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Analysis and Design of Reconfigurable Spiral Antenna for RF Interference Mitigation

Kunysz, Waldemar

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doctoral thesis

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Abstract

Pattern reconfigurable antennas represent an antenna design class that differs from classical fixed-form, fixed-function antennas in that they can reconfigure and adapt their radiation pattern to fit the requirements of a time varying system. The explosion of the variety of satellite applications in the last two decades demands novel, compact antenna designs that allow the capture of ultra-week signals in the presence of RF interference and multipath. Advances in electromagnetic simulation tools and RF/microwave processing technologies have enabled the design of new compact antennas that can be integrated with high quality RF and microwave active and passive components. This dissertation introduces the concept of pattern reconfigurable antenna in a small form factor using novel reactively loaded spiral antennas. The proposed design allows mitigation of reception of intentional or unintentional RF interference sources in the receivers designed for weak (i.e. satellite signal) reception. Two new antennas capable of reconfigurable pattern control are introduced: a dual polarized spiral antenna and x-spiral (cross-spiral) antenna. The novel approach to the creation of sharp nulls in the antenna radiation pattern is presented. The new antenna designs solved various problems associated with classical approach to spiral antenna design and lead to improvements such as lower overall Axial Ratio and higher efficiency/gain at lower end of operating frequency bandwidth. The relevant background work is described and then the design details, computer simulations and measured experimental results are given.
Acknowledgements

I would like to thank many people who have made the completion of this dissertation possible. They have given both personal and technical guidance during my PhD work.

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Most importantly, I would like to thank my wife and children for their continuous love. The ability to pursue my dreams would not have happened without their support and understanding. This dissertation is dedicated to my wife Shawna, her limitless patience and support made all my effort worthwhile through all stages of my engineering education and career spanning almost 30 years.
Dedication

This thesis is dedicated to my wife Shawna who has endured my educational never-ending journey.
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<tr>
<td>ABF</td>
<td>Aerial beam forming</td>
</tr>
<tr>
<td>AF</td>
<td>Antenna Factor</td>
</tr>
<tr>
<td>AOA</td>
<td>Angle of Arrival</td>
</tr>
<tr>
<td>Aperture</td>
<td>the largest dimensions of the antenna array, usually expressed in wavelengths</td>
</tr>
<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
</tr>
<tr>
<td>Axial Ratio</td>
<td>The ratio of the major to minor axes of a polarization ellipse [1]. Axial Ratio pattern is a graphical representation of the axial ratio of a wave transmitted (or received) by the antenna over given radiation pattern cut.</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>The IEEE (Institute of Electrical and Electronics Engineers) standard [1] defines the bandwidth of an antenna as “the range of frequencies within which the performance of the antenna, with respect to some characteristic, conforms to a specific standard.” In this dissertation, the bandwidth is defined for impedance and radiation pattern separately. The impedance bandwidth is defined for a VSWR less than 2. The radiation pattern bandwidth, however, is hard to define with a specific criterion, so it is simply taken to be the frequency range within which the patterns are acceptable for a specific application.</td>
</tr>
<tr>
<td>Beamwidth</td>
<td>In the plane of the antenna, beamwidth is the angular width of the directivity pattern where the power level of the received signal is down by 50 percent (3 dB) from the maximum signal in the desired direction of reception.</td>
</tr>
<tr>
<td>C/A code</td>
<td>The GPS Coarse/Acquisition (C/A) ranging code freely available to the public.</td>
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<tr>
<td>CDMA</td>
<td>Code division multiple access</td>
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| Co-pol   | Co polarization, That polarization which the antenna is
intended to radiate (receive) [1].

Cross-polarization. Radiation pattern corresponding to the polarization orthogonal to the co-polarization [1].

CRPA
Controlled Radiation Pattern Array antenna

DBF
Digital Beam forming

dBm
Abbreviation for the power ratio in decibels (dB) of the measured power referenced to one milliWatt (mW).

dBc
Abbreviation for the power ratio of a signal to a carrier signal, expressed in decibels.

dBic
The directive gain of a circularly polarized antenna, expressed as the ratio, in decibels, of the antenna's directivity to that of an isotropic antenna with the same polarization characteristic.

DF
Direction Finding, the process of determining the AOA

DME
Distance Measuring Equipment

Driven Element
A radiating element connected directly to the feed line of an antenna

ESPAR
Electronically Steerable Passive Array Radiator

FFT
Fast Fourier Transform

Gain
Signal level produced (or radiated) relative to that of an isotropic radiator. Gain is closely related to directivity, which in turn is dependent upon the radiation pattern.

GNSS
Global Navigation Satellite System

GPS
Global Positioning System

Galileo
European equivalent of GPS system

Glonass
Russian Federation equivalent of GPS system

IF
Intermediate Frequency

Input impedance
Conjugate of load impedance into which a receiving antenna will deliver maximum power.

LHCP
Left Hand Circular Polarization, see sense of polarization

MBF
Microwave beam forming

MVDR
Maximum Variance Distortionless Response. The weight
vector obtained by solving the following constrained optimization problem: Minimize the output power of the beam former subject to the constraint that the gain at the AOA of the desired signal is unity

**MLM** Maximum Likelihood Method

**MUSIC** Multiple Signal Classification algorithm

**Null Steering** An antenna having in its radiation pattern one or more directional nulls that can be steered, usually electronically [1].

**Parasitic Element** A radiating element that is not connected to the feed lines of an antenna and that materially affects the radiation pattern or impedance of an antenna, or both [1].

**Phase Center** The location of a point associated with an antenna such that, if it is taken as the center of a sphere whose radius extends into the far-field, the phase of a given field component over the surface of the radiation sphere is essentially constant, at least over that portion of the surface where the radiation is significant [1].

**PCO** Phase Center Offset of the antenna (mean value)

**Polarization pattern** (1) The spatial distribution of the polarizations of a field vector excited by an antenna taken over its radiation sphere.

(2) The response of a given antenna to a linearly polarized plane wave incident from a given direction and whose direction of polarization is rotating about an axis parallel to its propagation vector; the response being plotted as a function of the angle that the direction of polarization makes with a given reference direction [1].

**RF** Radio Frequency

**RFI** Radio Frequency Interference (Interferer)

**RHCP** Right Hand Circular Polarization, see sense of polarization

**SATCOM** Series of various type communication satellites

**Sense of polarization** For an elliptical or circularly polarized field vector, the sense
of rotation of the extremity of the field vector when its origin is fixed. When the plane of polarization is viewed from a specified side, if the extremity of the field vector rotates clockwise [counterclockwise] the sense is right-handed [left-handed]. For a plane wave the plane of polarization shall be viewed looking in the direction of propagation [1].

*SINR*  Signal to Interference and Noise Ratio  
*SNR*  Signal to Noise Ratio  
*Spatial Filtering*  Attenuating unwanted signals and passing through desired signals in 3D space using reconfigurable antenna radiation pattern prior to feeding the antenna output signal to a receiver  
*STAP*  Space-time adaptive processing  
*Radiating Element.*  A basic subdivision of an antenna that in itself is capable of radiating or receiving radio waves  
*VSWR*  Voltage Standing Wave Ratio  
*VOR*  VHF omnidirectional radio range  
*TACAN*  Tactical Air Navigation

<table>
<thead>
<tr>
<th>Symbols</th>
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<tbody>
<tr>
<td>AF</td>
<td>Array Factor</td>
</tr>
<tr>
<td>α</td>
<td>Attenuation constant</td>
</tr>
<tr>
<td>β</td>
<td>Phase constant</td>
</tr>
<tr>
<td>θ</td>
<td>Elevation angle</td>
</tr>
<tr>
<td>φ</td>
<td>Azimuth angle</td>
</tr>
<tr>
<td>D</td>
<td>Antenna Directivity</td>
</tr>
<tr>
<td>G</td>
<td>Antenna Gain</td>
</tr>
<tr>
<td>γ</td>
<td>Propagation constant</td>
</tr>
<tr>
<td>λ₀</td>
<td>Free space wavelength</td>
</tr>
<tr>
<td>λₔ</td>
<td>Guided wavelength</td>
</tr>
<tr>
<td>Z₀</td>
<td>Characteristic Impedance</td>
</tr>
<tr>
<td>Zₔ</td>
<td>Load Impedance</td>
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\( k_0 \)  Free space wave number
\( Z_L \)  Load Impedance
\( \eta \)  Antenna Efficiency
\( \Gamma \)  Reflection coefficient
\( M \)  Number of radiation modes in spiral antenna
\( N \)  Number of spiral arms
\( \omega \)  Angular frequency
\( c \)  Speed of light
Chapter One: Introduction

The objective of this thesis is to develop a novel antenna for high-end applications that allows enhanced performance in a hostile environment of intentional and/or unintentional RF interference while reducing the cost and complexity of the overall antenna/receiver system. The roots of this approach are taken from the Electronically Steerable Parasitic Array Radiator (ESPAR) antenna with several novel modifications in order to obtain a good balance between performance versus size, cost, weight, complexity and power consumption. All ESPAR research in the last two decades was focused on linearly polarized systems that provided a reconfigurable radiation pattern in one plane (Azimuth/Phi) and very limited beam/null scan ability in the other plane (Elevation/Theta). The main aim of this proposal is to develop a wideband circularly polarized ESPAR type antenna that has the null scan capability in 3D hemispherical space, while being small and compact. Extensive search of published papers indicates that such system has not been yet analyzed, designed or researched.

The accuracy of any receiving system is largely dependent on four factors: amount of received multipath signal, level of RF jamming, antenna characteristics and type of signal processing employed. The first three factors can be controlled by the proper selection of the antenna used for signal reception. In some applications such as GPS, five and seven element arrays are used for military applications where signals generated by the array are processed by multiple receivers. Nulls and/or beams are formed in the digital domain in order to remove the interfering RF or multipath (delayed replica) signal. RF amplifiers and frequency converters are needed for each antenna element; hence the necessary size of the analog circuits is actually increased over that of
an analog phased array. There is a tendency to eliminate analog circuits as much as possible by means of analog/digital (A-D) conversion at the highest frequencies possible. The application of this concept to phased arrays is denoted as digital beamforming (DBF). In DBF, the antenna radiation pattern is digitally controlled thus providing great flexibility to implement various beamforming schemes. However, DBF involves several (as many as the number of antenna elements) front end systems coupled with corresponding high speed A-D converter circuits. Hence, if the beam former is digitized, the antenna system will require a large processing capability, significant amounts of power and occupy a large space. The problem becomes more serious as the number of antenna elements increases. This ironic circumstance is a significant barrier to cost reduction of the systems. The problem becomes more serious as the number of antenna elements increases. DBF performance quickly degrades or become completely unusable in the presence of high power and wideband interfering signals due to saturation of RF front ends. Although DBF by itself is a digital technology, the high cost and limited performance are critical reasons why this architecture has not proliferated in civilian or consumer use.

A microwave beamforming (MBF) is used where only one frequency converter and A-D converter set is required or available. MBF is accomplished with the use of variable phase shifters or RF delay lines and power combiners. The large size, weight, cost and power consumption associated with MBF systems have prevented them from widespread deployment with few exceptions (i.e. various military applications). MBF can be very resilient to interference as compared to DBF; however it does not provide a flexibility of generating and steering multiple beams or nulls like DBF.
Another approach is to carry out signal processing in the reconfigurable radiation pattern of the antenna, sometimes referred to as aerial beamforming (ABF). A single set of high-frequency amplifiers, a frequency converter, and other receiver circuitry is needed in the ABF system. ESPAR antenna is the most common ABF concept realized in practice. Research in ESPAR technology is relatively new and many topics have been left unresolved such as antenna hardware design, beam and null control in two-dimensional (2D) and three-dimensional (3D) space, direction finding, antenna self-orientation, polarization diversity, and optimization methods to control parasitic elements. ESPAR antennas are relatively large and bulky with a high vertical profile and large footprint. The occupied space for the 7-element ESPAR monopole antenna is in the order of half wavelength in diameter and quarter wavelength in height [71]. At a frequency of 1 GHz this would correspond with antenna dimensions of 150 mm (diameter) and 150 mm (height, including ground plane).

The aim of this thesis is to design a wideband circularly polarized ESPAR type antenna in a compact physical form. The number of driven elements will be kept to only one and two in order to achieve better control of spatial null placement in 2D angular space and to preserve circular polarization of the synthesized signal, and keep the cost as low as possible. While this increase in complexity, as compared to a single driven ESPAR antenna, is necessary in order to achieve the desired system performance, it still is a major improvement in lowering of the complexity when compared to classical analogue and digital beamforming systems. The proposed system provides significant reduction in size, weight, cost, power consumption and dynamic range for the input SNR. Mutual coupling will be intentionally used as a design parameter, which is in a striking
contrast to conventional array systems where mutual coupling is one of the main causes of performance deterioration and is avoided. The main issue will be the understanding of the ABF behaviour while changing various antenna parameters (i.e. excited phase gradient, reactive loads, number of excited spiral arms, etc.) due to mutual coupling occurring between ESPAR antenna elements. The known antenna array design theory and control algorithms cannot be used directly for this application.

Wideband scanning arrays are difficult to design because of conflicting requirements. The difficulty arises first with the choice of a wideband element, which is typically physically large compared to narrowband elements. Element size is important, since it limits how closely these elements can be placed together in the array formation. Element spacing of $\lambda/2$ limits the maximum scan angle and visibility of grating lobes in all designs. A broad beamwidth must exist in both planes, since scanning is desired in all directions of $(\theta, \phi)$. To meet this requirement various small, wideband antennas have been developed in the past such as sinuous (dual linear polarization), equiangular and Archimedean spiral antennas (circular polarization), Vivaldi (linear polarization) antenna and others. It is the aim of this thesis to explore circular polarized wideband antenna elements for ESPAR application.

1.1 Motivation

There have been significant developments of satellite based applications (i.e. GPS, Galileo, Glonass, Sirius Satellite Radio, etc.) in the last decade. Those industries, and others, generated a need for re-configurable circularly polarized pattern antennas that allow for creation of deep nulls pointing towards interfering signal or multipath generated
replicas of the original signal. The satellite based reception has many benefits (i.e. global reach); however the Achilles’ heel of these systems is their susceptibility to RF jamming, self-interference and multipath. Many forms of proprietary digital and RF processing have been implemented (with various degree of success) to allow for maintaining a lock on the signal in the presence of the interfering signal. For military application digital phased arrays have been designed where null forming is done in the digital domain. However, neither approach works when the receiver RF front end is saturated in the presence of a very strong jammer signal. The only solution to such condition is a reconfigurable pattern antenna that can create RF deep nulls towards the interfering signal and hence prevent the receiver front end saturation.

There are various methods of designing reconfigurable pattern antennas (i.e. phased array antennas). The typical designs are often large and expensive, requiring complicated RF feed systems connected to a multiple channel receiver where extensive digital signal processing (e.g. MUSIC, MLM) is employed. This thesis proposes a simple method of creating small, wideband reconfigurable antennas for reception of circular and/or elliptically polarized signals using a single channel receiver. The idea is similar to ESPAR (Electronically Steerable Passive Radiator Array) with several modifications. The ESPAR antenna was originally developed by Harrington [2] and later by Ohira [3]. The main purpose of ESPAR antennas was to provide a cost-effective antenna platform for wireless communication systems. In most cases ESPAR antennas are linearly polarized. The recently designed ESPAR antenna using circularly polarized patches was proposed by Kato and Kuwahara [4]. In all reported cases ESPAR antenna arrays are large and bulky. In addition the beam/null control in elevation (theta) plane is limited as
opposed to full coverage in the azimuth (phi) plane. The antenna proposed in this thesis is targeted for satellite signal reception (circularly polarized) and has the ability to create narrow spatial nulls to cancel any intentional or unintentional interfering signals. It allows for a full control of null placement in both azimuth and elevation planes above antenna horizon. The antenna also provides ability to change its polarization sense (from left-hand to right-hand polarization or to linear polarization).

Spiral antennas have an inherent “built-in” handedness that allows transmitting or receiving circularly or elliptically polarized signals. The presence of multiple spiral arms provides multiple degrees of freedom for the designer. For example, a four arm spiral antenna has eight possible entry points as excitation ports. In normal applications between one to four ports are excited leaving the remaining “entry” ports for other usage. One such usage is to employ them as “reactively” loaded passive ports to establish a new antenna behavior (i.e. modified antenna pattern).

In order to create deep nulls off antenna boresight, various non-conventional excitation methods need to be introduced (as described in later sections of this thesis). Most spiral antenna designs exhibit relatively high cross-polarization levels that can be amplified when adding “reactively” loaded ports. Two novel spiral antenna designs are proposed in this thesis that allow to create a deep null in the radiation pattern while minimizing the level of cross-polarization levels.

1.2 Dissertation Overview

This dissertation is organized into three parts. The first part consists of the first three chapters and is a review of previous work and necessary background information.
The motivation and overview of the dissertation is presented in Chapter 1. A review of existing satellite systems and RF interference is described in Chapter 2. Chapter 3 provides a description of the current state of reconfigurable pattern antenna research.

The second part of the dissertation is contained in Chapter 4. It provides a detailed analysis of two novel antenna design concepts and associated feeding mechanisms that allow creation of deep nulls in otherwise omnidirectional radiation patterns.

Chapters 5 and 6 present the results of specific investigations using simulations and experimental methods. Chapter 5 presents simulation results of novel spiral antennas using various feeding mechanisms and reactive loading used to control the steering and creation of deep nulls with very narrow spatial angular area. The investigation was carried out using computer simulations performed with FEKO and HFSS commercial EM software packages. Chapter 6 presents measurements for the antennas described in Chapter 4 and simulated in Chapter 5. Conclusions are then presented in Chapter 7.

1.3 Contributions

The major contributions of this dissertation are twofold. First, the new concept of narrow spatial width nulls in a circularly polarized antenna pattern is developed. Until now, the primary areas of interest in ESPAR reconfigurable antenna designs have been confined to linearly polarized circular monopole-type array antennas. The second significant contribution presented here encompasses two new spiral antenna designs created to achieve sharp nulls in the otherwise omnidirectional upper-hemisphere circularly polarized pattern. The new designs developed in this dissertation are the
crossed 4-arm spiral antenna and dual-polarized spiral. All these antennas exhibit the desired null control while maintaining nearly constant performance over all other critical electrical characteristics.

1.4 Publications

One paper was presented at an IEEE recognized conference and published in the proceedings. Another paper was submitted to IEEE Transactions on Antennas and Propagation. Two additional papers are under preparation for submittal to IEEE Antennas and Propagation Letters and to a FEKO student competition that will take place in November 2013. A book was published in October of 2012 with three co-authors. The detailed list of these publications is given below.


“Novel Dual Polarized Spiral Antenna” to be submitted in June, 2013 to IEEE Antennas and Propagation Letters.
“Novel Method of Mitigation Reflections from Antenna Spiral Arm Ends”, to be submitted for FEKO 2013 Student Competition.

Chapter Two: **Background**

2.1 Current Satellite Systems Survey

2.1.1 Global Navigation Satellite Systems

Currently, there are two fully operational geo-spatial positioning satellite systems with global coverage. These are GPS (launched and operated by USA) and GLONASS (launched and operated by the Russian Federation), see Fig. 1. Additional new systems are being developed and deployed mainly by the European Union (Galileo satellite system) and China (Beidou satellite system). Other countries are also developing similar systems, however with limited regional coverage and using stationary satellites. The four systems mentioned above provide (or will provide when fully operational) full global coverage. These systems allow small devices (like smart phones) to determine their 3D position to within few meters. These systems have also found wide acceptance in various timing applications (i.e. power grid synchronization, cell phone timing service, etc.).

The global coverage is achieved with a constellation of typically 24-40 satellites orbiting at a distance of 20,000 km above Earth surface. The GNSS systems operate in multiple frequencies and channels spanning from 1164 to 1610 MHz, see Fig. 2 and Fig. 3. For example the GPS L2 band is located in the frequency band of 1227.6 ±10 MHz while L1 band is located in the frequency band of 1575.42±10 MHz (C/A code occupies 2 MHz). The limited output power of the satellites (order of 30 Watts) and large propagation loss (order of 183 dB) causes the received signal to be very weak and buried beneath the thermal noise. The system works very well in normal non-interfering conditions, specifically in the protected upper frequency band (1560-1610 MHz) where no other systems are allowed to operate. The performance of these systems quickly
deteriorates in the presence of RF interference, whether intentional or not. The GPS system is very prone to jamming during the C/A code acquisition phase where conventional receiver technology has only limited jammer tolerability (J/S - 27 dB) [75-77].

Fig. 1 GPS + Glonass satellite constellation (visual concept)

Fig. 2 GPS and Galileo Spectrum Allocation [73]

Fig. 3 Glonass Spectrum Allocation [73]
The GNSS system is evolving from a dual GPS frequency system to multi-channel, multi-frequency and multi-constellations satellite networks that provide an enhanced positioning accuracy, availability, number of available satellites, redundancy etc. Unfortunately, due to bandwidth restrictions, the frequency separation is limited hence the system is still vulnerable to wideband RF jamming.

