Design and Evaluation of V/UHF Satellite Communication Antennas for Naval Applications

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Abstract

In this master thesis, compact antenna design aimed towards naval applications have been designed, analyzed and evaluated. There is a recent interest in the development of compact antennas to be used for smaller submarine models, and with a smaller hull on a submarine, communication and antenna systems must be adapted and minimized, which limits antenna design. With two limiting cylindrical volumes with maximum allowed dimensions $r = 10 \text{ cm}$, $h = 50 \text{ cm}$ and $r = 5 \text{ cm}$, $h = 90 \text{ cm}$, the antennas would operate on the upper to lower V/UHF band, radiate hemispherically and have a high RHCP purity. It was found that the most appropriate antenna structure for both volumes was QHA design. After the design and analysis process was completed, it was concluded that the shorter antenna design could meet all requirements set while the longer antenna design did not meet all requirements but could still establish a good communication link on the higher frequencies. Antenna prototypes based on the produced design were constructed and measured and, despite minor deviations, verified that the results obtained from this thesis were reliable.

Index terms — Antenna design, Antenna simulations, QHA, Submarine communication, SatCom, MUOS, Legacy
Sammanfattning

I denna mastersuppsatsen har kompaktta antenner riktade mot marina applikationer designats, analyserats och utvärderats. Det finns ett intresse för utvecklingen av kompaktta antenner som ska användas för mindre ubåtsmodeller och med ett mindre skrov på en ubåt måste kommunikations- och antennsystemen anpassas och minimeras därefter, vilket begränsar antenndesignen. Med två begränsande cylindriska volymer med maximalt tillåtna dimensioner $r = 10$ cm, $h = 50$ cm samt $r = 5$ cm, $h = 90$ cm, skulle antennerna verka på det övre till lägre V/UHF-bandet, stråla hemisfäriskt och ha en hög RHCP-renhet. Det konstaterades att den lämpligaste antennstrukturen för båda volymer var QHA-design. Efter att design- och analysprocessen slutförts drogs slutsatsen att den kortare antenndesignen kunde uppfylla alla krav som ställdes medan den längre antenndesignen inte uppfyllde alla krav men fortfarande kunde upprätta en bra kommunikationslänk på de högre frekvenserna. Antennprototyper baserade på de framtagna designerna konstruerades och mättes, trots mindre avvikelser, verifierade att de erhållna resultaten från denna avhandling var tillförlitliga.

Nyckelord—Antenndesign, Antennsimulering, QHA, Ubåtskommunikation, Satellitkommunikation, MUOS, Legacy
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Abbreviations

AMHA - Axial Mode Helix Antenna
AR - Axial Ratio
AUT - Antenna Under Test
BHA - Bifilar Helix Antenna
BW - Beam Width
CP - Circular Polarization
DAMA - Demand Assigned Multiple Access
DL - Downlink
HP - Horizontal Polarization
LGA - Low-Gain Antenna
LHCP - Left Hand Circular Polarization
LoS - Line of Sight
LP - Linear Polarization
MUOS - Mobile User Objective System
NMHA - Normal Mode Helix Antenna
QHA - Quadrifilar Helix Antenna
RAF - Radio Access Facilities
RCS - Radar Cross Section
RF - Radio Frequency
RHCP - Right Hand Circular Polarization
SatCom - Satellite Communication
SGA - Standard Gain Antenna
SP QHA - Self-Phased Quadrifilar Helix Antenna
TEM - Transverse Electromagnetic
UHF - Ultra High Frequency
UL - Uplink
VHF - Very High Frequency
VP - Vertical Polarization
VSWR - Voltage Standing Wave Ratio
Nomenclatures

Bold letters denotes vectors and bold letters with circumflex denotes unit
vectors. Hard brackets around a variable indicates a matrix. Indexing on
variables will occur in the report which will have the same dimensions as
the non-indexed units. The dimensions are not necessarily expressed in SI
base units.

