080	106.2	0.762	-28.5
0.468	-119.1	1.206	75.0	0.093	110.5	0.759	-31.2
0.442	-127.3	1.139	70.0	0.113	113.5	0.757	-34.0
0.422	-135.5	1.079	65.4	0.135	115.3	0.754	-37.0
0.405	-143.6	1.025	61.3	0.159	116.0	0.752	-40.1
0.392	-151.6	0.975	57.5	0.185	115.8	0.749	-43.3
0.383	-159.6	0.929	54.0	0.214	115.0	0.745	-46.7
0.374	-167.6	0.886	51.0	0.246	113.6	0.741	-50.3
0.372	-175.4	0.846	48.2	0.279	111.8	0.735	-54.1

END

OPT

FETCG F=46MHZ ME21=1500 ME11

TERM=.1

END

Gradient, Random or Quit? (G/R/Q):

G

Initial Variables and Gradients

Variables	Gradients
(1): 4.5000	(1): 121.14
(2): 5.0000	(2): 19.712
(3): 0.2000 PF	(3): 1.4973

Err. F.= 8.472

-----8888-----

Number of iterations? (N>0):

10

Print best values each iteration? (Y/N):

Err. F.= 1.592

```

Err. F. = 0.568
Err. F. = 0.373
Err. F. = 0.338
Err. F. = 0.134
(1): 4.9571
(2): 4.8329
(3): 0.17072 FF
Err. F. = 0.109
(1): 0.17088
(2): 0.15378
(3): 0.21309E-01

```

Gradient termination
 CFL time = 16.86 Secs.

172 Function evaluations.

CIRCUIT: FETCG
 S-MATRIX, ZS = 50.0+J 0.0 ZL = 50.0+J 0.0

Freq	S11		S21		S12		S22		S21
GHz	Mag	Ang	Mag	Ang	Mag	Ang	Mag	Ang	dB
2.00000	0.107	163.6	0.029	33.5	0.000	-11.9	0.997	-12.9	-30.71
3.00000	0.125	154.9	0.127	20.5	0.000	-3.4	0.989	-20.9	-17.93
4.00000	0.153	144.1	0.468	-2.1	0.001	-4.6	0.965	-32.0	-6.59
5.00000	0.192	124.6	1.894	-43.4	0.004	-24.5	0.861	-53.7	5.55
6.00000	0.086	78.9	5.403	-154.3	0.016	-113.3	0.115	2.2	14.65
7.00000	0.120	148.1	3.081	126.0	0.011	-170.0	0.559	-1.8	9.77
8.00000	0.180	144.6	1.941	85.5	0.009	172.7	0.657	-7.0	5.76
9.00000	0.234	140.0	1.360	57.8	0.008	167.9	0.725	-11.4	2.67
10.00000	0.290	134.7	1.030	36.5	0.008	168.7	0.769	-16.2	0.25
11.00000	0.345	128.9	0.827	18.4	0.008	172.6	0.795	-20.7	-1.65
12.00000	0.404	122.9	0.693	2.1	0.009	178.0	0.810	-24.9	-3.18
13.00000	0.462	116.9	0.599	-13.3	0.010	-175.3	0.817	-28.9	-4.45
14.00000	0.520	110.6	0.533	-28.5	0.012	-167.8	0.820	-32.5	-5.47
15.00000	0.579	104.1	0.483	-43.7	0.015	-160.8	0.819	-35.8	-6.32
16.00000	0.634	97.6	0.447	-59.3	0.019	-153.4	0.815	-38.9	-6.99
17.00000	0.683	90.9	0.421	-75.1	0.026	-152.5	0.809	-41.7	-7.51
18.00000	0.725	84.0	0.404	-91.5	0.036	-152.7	0.802	-44.2	-7.87