2.1.2 Global Communication Satellite Systems

There is a new trend towards global information networks offering flexible multimedia information services to users on demand, anywhere, anytime. Future service will include any type of data communication services such as Internet access, data transfer, global banking, video/audio real-time transfer, etc. This need to support bandwidth intensive multimedia services places new and challenging demands on satellite systems and networks. There is a significant push to use satellite communications (SATCOM) networks to provide such services as an emerging fourth-generation (4G) communications system.

Presently a few current SATCOM systems employ advanced coding algorithms coupled with necessary bandwidth that can support a modern high speed communication system. However, several major systems are in the planning stages and most future systems plan to employ IP, ATM, or a combination of both in order to meet ever increasing demand for speed and bandwidth of advanced communication systems. Some systems, such as RapidEye, have constellations with many low earth orbit (LEO) satellites, while others plan to operate with as few as two to four satellites in the geosynchronous earth orbit (GEO).
Most of the new and future planned systems operate at Ka-band. The key reason for the use of Ka-band is the availability of bandwidth to support broadband multimedia communication (i.e., capacity). The main difference between positioning satellite systems and communication data satellite system is in the very high data transfer rate of the latter. The common denominator for all satellite systems is low power level of transmitted signals, therefore making these systems vulnerable to any interference.

2.2 RF Interference

The allocated GNSS frequency bandwidth is surrounded with signals from hundreds of various wireless systems that can cause unintentional RF interference. The main focus of this thesis was to develop novel antenna system concepts that would enable mitigating RF interference in the frequency range of 1-2 GHz. Intensive simulation and measurement campaigns were conducted in that frequency band.

2.2.1 Background

It is easy to find out how to jam a GPS or any other satellite based system signal. Many products and reference designs are readily available on the internet. Not only are jammers relatively easy to construct but they are also readily available at many internet sites. See a sample Google search on the subject below, see Fig. 4. A one Watt jammer can be built relatively “cheap” and in a small size factor. In addition, the GPS L1 C/A code can be spoofed by an even lower power jammer. So generally, a GPS receiver cannot be expected to reliably acquire the C/A code in any RF hostile environment.
There are also unintentional, accidental interference sources such as welding kits, television transmitters, radars, communication satellites, or electronic products that generate harmonics that fall into GPS band (a good examples are a Radio-Shack TV tuner (booster) that jammed an event in San Francisco bay area in 1990 or TV station interference in Hudson, Massachusetts [72]).

A typical TV station transmits power at a 10 KW (+70 dBm) level. The required spurious emission of -60 dBc yields an effective radiated power of +10 dBm at any out-of-band frequency. There are certain TV channel combinations whose 3rd harmonic fall directly inside the GPS L1 frequency. To determine the maximum radius distance where a GPS receiver is not capable of tracking satellites, we need to compute the propagation loss that would yield jamming signals equivalent to the GPS receiver acquisition threshold level. Most manufactures do not give out the actual specifications of their tracking loops, however we can use an approximation of a good tracking loop (i.e. 44.6 dB-Hz achieved with a third-order PLL, and a 20ms pre-detection integration time). The
corresponding minimum guaranteed received GPS signal would be -129.6 dBm. Hence
the threshold jamming power needed at the receiver is

\[ J_r = J/S + S_R = 44.6 - 129.6 = -85 dBm \]  \hspace{1cm} (1.1)

The line of sight distance to the transmitting television antenna where the power
level will be at the jamming threshold can now be found using standard propagation
equations. Since the jamming frequency is assumed to be in-band interference, the power
loss due to front end receiver filtering (L_f) can be assumed to be zero. The receiver
antenna gain (G_R) is assumed to be –4.5 dBi at the horizon level (theta=90°) due to the
fact that this is where TV station interference would be created.

\[ L_p = 20log\left(\frac{4\pi R}{\lambda}\right) = ERP_j - J_r + G_R - L_f = 10 + 85 - 4.5 - 0 = -90.5 \]  \hspace{1cm} (1.2)

Solving for the distance from the receiver to the jammer to attenuate the signal to the
threshold limit, yields

\[ D_{min} = 2.7 km \]  \hspace{1cm} (1.3)

An increased power level of a TV station to 5 MW would yield the minimum
distance of 11.3 km where GPS receivers would not function! Fortunately most TV
stations exceed FCC recommendations for out-of-band spurious emissions and are being
replaced with digital TV stations that provide much better out-of-band performance.
2.2.2 Interference Sources

In this section we will provide examples of proven interference cases and possible cases scenarios that may occur.

2.2.2.1 LightSquared GPS Interference case

LightSquared is a corporation involved in delivery of mobile voice and data services using satellite data links to North America since 1995. These satellites (MSAT-1 and MSAT-2) operate in the band of 1525-1559 which is adjacent to the L1 GPS frequency band. The company received FCC authorization in 2004 to use this spectrum to build a nationwide 4G-LTE wireless broadband terrestrial network integrated with satellite coverage. The plan was to cover all parts of the U.S. with wireless service, with the intent that even remote rural areas would have 4G services via satellite when its standard terrestrial service would not reach customers. The idea was to combine existing mobile satellite communications services with a ground-based wireless communications network that uses the same radio spectrum as the satellites. Initially, the Federal Communications Commission (FCC) gave LightSquared conditional approval to build out its ground-based wireless network.

The controversy amid the GPS community arose when it became aware that base stations of the LightSquared network would transmit signals in a radio band immediately adjacent to the GPS frequencies. The GPS community was rightly concerned that LightSquared's ground-based transmissions would overpower the relatively weak GPS signal from space.
Although LightSquared would have operated in its own radio band, that band was so close to the GPS signals that most GPS devices would pick up the stronger LightSquared signal and become overloaded or jammed. The original plan to build up to 40,000 ground stations transmitting radio signals one billion times more powerful than GPS signals as received on Earth meant that 40,000 'dead spots' - each a few kilometers in diameter - would disrupt the vitally important services GPS provides.”

The issue was resolved by extensive interference testing in the Las Vegas area conducted in 2011. It was shown beyond doubt that LightSquared base stations caused GPS denial of service ranging from 100m to a few km for various type of GPS receivers (from military, to commercial or consumer products). FCC decided to revoke the experimental spectrum license granted earlier to LightSquared Company. The company had no choice but to kill its plans to deploy the network.

2.2.2.2 Digital and Analog Television

Analog television broadcasts occupy three frequency bandwidths (54-88 MHz, 174-216 MHz and 512-806 MHz). The lower frequency channel can transmit power at 100 kW while the upper frequency channels can transmit signals at 5 MW power. The analog TV is required to meet -60 dBc spurious emission requirements, which is not sufficient to guarantee non-interference to GPS satellite systems. Analog television has been gradually replaced with digital television that provides better control of out-of band emissions. In addition GPS receivers are used to synchronize the overall TV transmission network. No interference has been reported to date in case of digital TV.
2.2.2.3 Amateur Radio

Amateur radio operates in many frequencies spanning from 1 MHz to 300 GHz with power levels of up to 1500 Watts and the same spurious emissions levels as analog TV (-60 dBC). A lot of amateur radio equipment is experimental in nature hence the probability of producing unintentional harmonics is high. They can cause GPS signal degradation from few hundred meters to a couple of km in the worst case.

2.2.2.4 Wireless ISM Devices

Unlicensed ISM communication bands (902-928 MHz, 2400-2483.5 MHz and 5725-5875 MHz) are increasingly filled with various wireless applications (i.e. baby monitors, Wi-Fi laptop connectivity, power grid wireless meters, security video monitoring, etc.). Much of the undesired radiation comes from increasingly higher frequencies at which the microprocessors are clocked in these devices. In some cases, a large number of these devices concentrated in a small physical area can increase the thermal noise floor of a GPS receiver and significantly reduce its performance (even without causing any direct RF interference). This is the main reason that commercial airlines enforce procedures for not using these devices during takeoff and landing.

2.2.2.5 DME and TACAN

DME and TACAN are high power pulsed distance measurement equipment type radars. Aircraft use DME or TACAN to determine their distance from a land-based transponder. These ground stations are typically co-located with VOR’s. A typical DME
ground transponder transmits with a 1 kW peak pulse output on the assigned UHF channel. DME system operates in the 1025-1150 MHz band while TACAN operates in the 960-1215 MHz band. The DME channels are located 25 MHz below E5 Galileo and L5 GPS channels while TACAN basically overlaps these two bands. In addition TACAN frequency band is separated by only 12 MHz from L2 GPS channel. There is no reported interference at L1 GPS frequency; however there are reports of interference at the newly assigned L5 Galileo and GPS bands [73, 74]. See a US map (Fig. 5) of computed signal degradation of L5 GPS and E5A Galileo signal due to interference from DME/TACAN systems.

Fig. 5 GPS L5 and Galileo E5A signal reduction due to DME/TACAN interference (courtesy of Mitre Corporation, Approved for Public Release, Case No: 05-1354)
2.2.2.6 Radar and other military systems

There are various radar systems located either below the L2 and L5 GPS/Glonass and Galileo frequencies or between the L1 and L2 bands (spectrum of 1250-1400 MHz allocated for military radar applications). A sample of these systems is shown in Fig. 6 and Fig. 7. These high power systems can easily disable GPS signal reception for any receiver at a large radial distance.

Fig. 6 Military Radar systems (courtesy of Mitre Corporation, Approved for Public Release, Case No: 05-1354)). Lower images display the format of radiated signal pulses in time domain (x-axis time increments are in μs)
2.2.3 RF Interference Mitigation Methods

There are numerous methods of RF interference mitigation. The most common techniques of interference suppression/rejection are beam steering, null steering, signal cancellation, polarization filtering, frequency incision, tapped-delay transversal lines, FFT based adaptive filters and signal processing. With the exception of polarization filtering, frequency incision, and possibly adaptive signal processing methods, most of these techniques require multiple RF channels and antenna elements or phased arrays to successfully eliminate interfering signals. Other methods that start to gain popularity are integrated Inertial/GPS systems (designed and manufactured by Raytheon Systems), and various analog based solutions.
Phased-array systems are capable of providing interference rejection and/or suppression. Phased-array antenna systems can be very effective in mitigating the impact of one or several interfering sources. Interference suppression in conventional adaptive phased-array systems is achieved by summing the weighted outputs from at least two antenna elements. A digital part of the receiver determines a complex set of weights for each received signal. The DBF computes weights so that the effective power of the interference in the final output will be significantly reduced and the desired signal strength will be enhanced or maintained at the same level. The antenna elements are designed to have wide pattern coverage and are offset from each other in position and/or orientation. These offsets have to be large enough so that there are sufficient signal phase differences among the individual element outputs. For adequate spatial filtering, element separations ranging from 0.3 to 0.5 wavelengths are required. The drawback of phase arrays is that multiple matching networks, filters, LNA’s and down-converters are required for each front-end of the receiver. An example of such system is a CRPA antenna used by US military, see Fig. 8. These antennas are capable of steering a single -40 dB null or multiple -20 to -25 dB nulls towards the interfering signals. The minimum cost of 10,000 $US prevents wide acceptance and has been restricted for military use only.
2.2.3.1 Spectral Filtering

Tuned band-reject and adaptive digital signal processing are used more often in modern GPS receivers. The filters are designed to adaptively attenuate in-band and out-of-band interference signal. The main disadvantage is reduced acquisition time of the GNSS satellites and inability to deal with wideband jammers.

2.2.3.2 Temporal Filtering

Temporal filtering is performed at intermediate frequency stage (IF) using digital signal processing. A FFT of the incoming signal is performed. Any strong signal above the thermal noise floor is removed by an adaptive FIR filter. An inverse FFT transform yields a “cleaned-up” original signal with most of the interfering energy removed. Similar to Spectral Filtering, this method cannot detect and mitigate broadband RF interference.
2.2.3.3 Inertial/GPS Systems

Initially these systems were designed for military applications (tactical grade) with secret classification equivalent to a hydrogen bomb; hence available technical information on high-end products is still limited. Recently, many commercially developed systems have been released to the general market (i.e. Trimble, NovAtel, Ashtech, etc.). J/S improvement of 8-15 dB has been reported. Experiments have shown an improvement in GPS code tracking of about 10 to 15 dB in the presence of wideband interference.

2.2.3.4 Analogue Based solutions

FFT based adaptive filters remove bins with power above thermal noise. These methods are suitable for narrowband jammers (i.e. order of few kHz) but provide no mitigation against broadband jammers (more than 1 MHz). Analogue and spatial cancellers offer the best mitigation performance against all jammer types. These analogue systems manipulate the antenna received gain pattern to reject the jammer or interference signal in that direction only while allowing the satellite signals arriving from other directions to be received in the normal way.

2.3 Summary

GPS dual frequency (L1, L2) operation was initially designed to operate as a secret US military positioning service “hidden” in the presence of the thermal noise and hence undetectable by the enemy forces. The initial advantage of GPS signal being buried in the noise became its undoing in the presence of intentional or unintentional RF interference.
Once the presence was of the GPS system was “discovered”, international efforts were made to protect the GPS L1 band from any RF interference or other signal spectrum co-location efforts. The GPS L2 band, however, is located in the unprotected band where it is subject to RF interference from various systems located in the adjacent spectrum. In fact, the GPS L5 band has to share the spectrum with the TACAN military distance measurement system.

Recent pressure to re-allocate RF spectrum to new wireless services prompted the FCC to allow LightSquared to design and deploy high-power terrestrial communication base stations by issuing a spectrum license just below GPS L1 band. The FCC made the decision to revoke the same spectrum license to LightSquared in 2012 soon after industry and military complaints about the level of interference generated by this system.

There are many future satellite communications systems being designed at this moment. There is a need for a simple and cost effective solution to RF interference not only for GPS or GNSS systems but for any satellite based services. The received satellite signal will be relatively weak due to large propagation and atmospheric losses and finite RF power available on board of a typical satellite. Specifically, a simple solution is required for small platforms with finite computational capacity and power. The object of this dissertation is to develop a new antenna technology with reconfigurable radiation patterns to mitigate RF interference through placement of a deep null towards the jamming signal. The new solution should ideally provide wide bandwidth of operation, small foot-print and planar structure for ease of integration in a final product.
Chapter Three: Previous Work

This chapter presents a review of the previous work related to the subject of reconfigurable antennas and provides background information for the new designs developed in this dissertation. More specifically, this Chapter gives an overview of the current state of reconfigurable antennas - a relatively recent objective in antenna design.

3.1 Introduction

An antenna that possesses the ability to modify, in real time, its characteristics, such as operating frequency, polarization or radiation pattern, is referred to as a reconfigurable antenna. Reconfigurable antennas were first introduced in 1956 [3-5], through an intentional redistribution of the currents or, equivalently, the electromagnetic fields of the antenna’s effective aperture, resulting in reversible changes in the antenna impedance and/or radiation properties [6]. The reconfiguration of an antenna may be achieved through many techniques. Some designs employ various circuit controls while others rely on mechanical alteration of the structure such as rotating or bending of one or more of its parts [7]. Other approaches are to energize different antenna parts at different times, reconfiguring the feed networks or appropriately exciting the antenna arrays [8]. All such approaches have significantly contributed to the evolution of reconfigurable antennas during the last decade. Recently, electrically-actuated switches and variable capacitors such as PIN diodes, RF MEMS and varactors have been used to achieve reconfigurable pattern antennas.
Reconfigurable antennas can be classified into six main groups based on the reconfiguration technique applied:

- Antennas using switches
- Antennas using reactive elements such capacitors, varactors or inductors
- Antennas using physical angular alteration
- Antennas using different biasing networks
- Antenna arrays
- Antennas using reconfigurable feeding networks

Reconfigurable antennas have the potential to add substantially to the degrees of freedom and the functionality of mobile communications and phase array systems. Many reconfigurable antenna designs can be readily found in the literature. However most reconfigurable antennas introduced so far concentrate on changing their operating frequencies while maintaining their radiation characteristics. This type of antenna is referred to as a frequency reconfigurable antenna. Modern communication systems demand transmitters and/or receivers with multi-band operation; as a result, numerous techniques for achieving frequency reconfigurability have been proposed in the literature. For instance, Weedon [49], reported a reconfigurable multi-band antenna integrated with radio frequency micro electro mechanical systems (RF MEMS) switches for applications at drastically different frequency bands, such as communications at the L band (1-2 GHz) and synthetic aperture radar (SAR) at the X band (8-12.5 GHz).

Another example of the frequency reconfigurable antenna is described by Qian [10]. It consists of a linear array of microstrip-based leaky-mode antennas. The work on the microstrip-based leaky-mode antennas was originally done by Menzel, and further
investigated by Oliner [9]. The operating frequency is controlled by the state of the switch operation.

3.2 Reconfigurable Pattern Array Antennas

Electronically reconfigurable multimode direction-finding antennas have been found to provide excellent capability for interference suppression. The initial purpose of reconfigurable antennas was to provide DF capability. It was discovered that a multimode spiral antenna (Corzine [78]) could easily provide such a capability. In addition, interference suppression can also be achieved through a set of variable control loads that are adjusted in response to a reference signal or power level in the frequency band of interest for spread spectrum systems.

An exemplary single-aperture direction-finding antenna includes the multimode type spiral antenna and its planar, conical or slotted variations such as the square spiral, Archimedean spiral, equiangular spiral, and the logarithmic spiral. A typical military DF antenna system has a feed network and appropriately placed control ports embedded in its aperture with RF front-end, microcontroller, feedback control electronics, and controllable loads attached to the control ports.

3.2.1 Beam/Null Steering Antennas

Beam/null steering capable arrays are generally classified as: (1) phased arrays, (2) switched-element arrays, and (3) parasitic arrays. Phased arrays were the most popular in the past when various types were designed and commissioned for military and government use. Phased array antennas are complex to build, expensive, heavy [12] and
therefore their use has been very limited. Parasitic arrays have had a wider deployment due to their low cost but have many limitations that prevent them from having a wider acceptance.

The parasitic array with beam steering capability was originally patented by Yagi, and its low-cost merits has been researched and explored by many (e.g., [13]). Most designs to date are narrowband, typically employing monopole antennas as the driven and parasitic array elements resulting in a high profile.

The analog approach to create an alternative architecture of phased array antennas is a re-emerging technology. The history of practical analog beam-forming antennas dates back to the Butler matrix [14] which consisted of hybrids as fixed phase shifters. Since beam steering is operated by employing selective RF switches the steering angle is only discrete. An alternative approach for beam steering using varactor-tuned passive radiators is proposed by Ohira and Gyoda [15]. By adding varactors to the passive radiators, beam steering is facilitated by controlling the phases of the currents in the passive radiators. To estimate the value of the capacitances required to focus the beam or null in a particular direction, various optimization techniques are employed [16].

3.2.2 Single Arm Spiral Antennas

Variations were introduced in the current flow of the outer circumference of the spiral antenna by applying short circuits at different points on this section of the antenna [59]. One-point and two-point short-circuits were used to generate nulls at limited range of elevation angles (highest theta angle of 42° was achieved). It was also noted that circular
polarization sense was lost after insertion of short circuits. Similar results (linear polarization with limited null scan angles) were obtained in [60].

3.2.3 ESPAR

Electrically steerable passive array radiator (ESPAR) antennas have been investigated for low-cost, small analog adaptive beam forming [17]. ESPAR antennas are constructed from dipole elements without a ground plane, or from monopole elements with a ground plane. Each passive element located around an active element is loaded by capacitive variable elements such as a varactor diode. Each element consists of a quarter-wavelength vertical monopole fixed to a base plate and a variable capacitor. For far-field analysis, each reactive monopole is treated as a dipole with a center capacitor. The active radiator is connected to a voltage source, see Fig. 9.

![Typical Equivalent ESPAR antenna model](image)

Fig. 9 Typical Equivalent ESPAR antenna model [71]

The ESPAR antenna can form an adaptive beam and maximize the receiving signal level by searching the reactance value set that yields a maximized correlation
coefficient between the received signal and the known reference sequence. The main design issues of ESPAR antennas are their large and bulky size and the fact that the global optimum solution cannot be easily obtained.

A newer approach in ESPAR antenna development was to use patch elements instead of monopole antenna elements [18]. This is the first ESPAR antenna that can generate circular polarized waves over a wide angular space.

3.2.4 Reconfigurable Arrays

Antenna arrays can employ several methods to achieve reconfigurable designs. Dynamic reconfigurable elements can be used instead of conventional fixed antennas as the array elements. Or the array can use conventional fixed elements and the feed section phase-controllers might employ reconfigurable elements.

Fig. 10 Flared array and time-delay beam/null steering
A reflective reconfigurable aperture that uses an array of fixed flared-notch antennas (see Fig. 10) is presented in [19], [20], and [21]. The flared notch antennas have MEMS switches placed along the notch feed section that allows the length of the feed section to be changed. This alteration of feed length introduces a time delay in the received energy. The time delay in the individual array elements allows the array to be built of Reconfigurable Elements.

Reconfigurable elements represent antennas that radiate with one or few primary radiating elements. That primary element is reconfigured via switches or some other variable element to provide parameter control. The antennas listed in this subsection all function in this manner (lumped reconfigurable elements). The reconfigurable elements can then be used to design and build a reconfigurable array antenna in order to have ability to steer the null or beam in any desired spatial point.