$\alpha$ - Pitch angle [Degree]
$\Gamma$ - Reflection coefficient [dB]/[Dimensionless]
$\gamma$ - Depth of pattern minimum for a QHA [dB]
$\delta$ - Penetration depth [m]
$\varepsilon$ - Electric permittivity [F/m]
$\eta$ - Antenna efficiency factor [Dimensionless]
$\hat{\theta}, \theta$ - Altitude coordinate [Degree]
$\kappa$ - Helix radius to spacing distance ratio [Dimensionless]
$\lambda$ - Wavelength [m]
$\mu$ - Magnetic permittivity [H/m]
$\rho$ - Electric resistivity [\Omega\cdot m]
$\sigma$ - Transmission coefficient [dB]/[Dimensionless]
$\phi$ - Phase component [Degree]
$\hat{\varphi}, \varphi$ - Azimuth coordinate [Degree]
$\omega$ - Angular frequency [Hz/rad]
$\mathbf{A}, \mathbf{A}$ - Magnetic potential [V\cdot s/m]
$C$ - Capacitance [F]
$c$ - Speed of electromagnetic waves in vacuo [m/s]
$D$ - Conductor depth [m]
$\hat{d}, d$ - Traveling distance coordinate [m]
$\mathbf{E}, \mathbf{E}$ - Electric field [V/m]
$f$ - Frequency [Hz]
$G$ - Antenna gain [dBi]/[dBm]
$\mathbf{H}, \mathbf{H}$ - Magnetic field [A/m]
$h$ - Helix height [m]
$\hat{I}, I$ - Electric current [A]
$j$ - Complex imaginary unit [Dimensionless]
$k$ - Wave number [1/m]
$L$ - Inductance [H]
$\ell, \ell$ - Helix element length [m]
$N$ - Number of turns [Dimensionless]
$P$ - Power [W]
$p$ - Cylindrical coordinate parameterization variable [m]
$R$ - Resistance [\Omega]
$r$ - Helix radius [m]
$[S]$ - Scattering parameter [Dimensionless]
\( s \) - Helix spacing distance [m]
\( t \) - Unit of time [s]
\( V \) - Voltage [V]
\( v \) - Phase velocity [m/s]
\( w \) - Conductor width [m]
\( W_{\hat{x}},W_{\hat{y}} \) - Energy flux of an electromagnetic field [W/m^2]
\( \hat{x},x \) - Cartesian coordinate [m]
\( \hat{y},y \) - Cartesian coordinate [m]
\( Z \) - Impedance [\( \Omega \)]
\( \hat{z},z \) - Cartesian coordinate [m]
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12.1 Summary and Conclusion
12.2 Future Work
1 Introduction

There is a recent interest in the development of compact antennas for naval use. In larger submarines, there usually is a sail where the communication masts can be raised and lowered. The restrictions in regards to allocation of space for the antenna elements is not as critical in those cases. A sail as such may be absent on smaller ships, causing a challenge for the antenna element to fit within the ship in a suitable way. One suggested solution presented in order to circumvent the problem with the lack of space for the antenna system in a small-vessel submarine is to use a mast configuration which is vertically foldable and mounted on the outer rim of the ship. The idea is that the antenna will be folded up when used and folded down when not used, see Figure 1. Using a configuration as such do not set requirements for hull depth in the same way since the antenna elements does not need to be lowered and accommodated inside the ship. With a smaller hull on a submarine, the communication- and antenna systems of such a craft will have to add here to the limitations this will impose on the design of the system in terms of volume.

A small-vessel submarine, meant to accommodate only a few crew members, have the advantages of having a lower hydroacoustic signature and radar cross-section (RCS), than a normally sized submarine due to its smaller area [1][2], making it suitable for surreptitious operations. A small vessel also enables the possibilities for the submarine to being launched through the air by helicopter or through water by another submerged submarine [3].

Figure 1: Proposed configuration for a foldable antenna mounted on the rear of a submarine. Illustration courtesy of James Fisher Defence Sweden.

The properties of saltwater and its permittivity makes it act as a good conductor, and the path losses generated will be too great to withhold a communication link on higher frequencies while a submarine is submerged [4]. Therefore, the ship, or at least its antenna elements via e.g. a buoy\(^1\), must rise to the surface in order to establish a link when communicating [5]. On the open sea,

\(^1\)Flotation device.
the distance between transmitter and receiver might be beyond the line of sight. Using a satellite to relay the signal, the communication distances can be increased greatly [6].