CIRCUIT: FETCG

Free		K	S-- MAX GAIN --S		S-- CONJUGATE MATCH GAIN --S			
GHz	SIGN		AVAILABLE	STABLE	SOURCE		LOAD	
			dB	dB	Mag	Ang	Mag	Ang
2.00000	999.900	+	0.00		0.107	-163.6	0.997	12.9
3.00000	510.371	+	-1.14		0.125	-154.5	0.989	20.9
4.00000	88.151	+	5.14		0.151	-142.3	0.965	32.0
5.00000	15.593	+	11.53		0.173	-118.8	0.860	53.7
6.00000	5.818	+	14.75		0.096	-80.0	0.123	-2.9
7.00000	9.983	+	11.44		0.093	-151.2	0.556	1.7
8.00000	16.027	+	8.34		0.171	-150.4	0.655	6.7
9.00000	21.020	+	6.15		0.233	-144.0	0.725	11.2
10.00000	23.642	+	4.53		0.291	-137.6	0.770	16.0
11.00000	24.570	+	3.25		0.347	-131.2	0.795	20.5
12.00000	23.939	+	2.24		0.406	-124.9	0.811	24.7
13.00000	22.101	+	1.40		0.465	-118.7	0.818	28.4
14.00000	19.149	+	0.76		0.522	-112.3	0.821	32.2
15.00000	15.557	+	0.28		0.580	-105.9	0.820	35.4
16.00000	11.730	+	-0.01		0.636	-99.4	0.816	38.3
17.00000	8.346	+	-0.11		0.688	-93.0	0.812	40.8
18.00000	5.718	+	-0.01		0.736	-86.5	0.810	42.7

CIRCUIT: FETCG

STABILITY CIRCLES

Free	LOCATION	RAD	STB	LOCATION	RAD	STB	GA	MAX	K	San
GHz	Mag	Ang	REG	Mag	Ang	REG	dB	dB	BI	
2.00000	9.350	-163.6	0.010	1.003	12.9	0.000	OUT	0.0	999.90	+
3.00000	8.022	-154.5	0.062	1.011	20.9	0.000	OUT	-1.1	510.37	+
4.00000	6.635	-142.3	0.244	1.036	32.0	0.000	OUT	5.1	88.15	+
5.00000	5.985	-118.8	1.079	1.163	53.7	0.011	OUT	11.5	15.59	+
6.00000	53.143	-80.0	47.633	16.058	-2.9	11.232	OUT	14.7	5.82	+
7.00000	13.071	-151.2	8.067	1.809	1.7	0.113	OUT	11.4	9.98	+
8.00000	6.041	-150.4	1.071	1.528	6.7	0.042	OUT	8.3	16.03	+
9.00000	4.340	-144.0	0.420	1.380	11.2	0.022	OUT	6.2	21.02	+
10.00000	3.452	-137.6	0.230	1.300	16.0	0.015	OUT	4.5	23.64	+
11.00000	2.888	-131.2	0.149	1.257	20.5	0.012	OUT	3.3	24.57	+
12.00000	2.469	-124.9	0.106	1.234	24.7	0.011	OUT	2.2	23.94	+
13.00000	2.157	-118.7	0.082	1.223	28.4	0.011	OUT	1.4	22.10	+
14.00000	1.919	-112.3	0.070	1.219	32.2	0.013	OUT	0.8	19.15	+
15.00000	1.727	-105.9	0.064	1.221	35.4	0.016	OUT	0.3	15.56	+
16.00000	1.577	-99.4	0.063	1.227	38.3	0.021	OUT	0.0	11.73	+
17.00000	1.459	-93.0	0.067	1.234	40.8	0.031	OUT	-0.1	8.35	+
18.00000	1.367	-86.5	0.075	1.239	42.7	0.047	OUT	0.0	5.72	+

Gradient, Random or Quit? (G/R/Q):