3.2.5 Planar Reconfigurable Slot

The reconfigurable slot antenna presented in [22] is an electronically tunable planar VHF slot antenna. It consists of a microstrip fed resonant slot structure loaded with a series of PIN diode switches. The resonant frequency of operation is selected by varying the length of the radiating slot and thus changing its electrical length. The length of the slot is altered by biasing the PIN diode switches along the slot length. The useful bandwidth operation of this antenna is from 500 to 900 MHz. The design can be scaled to other frequencies of operation.
3.2.6 MEMS Reconfigurable Vee

The MEMS reconfigurable Vee antenna is a printed antenna that uses dynamic actuators to alter the radiation characteristics of the antenna. Fig. 11 shows the geometry of the reconfigurable Vee antenna. The antenna presented in [23, 24, and 25] uses MEMS actuators to alter the geometry of the Vee dipole. Push-pull actuators connected to the dipole arms enable reconfiguration of the Vee pitch angle. Each arm of the Vee is independently controlled, allowing the antenna to both steer the antenna beam and alter the beam shape. Symmetrical movements of the Vee actuators result in a widening and narrowing of the Vee angle, $\alpha$, and thus widen and narrow the main beam. The main beam may be steered off broadside by moving each Vee actuator by different distances.

Fig. 11 The Reconfigurable VEE antenna
The VEE antenna has a limited scan range (demonstrated as 48° of main beam shift from broadside).

### 3.2.7 Reconfigurable Dime and Q-dime Antennas

The dime antenna [26, 27, and 28] and its building block q-dime element are broadband multilayered stacked circular patches, see Fig. 12. The dime antenna is constructed as either a single monolithic element or as a combination of four quarter dime (q-dime) elements. The antennas are electrically smaller than other type antennas (radius < 0.2λ and height < 0.5λ). Broadband operation is accomplished by creating two degenerate modes via two cylindrical radiating slots in the patch structures. MEMS switches located within the geometry of the patches allow the antenna to alter the working frequency of the design. The switches are successively turned on to tune the frequency of the antennas. The antennas are also able to control polarization and radiation pattern shape by activating the MEMS switches.

![Reconfigurable Dime antenna geometry](Fig. 12)

Fig. 12 Reconfigurable Dime antenna geometry
3.2.8 Reconfigurable Microstrip and Leaky Mode Patch

Leaky-wave radiation can also be facilitated by the presence of higher order modes on microstrip periodic structures [31]. A combination of a typical microstrip patch antenna and a leaky-wave antenna was presented in [29], [30] as shown in Fig. 13. The antenna shown in Fig. 13 consists of a high-order mode launcher connected to a length of X-band leaky mode microstrip antenna. The mode launcher is composed of a microstrip impedance transformer, 180° phase shifter and an even-mode suppressor. The leaky-mode antenna is then connected via PIN diodes or MEMS switches to a conventional C-band microstrip patch antenna. When the switches are deactivated the antenna operates as a conventional patch antenna. Activating the switches causes the patch structure to become part of the leaky-mode structure and the antenna radiates as a leaky-mode antenna [29].

![Diagram](image.png)

Fig. 13 The reconfigurable leaky mode patch antenna
3.2.9 MEMS Mechanical Beam Steering

The antenna system presented in [32] addresses the problem of scan loss inherent to electronically steered antennas. Scan loss is a broadening of the array main beam and the associated reduction of array gain seen as the main beam is steered off broadside. The system overcomes this loss by steering the antenna electromechanically via a MEMS structure as opposed to a strictly electrical scan as in a phased array. The MEMS antenna system is capable of scanning the main beam over 60° with no measurable scan loss.

3.2.10 Distributed Radiators

The distributed radiators class represents antennas that radiate primarily by combining sub-elements together to form a larger discontinuous radiating structure. The operation of this class of antennas is distinct from conventional arrays because the individual elements that make up the radiator effectively radiate as a single unit with a single feed. The sub-elements do not act as autonomous radiating structures. The sub-elements control the current distribution on the aperture of the structure and thus control the radiation properties of the antenna. Various arrangements of these sub-elements allow the overall structure to have different radiation patterns. The reconfigurable nature of the switched elements allows the antenna to change its functionality by activating and deactivating specific switches.

The micro-switched aperture developed by Georgia Tech Research Institute (GTRI) was reported in [33, 34]. The aperture investigated by GTRI consists of small metallic conducting pads connected by MEMS switches. The MEMS switches are activated by control chips that are placed on or near the radiating aperture. Each control
chip directs either a single switch or a small functional cell of switches. The null creation was not a primary objective in these antenna designs, hence their suitability for low profile, small platform system providing circular polarization is not known at this point.

3.3 Reactive loads, nonlinear loads (diodes) and switches

The antenna radiation pattern can be changed by introducing various reactive and nonlinear loads in various parts of antenna aperture. Another method is to switch in and out various parts of antenna aperture that is connected to the excitation ports. The subsequent sections describe various components that can be used to implement these methods.

3.3.1 PIN Diode Switches

The PIN diode [35] switch is popular in microwave circuit applications due to its fast switching times, high isolation and relatively high current handling capabilities. The PIN diode can operate at speeds orders of magnitude faster than mechanical switches and can be placed in packaging measuring a fraction the size of mechanical RF switches. Switching speeds of less than 100 ns are typical. An important quality for RF applications is the fact that it can behave as an almost pure resistance at RF frequencies. This resistance may be varied over a range of approximately 1Ω to 10 kΩ by biasing with a dc or low frequency current [36]. The bias current required for on state operation is normally on the order 10 mA.

3.3.2 FET Switches

Before the emergence and common application of gallium arsenide field effect transistors (GaAs FET) as microwave switches in the mid-1980s, the PIN diode was
primarily the only high speed alternative available. However, advances in semiconductor processing techniques and design innovations have moved GaAs FETS as well as other microwave monolithic integrated circuits (MMIC) into popular use for RF switch applications. The first use of FETs as microwave switches was reported by Liechti in 1976 [37].

FET switches possess a number of characteristics that make them more attractive than PIN diode switches. First, control biasing is isolated from the RF signal path. Thus, only minimal choke separation is required to ensure no dc leaks into the RF signal. Second, the biasing power requirements are much lower for FET switches than for PIN diodes. Gaspari reports [38] the possibility of zero dc power dissipation in biasing the switch. Typically however, the gate bias current is less than 10 μA. The third key advantage of FETs over PIN diode switches is switching speed. Large electron mobility and drift velocity in GaAs devices translates directly to very fast operating switches. FETs can typically switch on the order of a few nanoseconds while PIN diodes are normally limited to a hundred nanoseconds. This combination of lower power draw and faster switching speed makes the FET switch preferable on these bases compared to a corresponding PIN diode switch.

RF FET switches tend to operate over a broader bandwidth than PIN diode switches but the associated insertion losses for FET switches are comparatively higher than PIN diode implementations. FET switches tend to suffer from increased insertion losses at frequencies over 1 GHz and reduced isolation in the off state. Typically 1-2 dB of insertion loss and 20 to 25 dB of isolation in the off state are seen at these higher frequencies.
PIN and FET switches are unsuitable for reconfigurable antenna design where a large number of switches may be employed and individual device losses have a cumulative impact on overall antenna performance. Device deficiencies including narrow bandwidth, comparatively low isolation and high insertion loss, and finite power consumption make them unattractive for use in many reconfigurable applications. Additionally, the non-linear nature of solid-state semiconductor switches always has the potential to introduce undesirable inter-modulation and harmonic products into the RF signal path.

3.3.3 MEMS Switches

The RF micro electromechanical systems (MEMS) have moved into the forefront of reconfigurable antenna design because of their potential to overcome the limitations imposed by conventional RF switches as previously described; conventional PIN diode and FET switches have seen limited use in RF and microwave antenna design.

MEMS large isolation and low insertion loss characteristics result in a switch that is very close to an ideal switch for RF applications. The low actuation power consumption, small feature size and extremely wide bandwidth makes MEMS switches ideally suited to reconfigurable antenna applications.

Active MEMS devices can function as variable capacitors, resistors, inductors, filters, resonators and switches. The MEMS RF switch is the most applicable for reconfigurable antenna applications due to its linearity and lower noise performance compared to switching methods described previously. As with other RF switch
technologies, RF MEMS switches normally are designed in either series or shunt topologies. Shunt switches are commonly used with coplanar waveguide structures and function by shorting the RF signal path to the coplanar ground lines. A movable MEMS shunting bridge is placed between the ground lines and suspended above the signal trace. Activating the switch pulls the shorting bridge down and into contact with the signal line which shorts the RF signal path [39]. Series MEMS switches are favorable for use in microstrip topologies and hence applicable to spiral antennas.

Insertion loss is given for the conduction or on-state of the switch and is normally specified by the S-parameter coefficient S21 in decibels. Efficient on-state switch transmission requires small insertion losses. Also specified by the S-parameter transmission coefficient S21 is switch isolation. Isolation is defined for the non-conduction or off-state and represents the coupling between the input and output points. A high level of isolation is necessary to block RF energy from propagating through the switch when it should be off. The contrast between insertion loss and isolation provides a quick measure of the quality of the switch. The off state isolation of the simple dc-contact series MEMS switch can be derived from its transmission coefficients and is given by

\[
S_{21} = \frac{2j\omega C_u Z_0}{1 + 2j\omega C_u Z_0}
\]

(3.1)

where \( C_u \) is the up state (off) capacitance of the switch and \( Z_0 \) is the characteristic impedance of the transmission line. Fig. 14 shows isolation versus frequency for various off-state capacitance of MEMS switch. 20 dB isolation is achieved up to 2 GHz and it
slowly degrades with frequency. This can limit its application with wideband antennas or when the operating frequency is above 10 GHz.

![Isolation versus off state capacitance (typical dc-contact MEMS switch)](image)

Fig. 14 Isolation versus off state capacitance (typical dc-contact MEMS switch)

Limited isolation at high frequencies and limited switching speed (order of magnitude lower than PIN diode switches) is the main limiting factor for wide acceptance of the MEMS switches at the present time.

A reconfigurable array design based on TEM horn elements is presented in [41]. The array was designed to achieve an effective bandwidth of 10:1 over the 2-20 GHz range. To attain the 10:1 bandwidth the array uses MEMS switches to control the antenna ground plane. A switchable reflective/transmissive ground screen was numerically
demonstrated to offer the necessary reflective control. The ground screen was constructed from closely spaced wires connected together with MEMS switches. The MEMS switches allow the screen inter-element spacing to be varied, and consequently, the screen can be configured such that it is transparent or opaque to selected frequencies.

A method is presented in [42] for control of reconfigurable array antennas. The method requires phase-only control of quantized MEMS phase shifters. A 150,000 element space-based radar is presented in [43] that use MEMS phase shifters. The weight of the electronically scanned array was reduced by employing the MEMS phase shifters in place of conventional phase shifters.

3.3.4 Mechanical Switch Methods

Mechanical switches are not practical for reconfigurable antenna applications due to their large size. However, they have been used extensively in high power RF applications. Mechanical switching is normally accomplished by breaking the conducting path of a transmission line within the switch. The electrical performances of mechanical switches are often excellent providing an insertion losses of less than 0.1 dB, and isolations of over 70 dB are common [44]. The excellent matching characteristics and high power handling capabilities make them the only suitable alternative for many high power applications such as radars or broadcasting. The other major drawback of mechanical RF switches is long transition times associated with mechanically actuating a switch. The moving switch components parts have a mechanical resonant frequency that naturally limits the speed at which the switch can change states. Typical switching speeds for mechanical switches are on the order of 2 ms [44].
3.3.5 Physical angular alteration

The final identified method of reconfigurable antenna design is smart geometry reconfiguration. This method modifies only critical parameters of the antenna radiating structure to achieve the desired reconfigurable performance. It can be implemented with considerably fewer control elements leading to reduced design complexity. The primary disadvantage of this method is that the underlying physics of the particular antenna must be known in order to take advantage of minor geometry modifications to achieve the reconfigurable goal. The amount of reconfigurability is therefore limited by the electrical characteristics of the antenna geometry.

3.4 Space Time Adaptive Processing (STAP)

STAP is a two-dimensional filtering technique using a phased array antenna connected to multiple RF channels. Applying the statistics of the interference environment, an adaptive STAP weight vector is formed. This weight vector is then applied to the coherent samples received by the phased array antenna. Temporal tabs are used to eliminate CW jammers while reserving the spatial nulling capabilities to deal with broadband jammers.

All weights must be calculated in parallel with intensive signal processing involving inversion of large matrices. This is one of the most promising technologies in mitigating RF interference (CW and broadband).
3.5 Frequency Adaptive Processing (FAP)

The STAP process can also be implemented in the frequency domain since frequency-domain implementations are often more efficient than corresponding time-domain implementations. Hence a technique known as frequency adaptive processing (FAP) is used alternatively to STAP. It offers a reduction of computation at the expense of small performance degradation due to approximation processes. It still offers a very effective method against large CW jammers, while minimizing the performance degradation elsewhere.

3.6 Summary

The ideas presented in this dissertation are novel and have not been discussed in existing literature. The work that is most closely related to research reported here was done on single arm rectangular spiral antennas [59-60], however the method of tilting the antenna pattern was more targeted to move the main beam than to generate any deep nulls and destroyed the circular polarization sense. It will be shown that the methods described in this thesis yields very deep, angularly small nulls while preserving intended circular polarization sense of antenna radiation pattern.

There are multiple ways to implement reconfigurable pattern antennas through the means of switching in and out various parts of antenna element structures, reactive and/or resistive elements and components. Proper analysis of these methods would require a separate research investigation in order to determine their net effects and suitability to work with proposed antenna solutions described in following chapters. It was decided to use discrete reactive components in the research work described in this dissertation in
order to reduce the complexity of required research work without losing generality as the design can be modified to use other switching methods.

The material covered in this Chapter is important from the point of view of null steering required for future work in order to have a fully functional system.
Chapter Four: **Novel Null Steering Spiral Antennas**

4.1 Introduction

At present, the GNSS satellite constellation consists of 50-60 satellites while future deployment will reach 80-120 satellites. An omnidirectional uniform antenna radiation pattern is an optimum and necessary condition for the GNSS receiver to provide a simultaneous tracking of constantly moving satellites. The application of beam steering concept in this case is not very practical and therefore not explored in this dissertation. As described in previous chapters there is a need to protect GPS and GNSS receivers by having a means of generating “spatially narrow” nulls in the antenna radiation pattern that can significantly reduce the power of an interfering RF signal or multiple signals. The radiation pattern needs to be reconfigurable (i.e. ability to steer the null in various spatial points) to accommodate the fact that receiving antennas might be mounted on a moving platform and the source of an interfering signal will be unknown.

Ideally, we would like to generate multiple (at least one or two) nulls in the otherwise omni-directional, hemi-spherical, circularly polarized antenna radiation pattern that can be reproduced over wide bandwidth with relatively stable phase center location (cm level). The phase center performance is an important requirement for high accuracy wireless positioning systems. Spiral antennas seem to be a natural candidate choice as they have a potential to meet all of these tough requirements. Spiral antennas are known to be wideband, provide circular polarization and have reconfigurable patterns that depend on the excitation mode of operation.
4.2 Spiral Antenna Background

The spiral antenna was introduced as a frequency independent antenna in 1950’s. Two-arm spiral antennas are formed from a spiraled two-wire transmission line that gradually transforms itself into a radiating structure. The equiangular spiral antenna published by J. D. Dyson in 1959 [45] and Archimedean spiral antennas [49, 58] are the most well-known spiral antennas. Spiral antennas have broad, about a 10:1 bandwidth, providing circular polarization in low-profile geometry [46]. In 1982, R. H. Duhamel invented the sinuous antenna, which provides dual-linear polarization in addition to wide bandwidth in a compact, low-profile geometry [46]. While the sinuous antenna is more complicated than the spiral antenna, it provides dual orthogonal linear polarizations so that it can be used for polarization diversity or for transmit/receive operation. Also, the two two-arm pair outputs can be combined to produce simultaneous LHCP and RHCP.

The spiral antenna design and its operation have been explained in many references. A good overview is given by J. Shelton [61] and R. Corzine [78]. The sinuous antenna does not provide sufficient flexibility required for reconfigurable pattern applications. The best performance of a sinuous antenna is achieved when its arms are excited from the inside spiral arm ends as opposed to outside spiral arm ends. However, the space between inside arms is very limited, hence it is difficult to implement a feed structure and/or reactive loads for multi-arm (4 or more arms) spiral antennas.

The Archimedean spiral antenna has the form of a self-complementary antenna, in which arm metallization width and the space between are equal. In theory, the self-complementary antenna has about 188Ω input impedance [47]. However, the experimental data revealed that the impedance ranges from 120Ω to 200Ω [48]. It is
generally well known that the Archimedean spiral antenna produces broad main beams perpendicular to the plane of the spiral. The broad bidirectional main beam can be converted to a unidirectional beam by backing the spiral with a ground plane, which is described in many patents and implemented in many products. In typical commercial products, absorbing material is loaded around the edge of the cavity or the cavity is completely filled to reduce the resonant effects and provide a unidirectional beam pattern. The typical performance parameters for the cavity-backed Archimedean spiral have a peak gain of between 4-6 dBiC, an Axial Ratio of 1-3 dB on boresight and bandwidth of 10:1. The input impedance can range between 100Ω to 188Ω. The performance of the equiangular spiral is similar to that of the Archimedean type spiral antenna.

Spiral antennas are attractive for communication applications where broadband characteristics with respect to both input impedance and radiation pattern are required. There have been extensive investigations regarding radiation characteristics of spiral antennas with different geometrical shapes such as circular, rectangular and eccentric [50, 51]. These antennas are mainly used to radiate a circularly polarized wave forming either an axial beam—normal to the plane of the spiral—or tilted beam pattern—off-normal to the plane of the spiral [52]. The single-arm spiral, which is used in this work, has the advantage of not requiring a balun circuit between the spiral and the feed line.

4.2.1 Spiral Antenna Basic Theory

Theory [45,83-93] shows that a finite multiport antenna with N pairs of feed terminals (i.e., ports) will produce exactly N linearly independent “radiation” modes whose linear combinations may describe any possible transmit or receive pattern for that
array (assuming the terminal port excitations produce linearly independent far zone fields, as is nearly always the case in practice). The scattering network description together with the mode functions can be used to completely characterize the transmitting and receiving properties of an antenna. A different modal description was developed in the 1960s by Garbacz [109] and later developed for antenna arrays [53]. The newer approach was to determine a set of orthogonal excitations that effectively eliminate mutual coupling between elements.

It is often convenient to decompose an arbitrary excitation into a set of modes, also referred to as eigenvectors. For circularly polarized antennas it is preferable to use exponential functions to define and express these modes. For an N-arm spiral antenna, to excite a mode $m$, the subsequent spiral arms will be excited with $2\pi m/N$ $(m=1, \ldots N-1)$ phase gradient at the antenna terminal, and the azimuth pattern is the summation of $p$ terms in the form of $e^{-j(m+pN)\phi}$.

The analytical description of a multiport antenna has typically been provided by the terminal port impedance, admittance, or scattering matrix together with radiation patterns obtained by successively exciting each consecutive port with the other ports terminated in known loads. Linear combinations of the resulting field patterns are used to obtain desired beam shapes. For the receiving case, equivalent circuits have been developed that allow calculation of the received signals for a multiport antenna under plane wave incidence. For example, [54] provides a rigorous derivation of such an equivalent circuit in the frequency domain and [55] provides the same in the time domain. Alternatively, [56] describes an N element multiport antenna by including a test antenna and considering the resulting N + 1 element antenna system.
4.3 Radiation Modes

The radiation modes of spiral antennas have been described in many papers [49, 79-82]. The multi-arm spiral antenna radiates from discrete regions of its aperture. The spiral arm currents must be in phase (or close to it) in these discrete regions, often referred to as \textit{active regions} [83-85].

An active region can be explained using the Fig. 15. The spiral antenna can be composed of two contours described by the curve $r(\theta) = a\theta$ associated with Archimedean type spiral antenna or by the curve $r(\theta) = e^{a\theta}$ associated with equiangular spiral antenna.

![Fig. 15 Spiral Arm Active Region](image)

Fig. 15 Spiral Arm Active Region. At point A, the currents in two arms are approximately $180^\circ$ out of phase, hence their radiation cancelled out. At point C, the currents in the two arms are now approximately in phase, their respective radiation adding up in broadside. The region near and adjacent to points C and B (shown in gray) is referred to as the active region of spiral antennas. The circumference distance between point B and C is half wavelength.
The vector $\vec{r}_s(\theta)$ is the position vector of the first spiral arm while the vector $\vec{r}_t(\theta)$ is the position vector of the second spiral arm. The relation between two vectors is:

$$\vec{r}_t(\theta) = -\vec{r}_s(\theta)$$  \hspace{20pt} (4.1)

Let’s assume that the current source is used to excite both arms using time harmonic excitation. A steady-state current in each spiral arm will have the same time variation adjusted for phase shift associated with distance travelled along the arm’s path. The current at point $\vec{r}_\pm(\theta)$ can be described by:

$$I_\pm(\theta, t) = i_\pm(\theta)e^{i\omega t}$$  \hspace{20pt} (4.2)

where $i_\pm(\theta)$ is a complex valued function of $\theta$.