1.1 Problem Formulation

Two antenna designs suited for a foldable communication system needs to be proposed and evaluated for a small-vessel submarine currently in development at this point in time. The submarine under construction has a hull with a total length of 10.45 m and main radius of 1.7 m. In the present case, there are two available design options with two different cylindrical volumes where the antenna elements can fit. One setup has a shorter mast and space for a radially narrow but axially longer antenna element with maximum radius of 5 cm and maximum height of 90 cm, or another setup where the mast is longer and elevates the antenna element, as in Figure 1, where the critical dimensions are flattened axially but extended radially with maximum radius of 10 cm and maximum height of 50 cm. The main intent for both designs is to use the antenna system to establish a communication link with the satellite system MUOS/Legacy which operates on frequencies between $f = [244-380]$ MHz. Since the submarine is supposed to operate all over the seas, it is desired that the antennas will radiate radio waves with a hemispherical beam shape in order to be able to establish a communication link regardless of position between submarine and satellite. The relative position between transmitter and receiver may vary significantly and may be completely unknown and. To avoid losses associated with this, circular polarization is used with MUOS/Legacy. Thus, circular polarization is set as a requirements for the designs.

The emphasis with the project will be on design parameters and on product development; to design a passive antenna element without power supply, find most suitable geometry and optimize the affecting parameters, and to explain the technical limitations of chosen designs. It will be evaluated if there are antenna constructions that can satisfy the given technical requirements, and to which extent, without exceeding the maximum volume for the two available antenna options. In conclusion, it will be compared which design seems most suitable for this application.

Simulations of the antenna design models will be performed in a time domain full-wave solver. Antenna prototypes based on presented theory in the thesis will be constructed and the far-field characteristics of the prototypes will be measured using slanting field measurements and direct illumination measurements.
1.2 Technical Specifications

The project scope is to design and evaluate antenna elements applied for naval use with following technical specifications:

- **Size constrains for the antenna elements**
  - Design alternative 1: radius = 10 cm, height = 50 cm. Will be referred to as *short antenna*
  - Design alternative 2: radius = 5 cm, height = 90 cm. Will be referred to as *long antenna*

- **Frequency Ranges, Acceptance VSWR ≤ 2**
  - $f = [244-270]$ MHz (corresponding to SatCom system Legacy’s down-link)
  - $f = [292-320]$ MHz (corresponding to SatCom system Legacy’s uplink and MUOS’s uplink)
  - $f = [360-380]$ MHz (corresponding to SatCom system MUOS’s down-link)

- **Hemispherical coverage over the horizon with isotropic gain requirements**
  - $G_i(\theta, \varphi) \geq 1.5$ dBi at $\theta = 0^\circ$
  - $G_i(\theta, \varphi) \geq 0$ dBi at $\theta = \pm 65^\circ$

within the frequency ranges

- **Right hand circular polarization**
  - $\text{AR}(\theta, \varphi) < 5$ dB when $-65^\circ < \theta < 65^\circ$

within the frequency ranges

These requirements are set to be met by simulated antenna models. Antenna prototypes based on the simulation model design will be constructed, measured, and analyzed. However, the main focus of this thesis report and the decision whether the antenna designs meet the requirements set will be based on the outcome from the simulation models.

1.3 Delimitations

This thesis will extend the full time over an entire semester, the equivalent of 30 European Credit Transfer and Accumulation System (ECTS), and limitations on the project will be set thereafter to have a reasonable scope. For this thesis, following assumptions and limitations were set:

- No signal processing will be performed
- Efficiency of the antenna will not be regarded
- Ideal components will be assumed in matching network calculations
- Near field effects due to a covering radome will not be evaluated in detail. A radome will, however, be included in simulation models
The mechanical construction of the foldable antenna system and possible problems, limitations, and drawbacks in regards to said construction will not be regarded.

1.4 Business Involvement

The project was designed by and carried out at Combitech AB. Combitech is a technology consulting company focusing mainly on telecommunication, information security, and system development, and is an affiliated company of the Saab Group. The design evaluation this thesis was based on was ordered by Comrod AS, a Norwegian antenna manufacturing company. The submarine manufacturer in mind for this thesis was James Fisher Defence Sweden who currently are developing small-vessel submarines, and who sees a field of application for the foldable antenna system.