SUPER-COMPACT Version 1.0 f 002 10/17/85

22-NOV-85 10:46:09

* FIN

CN: 600PF

BLK

TWO 1 2 3 FET300 X10

TRL 2 22 W=36UM P=75UM SUB2

TRL 22 33 W=18UM P=120UM SUB2

SPI 33 12 N=4 W=10UM S=6UM R=16UM SUB

TFL 12 18 W=40UM P=60UM SUB2

TFC 17 18 L=91.3UM W=91.3UM CS-CN SUB Q=20 F=80KHZ

IND 18 L=0.3MM

TRL 13 17 W=18UM P=30UM SUB

IND 10 8 L=0.3MM

TRL 9 1 W=36UM P=60UM SUB2

TRL 3 44 W=30UM P= 96UM SUB2

SPI 44 9 N=4 W=6UM S=6UM R=12UM SUB

IND 9 L=0.3MM

TFC 44 88 L=91.3UM W=91.3UM CS-CN SUB Q=20 F=80KHZ

TRL 88 99 W=20UM P= 39UM SUB

IND 99 L=0.2MM

TFC 22 55 L=91.3UM W=91.3UM CS-CN SUB Q=20 F=80KHZ

RES 55 14 R=500OHM

TRL 55 5 W=18UM P= 45UM SUB2

TFC 14 13 L=91.3UM W=91.3UM CS-CN SUB Q=20 F=80KHZ

IND 13 L=0.2MM

TWO 5 6 4 SSTAGE

TRL 4 W=120UM P=120UM SUB2

SPI 64 15 N=4.25 W=10UM S=6UM R=16UM SUB

TRL 15 16 W=40UM P=160UM SUB2

IND 16 L=0.3MM

TFC 16 19 L=91.3UM W=91.3UM CS-CN SUB Q=20 F=80KHZ

TRL 19 20 W=50UM P=80UM SUB2

IND 20 L=0.2MM

TRL 6 64 W=30UM P=150UM SUB2

TFC 64 77 L=719.75UM W=18UM CS-CN SUB Q=30 F=80KHZ

TRL 77 78 W=40UM P=270UM SUB2

IND 78 7 L=0.3MM

FETC8: 2POR 10 7

END

FREQ

*STEP 2.90KHZ 3.10KHZ 100KHZ

*STEP 100KHZ 10KHZ 250KHZ

STEP 5.50KHZ 6.00KHZ 500KHZ

*STEP 5.70KHZ 6.60KHZ 1000KHZ

END

OUT

PRI FETC8 S

END

DATA

SUB1: MS H=250UM ER=12.6 MET1=AW 1.0UM

SUB2: MS H=250UM ER=12.6 MET1=AW 1.5UM

FET300:S

26KHZ	0.971	-24.8	1.994	157.0	0.032	75.8	0.828	-8.3
30KHZ	0.940	-34.5	1.915	144.2	0.046	69.3	0.817	-12.1
40KHZ	0.902	-47.5	1.829	134.0	0.068	63.5	0.804	-15.7
50KHZ	0.862	-57.7	1.717	126.4	0.069	58.5	0.791	-19.0
60KHZ	0.822	-67.0	1.614	118.0	0.077	54.2	0.778	-22.1
70KHZ	0.785	-75.4	1.515	110.0	0.083	50.5	0.764	-25.0
80KHZ	0.750	-83.5	1.422	102.4	0.089	47.5	0.754	-27.8
90KHZ	0.720	-90.8	1.337	95.0	0.093	45.0	0.747	-30.5
100KHZ	0.693	-97.6	1.259	89.5	0.096	42.9	0.739	-33.1
110KHZ	0.670	-103.8	1.189	83.5	0.099	41.2	0.733	-35.7
120KHZ	0.649	-109.6	1.125	78.0	0.101	39.9	0.727	-38.3
130KHZ	0.632	-115.0	1.067	72.8	0.103	39.0	0.722	-40.9
140KHZ	0.617	-120.0	1.015	67.8	0.105	38.2	0.718	-43.5
150KHZ	0.603	-124.8	0.967	63.2	0.106	37.8	0.715	-46.0
160KHZ	0.592	-129.3	0.924	58.7	0.107	37.5	0.712	-48.6
170KHZ	0.583	-133.5	0.884	54.5	0.108	37.4	0.710	-51.2
180KHZ	0.574	-137.5	0.847	50.5	0.109	37.5	0.707	-53.8