Since the spiral is modeled with no outer truncation, the current may be assumed to consist of an outward traveling wave only (no reflection from outside truncation). If the phase velocity of the wave is assumed to be a constant, $v_p$, the phase of the current on the spiral at any point can be described by the travel time ($t_d$) from the origin assuming constant phase velocity ($v_p$). This travel time is $s\pm(\theta)/v_p$, where $s\pm(\theta)$ is the distance along the spiral path from the origin to the point $\vec{r}_\pm(\theta)$. The calculation of active region shown below is given for equiangular case due to closed-form solution available when solving for angle $\theta$ where currents on the opposite arms are in phase. This distance along the spiral arm can be computed as.
\[ s_\pm(\theta) - s_\pm(0) = \int_0^\theta \left| \frac{r_\pm(\theta)}{d\theta} \right| d\theta \]

\[ = \int_0^\theta \left( \sqrt{\left( \frac{dr(\theta)}{d\theta} \right)^2 + r^2} \right) d\theta \]

\[ = (e^{at} - 1)\sqrt{1 + a^{-2}} \] (4.3)

Assuming that the feed diameter is a fraction of operating wavelength and that there is a 180° phase excitation offset between two arms, implies that currents shown at location A are 180° out of phase:

\[ \angle l_+(\pi, t) - \angle l_-(0, t) \approx \angle l_+(0, t) - \angle l_-(0, t) = \pi \] (4.4)

The radiation contributions of two adjacent, closely spaced currents at location A will tend to cancel in the far field. Initially, as the current travels outward from the excitation point, all arms are out-of-phase and closely located, and little or no radiation occurs from antenna aperture. The cancellation will continue as long as the path traveled by both current is electrically small. As the currents travel along the spiral arms, the phase difference with respect to initial phase gradient starts to converge due to extra distance that each current had to travel. As \( \theta \) increases, there will be a point \( r_\pm(\theta) \), where

\[ s_-(\theta_1 + \pi) - s_-(\theta_1) = \frac{\lambda}{2} \] (4.5)

Spiral arms currents become in-phase at electrical distances large enough to allow the antenna to radiate. The point where this occurs can be found by solving (using 4.3 and 4.5)

\[ \theta_1 = \frac{1}{a} \left( \frac{\frac{\lambda}{2(e^{a\pi} - 1)\sqrt{1 + a^{-2}}}}{2(e^{a\pi} - 1)\sqrt{1 + a^{-2}}} \right) \] (4.6)
The simplest way to illustrate how to compute the active region area is to use an example of spirals with very tight winding. It is apparent that letting $a$ tend to zero, we can obtain an approximate radiation condition by equation $r(\theta) = e^{a\theta}$ and (4.6).

$$r(\theta_1) = e^{a\theta_1} = \frac{a\lambda}{2(e^{a\pi} - 1)\sqrt{1 + a^2}}$$

$$\approx \frac{a\lambda}{2(-1 + 1 + a\pi + \ldots)1} \approx \frac{\lambda}{2\pi}$$

$$2\pi r(\theta_1) \approx \lambda$$

(4.7)

A tightly wrapped spiral antenna will radiate in a circular band of circumference $\lambda$. A simulated current distribution and Far-Field normalized response are shown in Fig. 16. The left column represents the current distribution on the 4-arm tightly wound (1mm space between adjacent arms with 30 turns) spiral antenna. The areas of active region are clearly visible when relative currents in all arms are close to zero degrees (light green color). The active region forms a ring whose location depends on the operating frequency and excited mode. From the right column in Fig. 16 one can observe that the active region occupies a significant portion of antenna aperture. For frequency of 1 GHz (left upper corner) the active region occupies circumference between $\lambda \pm \lambda/2$ with radiation of at least -3 dB or higher with respect to the maximum radiation.
Fig. 16 Active region (left column) corresponding to Mode 1 of tightly wound 4-arm spiral at 1GHz (top right), 2GHz and 3 GHz (bottom left). The right column shows corresponding Far Field strength as a function of active region circumference in terms of wavelength.
A distinct radiation pattern corresponds to each mode of radiation. The number of modes of operations is directly proportional to the number of spiral arms. The mode $m$ is related to the phase gradient at which spiral arms are excited. The phase gradient at the antenna input terminals for a given transmission mode $m$ is given by:

$$\psi = \frac{2\pi m}{N} \quad m = 1, \ldots, N - 1$$  \hspace{1cm} (4.8)

where $N$ is the number of spiral arms.

This *active region* of radiation corresponds to the area located near the circumference of $m \cdot \lambda$. If the antenna aperture is large enough another, *active region* will occur at a circumference of

$$(kN + 1)m \cdot \lambda, \text{ where } k \geq 0 \text{ integer}$$  \hspace{1cm} (4.9)

This is clearly evident again in Fig. 16 in the right column. At higher frequencies (2 GHz and above) a second *active region* starts to emerge at circumference of $5\lambda$. The reason the second active region is not shown on the left column of Fig. 16 is that phase magnitude has not been unwrapped (the phase magnitude is a multiples of $2\pi$ at this location).

In the case of a 2-arm spiral antenna having an aperture of radius $r = 7\lambda/2\pi$, the *active region* for the $1^{\text{st}}$ mode ($m=1$) will occur at circumference of $1 \cdot \lambda, 3 \cdot \lambda, 5 \cdot \lambda, \text{ and } 7 \cdot \lambda$, while in the case of an 8-arm spiral antenna only one active region will be present at a circumference of $1 \cdot \lambda$ ($m=1$ mode). There will be a significant difference in radiation patterns of the $1^{\text{st}}$ mode between 2-arm and 8-arm spiral antennas. The radiation of the $1^{\text{st}}$ mode for a 2-arm spiral antenna at circumferences of $3 \cdot \lambda, 5 \cdot \lambda, \text{ and } 7 \cdot \lambda$ is equivalent to radiation of $3^{\text{rd}}, 5^{\text{th}}$ and $7^{\text{th}}$ modes. This is not the case for a 8-arm spiral
antenna which will radiate only a pure 1st mode radiation pattern. Using a multi-arm approach to spiral antenna design allows maintaining the same or similar radiation pattern over wide frequency bandwidth. As an example, refer to Fig. 17, which demonstrates that the first two active regions for 1st radiation mode \((m=1)\) of the 4-arm spiral antenna occurs at a circumference of \(1 \cdot \lambda\) and \(5 \cdot \lambda\). Designing an antenna with a maximum circumference dimension of less than \(5 \cdot \lambda\) will ensure that only the fundamental radiation mode is excited and hence no contamination of radiation pattern by higher order modes is observed.

The concept of modes on a spiral antenna was first realized by early inventors (see e.g.: Edwin Turner and John Dyson [45]), who designed a two-arm spiral antenna with a single fundamental mode of operation. Early researchers found that exciting a 2nd mode proved to be impossible with a two-arm spiral antenna. Introduction of multi-arm spiral antenna designs by Paul Shelton in 1960 [61] allowed an antenna to be excited in higher modes. The modes get excited by providing phase shifted signals that have phase gradients as per Eqn. (4.8)

The radiation pattern for a given mode \(m\) is identical to mode \(N-m\) except for the reversal of antenna pattern polarization. The fundamental modes \(m=1\) and \(m=N-1\) provide an omnidirectional pattern with peak directivity on antenna boresight but with opposite polarization. The remaining modes have a deep narrow null on the antenna boresight. The highest mode is generally not considered useful since it provides an unbalanced system (vector summation of the initial phase does not end up as zero) that causes power to be reflected back to the antenna feed line system. The presence of
various radiation modes provides a lot of flexibility in designing pattern reconfigurable antennas with added ability to steer a null towards the interfering signal.

4.3.1 Non-Integer Radiation Modes

The N-arm spiral antenna can also be excited with a slightly unbalanced higher order mode \( m \), normally associated with an \( M \)-arm spiral antenna, where \( M \) is greater than \( N \).

\[
\psi = \frac{2m\pi}{M} \quad m = 1, \ldots, M - 1, M > N \tag{4.10}
\]

This approach takes advantage of the fact that the effective radiation area (also referred to as the active region) around the circumference of \( m \times \) wavelength \( (m\lambda) \) is quite large and can be truncated. As shown in Fig. 17, the maximum radiation for the fundamental mode \( (m=1) \) does indeed occur at antenna circumference of one lambda \( (1.0 \times \lambda) \), however there is also a significant radiation within the area of \( a=0.5 \lambda \) to \( 1.5 \lambda \).

Fig. 17  Maximum Far Field (FF) radiation along azimuth rotation angle \( \varphi \), four-arm tightly wound spiral antenna, \( dr=2 \text{ mm} \), starting radius=60mm, freq=2 GHz
It can be observed that the shape of the *active region* is symmetrical around the antenna circumference (1*λ in Fig. 17) that corresponds to the excited radiation mode. If the *active* region is truncated on one side then the radiation contribution will be “uneven” resulting in a (non-symmetrical) pattern squinting. This can be implemented in two ways: first by exciting the antenna with an unbalanced higher order mode or, secondly, by truncating the antenna aperture. There will, of course, be some gain degradation (i.e. a 50% reduction of radiation area will translate to 3 dB loss), however we will gain the ability to steer the main beam or null off boresight (when exciting in mode 2) to another elevation angle θ. This is evident in Fig. 18 which shows various planar cuts of the normalized radiation pattern of a 4-arm spiral antenna excited in two different modes. Mode 1,4 is the 1st fundamental mode associated with a 4-arm spiral (phase gradient of 0°-90°-180°-270°) while a non-integer Mode 1,8 is the 1st fundamental mode associated with an 8-arm spiral (phase gradient of 0°-45°-90°-135° for 4-arm spiral).
Fig. 18  Normalized Far Field of four-arm spiral antenna excited using integer Mode 1, 4 (red curves) and non-integer Mode 1, 8 (black curves). Families of curves represent planar cuts for $\phi$ changing from 0 to 360°.

Each line in Fig 18 represents a separate vertical cut at various phi angles. The black curves indicate that at some phi angle a deep null (-47 dB) is present at theta angle near $\pm 40^\circ$. 


Fig. 19 *Active* regions of four-arm spiral antenna excited using integer Mode 1, 4 (red curve) and non-integer Mode 1, 8 (black curve). Spiral arm growth dr=4mm

Fig. 19 shows that exciting a 4-arm spiral with non-integer Mode 1, 8 causes the *active region* to be shifted towards the center of the antenna. This also causes a partial excitation of the 2nd mode from current present at antenna region bounded by circumference of 4-4.5 wavelengths. The presence of an unbalanced phase gradient in a shifted 1,8 *active region* will create a null in the radiation pattern as evident in Fig. 18 at Θ=±40°. Mode 1, 4 has an *active region* in the expected area of a circumference of one wavelength that correspond to 1st mode of radiation and hence no null.

Adaptive null forming is the process of minimizing the S/N of the received signal coming from a particular spatial angle(θ₀, φ₀). In our case the main interest is to adjust weights in order to minimize the ratio of an interfering signal (J/S) to the desired set of
satellite signals. The receiver can track the J/S (Jammer to desired Signal ratio) level while dynamically adjusting the parameters that control the location, size and depth of the null. A various forms of blind search algorithms can be employed to accomplish that task. Further research would be required to find an optimum implementation of such approach. Basic null theory and approach to create such null (without adaptive steering algorithm) is given next.

4.4 Null Forming

An antenna array system is composed of a collection of spatially separated antenna elements, the outputs of which are combined by a multiple or single input receiver. The antenna array system has the ability to dynamically adjust the combining mechanism in order to improve system performance compared with single antenna receivers. In the array antenna pattern, a null occurs where there is a reversal in the phase of the radiation pattern [98,100,101]. This method is based on the minimization of the output power of the array, obtained with a small perturbation of the phases of the radiating elements. The pattern distortion is usually very small, and depends on the amount of the phase perturbations. The depth of the null depends on the slope of the phase reversal [97].

4.4.1 Basic Null Theory

Generating nulls in the radiation pattern while maintaining uniform gain in any directions away from the null (in the whole upper hemisphere of the antenna) is not an easy design task. The spatial null width is typically achieved at the expense of increased physical dimensions of the antenna; however in our approach we keep the antenna size
constant while changing feeding and loading mechanism to achieve the objective of creating narrow nulls.

There are two types of nulls; a tangential null and a zero crossing null [97-101]. A tangential null occurs in the far-field pattern that has the same phase sign before and after the null achieves its minimum value. The far-field pattern with a zero-crossing null exhibits an abrupt phase reversal across the null. The zero-crossing null is narrower than the tangential null for a given physical dimension of the array. The zero-crossing nulls have been successfully used in the past for DOA and tracking (i.e. monopulse, difference patterns) applications.

One method of obtaining a zero crossing null pattern is to have two distinct radiation patterns due to two separate currents flowing in the antenna aperture. The second current is an attenuated version of the original current excited at the antenna terminals and reflected back into antenna aperture by its reactive load. This method is explored in detail in subsequent sections of this dissertation.

To generate a beam pattern from a spiral array the correct electrical phase compensations must be applied to the elements, so that the radiations from all elements will be reinforced in a given direction. For a spiral array, the required electrical phase compensation varies as a cosine function with rotation angle phi. The resulting directional pattern has been termed 'beam cophasal' in several papers [102], and for the case of omnidirectional elements and no amplitude tapers, takes the form of a zero-order Bessel function. If the excitation to the \( n \)-th element of an antenna array is \( i_n \), the far-field \( k \)-th mode radiation pattern for the array is
\[ D_k(\theta, \phi) = \sum_{n=0}^{N-1} i_n e^{j\xi \cos(\phi - \nu_n)} \]  
\[ \nu_n = k \frac{2\pi n}{N} \]

with k-th sequence excitation where \( i_n = e^{j\nu_n} \) the far field pattern is

\[ D_k(\theta, \phi) = \sum_{n=0}^{N-1} e^{j(n\nu_n + \xi \cos(\phi - \nu_n))} \]  
\[ \xi = \text{Bessel function of the k-th order} \]

Higher-order mode phase-sequence patterns have a zero along the vertical axis of the array, and an improved gain in the horizontal plane. To generate a beam-cophasal radiation in the direction \((\theta_0, \phi_0)\) the excitation of the n-th element is

\[ i_n = e^{-j\frac{n}{2} \sin \theta_0 \cos(\phi_0 - \nu_n)} \]  
\[ \xi = \text{Bessel function of the k-th order} \]

The corresponding far-field radiation beam pointing pattern for such an excitation is given by

\[ D_b(\theta, \phi) = \sum_{n=0}^{N-1} e^{jka \{\sin \theta \cos(\phi - \nu_n) - \sin \theta_0 \cos(\phi - \nu_n)\}} \]  
\[ a = \text{constant} \]

Again, if the inter element spacing is small enough, the array pattern in the horizontal plane will become
\[ D_b(\theta, \phi) = J_0 \left| 2k \sin \frac{1}{2} (\phi - \phi_0) \right| \]  

(4.16)

The appropriate weighting coefficients must be applied to the pattern functions in order to create a null in the direction \((\theta_0, \phi_0)\) by the superposition of the two modes of excitations shown above. The following condition must be satisfied:

\[ AD_b(\theta, \phi) + BD_k(\theta, \phi) = 0 \quad \theta = \theta_0 \quad \phi = \phi_0 \]  

(4.17)

The null is therefore created by passing the signal from the “input feed” through an appropriate phasing network that will combine in radiated space with an attenuated signal from the “reflective feed” created by the network of reactive loads at the opposite end of spiral arms. Additional analysis of null creation using non-integer mode excitation is discussed in Appendix A.

### 4.4.2 Null Implementation

The relationship between the reactive load configuration and adaptive nulling is not directly obvious, but in short it is related to the uniqueness of the resulting steering vectors for given load values. The solution to the problem is not well defined. Extensive experimentation and simulations were carried out to find an optimum antenna configuration. There is a possibility that better solutions may be found, since a finite number of experiments and simulations were carried out. The methods implemented are described in later Chapters.
4.4.2.1 Reactive Loads

The idea of having one (or only a few) array antenna element(s) driven and the other ones acting as parasitic elements is not new. It may be further noted that parasitic elements do not necessarily have to be only short circuited. Instead, the parasitic elements may be reactively (capacitively or inductively) loaded. By choosing the correct reactive loading of the parasitic array antenna elements, the array antenna beam/null may be directed into a desired direction.

The impedance of an element attached to an antenna at a control port will impact the currents that flow on the antenna. Control circuits can be electrically connected to the antenna element at the appropriately located control ports. Since the radiation pattern of an antenna is determined from the currents flowing on the antenna, attaching lumped impedance components (referred to as "reactive loads") to the antenna will affect the radiation (reception) pattern of the antenna. A load can be as simple as a resistor, capacitor, inductor, or a coaxial line of appropriate length. Alternatively a load can be a more complex, variable circuit or device such as a varactor. The purpose of reactive loads is to disrupt the current flow of the antenna in a desired manner for either interference rejection or direction-finding. Fixed impedance loads were used during the research stage in order to validate the undertaken concept. If electronically adjustable impedance, rather than a fixed one, is attached to the antenna, the tuning and radiation (reception) pattern of the antenna can be controlled by an external signal.

We can have various combinations of reactive load connections with the spiral arms. When two ports are used to excite the 4-arm spiral antenna that leaves six available ports (locates at spiral arm ends) which can give us 6! = 720 different connection
combinations. It would take a long time to test or simulate each combination. Adding different values of reactance would increase this problem tremendously. To simplify the problem, let’s consider our antenna as an 8-port network with 4 ports on the left hand side representing the outside spiral arms connection while 4 ports on the right hand side represent the inside arms connection, see Fig. 20. The box in the middle represents the spiral antenna itself.

![Network representation of the spiral arm antenna](image)

**Fig. 20** Network representation of the spiral arm antenna

**Example 1:** Port 2 and 3 are excited with 90 deg. phase gradient using 3-dB quadrature hybrid, and no reactive loads are connected to the remaining pins, see Fig. 21.

![Network schematic of 4-arm spiral antenna fed from outside two ports (P2, P3) using 90° 3-dB directional coupler](image)

**Fig. 21** Network schematic of 4-arm spiral antenna fed from outside two ports (P2, P3) using 90° 3-dB directional coupler
Example 2: Ports 2 and 3 are excited with a 90 deg. phase gradient using a 3-dB quadrature hybrid. Ports 6 and 7 are terminated with reactive load using a 3-dB quadrature hybrid, see Fig. 22.

Fig. 22 Network schematic of 4-arm spiral antenna fed from outside two ports (P2, P3) using a 90° 3-dB directional coupler, with two inside ports (P6, P7) terminated with reactive loads using a 90° 3-dB coupler.

There are many other combinations. It was found through experimentation that the following combinations yields the best performance (referred to as a cross-spiral antenna). See Fig. 23.

Fig. 23 Network schematic of a 4-arm spiral antenna fed from outside two ports (P1, P2) using a 90° 3-dB directional coupler, with two remaining outside ports (P4, P4) terminated with reactive loads using a 90° 3-dB coupler. Inside ports connected using cross-over joint connection.
4.5 Spiral Antenna Design Approach

4.5.1 Single Arm Spiral Antenna

It has been shown that the single arm Archimedean spiral, which has a high bandwidth, can produce tilted beams by changing its arm length [57].

Single arm rectangular spiral antennas radiate a tilted beam from the active regions [58]. In [59, 60], it was demonstrated that a single element spiral antenna can provide a steerable radiation pattern under electronic control without need of a complex feeding network. The direction of the antenna beam is controlled through a set of switches shorting the spiral arm to the ground plane at selected points.

A single arm rectangular spiral antenna with switches was proposed for adaptive beam applications [57-60]. The switches, located beyond one wavelength loop on the spiral arm, make the antenna arm open circuited at the selected points. The switch excitation introduces a current variation with a new phase distribution over the antenna arm, leading to a change in the direction of the radiated beam. In an experimental antenna, the VSWRs for these configurations were found to be less than 2 within a 200 MHz frequency band centered at 3.3 GHz. The gains for all the configurations are found to maintain uniformity at 6.3 dBi, with a 0.8 dB variation.