1.5 Structure of the Thesis

This thesis has an arrangement as follows: Section 2 will address the terms and criteria needed in order to established and maintain a satellite communication link from a naval platform; which physical parameters need to be employed, and explain and clarify the origins of the parameters set in the scope. This section will also present a brief history of naval- and satellite communication. Next section, Section 3 will consider possible antenna design choices and state the advantages and drawback of each suitable antenna structure. Finally, the chosen antenna design will be present with accompanying motivation to why this design seemed to be the optimal solution for this application. Section 4 will present the theory of the chosen antenna structure and explain the relationship between the electrical and geometrical parameters. Also, the general theory of the expressions used throughout the thesis will be explained and derived if needed. Section 2, Section 3, and Section 4 corresponds to the information gathered during the literature review of the thesis. In Section 5, design, calculations, and dimensions for the antennas and matching network will be presented, which will be based on the theory presented in Section 4. The construction and measurement procedures of the antenna prototypes will thereafter be presented in Section 6 and Section 7. The next section, Section 8 will deal with presenting and analyzing the simulated and measured data gathered from the antenna models. Section 9 and Section 10 will present setup and results of the produced antenna designs when placed in a submarine-like simulation environment. Section 11 will discuss and determine the plausibility of the produced results and point out potential sources of error. Finally, in Section 12, conclusions and future work based on this thesis will be presented.
2 Satellite Communication for Naval use

This section will present a brief of the history of communication for submarines, and also clarify the conditions needed to establish a communication link when in a water-surrounded environment. Also, brief history and technical properties of satellite communication will be explained. Thereafter, the general electrical properties sought of an antenna aimed towards naval satellite communication applications will be presented, as well as the properties of the satellite systems which the antenna aims to be compatible with; the operating frequencies and coverage areas. The purpose of this section is to clarify the technical specifications set in this thesis.

2.1 History and Conditions for Radio Communication for Submarines

Long distance communication for submarines dates back as early as the mid-1800s where long-haul coaxial lines were placed across the ocean floor where the submarines could connect to fixed points and links to the mainland could thereafter be established. During the second world war, where both submarines and radio communication became key components in modern warfare, these fixed points were easy targets for the enemy to sabotage and a more mobile way of communicating with naval vehicles was sought [7]. The solution was found in combining the two techniques and use wireless telegraphy as the main way of communication [8]. Wireless transmission provided the advantage of mobility for the ships when establishing communication with mainland.

In a material with good conductivity, such as salt water, the penetration depth, or skin depth, $\delta$, of the radio frequency (RF) wave is proportional to inverse square root of operating angular frequency, $\delta \propto 1/\sqrt{\omega}$, where $\omega$ is the operating frequency, i.e. with increased operating frequency, the radio wave attenuates more rapidly when propagating through water. This means that in order to communicate using RF waves while submerged, low operating frequencies need to be used in the transmission [9]. With a lower operating frequency on the signal, the bandwidth of said transmission will be narrower and, therefore, the quality of the data transferred in the process will be lower. By communicating using frequencies on very low frequencies (VLF), $f = [3-30]$ kHz, or using an even lower band, submarines can communicate using RF waves while still submerged. However, the information sent will have to be in form of low-bit data, e.g. Morse code [9]. To communicate with a higher bit rate and higher quality, the communication system needs to be designed for higher frequencies and submarine, or at least the antenna element needs to rise up to the surface in order to establish a communication link and may, therefore, potentially reveal its position more easily. The trade-off in regards to mobile submarine communication is, therefore, either to stay submerged and only be able to communicate with information of low-bit capacity, or rise to the surface where the data transfer rate of the messages can be extensively larger but by doing so increases the risk of the position of the submarine to be revealed by enemy forces.

Direct Line-of-Sight (LoS) RF communication between transmitter and receiver is limited due to Earth’s curvature. Communication systems using satel-
lites was introduced in the 1960’s in order to bypass the limitations of point-to-point RF transmissions [10]. Global satellite systems were put into orbits around the Earth and were used as relays where the received signal was amplified and re-transmitted, and the coverage area of the transmission was, therefore, increased greatly. As long as the intended transmitter, or receiver, was located within the satellites covering area, and had the right antenna characteristics, a communication link could be established.