NOI RN 50

40KHZ 1.3 .5 103 .25

80KHZ 2.3 .46 144 .21

8000UM NOISE DATA

SSTAGE:S

26KHZ	0.941	-20.8	2.242	151.9	0.024	79.8	0.827	-7.2
30KHZ	0.881	-41.9	2.129	139.2	0.033	76.8	0.814	-10.4
40KHZ	0.813	-54.0	1.980	127.8	0.039	75.9	0.804	-13.3
50KHZ	0.745	-65.0	1.833	117.4	0.044	77.4	0.793	-16.0
60KHZ	0.682	-75.2	1.695	108.5	0.049	81.2	0.783	-18.5
70KHZ	0.626	-84.7	1.571	100.3	0.053	87.0	0.776	-21.0

```

800K 0.577 -93.7 1.442 93.0 0.060 93.7 0.770 -23.4
900K 0.534 -102.4 1.346 84.4 0.048 100.4 0.745 -25.9
100K 0.498 -110.8 1.281 75.5 0.080 104.2 0.742 -28.5
110K 0.448 -119.1 1.204 65.0 0.095 110.5 0.759 -31.2
120K 0.442 -127.3 1.139 70.0 0.113 113.5 0.757 -34.0
130K 0.422 -135.5 1.079 65.4 0.135 115.3 0.754 -37.0
140K 0.405 -143.6 1.025 61.3 0.159 116.0 0.752 -40.1
150K 0.392 -151.6 0.975 57.5 0.185 115.8 0.749 -43.3
160K 0.383 -159.6 0.929 54.0 0.214 115.0 0.745 -46.7
170K 0.376 -167.6 0.884 51.0 0.244 113.6 0.741 -50.3
180K 0.372 -175.4 0.846 48.2 0.279 111.8 0.735 -54.1
END
OPT
FETCB F=5.00K 6.00K MS21=1500 W=10 MS11 W=40000
FETCB F=6.10K 5.20K MS21=1500 W=5 MS11 W=1000
FETCB F=6.30K 6.50K MS21=1500 W=100 MS11 W=10
TERN=.1
END

```

CIRCUIT: FETCB
S-MATRIX, ZS = 50.0+J 0.0 ZL = 50.0+J 0.0

Free GHz	S11 Mag	Angle	S21 Mag	Angle	S12 Mag	Angle	S22 Mag	Angle	S21 dB
5.50000	0.172	83.0	3.812	-117.5	0.011	-78.4	0.519	-111.5	11.42
5.55000	0.164	79.9	3.786	-123.0	0.011	-83.4	0.475	-115.8	12.01
5.60000	0.155	74.9	4.139	-130.2	0.012	-88.6	0.430	-120.1	12.34
5.65000	0.145	74.2	4.248	-136.7	0.012	-93.8	0.382	-124.5	12.60
5.70000	0.134	71.9	4.349	-143.2	0.013	-99.1	0.333	-129.0	12.81
5.75000	0.122	70.1	4.441	-149.7	0.013	-104.3	0.285	-133.3	12.95
5.80000	0.110	69.0	4.484	-156.0	0.013	-109.4	0.237	-137.6	13.03
5.85000	0.098	69.0	4.500	-162.2	0.014	-114.3	0.191	-141.9	13.06
5.90000	0.087	70.1	4.490	-168.2	0.014	-119.1	0.147	-146.3	13.04
5.95000	0.077	72.7	4.459	-174.0	0.014	-123.6	0.105	-151.1	12.98
6.00000	0.068	77.1	4.410	-179.6	0.014	-127.9	0.067	-157.3	12.89
6.05000	0.061	83.3	4.348	-175.1	0.014	-132.0	0.032	-170.0	12.77
6.10000	0.056	91.3	4.274	-170.0	0.014	-135.9	0.011	-165.7	12.62
6.15000	0.054	100.3	4.193	-165.1	0.014	-139.5	0.034	-171.5	12.45
6.20000	0.054	109.4	4.107	-160.4	0.014	-142.8	0.040	-177.4	12.27
6.25000	0.054	117.5	4.018	-156.0	0.013	-146.0	0.085	-183.0	12.08
6.30000	0.059	124.3	3.927	-151.7	0.013	-149.0	0.100	-189.9	11.88
6.35000	0.063	129.5	3.834	-147.7	0.013	-151.8	0.129	-197.7	11.68
6.40000	0.068	133.6	3.745	-143.8	0.013	-154.4	0.149	-206.0	11.47
6.45000	0.074	136.6	3.655	-140.0	0.013	-156.9	0.167	-214.6	11.26
6.50000	0.079	138.8	3.567	-136.4	0.013	-159.2	0.184	-223.3	11.05
6.55000	0.085	140.5	3.481	-132.9	0.012	-161.4	0.201	-232.0	10.84
6.60000	0.090	141.7	3.398	-129.6	0.012	-163.4	0.216	-240.7	10.62
6.65000	0.095	142.5	3.316	-126.4	0.012	-165.4	0.231	-249.3	10.41
6.70000	0.100	143.1	3.237	-123.2	0.012	-167.2	0.245	-257.7	10.20
6.75000	0.105	143.6	3.161	-120.2	0.012	-169.0	0.258	-266.0	10.00
6.80000	0.110	143.8	3.087	-117.3	0.012	-170.6	0.271	-274.2	9.79