4.5.2 Archimedean Spiral

A tightly wound self-complementary Archimedean spiral antenna is shown in Fig. 24. The input impedance of a self-complementary antenna can be found using Babinet’s principle, giving [94]
\[ Z_{\text{metal}}Z_{\text{air}} = \frac{\eta^2}{4} \]  

(4.18)

where \( \eta \) is the characteristic impedance of the medium surrounding the antenna. For a self-complementary Archimedean spiral antenna in free space the input impedance should be

\[ Z_{\text{in}} = \frac{\eta_0}{2} = 188.5 \, \Omega \]  

(4.19)

The radius of each arm of an Archimedean spiral is linearly proportional to the rotation angle \( \phi \), and is described by the following relationships

\[ r_n (\phi + \frac{2\pi}{N} n) = a\phi + r_{\text{start}} \quad n = 0, \ldots, (N - 1) \]

and  \( \phi \in (0, 2\pi T) \)  

(4.20)

where \( r_n \) is the radius of the \( n \)-th spiral at a given angle \( \phi \), \( T \) is the number of turns and \( N \) is the number of spiral arms. The proportionality constant \( a \) is related to the radius expansion ratio \( dr \) as

\[ a = \frac{dr}{2\pi} \]  

(4.21)

where \( dr \) is the radius difference for a given arm after one full spiral turn.
Fig. 24 Self complementary tightly wound 4-arm Archimedean Spiral antenna with \( w=s=0.5\text{mm} \)

The proportionality constant \( a \) is also determined from the width of each arm, \( w \), and the spacing between each turn, \( s \), which for a self-complementary 2-arm spiral is given by

\[
a = \frac{s + w}{\pi} = \frac{2w}{\pi} = \frac{\text{dr}}{2\pi}
\]

Thus the spacing or width may be written as \((r_M \text{ is the maximum radius and } r_0 \text{ is the minimum radius of spiral antenna aperture})\)

\[
s = w = \frac{r_M - r_0}{2TN} = \frac{a\pi}{N}
\]
The Archimedean spiral antenna radiates from a region where the circumference of the spiral equals one wavelength for the 1\textsuperscript{st} radiation mode or M-multiples of wavelengths for the M\textsuperscript{th} radiation mode. This area of efficient radiation is called the active region of the spiral antenna. All spiral arms are fed with $2\pi M/N$ ($M=1,\ldots,N-1$) phase gradient at the antenna terminal, so when the circumference of the spiral is $M$ wavelengths, the currents on all arms of the spiral add in phase in the far field.

The low frequency operating point of the spiral is determined theoretically by the outer radius and is given by

$$f_{\text{low}} = \frac{c}{2\pi r_M} \quad \text{(4.24)}$$

Similarly the high frequency operating point is based on the inner radius giving

$$f_{\text{high}} = \frac{c}{2\pi r_0} \quad \text{(4.25)}$$

In practice the low frequency point will be greater than predicted (4.24) due to reflections from the end of the spiral. The reflections can be minimized by using resistive loading at the end of each arm or by adding conductivity loss to some part of the outer turn of each arm. Also, the high frequency limit may be less than found from (4.25) due to feed region effects. The sense of polarization is determined by wrapping fingers of the right hand around the axis of the spiral structure with the fingers pointing in the direction of circumferential current flow. The thumb points to the direction of RHCP radiation, the opposite direction yields LHCP radiation.
4.5.3 Four -Arm Spiral Antenna

The typical M-th mode radiation pattern of the 4-arm spiral antenna excited at 4 ports $P_i$ can be summarized as:

\[
\begin{bmatrix}
M_1 \\
M_2 \\
M_3 \\
M_4
\end{bmatrix} = \begin{bmatrix}
e^{-j\pi/2} & e^{-j\pi} & e^{-j3\pi/2} \\
e^{-j\pi} & 1 & e^{-j\pi} \\
e^{j\pi/2} & e^{j\pi} & e^{j\pi/2} \\
1 & 1 & 1 & 1
\end{bmatrix} \cdot \begin{bmatrix}
P_1 \\
P_2 \\
P_3 \\
P_4
\end{bmatrix}
\] (4.26)

The above matrix equation does not describe the radiation pattern but the mode of radiation pattern for a given ports excitation. All $M_i$ modes have a wide beam omnidirectional pattern with the exception of the $M_2$ mode. A deep null present on boresight of the $M_2$ mode has been used in the past for DF applications where a switching circuit allows alternating between the 1st and 2nd radiation pattern modes. Using sub-integer modes of excitation, the radiation modes matrix can be presented as

\[
\begin{bmatrix}
M_{k1} \\
M_{k2} \\
M_{k3} \\
M_{k4}
\end{bmatrix} = \begin{bmatrix}
e^{-j2\pi/k} & e^{-j3\pi/k} & e^{-j4\pi/k} \\
e^{-j4\pi/k} & e^{-j6\pi/k} & e^{-j8\pi/k} \\
e^{-j6\pi/k} & e^{-j9\pi/k} & e^{-j12\pi/k} \\
e^{-j8\pi/k} & e^{-j12\pi/k} & e^{-j16\pi/k}
\end{bmatrix} \cdot \begin{bmatrix}
P_1 \\
P_2 \\
P_3 \\
P_4
\end{bmatrix}
\] (4.27)

where $k > 4$ (and greater than N in case of N-arm spiral antenna). Since the gradient is smaller than in a typical case, the effective region starts earlier than in the typical case ($k=4$).

Another approach is to use partial excitation (one, two or three arms are excited) with remaining arms acting as parasitic elements terminated with the reactive loads (denoted as $L_3$ and $L_4$ in the equation 4.28).

\[
\begin{bmatrix}
M_{k1}^* \\
M_{k1}^* \\
M_{k1}^* \\
M_{k1}^*
\end{bmatrix} = \begin{bmatrix}
e^{-j2\pi/k} & e^{-j\delta_1} & e^{-j\delta_2} \\
e^{-j4\pi/k} & e^{-j\delta_1} & e^{-j\delta_2} \\
e^{-j6\pi/k} & e^{-j\delta_1} & e^{-j\delta_2} \\
e^{-j8\pi/k} & e^{-j\delta_1} & e^{-j\delta_2}
\end{bmatrix} \cdot \begin{bmatrix}
P_1 \\
P_2 \\
L_3 \\
L_4
\end{bmatrix}
\] (4.28)
4.5.3.1 Archimedean 4-arm Spiral Antenna (spiral growth \( d_r = 4 \text{mm} \)) – Type 1

Various spiral antenna prototypes were designed and built to conduct the measurement campaign. In all cases the number of spiral arms was fixed to 4, and the same maximum aperture radius of 60.5 mm was used. The targeted operating bandwidth was from 1-2 GHz to cover sufficiently all GNSS frequency bands. The design theoretically should extend to the 4-5 GHz range. The following parameters were used as variables in order to determine the desired antenna performance:

1) Phase gradient between adjacent arms

2) Location and number of reactive loads

3) Interconnection arrangements between reactive loads and spiral arm ends

4) Values of the reactive loads

Refer to Fig. 25 for the layout of one arm of an Archimedean Type 1 antenna.

Fig. 25 Archimedean 4-arm spiral antenna (single arm shown), \( d_r = 4 \text{mm} \)
<table>
<thead>
<tr>
<th>Design Parameter</th>
<th>Inside Curve</th>
<th>Outside curve</th>
</tr>
</thead>
<tbody>
<tr>
<td>Starting radius ($r_0$)</td>
<td>2.00 mm</td>
<td>2.50 mm</td>
</tr>
<tr>
<td>Ending radius ($r_{end}$)</td>
<td>60.00 mm</td>
<td>60.50 mm</td>
</tr>
<tr>
<td>Total spiral length (L)</td>
<td>2825.3 mm</td>
<td>2848.1 mm</td>
</tr>
<tr>
<td>spiral width (w)</td>
<td>0.5 mm</td>
<td></td>
</tr>
<tr>
<td>slot width (s)</td>
<td>0.5 mm</td>
<td></td>
</tr>
<tr>
<td>Number of turns (T)</td>
<td>14.5</td>
<td></td>
</tr>
</tbody>
</table>

Table 1 Design parameters of a 4-arm Archimedean antenna (type 1)

4.5.3.2 Archimedean 4-arm Spiral Antenna (spiral growth dr=8mm) – Type 2

The Type 2 Archimedean antenna (see Fig. 26) has twice the width of each arm, but half the number of turns. The planar area occupied is the same for both – Type 1 and 2 spiral antennas.

![Archimedean 4-arm spiral antenna](image)

Fig. 26 Archimedean 4-arm spiral antenna (single arm shown), dr=8mm
### Table 2 Design parameters of a 4-arm Archimedean antenna (type 2)

<table>
<thead>
<tr>
<th>Design Parameter</th>
<th>Inside Curve</th>
<th>Outside curve</th>
</tr>
</thead>
<tbody>
<tr>
<td>Starting radius ($r_0$)</td>
<td>2.00 mm</td>
<td>2.50 mm</td>
</tr>
<tr>
<td>Ending radius ($r_{end}$)</td>
<td>60.00 mm</td>
<td>60.50 mm</td>
</tr>
<tr>
<td>Total spiral length (L)</td>
<td>1414.3 mm</td>
<td>1437.0 mm</td>
</tr>
<tr>
<td>Spiral width (w)</td>
<td>1.0 mm</td>
<td></td>
</tr>
<tr>
<td>Slot width (s)</td>
<td>1.0 mm</td>
<td></td>
</tr>
<tr>
<td>Number of turns (T)</td>
<td>7.25</td>
<td></td>
</tr>
</tbody>
</table>

#### 4.5.4 Equiangular Spiral

An equiangular four-arm spiral antenna is also used to assess the feasibility of creating nulls while maintaining handedness of circular polarization. The equiangular spiral is a geometrical configuration whose parameter surface can be completely described by angles. It fulfills all the requirements for shapes that can be used to design frequency independent antennas as stated originally by Rumsey [56].

Since a curve along its surface extends to infinity, it is necessary to truncate the length of the arm to meet the finite size antenna geometry. The lowest frequency of operation occurs when the total arm length is comparable to the wavelength. For all frequencies above this, the pattern and impedance characteristics are frequency independent. An equiangular spiral array is defined by:

$$r = r_0 e^{a\phi} \quad (\text{inside curve})$$

$$r(\phi) = r_0 e^{a\phi + a} \quad (\text{outside curve})$$

An example of 4- arm equiangular spiral antenna is shown in Fig. 27.
The width of the arms also changes with angle. The gap to arm ratio for an N-arm spiral antenna is defined by

$$\frac{gap}{arm} = \frac{2\pi}{N\delta} - 1$$  \hspace{1cm} (430)$$

where $\delta$ is a rotation angle required to create the 2\textsuperscript{nd} edge of each spiral arm.

Two quadrature arms are driven with a 90° phase difference in order for the antenna to be predominately circularly polarized. The other two arms (parasitically coupled to driven arms) are loaded with a reactance on both ends or on one end of the spiral. The antenna is designed to cover the 1.15-1.65 GHz frequency range in order to be able to receive various GNSS signals such American GPS, European Galileo, Russian Glonass or Chinese Compass navigation satellite system, see Fig. 28.
Fig. 28 Equiangular 4-arm spiral antenna (single arm shown for clarity)

<table>
<thead>
<tr>
<th>Design Parameter</th>
<th>Inside Curve</th>
<th>Outside curve</th>
</tr>
</thead>
<tbody>
<tr>
<td>Starting radius ((r_0))</td>
<td>4.94 mm</td>
<td>5.19 mm</td>
</tr>
<tr>
<td>Ending radius ((r_{end}))</td>
<td>59.91 mm</td>
<td>60.91 mm</td>
</tr>
<tr>
<td>Total spiral length (L)</td>
<td>1100.8 mm</td>
<td>1157.2 mm</td>
</tr>
<tr>
<td>Spiral width (w)</td>
<td></td>
<td>0.254 mm</td>
</tr>
<tr>
<td>Slot width (s)</td>
<td></td>
<td>2.969 mm</td>
</tr>
<tr>
<td>Number of turns (T)</td>
<td></td>
<td>1.000 mm</td>
</tr>
<tr>
<td>Starting radius ((r_0))</td>
<td></td>
<td>8</td>
</tr>
</tbody>
</table>

Table 3 Design parameters of a 4-arm equiangular antenna
4.5.5 Feeding/Excitation Mechanism

The concept of employing reactive loads is based on exciting only a subset $k$ of all available $N$ arms ($k < N$). The remaining arms would be terminated in reactive components ($L$, $C$) which, in turn, would allow for controlling the phase shifts of the reflected waves. These phase shifts, in turn, will allow for steering the pattern’s main beam/null to a different spatial location. The “excited” and “reactively loaded” arms are connected together in order to increase the mutual coupling between them. Such a connection is easiest to implement in the center of the spiral arm antenna, which necessitates that antennas be fed from the outside. A couple of such arrangements for a four-arm spiral antenna are shown in Fig. 22 and Fig. 23.

Table 4 (see page 88) displays the phase gradient when spiral arms are open-ended and short-ended. Terminating open spiral arms with reactive loads provides a broad range of phase progression of the reflected wave. It will be shown that phase progression adjustment of the reflected currents will allow fine steering of the nulls created in the antenna radiation pattern while the coarse steering is achieved by changing phase progression values between adjacent arms using non-integer excitation modes as described in Chapter 5 and 6. The useful ranges of reactive loads are shown in Fig. 29.

 Capacitive loads between 0.5pF-15pF and inductive loads between 0.5nH-30nH provide almost a full $\pm 180^\circ$ range to adjust the phase of the reflected energy. The mechanism to control the null position utilizes varying reactive loading of the passive arms and sequential rotation of excited ports.
4.5.5.1 90° 3-dB hybrid coupler

A 3 dB, 90° degree hybrid plays an important role in the reconfigurable pattern spiral antenna presented in this thesis. Directional couplers are an important category of passive microwave circuits. They are implemented in a variety of ways resulting in a range of capabilities and constraints. Directional couplers can be used to sample propagating microwave energy for the purpose of monitoring or measuring. Directional couplers are also used to divide signals from a single channel into multiple channels in both small signal and large signal applications.

The most general definition for an ideal directional coupler describes it as a multi-port matched, lossless, reciprocal circuit with an infinite isolation between each pair of input ports. Four port directional couplers are usually represented schematically by two single line representations of transmission lines with a crossed line between them to represent the coupling between the lines, as shown in Fig. 30.

Fig. 29 Reflection Angles of various Reactive Loads (each curve represents distinct frequency from 1-4 GHz)
The ideal hybrid conditions are, see equation 4.31b.

\[ S_{11} = S_{22} = S_{33} = S_{44} = 0 \text{ (Matched condition)} \quad (4.31a) \]

\[ S_{23} = S_{14} = 0 \quad (Isolation \text{ condition}) \quad (4.31b) \]

\[ S_{ij} = S_{ji} \quad (Reciprocal \text{ condition}) \quad (4.31c) \]

The original coupler parameters [95] can be modified such that \( S_{12} \) is represented as a positive real number \( C_1 \) and \( S_{13} \) is then represented as a complex number with magnitude \( C_2 \) and phase \( \theta \). All of the other S-parameters are then determined in terms of \( C_1, C_2, \) and \( \theta \). The general coupler scattering matrix can be presented as

\[
S = \begin{pmatrix}
0 & C_1 & C_2 e^{j\theta} & 0 \\
C_1 & 0 & 0 & C_2 e^{j(\theta-\pi)} \\
C_2 e^{j\theta} & 0 & 0 & C_1 \\
0 & C_2 e^{j(\theta-\pi)} & C_1 & 0
\end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix}
0 & 1 & j & 0 \\
1 & 0 & 0 & j \\
j & 0 & 0 & 1 \\
0 & j & 1 & 0
\end{pmatrix} \quad (4.32)
\]

where specifically for an ideal 3dB hybrid coupler \( \Theta=\pi/2 \) and \( C_1=C_2=1/\sqrt{2} \). The magnitudes of the direct and coupled outputs are equal but the coupled output leading the direct port by 90°.
The 3 dB 90° hybrid coupler serves a double purpose. The main goal is to introduce a phase reversal of the reflected energy. Secondly, the hybrid provides a good isolation between different impedances as encountered in this case (between antenna and reactive loads). When both output ports are connected to identical load impedances, all reflected energy is cancelled at the input port and re-directed to the isolation port with 90° phase shift ($S_{41}=j$ for open ended loads and $S_{41}=-j$ for short ended). In case of equal reactive loading the total phase shift will be $90°+\delta_L$, where $\delta_L$ is the shift associated with the reactance load of the coupler $\Rightarrow \tan (\delta_L) = 4\Gamma = 4\left[(X_L - Z_0)/(X_L + Z_0)\right]$. Various cases are presented in Fig. 31.

Fig. 31  Signal reflection at an input ports with output ports terminated with various lossless terminations
Another approach is to short circuit one output port while open circuiting the other output port. This will cause reflected energy to have a 180° phase shift reversal. In the case where two different reactive loads are connected to the 3-dB hybrid all energy will still be reflected with 180° phase reversal (see proof below). In all the cases above, the reflected energy from the spiral arms (terminated with 3 dB 90° hybrid coupler) will contribute to the co-pol radiation rather than cross-pol radiation. The phase reversal of the reflected energy will later re-radiate as co-pol component since the arm handedness is also reversed in that case.

The analysis of a coupler terminated with two different reactive loads is presented next. Assume an inductive load is connected to port 2 and a capacitive load connected to port 3 of the 3-dB hybrid. The reflection coefficient of the inductive load connected to the ideal hybrid is

\[
\Gamma_L = \frac{jX_L - Z_0}{jX_L + Z_0} = \frac{Z_0^2 - X_L^2}{Z_0^2 + X_L^2} - j \frac{2Z_0X_L}{Z_0^2 + X_L^2} = re^{-j\delta_L} = 1e^{-j\delta_L}
\]

(4.33)

since

\[
r = \sqrt{\frac{(z_0^2-x_L^2)^2 + (2z_0x_L)^2}{(z_0^2+x_L^2)^2}} = \frac{(z_0^2+x_L^2)^2}{(z_0^2+x_L^2)^2} = 1, \quad \tan\delta_L = -\frac{2z_0x_L}{z_0^2-x_L^2}
\]

(4.34)

similarly in the case of capacitive load:

\[
\Gamma_C = \frac{-jX_C - Z_0}{jX_C + Z_0} = \frac{X_C^2 - Z_0^2}{Z_0^2 + X_C^2} - j \frac{2Z_0X_C}{Z_0^2 + X_C^2} = re^{j\delta_C} = 1e^{j\delta_C}
\]

(4.35)

\[
r = \sqrt{\frac{(X_C^2 - Z_0^2)^2 + (2Z_0X_C)^2}{(Z_0^2 + X_C^2)^2}} = \frac{(Z_0^2 + X_C^2)^2}{(Z_0^2 + X_C^2)^2} = 1, \quad \tan\delta_C = \frac{2Z_0X_C}{Z_0^2 - X_C^2}
\]

(4.36)

The energy reflected at ports 1 and 4 are:
\[ b_2 = \frac{a_1}{\sqrt{2}} + j \frac{a_4}{\sqrt{2}} \quad \text{and} \quad b_3 = \frac{a_4}{\sqrt{2}} + j \frac{a_1}{\sqrt{2}} \quad (4.37) \]

since \[ a_2 = b_2 \Gamma_L \quad \text{and} \quad a_3 = b_3 \Gamma_C \]

\[ b_1 = \frac{a_2}{\sqrt{2}} + j \frac{a_3}{\sqrt{2}} = \frac{b_2 \Gamma_L}{\sqrt{2}} + j \frac{b_3 \Gamma_C}{\sqrt{2}} = \frac{a_1}{2} (\Gamma_L - \Gamma_C) + j \frac{a_4}{2} (\Gamma_L + \Gamma_C) \quad (4.38) \]

\[ b_4 = \frac{a_3}{\sqrt{2}} + j \frac{a_2}{\sqrt{2}} = j \frac{b_2 \Gamma_L}{\sqrt{2}} + \frac{b_3 \Gamma_C}{\sqrt{2}} = j \frac{a_1}{2} (\Gamma_L + \Gamma_C) - \frac{a_4}{2} (\Gamma_L - \Gamma_C) \quad (4.39) \]

\[ \Gamma_1 = \frac{b_1}{a_1} = \frac{1}{2} (\Gamma_L - \Gamma_C) + j \frac{a_4}{2a_1} (\Gamma_L + \Gamma_C) = \frac{1}{2} (\Gamma_L - \Gamma_C) + \frac{j e^{jx}}{2} (\Gamma_L + \Gamma_C) \quad (4.40) \]

\[ \Gamma_4 = \frac{b_4}{a_4} = -\frac{1}{2} (\Gamma_L - \Gamma_C) + j \frac{a_1}{2a_4} (\Gamma_L + \Gamma_C) = -\frac{1}{2} (\Gamma_L - \Gamma_C) + \frac{j e^{-jx}}{2} (\Gamma_L + \Gamma_C) \quad (4.41) \]

And assuming that \[ a_4 = a_1 e^{jx} \]

With few simple algebraic manipulations, it can be shown that \[ 4 \Gamma_1 - 4 \Gamma_4 = 180^\circ. \]

With 90° phase gradient \[ a_4 = j a_1 \] (4.40 and 4.41) will simplify to

\[ \Gamma_1 = \frac{1}{2} (\Gamma_L - \Gamma_C) + \frac{j^2}{2} (\Gamma_L + \Gamma_C) = -\Gamma_C \quad (4.42) \]

\[ \Gamma_4 = -\frac{1}{2} (\Gamma_L - \Gamma_C) + \frac{1}{2} (\Gamma_L + \Gamma_C) = \Gamma_C \quad (4.43) \]

Again 180° phase reversal is present between ports 1 and 4.

\[ S_{41} = S_{14} = j \frac{e^{-j\delta_L} + e^{j\delta_C}}{2} = j \Gamma_1 \quad (4.44) \]

Numerical simulations were performed using a 4-port network using ideal 3 dB 90 hybrids with two output ports terminated with various values of capacitive and inductive reactive load, see Fig. 32 and Fig. 33.
Fig. 32  Simulated Reflection Coefficient of two input ports

Fig. 33  Matlab Simulated reflection RL amplitude and phase difference versus frequency

Similar amplitude response and exact phase response shown in Fig. 33 was obtained using commercial Agilent simulation software (see Fig. 34). The Matlab simulation used a signal flow approach with 3-dB hybrid represented as 4-port network (see diagram in Appendix B).
Fig. 34 Agilent ADS Simulated reflection coefficient amplitude and phase difference versus frequency with reactive loads of 20pF and 10 nH

The difference in amplitude response between Matlab and Agilent simulation was due to the use of ideal components in Agilent simulation. The main point of the simulation was to verify the 180° phase reversal of the reflected signal between two input ports.