One of the major advantages of satellite communication SatCom is that it enables point-to-multi-point communication; in systems with multiple satellites in orbits which can transmit signals between themselves, almost global, high-speed, coverage is possible [11] [12]. The efficiency of SatCom is in general lower compared to modern, stationary, undersea fiber communication systems such as landlines, underwater cables, and microwave transmission links. However, the mobility SatCom entails still makes it a preferable alternative for military submarine communication system [13] [14]. Frequencies on the upper very high frequency (VHF) band, \( f = [30-300] \text{ MHz} \), and ultra high frequency (UHF) band, \( f = [300-3000] \text{ MHz} \), is usually employed when communicating via SatCom [15].

2.2 Mobile Satellite Communication Antenna Properties

In order for a submarine to communicate and establish links with satellites regardless of its position on the sea, the antenna far-field pattern is often desired to have hemispherical characteristics, illustrated in Figure 2 where the strongest gain region is visualized with the red color and the gain decay thereafter follows the scale of the visible light spectrum. The trade-off when using a hemispherical, round covering gain pattern is that the gain peak is significantly lower compared to an antenna with a directed beam. However, by having a broad coverage, the need to pinpoint the satellite position using complex tracking systems is removed, which is the case when using a directional antenna. Using a tracking system would have increased visibility on radar tracking system and potentially could reveal current position of the ship [16] [17].

Submarine SatCom systems usually utilize circularly polarized (CP) antennas when communicating. This is due to the fact that the ship and satellite may have a relative rotation between themselves and if communicating using linearly polarized (LP) RF waves, a polarization mismatch between transmitter and receiver can cause significant losses in information. Communication using circularly polarized waves removes the need to align the transmitting and receiving aperture to each other [18]. Another main reason to use CP waves in SatCom applications is to avoid the effects of Faraday’s rotation. When RF waves are entering or leaving Earths atmosphere, the waves propagate through a layer called the ionosphere. The ionosphere consists of ionized, charged particles in form of plasma radiated from the sun which is trapped in Earths static magnetic field [19]. When radio signals are propagating through this layer, the waves will interact with the charged particles causing the waves to rotate in an unpredictable manner [20]. This rotation is called Faraday rotation and can result in severe transmission losses due to polarization mismatch in RF communication; if the waves unintentionally rotate when passing the ionosphere, great losses between transceiver and receiver can occur if the waves are LP. Using CP,
this issue is avoided since a rotation as such will only result in a phase shift for the RF waves [21].

Figure 2: Representation of a hemispherical radiation pattern where the main beam points upwardly.

2.3 Satellite Systems; MUOS and Legacy

The antenna designs in this project are intended to support the satellite system Mobile User Objective System (MUOS), a duplex point-to-multi-point satellite communication system primarily used for military applications. MUOS consists of four active satellites traveling in geosynchronous orbit on longitude positions 177° W, 100° W, 15.5° W, and 75° E respectively on a height of 37 000 km above Earth. Each satellite covers from latitude 65° N to latitude 65° W. The satellites communicates with radio access facilities (RAF) which are located on Hawaii (USA), Virginia (USA), Sicily (Italy), and Geraldton (Australia). With this positioning, each satellite is in view of two RAFs and each RAF can view two satellites [22]. Satellite coverage and RAF’s are shown in Figure 3.

Figure 3: Coverage of the MUOS/Legacy satellites. RAF positions included.
To bridge the gap between two generations of SatCom systems, an older payload called Legacy is included in MUOS. This combines the global internet protocol, net-centric networks of MUOS with Legacy’s circuit-based, closed networks [23]. MUOS consist of one uplink (UL) and one downlink (DL) where the allocated frequencies are on the lower UHF band. With the Legacy up- and downlink included, located in the upper very high frequency (VHF) band, the frequency range is spanning in total over $f = [244-380] \text{ MHz}$ [10]. The frequency bands allocated for the systems are visualized in Figure 4.