CIRCUIT: FETCB

Free GHz	K	S11 dB	MAX GAIN AVAILABLE dB	STABLE	SOURCE Mag	Angle	LOAD Mag	Angle
5.50000	8.945	+	13.08		0.145	-89.6	0.513	-111.6
5.55000	8.589	+	13.21		0.138	-78.2	0.449	-115.9
5.60000	8.390	+	13.30		0.130	-76.1	0.423	-120.2
5.65000	8.084	+	13.35		0.122	-74.2	0.375	-124.5
5.70000	7.942	+	13.38		0.114	-72.4	0.327	-129.0
5.75000	7.850	+	13.37		0.105	-71.9	0.279	-133.3
5.80000	7.829	+	13.32		0.094	-71.6	0.231	-137.6
5.85000	7.849	+	13.24		0.088	-72.1	0.186	-141.9
5.90000	7.911	+	13.17		0.080	-73.5	0.142	-146.3
5.95000	9.013	+	13.06		0.072	-76.0	0.102	-151.1
6.00000	8.149	+	12.93		0.065	-79.7	0.064	-157.3
6.05000	8.315	+	12.79		0.060	-84.8	0.030	-165.7
6.10000	8.509	+	12.63		0.056	-91.1	0.040	-177.4
6.15000	8.726	+	12.47		0.053	-98.3	0.053	-189.9
6.20000	8.964	+	12.30		0.052	-106.3	0.059	-206.0
6.25000	9.220	+	12.12		0.052	-114.0	0.083	-223.3
6.30000	9.491	+	11.95		0.054	-121.0	0.104	-240.7
6.35000	9.774	+	11.76		0.058	-127.0	0.127	-257.7
6.40000	10.073	+	11.58		0.061	-131.9	0.146	-274.2
6.45000	10.379	+	11.40		0.064	-135.9	0.164	-291.6
6.50000	10.694	+	11.22		0.071	-138.9	0.181	-309.0
6.55000	11.014	+	11.04		0.076	-141.3	0.197	-326.4
6.60000	11.340	+	10.86		0.081	-143.2	0.213	-343.8
6.65000	11.670	+	10.68		0.086	-144.6	0.227	-361.2
6.70000	12.002	+	10.51		0.091	-145.6	0.241	-378.6
6.75000	12.336	+	10.34		0.096	-146.4	0.255	-396.0
6.80000	12.670	+	10.17		0.102	-146.9	0.268	-413.4

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