4.6 Proposed Antenna Design

A typical multi-arm (N-arm) spiral array can have various modes of operation that depend on the phase gradient across its arms. As described in Section 4.2, the radiation pattern mode corresponds to, so called, active regions of the antenna aperture. These active regions correspond, in turn, to a circumference that is an integer multiple of the operating wavelength. The multiplying integer \( n \) is the excited mode at the antenna terminals. All modes of operation can be excited provided that the antenna aperture is large enough to support them. Each mode allows for generation of different antenna
pattern modes with \( n \) between 2 and N-2 and is capable of forming radiation pattern nulls. This dissertation proposes a way to excite non-integer modes of radiation that allow steering of the null away from antenna boresight to other elevation angles. Furthermore, a reactive loading of spiral arm ends offers additional ability to control the steering and hence the location of nulls in the antenna pattern.

4.6.1 Novel spiral antenna configurations

Spiral antennas should, in theory, radiate a circularly polarized field along its axis with a polarization sense corresponding to the winding sense of the spiral. However, the presence, on the spiral arms, of a current wave flowing in the opposite direction to the desired one increases cross-polarization levels and degrades circular polarization to elliptical and even linear polarization. Such a current can, for instance, be caused by reflections at the ends of spiral arms caused by their abrupt truncation. Various methods have been used to suppress the reflected energy. Creation of nulls will disturb the radiation pattern and cause negative effects such as reduced gain, increased AR, etc. New methods of improving antenna polarization purity is investigated in this dissertation and presented in subsequent sections of chapters 4-6.

4.6.2 Non-Integer mode excitation

For historical reasons, most spiral antennas have been designed with excitation ports located at the center (inside ends of spiral arms) of the antenna. In our designs we use external spiral arms for all port locations (excitation and reactive loading). The only difference between inside-fed and outside-fed spiral antennas is the polarization sense of
radiated fields, while offering lots of space to implement large number of ports in the case of a multi-arm spiral antenna.

The outside fed spiral antennas exhibit the same problem of reflected energy at the truncated spiral arm ends as inside-fed spiral arms. Figures 35-36 below illustrate the problem.

![Diagram of a LHCP polarized outside-fed four arm spiral antenna](image)

**Fig. 35** A LHCP polarized outside-fed four arm spiral antenna (assume radiation coming out of the page) with shallow metal cavity backing the spiral arms

Let us assume that a 4-arm spiral antenna is fed from the outside using $0^\circ, -90^\circ, 180^\circ, +90^\circ$ phase progression to create a Mode-1 LHCP radiation pattern that radiates out of the page towards the reader. Opened or shorted inner ends of spiral arms will reflect any incident energy (that was not fully radiated in the *active region*) causing the reversal of phase gradient across the spiral arms as shown in Fig. 36. It is interesting to note that the phase gradient of the reflected wave is the same regardless of whether the spiral ends
are opened or shorted. It can be shown that a similar argument applies if inductive or capacitive loads are attached to the inner arm ends.

![Diagram showing RHCP cross-polarized fields from open-ended (left side) and short-ended (middle) inside spiral arm ends, versus LHCP fields from combination of open and shorted arms (right side)](image)

**Fig. 36** RHCP cross-polarized fields from open-ended (left side) and short-ended (middle) inside spiral arm ends, versus LHCP fields from combination of open and shorted arms (right side)

<table>
<thead>
<tr>
<th>Phase Gradient</th>
<th>All opened arms, $\Gamma=+1$</th>
<th>All shorted arms, $\Gamma=-1$</th>
<th>One pair opened and another pair shorted</th>
</tr>
</thead>
<tbody>
<tr>
<td>$0^\circ$</td>
<td>$0^\circ$</td>
<td>$180^\circ$</td>
<td>$0^\circ$</td>
</tr>
<tr>
<td>$-90^\circ$</td>
<td>$-90^\circ$</td>
<td>$+90^\circ$</td>
<td>$+90^\circ$</td>
</tr>
<tr>
<td>$180^\circ$</td>
<td>$180^\circ$</td>
<td>$0^\circ$</td>
<td>$180^\circ$</td>
</tr>
<tr>
<td>$+90^\circ$</td>
<td>$+90^\circ$</td>
<td>$-90^\circ$</td>
<td>$-90^\circ$</td>
</tr>
</tbody>
</table>

Table 4 Reflected phase gradient between spirals arms (open and shorted-ended arms)

However, it was noticed that the reflected energy can still have correct phase gradient that will contribute to the co-pol radiation instead of unwanted cross-pol radiation if one pair of spiral arms are shorted while the other arms pair is left open-ended (Table 4). This provides a simple method of improving the axial ratio of spiral antennas. As far as we know, no such observation was made in any spiral antenna publications or
commercially made antenna products. There are two additional possible methods to avoid the phase gradient reversal that caused unwanted RHCP radiation.

Fig. 37  LHCP polarized outside-fed, inside cross-joined four arm spiral antenna (Type A)

The proposed antenna shown in Fig. 37 has two external ports excited in 0°, -90° phase gradient with inner opposite arms joined together. This arrangement allows for maintaining the same phase gradient of the back-travelling energy as forward-travelling energy. The remaining two external arms are connected to passive ports that can be either used to absorb that incoming reflected energy or reflect again using reactive loading. It is shown later in this work that reactive loading provides an ability to dynamically change the antenna radiation pattern. This arrangement introduces an antenna that has better polarization purity than if we had the reactive loads connected at the center of the antenna (i.e. connected to the inner arm ends)

Another approach is to connect two opposite sense spiral antennas in the center while again exciting them at the outer ends, see Fig. 38 below. The sense of handedness is left-handed for the top layer (black color) spiral antenna when excited from outside and
also left-handed for the lower layer (red color) when excited from the inside. This arrangement reduces the cross-polarization level of the reflected current at the arm ends, since the spiral arm handedness and phase gradient are opposite to each other. The spiral arm handedness and phase gradient coincide in a normal spiral antenna configuration, therefore providing efficient radiation of a cross-polarized signal.

Two spiral antennas are separated by a small vertical distance \( h \) (i.e. substrate thickness) and the inner arms are connected together using vertical vias. This configuration allows multiple polarization diversity similar to sinuous antennas. The difference is in the fact that the sinuous antenna is an equiangular type while this is an Archimedean type and provides lots of space to accommodate room for multiple ports. In our case we have eight available ports on the outside perimeter of the spiral antenna.

![Diagram of LHCP dual polarized outside-fed, inside joined four arm spiral antenna (Type B)](image-url)

Fig. 38  LHCP dual polarized outside-fed, inside joined four arm spiral antenna (Type B)
4.7 Cavity and absorber effects

Spiral antennas are typically backed by a lossy cavity, which restricts the radiation to one hemisphere and improves impedance bandwidth at the expense of a 2-3dB gain reduction due to the decrease in antenna efficiency. To increase the gain, shallow cavities with conducting ground planes have become more popular. These types of spirals have more gain but the axial ratio and pattern bandwidths are reduced compared to spirals backed by absorber filled lossy cavities.

Two types of shallow cavities (h=15mm and h=20 mm) were used to build various antenna prototypes. This translates to λ/20 to λ/5 in the operating range of the antenna (1-4 GHz). A layer of RF absorber was placed inside the vertical walls of the cavity to reduce the amount of reflections from side walls but not the bottom wall. The radiated energy from the spiral antenna towards the bottom of the cavity will be cross-polarized and it will reverse its polarization to co-polar after the reflection from the cavity. In actual reality there could be a multiple reflections but the underlying conclusions can be assumed to be the same. The reflected energy will add vectorially with the other half of energy transmitted away from the antenna in the co-pol mode. The vectorial addition will have a phase error offset that is equal to double the depth of the cavity. The phase error will range between 36° and 144°, which translate to reduced gain and increased cross-pol level at higher frequencies. This will limit the useful range of bandwidth by half unless a conical type cavity is employed to provide constant distance between bottom of the cavity and the antenna in terms of wavelength distance.
A shallow cavity is needed in order to produce an antenna with a wider bandwidth. The shallow ground plane increases the power reflected into higher modes, however that can be mitigated by increasing the number of spiral arms to move higher order modes to active regions beyond the physical dimensions of the antenna.

4.8 Antenna Phase Center Variations

4.8.1 Introduction

The electrical location of the mean phase center offset (PCO) of an antenna plays an important role in many modern scientific and engineering applications. It matters in such applications as ranging (positioning, attitude, surveying, etc.), imaging applications (radio telescopes, image radars), antenna arrays (i.e. null steering), and ultra wideband (UWB) with signal cohesion and group delay distortion caused by phase center variations (PCV). Typical antenna experiences horizontal and vertical phase center shifts that vary depending on the direction to the satellite, and the direction in which the antenna is pointing. This translates to ranging error and variation in computed position estimate. This unpredictable behavior does not tend to cancel during double-difference range processing. Its random nature also means that it cannot be effectively modeled, leading to additional errors in the baseline solution. The phase center location offset from known mechanical center is often clearly marked and documented in most high-end precision antennas. There are specific measurement sites\(^1\) devoted to calibration of high-end antennas and publishing phase center offset results on their web sites for general access. Placement of nulls in the antenna reconfigurable pattern will cause the phase center to shift. It is desirable to
minimize this shift while performing various null placements during RF interference mitigation.

The phase center is defined in the IEEE standards as: “The location of a point associated with an antenna such that, if it is taken as the center of a sphere whose radius extends into the far-field, the phase of a given field component over the surface of the radiation sphere is “essentially” constant, at least over the portion of the surface where the radiation is significant”.

Since nothing is perfect, this “real” measured phase sphere will have deviations from an “ideal” phase sphere centered on the mean phase of the antenna as shown in Fig. 39. Three terms that are important for GPS ranging applications are PCO (phase center offset), PCV (phase center variation), and ARP (antenna reference point).

Fig. 39. PCO model (as per Zeimetz and Kuhlmann, 2006) [110]

PCO is the mean location of the antenna phase center which normally does not coincide with ARP (a visible outside reference point on an antenna chassis). A typical antenna installation for surveying is made using the ARP point; therefore a correction
must be made to translate this point to PCO and vice-versa. Any signal received from a
given satellite in direction \( r_0 \) will experience a phase shift caused by PCV. For millimeter
(mm) ranging applications, such shift has to be compensated for, in order to achieve
precise (sub-millimeter) electrical position of the antenna. Such correction is computed
using (4.45)

\[
S_{ARP} = r + \text{PCO} \cdot r_0 + PCV(\theta, \phi) + \varepsilon
\]  

(4.45)

where \( \phi, \theta \) are azimuth and elevation angles towards the signal source. PCO is found by
minimizing the following cost function over all spherical points.

\[
\sum (PCV)^2 = \min
\]  

(4.46)

4.8.2 Phase Center Determination

There are several papers published on the subject of phase center determination.
They can be classified into: Second Derivative Method [62], Two Point Method [63],
Edge Diffraction Method [64], Differential Phase Method [65], Three Antenna Method

The Spherical Phase Method is a simple and practical method of computing the
phase center based on real antenna measurements (i.e. from an anechoic chamber) or
simulated antenna radiation patterns. The least squares method coupled with the spherical
phase expansion from its origin located at point PCO \((x,y,z)\) is shown in equation 4.47a.
A term “\( c \)” is added that can represent some fixed constant. A measured or computed
radiation phase pattern can be described as
Phase(\(\theta, \phi\)) = \frac{2\pi}{\lambda} (x \cdot \cos\phi\sin\theta + y \cdot \sin\phi\sin\theta + z \cdot \cos\theta) + c \quad (4.47a)

or in matrix form:

\[
F(\theta, \phi) = \frac{2\pi}{\lambda} \begin{bmatrix} \cos\phi\sin\theta & \sin\phi\sin\theta & \cos\theta & 1 \end{bmatrix} \begin{bmatrix} x \\ y \\ z \\ c \end{bmatrix} = \Rightarrow F = M \cdot P \quad (4.47b)
\]

Using the pseudo-inverse method we can find the phase center \(P\) from (4.47b), hence

\[
P = M^{-1} \cdot F = \begin{bmatrix} x \\ y \\ z \\ c \end{bmatrix} = \frac{\lambda}{2\pi} \begin{bmatrix} \cos\phi\sin\theta & \sin\phi\sin\theta & \cos\theta & 1 \end{bmatrix}^T \cdot F(\theta, \phi) \quad (4.48)
\]

We are now ready to assemble a solution matrix using measurements obtained from sampling the phase pattern of the antenna to obtain a mean value of PCO.

\[
\begin{bmatrix} x \\ y \\ z \\ d \end{bmatrix} = \frac{1}{2\pi} \begin{bmatrix} \cos\phi_1\sin\theta_1 & \sin\phi_1\sin\theta_1 & \cos\theta_1 & 1 \\ \cos\phi_2\sin\theta_1 & \sin\phi_2\sin\theta_1 & \cos\theta_1 & 1 \\ \vdots & \vdots & \vdots & \vdots \\ \cos\phi_N\sin\theta_1 & \sin\phi_N\sin\theta_1 & \cos\theta_1 & 1 \\ \cos\phi_1\sin\theta_2 & \sin\phi_1\sin\theta_2 & \cos\theta_2 & 1 \\ \vdots & \vdots & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots \\ \cos\phi_N\sin\theta_K & \sin\phi_N\sin\theta_K & \cos\theta_K & 1 \\ \end{bmatrix}^T \begin{bmatrix} \varphi_1(\theta_1, \phi_1)\lambda \\ \varphi_1(\theta_1, \phi_1)\lambda \\ \vdots \\ \varphi_N(\theta_1, \phi_1)\lambda \\ \varphi_N+1(\theta_1, \phi_2)\lambda \\ \vdots \\ \varphi_N-K(\theta_N, \phi_K)\lambda \\ \end{bmatrix} \quad (4.49)
\]

where \(\phi_i = 0: \frac{2\pi}{N}: 2\pi\), \(\theta_i = 0: \frac{\pi}{2K}: \frac{\pi}{2}\) (upper hemisphere)

One can add more columns on the right term to represent measurements made at several frequencies and correspondingly we will get a mean PCO solution for all listed frequencies in one step using (4.49). Once the mean PCO is determined, we can multiply back by the spherical phase expansion matrix to obtain an ideal spherical phase pattern anchored at the mean PCO determined from the real measured phase pattern. The
difference residual between the ideal and the measured phase pattern will be simply phase center variation (PCV).

4.8.3 PCO Visual Demonstration

![Ideal Phase Pattern at PCO(0,0,0)](image1)

![Ideal Phase pattern at PCO(0,0,1mm)](image2)

Fig. 40  Ideal Phase Pattern shifted along Z-axis from origin (0,0,0) to (0,0,1)

An imaginary perfect antenna with ideal phase pattern of zero degrees and located in the antenna origin (0,0,0) can be represented as a flat surface (see Fig. 40 left image). The same ideal pattern “shifted” to a new location (z=1) will take a new shape as shown on the right image). Transformations along other axes will translate the pattern as shown in the Fig. 41
Ideal Phase Pattern shifted along X-axis from origin (0,0,0) to (1,0,0)

Ideal Phase Pattern shifted along Y-axis from origin (0,0,0) to (0,1,0)

Ideal Phase Pattern shifted along X,Y,Z-axis from origin (0,0,0) to (1,1,2)

Measured phase pattern of AR25 antenna (note that this phase pattern shape is heavily weighted by large Z-axis offset of yet to be determined PCO value)

Fig. 41 Translation of Non-Ideal Phase Pattern along Z-axis

Fig. 41 shows a PCO translation of a real commercial AR25 GNSS (Leica product). This is a tall antenna, hence it has a large PCO offset in vertical dimension (along the z –axis) as seen on the right image. The accuracy of the PCO method was compared to a Spherical Harmonics method used by Geo++ Company when performing
phase center calibration on the same antenna. An agreement to within of 0.2mm was achieved for all frequencies, which was within measurement bias errors of both methods [108].

4.8.4 PCO Computation Example

Compute mean PCO(x,y,z) using Least Square Method

Determine ideal spherical phase pattern at determined PCO location

Subtract two patterns from each other to obtain phase center variation residual pattern

Fig. 42 Graphical representation of phase center computation
Fig. 42 represents graphically how the PCO is computed. The mean single point PCO is computed from the phase error residuals using the least square method.

4.9 Summary

Novel methods of creating deep (30-50 dB), angularly narrow nulls in the otherwise uniform hemi-spherical antenna radiation pattern have been introduced in this chapter. The methods to create nulls can be classified as follows:

- Excitation of k spiral arms (where k <N)
- Non-integer mode excitation for all arms or subset thereof
- Combination of non-integer mode excitation
- Reactive Loading of Spiral Arms
- Various inter-connection between spiral arms (i.e. cross-over spiral antenna)
- Various combinations thereof (listed above)

Detailed design parameters of three spiral antennas are also given. An analysis of null creation is presented in Appendix A for the digital spiral array antenna and in this chapter for the continuous case. The derivation of antenna phase center computation is also presented. The material covered in this section is important from the point of view of null steering which is required for future work in order to have a fully functional system.
Chapter Five: Simulations

The main purpose of the simulations was to determine what spiral antenna configurations would yield an circularly or elliptically polarized radiation pattern that includes at least one null and is relatively omnidirectional outside the null area. The null area should be minimized in order not to compromise reception from tens of satellites located randomly throughout the upper hemisphere of the antenna radiation pattern. The source of interferer can be assumed to be a point-source.

The first step was to determine the minimum number of spiral arms than must be excited to provide at least elliptically polarized radiation pattern and minimum number of spiral arms that must be terminated with reactive loads in order to get any control of the null position, its depth, etc. Various simulations were carried for 4 –arm spiral antenna with single, dual and triple arms excitation. Other topologies identified in Chapter 4 were simulated next (i.e. Cross-Spiral and Dual Polarized spiral).

FEKO electromagnetic commercial software was used to simulate various spiral-antenna topologies. The main design focus was on a planar four-arm Archimedean type spiral topology with various tight winding (i.e. spacing of 0.5-1.0 mm and trace width of 0.5-1.0mm and 15-30 turns).

5.1 A 4-arm spiral antenna with partial excitation

5.1.1 A 4- arm spiral with single arm excitation

One arm of a 4-arm spiral antenna is excited with remaining three arms left as parasitic elements. The arms are excited sequentially and plotted in azimuth plane (Fig.43) or vertical plane (Fig.44). The natural handedness of the spiral antenna allows
generating elliptically polarized radiation pattern with partial excitation of each spiral arms.

Fig. 43 Conical (Azimuth) cut of single arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz), at $\Theta=0^\circ$

Fig. 44 Vertical cut of single arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz)

As shown above in Fig. 43 and Fig. 44, exciting a single arm of the spiral antenna allows reconfiguration of the radiation pattern of the antenna. The resolution and coverage will depend on the specific antenna design parameters (i.e. number of spiral arms, number of excited arms, and number of turns for each spiral arm). Note that the beam in the vertical plane also is not symmetrical and rotates with each given excited arm around its normal axis.
### 5.1.2 A 4-arm spiral with dual arm excitation

![Conical (Azimuth) cut of dual arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz) at Θ=0°](image1)

**Fig. 45** Conical (Azimuth) cut of dual arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz) at Θ=0°

![Vertical cut of dual arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz)](image2)

**Fig. 46** Vertical cut of dual arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz)

Similarly to the previous case of single spiral arm excitation, dual arm excitation of a 4-arm spiral antenna (see Fig. 45 and Fig. 46) provides ability to reconfigure the antenna radiation pattern. The pattern is not omni-directional, hence it can provide a natural null in the radiation pattern that can be steered towards the interfering signal.
5.1.3 A 4-arm spiral with triple arm excitation

Fig.47 Conical cut of triple arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz) at θ=0°

Fig.48 Vertical cut of triple arm excitation of 4-arm spiral antenna (1, 2 and 3 GHz)

Triple arm excitation of a 4-arm spiral antenna (see Fig.47 and Fig.48) provides also an ability to reconfigure the antenna radiation pattern like was in the case of single and dual arm excitation. If only one arm is excited at the time, then the beam/null resolution in the phi conical plane is 90° for the 4-arm spiral antenna. This is the same case if any set of 2 or 3 arms of a 4-arm spiral antenna is excited, however the relative beam/null position of a single excited spiral arm versus a set of two or three will not be 90° but a much smaller value somewhere between 10°-30°. This beam resolution will be enhanced if more spiral arms are available in the antenna.
Partial excitation (not all spiral arms are excited) of any spiral antenna provides a very simple mechanism for creating nulls in an antenna radiation pattern. Due to rotational handedness of spiral arms, the radiation pattern will preserve elliptic and/or circular polarization of the radiation pattern. The null presence in the vertical plane is shown in Fig.49 and Fig.50. Partial spiral antenna excitations still maintain a relatively good co-pol to cross-pol level.
5.2 A 4-Arm Spiral Antenna Excitation Method

Spiral antennas can be excited either from inside or from outside. Most of the known antenna designs and published work has concentrated on the inner feeding mechanism, however in the case of multiport excitation and beam control the outside feeding mechanism is much easier to implement and control for undesired effects (i.e. feed coupling, etc.). As seen from Fig.51 and Fig.52 a 4-arm Archimedean spiral antenna can be either excited using inner spiral arm ends or outer spiral arm ends. This is an important distinction and finding, since we plan to place excitation and reactive loading circuitry on the outside perimeter of the antenna. This arrangement will be a necessity as the number of spiral arms can be now equal or larger than four. The classical approach was focused mainly on 2-4 arm spirals due to limited available space in the interior of the spiral arms. Additional benefit of outside feeding mechanism is possibility to extend the upper bandwidth of the antenna by filling the inner space (previously reserved for inner feeding circuit) with additional antenna winding.