![Figure 4: Frequency band use for the MUOS/Legacy satellites.](image)

The Legacy payload consists of 38 communication channels; 17 25-kHz communication channels and 21 5-kHz communication channels where the up and downlink allocates a bandwidth of 26 MHz each, while MUOS is divided into eight 5-MHz channels; four for the uplink and four for the downlink with a total bandwidth of 20 MHz for both up and downlink. The V/UHF frequency bandwidth is highly limited and is a standard use in military SatCom [24]. In order to increase communication capacity on these frequency bands, a dynamic bandwidth allocation services called Demand Assignment Multiple Access (DAMA) is utilized with MUOS. This service provides multiple access to a single V/UHF channel [12]. The DAMA information transfer process is based on a technique where packet-switching is used. The data is broken down into small parts and is transmitted in the form of separate packets instead of the entire message in a large package, which is the case when transmitting on the Legacy channels. Each packet contains the source- and destination address and is transmitted through a common outbound link. Using this package-switch technique, 500 users can share a single communication channel [12] and it matches user demands to available satellite time, minimizing delays in the transfers [11]. The 5-kHz DAMA protocol primarily supports multiple-user 75 bps to 2.4 kbps data communication while 25-kHz DAMA protocol supports data rates of 75 bps to 16 kbps for data and voice. Both systems use right hand circular polarization RHCP as polarization state on the antennas [25].
3 Antenna Design Evaluation

In this section, a review of suitable antenna types which are of sought compact antenna design will be conducted. The electrical and geometrical properties of each antenna type will briefly be presented together with references to publications where more information on said antenna design can be found. If the antenna design is rejected for this project, it will be stated on which grounds the decision was made, and also a motivation will be presented to why the chosen design is believed to have the most suitable antenna design for this specific application.

3.1 Compact Antenna Designs for Mobile Satellite Communication Systems

The trivial relation between velocity, frequency, and wavelength in a non-dispersive medium, such as air, is

$$\lambda = \frac{c}{f}$$

where $\lambda$ is operating wavelength and $c$ is the speed of electromagnetic waves in vacuo. Using this equation, it is found that operating wavelength varies between $\lambda = [0.79-1.23]$ m when communicating on the MUOS/Legacy channels. Expressing the given maximum antenna dimensions in terms of wavelength, the radius, $r$, and height, $h$, for the small antenna will vary between $r = [0.08-0.13] \lambda$, $h = [0.41-0.63] \lambda$, and for the long antenna, the dimensions will be $r = [0.04-0.06] \lambda$, $h = [0.73-1.14] \lambda$. Most mobile communication antenna structures are based on resonating and traveling wave principles and therefore have a geometrical size in the same order of magnitude the operating wavelength [26]. The maximum cylindrical size constraints is a critical factor in this project and limit the possible design approaches. The given requirements in regards to gain level and far-field pattern coincide with the definition of a low-gain antenna (LGA), where the maximum antenna gain, $G_i$, in general, is low, between $G_i = [0-4]$ dBi, but instead, there is a wide beam coverage [27]. Below follows a review of suitable LGA designs.

3.1.1 Crossed Dipole Antenna

The crossed dipole antenna, also called turnstile antenna, consists of two crossed dipoles where the relative phase between the dipoles is shifted by the phase component $\phi = 90^\circ$, see Figure 5. Using this setup, CP waves will radiate from the antenna structure and it will have a main lobe along the antenna axis, making the antenna suitable for SatCom applications [28]. However, as with the single dipole structure, each dipole will approximately need radially spacing on $r = 0.25 \lambda$ and critical dimensions will, therefore, be exceeded in that direction, making the crossed dipole antenna unfit for this case[29]. Some optimization can be made on the compact design parameters, as shown in [30], but maximum radial dimensions would still be exceeded. The related designs, inverted V-form cross dipole antenna [31] and crossed-drooping dipole antenna [32] also need a radius in the order of $r = 0.25 \lambda$. In the latter case, the conductors are tilted towards the ground and therefore take up less space radially [33], but
the maximum radial volume is still exceeded for both designs. Using a crossed-slot dipole antenna [34], which is based on Babinet’s principle, the radius for the cavity needs to be \( r = \frac{\lambda}{1.36\sqrt{\varepsilon}} \), where \( \varepsilon \) is the dielectric constant of the substrate and the dimensions of this antenna design will exceed set radial limitations.