Fig.51 Vertical co-pol (LHCP) cut of 4-arm spiral antenna fed from inside at 2 GHz using non-integer modes 1,32 to 16,32. Left image – terminated (100 Ohms) outside ends, center image – open outside ends, right image – shorted outside ends.
Fig. 52 Vertical co-pol (RHCP) cut of 4-arm spiral antenna fed from outside at 2 GHz using non-integer modes 1,32 to 16,32. Left image – terminated (100 Ohms), inside ends, center image – open inside ends, right image – shorted inside ends.

5.3 Cross–Spiral Antenna

Let us consider an example of antenna type 1 (see Fig. 53 and Fig. 54), described previously in Chapter 4. In the first simulation, the first two adjacent arms are excited with a 90° phase gradient (to create a Mode 1 type RHCP pattern) while the third outside arm is terminated with a 32pF capacitive load and the fourth outside arm is terminated with various inductive loads, see Fig. 55. In the second simulation the third outside arm is now terminated in a 32 nH load while the fourth outside arm is terminated sequentially with capacitive loads of 1pF, 4pF, 16pF and 32 pF, see Fig. 56. The reactive load pair of 32 pF and 32 nH creates the deepest null. This combination introduces a 196° phase shift in the reflected wave which is the closest to the ideal case of 180° phase shift required to reverse the sign of the phase gradient of the reflected wave.
Fig. 53 Cross-over arrangement of inner spiral arms

![Cross-over arrangement of inner spiral arms](image)

Fig. 54 Network schematic of Cross-Spiral Antenna

Fig. 55 and Fig. 56 represent the conical cuts at elevation angle of theta=90° (antenna horizon). The ability to create a null on the antenna horizon is very important in order to mitigate terrestrial based interfering signals. Deep nulls in order of 30-40 dB are present in the computed radiation pattern.
Fig. 55 Co-pol (RHCP) conical cut (theta=90°) at 1 GHz of a four arm spiral antenna type 1, excited in partial Mode 1 (two adjacent arms) with reactive loading of 32pF (3\textsuperscript{rd} arm) and 4nH, 8 nH, 16nH, 32nH (4\textsuperscript{th} arm)

Fig. 56 Co-pol (RHCP) conical cut (theta=90°) at 1 GHz of a four arm spiral antenna type 1, excited in partial Mode 1 (two adjacent arms) with reactive loading of 32 nH (3\textsuperscript{rd} arm) and 1pF, 4pF, 16pF, 32 pF (4\textsuperscript{th} arm)
The first simulated case was with mixed capacitive and inductive loads that terminated two spiral arms. In the next case, as shown in Fig. 57 and Fig. 58, both arms are terminated with inductive loads only. The radiation pattern exhibits the same characteristics with small variations of angular null position in the conical cut plane.

![Co-pol (RHCP) conical cut (theta=90°) at 1 GHz of a four arm spiral antenna type 1, excited in partial Mode 1(two adjacent arms) with reactive loading of 32 nH (1st arm – left side, 2nd arm – right side) and 4nH, 8pF, 16nH, 32 nH (1st or 2nd arm)](image)

The corresponding vertical cut is shown in Fig. 58.
Fig. 58 Co-pol (RHCP) vertical cut at 1 GHz of a four arm spiral antenna type 1, excited in partial Mode 1 (two adjacent arms) with reactive loading of 1 pF (3rd arm) and 4nH, 8 nH, 16nH, 32nH (4th arm).

Varying the inductance load allows the radiation pattern null to be steered in the vertical plane between theta angles of 70°-90° while maintaining a relatively stable azimuth (phi) angle. There is another null located at a higher elevation (theta) angle as shown in Fig. 59 whose 3D location is frequency dependent due to the frequency scaling nature of the spiral antenna and frequency phase scaling of the passive reactive type load. Similar observations can be made about the angular null position in the planar plane as seen in Fig. 59 and Fig. 60.
Fig. 59 Co-pol (RHCP) vertical cut of a four cross-arm spiral antenna (left side - phi=70° at 1 GHz, middle plot - phi=38° at 2 GHz, right plot phi=86° at 3 GHz

The null position (theta, phi) moves from (10°, 70°) at 1 GHz, to (15°, 38°) at 2 GHz and (30°, 86°) at 3 GHz. The null depth is in the order of 25-35 dB and occupies a relatively small conical angle, hence significantly attenuating the interfering signal while maintaining good signal reception everywhere else.

Fig. 60 Co-pol (RHCP) vertical cut of a four cross-arm spiral antenna (left side - phi=-24° at 1 GHz, middle plot - phi=80° at 2 GHz, right plot phi=-54° at 3 GHz
5.4 Dual polarized antenna

As a second example let us consider a dual polarized four arm spiral antenna (shown in Fig. 61). All four arms are excited from outside, using a non-integer mode (M=32) of phase gradient between adjacent spiral arms. The first 16 non-integer modes (1,32 to 16,32) are used to generate various radiation patterns for a given polarization (LHCP in this case). Higher order modes will generate similar patterns but with flipped polarization (RHCP).

![Fig. 61 Top View of Dual polarized Spiral antenna](image)

The simulated normalized radiation patterns at 1 GHz and 2 GHz are shown in Fig. 62 and Fig. 63. At the lower frequency, the beams and nulls can be steered through the entire upper hemisphere of the antenna. At higher frequencies the ability to steer the beam/null in a vertical plane becomes constrained as shown in Fig. 63. The addition of reactive loads allows one to overcome this constraint as shown in later sections.
Fig. 62 Co-pol (LHCP) vertical radiation pattern cut at 1 GHz of a dual polarized four arm spiral antenna, excited in partial Modes (1,32) to (16,32).

Fig. 63 Co-pol (LHCP) vertical radiation pattern cut at 2 GHz of dual polarized four arm spiral antenna, excited in partial Modes (1,32) to (16,32).
Fig. 64 Co-pol (LHCP/RHCP) vertical radiation pattern cut at 4 GHz of dual polarized four arm spiral antenna, excited in partial Modes (1,8) to (7,8).

Fig. 64 demonstrates the ability to change antenna polarization while maintaining the same vertical position (Theta=±25°) of the null in the radiation pattern. The efficiency (peak antenna gain) of Mode 7, 8 is lower than Mode 1, 8 due to the fact that the handedness of Mode 7, 8 is opposite to the physical handedness of the spiral antenna itself.

5.5 Summary

FEKO simulations confirmed the feasibility of the proposed methods in creating deep nulls in a circularly polarized antenna patterns. The null levels are in the order of 30-50 dB, which can virtually eliminate or greatly reduce the impact of RF interference on the tracking ability of a satellite receiver.

The null position can cover the entire hemi-spherical coverage for a given set of excitation phase gradients and reactive load values. Controlling the values of reactive
loading allows for “fine” steering of null placement in the radiation pattern. A phase gradient change or a change of spiral arms excitation allows for “coarse” steering of null placement in the radiation pattern. Increasing the number of spiral arms should improve the resolution at which the null position can be stepped between given sets of parameters used for null control.
Chapter Six: Practical Implementation & Measurements

6.1 Overview

All antenna pattern measurements were performed at the NovAtel Antenna Group's test facility. This facility maintains a calibration standard of the anechoic chamber to ensure that no significant reflections are present and that ambient temperature is maintained within one degree Celsius in order to prevent phase variations within RF cabling. Each antenna was mounted using a 16 inch long, hollow plastic cylinder that provided a mechanical interface between the phi rotation stage and the bottom of the antenna. The plastic tube was covered with an absorbing material along its length and its circumference. A large side cut-out in the plastic cylinder allowed for access to RF cables, and reactive ports terminations. The impedance measurements were performed with the AUT positioner pointed toward the farthest part of the chamber to further reduce any unwanted reflections and maintain repeatability between measurements.

The co-ordinate system of the anechoic chamber is depicted in Fig. 65 below. Antenna boresight is oriented towards an elevation angle at $\theta = 0^\circ$ and the antenna horizon at $\theta = 90^\circ$. Each axis rotation follows a left-hand rotation convention. The left-hand rotation convention reverses a sign convention compared to a more widely used right-hand rotation convention. This convention is specific to the NovAtel chamber and NSI (Near Field System) control and processing software. The proper angle orientation correction is applied during near-field to far-field transformation.
6.2 Equiangular 4-arm Spiral

6.2.1 Antenna Construction

The spiral antenna has maximum radius of 60.5 mm, see Fig. 67, and is backed by a shallow metal ground plane (reflector) at the distance of only 15mm (λ/12 at lower end of the band to λ/7 at upper end of the band). RF absorber (C-RAM MT-30, ¼ thick from Cuming Microwave) is used to form a slightly lossy cavity and is placed on the perimeter of the ground plane filling the space between antenna and ground plane. See cross-section in Fig. 66.
A phasing network (0°-90°) and reactive loads are located underneath the ground plane. These circuits are connected to the spiral antenna using vertical pins as shown in Fig. 66. Black pins indicate the connection to the phasing circuit (for the outside fed spiral), while green pins indicate connection of outside spiral arms to reactive loads and red pins indicate connection of inside spiral arms to reactive loads. This unit was assembled without any feed circuit, and all spiral arm ends were connectorized using SMA connectors. The coax feeds were attached to the back side of the cavity ground plane and then the feed becomes a single conductor connecting with spirals arm ends. This unit was used to measure S-parameters values of the 8-port network that the 4-arm spiral antenna represents. In addition various anechoic chamber measurements were performed for various configuration of the 8-port antenna network.
SMA connectors were connected with the ends of each of the spiral arms using vertical metal pins as shown in Fig. 68. The feed pins have a moderate amount of inductance, but the amount was not determined. SMA ports connected to the outside arm ends were labelled P1, P2, P3 and P4 while the inside arm ends were labelled P5, P6, P7 and P8. Points P1 and P5 represented the ends of the same spiral arm. Similarly pairs P2-P6, P3-P7 and P4-P8 represent other three spiral arms. Note that the ‘J’ designator has also been used (in early measurements) to denote antenna ports and is equivalent to ‘P’ designators throughout this thesis.

Fig. 67 Four-Arm Equiangular Spiral Antenna
The above described feeding arrangement (see Fig. 68 and Fig. 69) allows the antenna to be either right-hand or left-hand circularly polarized. The number of reactive terminations will give us the necessary means to create narrow nulls in the main lobe of the radiation pattern.
6.2.2 Radiation Pattern measurements

Fig. 70 Measured radiation pattern and PCO of 4-arm Equiangular spiral antenna (left image – C=4.7 pF, right image – C=10.0 pF) with schematic shown above.
The co-pol radiation pattern and the PCO for the circuit are shown in Fig. 70. The circuit diagram represents the equiangular spiral antenna fed from the inside using a 90° phase gradient between P6 and P7 to excite the LHCP mode of radiation. Two sets of equal capacitive loads are connected to the inside ports P5 and P8 using another 90° 3-dB hybrid coupler. The outside ports are left open-circuited. The PCO is computed for a radiation pattern bounded by the cone angle Θ whose angle is incrementally changed (starting with θ = 90°, then θ = 87° and so on until the single point spatial point on antenna boresight is reached). The mean PCO is shown as a circle for each given Θ angle range (azimuth angle of φ = 2π is used for each PCO calculation). The PCO offset from the geometrical center of the antenna varies between 5-30 mm, which is a small price to pay for the ability to suppress a strong RF interference (30-50 dB levels of suppression). See Fig 71.
Fig 71 Equiangular Spiral 4-arm antenna radiation patterns with corresponding network connections (image above)

The image on the right (Fig 71) was measured at C1 (Compass satellite channel of 1560 MHz), while L1 is the GPS frequency of 1575.42 MHz (image on the left side). The radiation pattern shown on the left image is LHCP (as opposed to RHCP shown on the right image) due to feeding the spiral structure from the inside as oppose to the outside.
(right image). Exactly the same RHCP pattern can be obtained if a reversed antenna handedness and phase gradient is applied to the same internal spiral antenna ports.

![Diagram](image1.png)

![Diagram](image2.png)

Fig 72 Cross-Spiral 4-arm antenna radiation patterns with corresponding network connections (image above)

Fig 72 (both sides) represents measurements made at GPS L1 frequency of 1575 MHz..

Note the deep nulls that were created in the radiation pattern from theta between -20° to -
80° in the azimuth (phi) sector of -40° to -60°. The presence of the null on the (left image) causes the PCO to shift 30 mm away from the geometrical center of the antenna. Some degradation in phase center offset is measured and calculated. This is a relatively small trade-off for having the ability to create sharp nulls.

6.2.3 S-parameter Measurements

The antenna 8x8 S-parameter matrix was measured across the frequencies of interest (1100-1800 MHz). One unit was modified so that SMA connectors mounted directly to the internal pins (on the back side of the antenna cavity). There were eight ports attached to 4 spiral arms ends (inside and outside ends). Various Matlab simulations were performed to optimize the best antenna configuration from the network point of view using T-parameter cascaded system networks. This approach allows the attachment of various components to the antenna like feed circuits, networks, various loads, hybrids, etc.

Fig. 73 displays the input reflection coefficient of the outside spiral arm ends (left image) and the inside spiral arm ends (right image). Note that the average return loss of the inside arm ends is better (lower value) than outside arm ends. This is due to the fact that outside spiral arm traces are wider than inside spiral arm traces and therefore have larger impedance discontinuity with 50 Ohm measuring equipment.
The Archimedean type antenna maintains the same trace width, hence it represents a lower impedance mismatch at the outside spiral arm ends than those of the equiangular type spiral antenna. Hence, an Archimedean type antenna design was subsequently used in this research.

Using S and T parameter transformations we can compute the return loss of the antenna when two outside ports are connected to open and short circuits through the 3 dB hybrid coupler, see Fig. 74. It is worth noticing that the return loss is almost identical as shown in the previous figure, indicating there is a good isolation between the adjacent arms despite their close proximity.
Fig. 74  Measured $S_{11}$ of the input excitation port

The return loss curves shown in Fig. 73 and Fig. 74 are not constant over the frequency range as suggested by the theory. The periodic nature of these graphs indicates that impedance does change with frequency, due to the effect of the shallow cavity placed underneath the antenna. The depth of the cavity is from $\lambda/12$ to $\lambda/17$ over the bandwidth of the spiral antenna.

Mutual coupling is very important in our case, since we want to use coupled spiral arms as means of changing the radiation pattern of excited arms.
Fig. 75 Cross-coupling of adjacent ports of the Equiangular 4-arm Spiral Antenna

Again, very good repeatability is shown in both figures (shown in Fig. 75). It is worth noticing that some frequencies have higher mutual coupling than others and therefore, these frequencies would allow for easier control of antenna radiation through means of mutual coupling. In particular, use of inside ports seem to be well suited for this purpose (i.e. placement of reactive loads), while outside ports can be used for excitation of the antenna.

S-parameter measurements for a given arm between the inside and outside port will yield a sum of radiation loss and insertion loss, see Fig. 76. Again, very good repeatability is shown between all four arms.

The Fig. 76 indicates that there is not much (5%) residual power left at the output of each end of a given arm. Most energy is radiated, dissipated as heat and/or reflected from each port (due to 50-180 ohm interface mismatch).
Fig. 76 Combined Insertion and Radiation Loss of the Equiangular 4-arm antenna

6.2.4 Antenna Impedance

The theoretical antenna impedance is in the order of 180 Ohms using Babinet’s principle. The S-parameter measurements were performed using 50 Ohm measurement system (including loads on non-measured ports), and therefore the reflection coefficient technically should be:

$$20 \log \left( \frac{Z_s - Z_L}{Z_s + Z_L} \right) = 20 \log \left( \frac{180 - 50}{180 + 50} \right) = -5 \text{ dB}$$ (6.1)

The 90° 3dB hybrid coupler provides good isolation between the feeding circuit and the antenna and its terminals (vertical pins). The input impedance is found from the N-port S-parameters using the following equation (I is an identity matrix and S is the S-parameter matrix of 4-arm spiral antenna)
The impedance is plotted in Fig. 77. The impedance does converge to 100 Ohms at higher frequencies, however at low frequencies it exhibits large value oscillations, likely caused by the shallow cavity and increased excitation of higher order modes. We would expect that increasing the number of arms should reduce how many higher order modes are excited and hence provide more uniform impedance with frequency.
6.2.5 Port Isolation

![Graph showing Port Isolation](image)

Fig. 78 Four arm spiral antenna fed from the outside (P1, P2)

The addition of a directional coupler to the excitation outside Ports 1 and 2 significantly improves isolation between various ports. This is due to the inherit properties of the 90°, 3dB coupler described in the previous chapter. The blue curve in Fig. 78 represents normal direct excitation while the red curve represents excitation of two arms using a direction coupler while the remaining two arms is either open or shorted. The black curve is a variation of the previous case where open and short conditions of non-excited spiral arms are replaced with reactive loading. It can be seen the Cross-Spiral antenna (labelled as X-config in Fig. 78) provides much better isolation that the classical spiral antenna (original configuration). The amount of reflections generated within Cross-Spiral antenna is minimized since the response is very similar.
with or without the presence of 3dB 90° hybrids (used to isolate the reflections in classical antenna case).

6.3 Archimedean crossed 4-arm Spiral (spiral growth dr=4mm)

A 4 arm X-spiral Archimedean antenna with crossed interconnection of inside arms is shown in Fig. 79. The mechanical configuration of the cavity and the pins (located within the cavity) is the same as for the equiangular spiral antenna case described in the previous section. For detailed design parameters refer to Section 4.4.4.1. Two opposing inner arms are connected on top of the PCB while the other pair is connected together on the opposite side of the antenna PCB.

![Image of Archimedean X-Spiral 4-arm antenna (dr=4mm)](image)

Fig. 79 Archimedean X-Spiral 4-arm antenna (dr=4mm)

This design was based on the results of extensive optimization done on an 8-port network formed by the 4 arm spiral antenna. Since the inner arms are connected, the antenna now represents a 4-port network. Two ports are used for antenna excitation while the other two ports are used to reactively load the structure with discrete capacitors or
inductors. Designators J and/or P are often interchanged when used to describe antenna ports in various simulation and measurement plots. The 3 dB 90° hybrid coupler (attached to ports 1 and 2) is required to generate the 90° phase gradient necessary to excite a circularly polarized partial mode 1 in the spiral antenna.

Fig 80 Archimedean Spiral 4-arm antenna radiation patterns with corresponding antenna network configuration (top image)
Fig. 81  Cross-Spiral antenna (dr=4mm) Measured Radiation Pattern (co-pol) at frequency of 1675 MHz with various reactive loads

<table>
<thead>
<tr>
<th>Case</th>
<th>J3</th>
<th>J4</th>
<th>Null Location &amp; Depth</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>2.2 pF</td>
<td>5.6 nH</td>
<td>Depth: -33 dB, Theta: -27°, Phi: -24°</td>
</tr>
<tr>
<td>B</td>
<td>3.3 pF</td>
<td>5.6 nH</td>
<td>Depth: -35 dB, Theta: -24°, Phi: -24°</td>
</tr>
<tr>
<td>C*</td>
<td>open</td>
<td>5.6 nH</td>
<td>Depth: -39 dB, Theta: -54°, Phi: +24°</td>
</tr>
<tr>
<td>D*</td>
<td>open</td>
<td>4.7 nH</td>
<td>Depth: -37 dB, Theta: -84°, Phi: +30°</td>
</tr>
<tr>
<td>E</td>
<td>1.0 pF</td>
<td>10.0 nH</td>
<td>Depth: -35 dB, Theta: -81°, Phi: -18°</td>
</tr>
<tr>
<td>F</td>
<td>2.2 pF</td>
<td>10.0 nH</td>
<td>Depth: -35 dB, Theta: -78°, Phi: -21°</td>
</tr>
<tr>
<td>G*</td>
<td>4.7 nH</td>
<td>4.7 pF</td>
<td>Depth: -44 dB, Theta: -75°, Phi: -3°</td>
</tr>
<tr>
<td>H</td>
<td>short</td>
<td>open</td>
<td>Depth: -40 dB, Theta: -78°, Phi: -12°</td>
</tr>
</tbody>
</table>

Table 5 Null location summary from Fig 81

Cases denoted with asterisk (*) indicate where excitation ports and reactive loads were moved by 7mm closer to the vertical pins that excite the spiral arms, see Table 5. A progression of null positions in the radiation pattern is shown in Fig. 81 for various discrete values of reactive loads attached to ports 3 and 4. The antenna configuration from Fig 80 was used to perform these measurements. The shift of current phase distribution (approx. 16°) at each arm translated to a position shift of the null in theta.
plane from -24° to -54°/84°. For some cases the null position is not sensitive to change in reactance load (i.e. case A versus case B) while for other cases a significant change in null position is observed.

Note that null steering in azimuth (phi) plane and elevation (theta) plane is possible for fixed phase gradient excitation (90° between ports 1 and 2 in this case). A similar measurement for a lower frequency of 1125 MHz is shown in Fig. 82.
Fig. 82 Cross-Spiral antenna (dr=4mm) Measured Radiation Pattern (co-pol) at frequency of 1125 MHz with various reactive loads
### Table 6 Null location summary from Fig 82

<table>
<thead>
<tr>
<th>Case</th>
<th>J3</th>
<th>J4</th>
<th>Null Location &amp; Depth</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>Depth</td>
</tr>
<tr>
<td>A</td>
<td>5.6 pF</td>
<td>5.6 nH</td>
<td>-38 dB</td>
</tr>
<tr>
<td>B</td>
<td>1.0 pF</td>
<td>10.0 nH</td>
<td>-41 dB</td>
</tr>
<tr>
<td>C*</td>
<td>open</td>
<td>2.7 pF</td>
<td>-36 dB</td>
</tr>
<tr>
<td>D*</td>
<td>open</td>
<td>4.7 nH</td>
<td>-37 dB</td>
</tr>
<tr>
<td>E</td>
<td>open</td>
<td>10.0 nH</td>
<td>-36 dB</td>
</tr>
<tr>
<td>F*</td>
<td>3.9 pF</td>
<td>3.9 pF</td>
<td>-38 dB</td>
</tr>
<tr>
<td>G</td>
<td>2.2 pF</td>
<td>5.6 nH</td>
<td>-34 dB</td>
</tr>
<tr>
<td>H*</td>
<td>open</td>
<td>0.5 pF</td>
<td>-36 dB</td>
</tr>
<tr>
<td>I*</td>
<td>open</td>
<td>1.2 nH</td>
<td>-39 dB</td>
</tr>
<tr>
<td>J</td>
<td>short</td>
<td>open</td>
<td>-41 dB</td>
</tr>
</tbody>
</table>

As above cases denoted with asterisk (*) indicate where excitation ports and reactive loads were moved by 7mm closer to the vertical pins that excite the spiral arms, See Table 6. The shift of current phase distribution (approx. 16°) at each arm translated to a position shift of the null in theta plane from 90° to 84°, see Fig. 82. For some cases the null position is not sensitive to change in reactance load (i.e. case A versus case G or case C* versus case H*) while for other cases a change in null position is observed.

The null steering is much more difficult near the low end of the bandwidth. However decent nulls are created at low elevation angles, which is still helpful since most RF jamming signals originate on the antenna horizon.
6.3.1 Archimedean Antenna Impedance

As mentioned in the previous section, the theoretical antenna impedance is in the order of 180 Ohms. The 90° 3dB hybrid coupler provides good isolation between the feeding circuit and the antenna and its terminals (vertical pins). The input impedance is found from the N-port S-parameters using Eq. 6.2 and is plotted in Fig. 83.

Fig. 83 Impedance of crossed Archimedean 4- arm spiral antenna fed from outside

Fig. 84 Impedance of crossed Archimedean 4- arm spiral antenna fed from outside (left image – a wider frequency range than Fig. 83, right image – antenna impedance dependence on reactive load impedance)
The impedance plotted in Fig. 83 and Fig. 84 (left image) is shown for two types of 4-arm Archimedean spiral antennas. One antenna has a much tighter winding of spiral arms (growth rate per one full revolution $dr=4\text{mm}$) than the other spiral antenna ($dr=8\text{mm}$). Each antenna occupies the same aperture area, backed by the same metal cavity with an RF absorber placed at the outside of the cavity perimeter. It can be seen that the antenna with the tighter spiral arm winding has lower impedance variation than the other antenna.

The addition of a 3dB 90° hybrid allows one to isolate the antenna from the measurement equipment or the antenna feed circuit as shown in Fig. 84 (right image). The case with no hybrid (red curve) has an impedance that gets lower in values with frequency while in the case of hybrid present, the impedance oscillate around a constant value of 100 Ohms.

**6.4 Axial Ratio**

To provide an antenna that exhibits good performance with respect to given circular polarization handedness, the axial ratio is critical. The higher the axial ratio, the more elliptical the polarization, and the lower the gain with respect to the desired circular polarization mode. This parameter is the result of design and process control.

An introduction of a null in the radiation pattern can cause severe degradation of Axial Ratio in the overall radiation pattern (not just in the close proximity to the null area). Fig. 85 demonstrates that Cross-spiral antenna excited from the outside perimeter of the antenna has a superior Axial Ratio performance when compared to the classical spiral antenna excited from the inside perimeter of the antenna.
Fig. 85 Co-pol, cross-pol and Axial Ratio of 4-arm spiral antenna excited from inside (left column) and 4-arm Cross spiral antenna excited from outside (right column). All patterns computed at 1675 MHz
Top row in Fig. 85 and Fig. 86 represents co-pol (RHCP) pattern, the middle row represents cross-pol (LHCP) pattern while the bottom row represents the corresponding Axial Ratio pattern. Comparing the left column in Fig. 85 with the right column we can observe that there is a significantly lower level of cross-pol and Axial Ratio in the right column (Cross-spiral antenna case). The cross-pol radiation pattern in the case of classical spiral antenna excited from inside is relatively high due to reflections from the outside spiral arm ends and it does not change with much with the presence of the null. In case of Cross-spiral antenna, the cross-polarization level is much lower (due to the natured of novel antenna approach) and introduction of null in the main co-pol pattern has a corresponding lower cross-pol level in the same region. Further analysis of Fig. 85 shows that classical spiral antenna exhibit a polarization handedness flips in some areas surrounding the null location. This is not case with Cross-spiral antenna where the polarization handedness is maintained throughout the entire upper radiation hemisphere.

Fig. 86 represents Axial Ratio of Cross-spiral antenna under two different reactive load cases. Ports J1, J2 are fed with 90° phase difference while J3 is open ended and J4 is terminated with 5.6 nH (left column) and 3.3 pF (right column). We can observe that Axial Ratio is similar for both cases, regardless of the null position in the radiation pattern.
Fig. 86 Co-pol, cross-pol and Axial Ratio of 4-arm Cross-spiral antenna excited from outside with J3 open ended and J4=5.6 nH (left column) and 3.3 pF(right column). J3All patterns computed at 1675 MHz
6.5 Summary

Measurements conducted in the anechoic chamber confirmed the feasibility of the proposed methods in creating deep nulls in circularly polarized antenna patterns. The null levels are in the order of 30-50 dB, which agrees with FEKO simulations.

Various Matlab simulations were performed to optimize various antenna configurations and corresponding reactive termination values from the network point of view using T- parameter cascaded system networks. It was demonstrated that attaching various components to the antenna spiral arms is feasible without destroying its main characteristics such as bandwidth, peak gain and polarization purity.

Some degradation in phase center offset was measured when nulls are present in the antenna radiation pattern. The small phase center offset (cm level) and reduced peak gain (3-4 dB) is a small trade-off for having an ability to create sharp nulls.

The addition of a directional coupler significantly improves isolation between various ports. This is due to inherit properties of the 90° 3dB coupler described in the previous Chapter.

Cross-spiral antenna shows promising results of low Axial Ratio in the presence of null in the radiation pattern. This novel spiral antenna concept solves an old problem of unwanted radiation from the currents that are normally reflected form the spiral arm ends. Existing methods used to suppress this radiation are not very successful and cause reduced gain and bandwidth of the antenna.
Chapter Seven: **Conclusions**

7.1 **Summary**

The concept of radiation pattern control through reactively controlled spiral antennas is introduced and explained in this dissertation. The primary contributions to the field, however, are the new reconfigurable antenna designs. These designs include the crossed spiral antenna and dual-polarized spiral antenna. Additionally, detailed analysis and explanation of how the antennas operate is presented through simulations and measurements (cross spiral antenna only).

Three methods have been identified for achieving reconfigurability of antenna designs and operation. These three methods are reactive loading of spiral arm ends, non-integer mode excitation, partial excitation and various combinations thereof. The reactive loading is implemented through a large array of switchable discrete reactive components (capacitors and inductors). Non-integer mode excitation requires a phase shift network to generate a required phase difference (phase gradient) across “active” spiral arms. In most cases this translates to a 90° 3-dB hybrid connected to two arms of a multiple (4-arm) spiral arm antenna. Partial excitation creates unsymmetrical patterns by causing only a subset of spiral arms to be excited when using integer mode excitation.

A thorough understanding of the operation and design of reconfigurable antennas is necessary for the development of new and innovative antenna designs and applications. The ability to control the radiation pattern at any frequency and the ability to create deep, spatially narrow nulls have many useful applications especially in systems that operate in
an acquire-and-track operation. The added ability to filter out interfering signals using the antenna pattern will aid in reducing the complexity and cost of RF processing subsystems. Generally, the ideas presented in this thesis will have beneficial implications for communication systems that operate in hostile signal environments. Military communication systems in particular can be subject to intentional jamming signals. Performing RF signal rejection at the antenna level can prevent damaging or interfering signals from ever reaching sensitive internal components.

7.2 Contributions

The following is the list of contributions of this thesis:

- The new concept for achieving narrow spatial width nulls in a circularly polarized antenna pattern was developed [103].
- Two new spiral antenna designs have been created capable of creating a sharp null in the otherwise omnidirectional upper-hemisphere circularly polarized pattern [104,105].
- The new concept of reactive loading spiral antenna arms to improve circular polarization handedness of antenna radiation pattern and null forming capability is found. [103-105].
- A Spherical Phase Method for antenna phase center determination was developed. This was presented and published in [11].
7.3 Future Work

The proper switching mechanism to adjust the value of reactive loads and provisions of proper phase gradients will be required to have a fully functional antenna. In addition a novel adaptive null steering with knowledge of antenna orientation with respect to its surroundings will also be required.

Controlled impedances can also be introduced within the antenna spiral structure, not just the spiral arm ends. This will bring another degree of freedom that may enhance the overall nulling performance of the spiral antenna.

More research is needed to determine how more nulls can be generated in the antenna pattern and what the fundamental limits are for a given antenna design approach. In particular an algorithm to find out the relationship between null depth and placement versus load reactances would be beneficial for practical implementation of this type of antenna.

A research to determine algorithm for blind search of an interfering signal or jammer will also be required. This would require a mechanism to lock on the jammer and maintain the lock regardless of antenna orientation and relative movement between jammer and antenna.
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APPENDIX A: DISCRETE SPIRAL ARRAY AF

This section provides an analysis for discrete spiral array antennas. This is an exercise to show how nulls can be created using discrete spiral antennas with sub-integer mode excitation. Similar performance of null creation can be expected in an analogue continuous multi-arm spiral antenna. The graph (see Fig. 87) describes the typical geometry of a two-arm spiral planar array. Each arm consists of N points (equally spaced).

![Fig. 87 Two-Arm Spiral Geometry](image)

Following similar derivation for a circular array [61] with small modifications we have

\[ \hat{a}_{\rho_n} = \hat{x} r(\phi_n) \cos(\phi_n) + \hat{y} r(\phi_n) \sin(\phi_n) \]  

(A1)
\[ \hat{a}_{\rho_n} = \hat{x} r(\varphi_n)\cos(\varphi_n + \pi) + \hat{y} r(\varphi_n)\sin(\varphi_n + \pi) \quad (A2) \]

\[ \hat{r} = \hat{x} \sin(\theta)\cos(\varphi) + \hat{y} \sin(\theta)\sin(\varphi) + \hat{z} \cos(\theta) \quad (A3) \]

where the first two equations describe the planar geometry of the 1st and the 2nd spiral arm. The radius to each point along the spiral will grow accordingly, depending on the spiral geometry

\[ r(\varphi_n) = a \varphi_n \quad (Archimedean) \quad (A4) \]

\[ r(\varphi_n) = e^{a \varphi_n} \quad (Equiangular) \]

The Far-Feld (\(R \gg \max(r(\varphi_n))\)) distance \(R\) is calculated as

\[ R_n = r - r(\varphi_n)(\hat{\alpha}_\rho \cdot \hat{r}) \cong r - r(\varphi_n) \cos(\psi_n) \quad (A5) \]

A projection of the 1st arm in the Far-Field (FF) will be given by

\[ \hat{a}_{\rho_n} \cdot \hat{r} = r(\varphi_n) \sin(\theta) [\cos(\phi) \cos(\varphi_n) + \sin(\phi) \sin(\varphi_n)] \]

\[ \hat{a}_{\rho_n} \cdot \hat{r} = r(\varphi_n) \sin(\theta) \cos(\phi - \varphi_n) \quad (A6a) \]

Similarly, the 2nd arm can be represented as

\[ \hat{a}_{\rho_n} \cdot \hat{r} = r(\varphi_n) \sin(\theta) \cos(\phi - \varphi_n + \pi) \]

\[ = -r(\varphi_n) \sin(\theta) \cos(\phi - \varphi_n) \quad (A6b) \]

The Electric field \((E)\) in the FF will be given as

\[ E(r, \theta, \phi) = \sum_{m=1}^{2} \sum_{n=1}^{N} \frac{e^{-jkr}}{r} A_n^m e^{jkr(\varphi_n)\sin(\theta)\cos(\phi - \varphi_n + \beta_n)} \quad (A7) \]

\[ A_1^1 = l_n e^{j\alpha_n} \quad \text{and} \quad A_2^2 = l_n e^{j(\alpha_n + \delta_m)} \quad (A8) \]
where \( \alpha_n \) is the phase progression along each spiral arm, \( \beta_m \) is the physical angular separation between spiral arms, and \( \delta_m \) is the phase excitation for a given arm. The mode of operation of the spiral array antenna is controlled by \( \delta_m \). \( I_n \) is the amplitude excitation of the \( n \)-th spiral element.

Removing the \( \frac{e^{-jkr}}{r} \) factor (its amplitude is constant) from (A7) we can derive now the total Array Factor (AF) for two arm spiral antenna case (\( \beta_1=0, \beta_2=\pi \)) as

\[
AF = \sum_{n=1}^{N} I_n e^{j\alpha_n} \left[ e^{jkr(\phi_n) \sin(\theta) \cos(\phi - \phi_n)} + e^{j\delta_2} e^{-jkr(\phi_n) \sin(\theta) \cos(\phi - \phi_n)} \right] 
\]  
(A9)

which can further be simplified, since \( \delta_2 = \pi \)

\[
AF = \sum_{n=1}^{N} I_n 2j \sin[kr(\phi_n) \sin(\theta) \cos(\phi - \phi_n)] e^{j\alpha_n} 
\]  
(A10)

**A.1. Maximum and Minimum Conditions**

The maximum radiation towards the FF point in \((\theta_0, \phi_0)\) direction is satisfied when

\[
\alpha_n + kr(\phi_n) \sin(\theta_0) \cos(\phi_0 - \phi_n - \beta_m) + \delta_m = 2p\pi, \\
p = 0, \pm 1, \pm 2, ... 
\]  
(A11)

while the minimum radiation towards the FF point in \((\theta_0, \phi_0)\) direction is satisfied when

\[
\alpha_n + kr(\phi_n) \sin(\theta_0) \cos(\phi_0 - \phi_n - \beta_m) + \delta_m = \frac{p\pi}{2}, \\
p = \pm 1 
\]  
(A12)

It can be shown that the AF with steering phase vector \( \alpha_n \) is given by

\[
AF = 2 \sum_{n=1}^{N} I_n \sin[kr(\phi_n) \cos(\phi_0)] e^{-jkr(\phi_n) \cos(\phi_0)} 
\]  
(A13)

\[
cos(\phi_{on}) = \sin(\theta_0) \cos(\phi_0 - \phi_n) 
\]
\[ \cos(\psi_n) = \sin(\theta) \cos(\varphi - \varphi_n) \]

**A.2. AF for M-arm spiral array**

By inspection, equation (A13) can be extended to compute the AF for an M-arm spiral array:

\[
AF = \sum_{m=1}^{M} \sum_{n=1}^{N} I_n e^{-j[kr(\varphi_n) \sin(\theta) \cos(\varphi_n - \beta_m) + \alpha_n + \delta_m]}
\]  
(A14)

\[ \alpha_n = dL_n \frac{2\pi}{\lambda_y} \quad ; \quad \beta_m = \frac{2\pi m}{M} \]

\[ \delta_m = \frac{2\pi m}{M} s \]

*where s – mode number,*

\[ dL_n – \text{distance along the spiral arm} \]

A given spiral arm is discretized using N isotropic radiating points. A large value of N will allow the solution to merge with the continuous spiral arm. The phase progression along each arm is taken into account.

**A.3. Null Patterns**

Spiral arrays are characterized by the efficient utilization of antenna real estate space. It can be shown that many interesting properties can be achieved when various antenna configurations are implemented. The main focus is placed on the dynamic pattern configuration for a given frequency of operation.

**A.3.1. Integer Modes of Excitation**

The radiation pattern of an M-arm spiral antenna will change with relative phase gradient applied to each arm. These are M order modes, which are associated with M-arm spiral
antenna.

\[
\delta_m = \frac{2\pi}{M} m \quad m = 0, \ldots, M - 1
\]  

(A15)

Useful modes of operation are from \(m=1\) to \(M-1\). Modes that impose a \(0/2\pi\) phase gradient across spiral arms create an unbalanced system, causing significant antenna performance degradation, and therefore are not used in practice. The first and last mode \((m=1\) to \(M\) with \(M>3\)) provide a hemispherical, single, wide beamwidth radiation pattern, while other modes provide a null on antenna boresight. Wide beamwidth radiation modes are used for omnidirectional coverage (i.e. satellite reception) where circular polarization is needed. The higher modes radiation patterns are used for target tracking and anti-jamming scenarios. Relatively complicated hardware is used to implement the tracking and nulling capabilities. In general, applications that use integer modes of excitation for spiral arms are widely published in the literature.

**A.3.2. Fractional (Non-Integer) Modes of Excitation**

This section addresses a novel method of generating anti-jamming capabilities by introducing non-integer modes of excitation and by changing the spiral configuration to create a deep null on the antenna horizon.

\[
\delta_m = \frac{2\pi}{M} m \quad m \in \text{Real (Non Integer)}
\]  

(A16)

These modes of excitation are equivalent to beam/null steering by providing a desired
phase gradient between spiral arms that meets the following criteria:

\[ \delta_m = dL_n \frac{2\pi}{\lambda_g} - k r (\varphi^m_n) \sin(\theta) \cos(\phi - \varphi_n - \beta_m) \]  \hspace{1cm} (A17)

The null can be steered off from the antenna boresight to other elevation angles due to a given non-integer mode of excitation. An example of mode-2 radiation pattern whose boresight null is steered off towards the horizon is shown in Fig. 88. The Mode-2 phase gradient \( \delta_m \) was multiplied by a factor of 0.99.

Fig. 88 3D amplitude pattern (top view projection) of 4-arm spiral array, \( \delta_m=0.99 \pi \)

The x and y axes in Fig. 88 are defined as:

\[ X = \sin(\theta) \cos(\varphi), \quad Y = \sin(\theta) \sin(\varphi) \]  \hspace{1cm} (A18)

Rotating the reference phase point among the spiral arms allows moving the null in the azimuth direction. This provides full control of the null that is created in the antenna pattern. Though the radiation pattern will not be perfectly circularly polarized, (it will be elliptically polarized) it will however provide ability to null an interfering signal without saturating the front end of the receiver.
APPENDIX B: S-PARAMETERS OF 4 PORT NETWORK

Generalized scattering parameters have been defined by K. Kurokawa [68]. The S-parameters describe the interrelationships between normalized complex voltage wave incident on and reflected from a given i-th port. They are defined in terms of terminal voltage $V_i$ and terminal current $I_i$ for any arbitrary reference impedance $Z_i$ as follows:

$$ a_i = \frac{V_i + Z_i I_i}{2\sqrt{|Re(Z_i)|}} \quad \text{and} \quad b_i = \frac{V_i - Z_i^* I_i}{2\sqrt{|Re(Z_i)|}} $$  \hspace{1cm} (B1)

Fig. 89 Flow Graph of 4-port S-parameter network with attached loads ($\Gamma$’s)

Flow graphs make S-parameter calculations much simpler. Each port is represented by two nodes. Node $a_n$ represents the wave coming into the device from another device at port n and node $b_n$ represents the wave leaving the device at port n. The
complex scattering parameters are represented as multipliers on branches connecting the nodes within the network. The analysis of signal analysis by means of signal flow graph is covered in [69,70]. A signal flow diagram of 4-port network is shown in Fig. 89.