![Crossed dipole antenna setup seen from above. The non-phase shifted dipole conductors is visualized with the green lines and the quadrature phase shifted dipole conductors is visualized with the blue lines. A simplified feed network is included.](image)

**Figure 5:** Crossed dipole antenna setup seen from above. The non-phase shifted dipole conductors is visualized with the green lines and the quadrature phase shifted dipole conductors is visualized with the blue lines. A simplified feed network is included.

### 3.1.2 Spiral Antenna

A wire antenna structure which radiates a hemispherical far-field with high polarization purity is the spiral antenna. The most common designs are Archimedean spiral visualized in Figure 6 [35], logarithmic spiral [36], and equiangular spiral [37]. Antenna structures which are circularly symmetric will naturally generate CP waves [38]. The spiral antennas are so-called frequency independent which means that they can operate over large frequency bandwidths. However, the frequency for the spiral antennas is usually in the order of \( f = [1-18] \text{ GHz} \) [39]. For frequencies in the V/UHF band, it was shown in [40] that the required radius of the different spiral antennas will be too great for the constructions to fit inside the allocated spaces.

The conical logarithmic spiral is another LGA spiral design where the conductor radius is decreasing while rotating upwardly around the antenna axis, allocating a total volume of conical shape [41]. This antenna design also has the sought properties of hemispherical radiation pattern and a high polarization purity along the antenna axis. However, as in the case of the other spirals, the allocated radial space is too narrow and neither of the antenna design volumes in this project will have room for such antenna type.
3.1.3 Patch Antenna

With two feed ports separated relatively in phase by $\phi = 90^\circ$, or using a circular patch as in Figure 7, it is possible to design a microstrip patch antenna which has sought features in regards to radiation pattern and circular polarization, as demonstrated in [42]. A microstrip patch is, however, a resonant antenna which, as stated previously, means that it must have a length comparable to half a wavelength at the relevant operation frequency and it is therefore not applicable to this case. Also, the impedance bandwidth of a microstrip patch antenna is narrow and covering the whole frequency span would have been difficult [43].

3.1.4 Monofilar Helix Antenna

The monofilar antenna structure, Figure 8, with the two most common modes axial mode helix antenna and normal mode helix antenna, radiate CP waves and are used for SatCom and broadcasting [44] [45]. However, the characteristics of these structures are not suited for this application: The axial mode helix antenna radiates with a high gain, narrow-beamed characteristic aligned with the antenna axis and is more suited when pinpointing satellites rather than when
wide beam coverage is desired. The normal mode helix antenna radiates along the normal direction of the antenna, thus creating an omnidirectional radiation pattern, also not wanted for this application. Both monofilar helix antennas require space in the radial direction which would exceed set limitation and also need to be connected to a ground plane, which contributes to an increased volume of the structures. General properties of these structures will be presented in Section 4.1.

Figure 8: Conductor shape of a monofilar helix antenna.

3.1.5 Chosen Design: Quadrifilar Helix Antenna

It was found that the most suitable antenna structure for both design options would be to have a helical geometry with a quadrifilar set of antenna conductors, see Figure 9. The quadrifilar helix antenna structure radiates a broad beam along the antenna axis with high circular polarization purity. The antenna dimensions are radially small compared to operating wavelength, and the cylindrical volume the antenna elements allocates when operating on the MUOS/Legacy channels coincides with the set size constraints. The quadhelix antenna design is common when it comes to submarine communication systems and has been used for submarine applications since the 1970’s [46]. The quadri- filar helix antenna has the ability to generate a backward wave, removing the need for a ground plane which is of great advantage when volume minimization is of high priority [47]. A more in-depth description of the quadrifilar helix structure and its properties will be presented in Section 4.2.
3.1.6 Combination of Antenna Elements

Some design ideas using multiple antenna elements combined were considered but eventually abandoned. A suggested option for the short antenna was to use a design consisting of an axially aligned dipole construction radiating in the azimuth direction, enclosed by an axial mode helix antenna radiating in the axial direction, see Figure 10. The idea was to have an electronic switch device, enabling the possibility to switch between the two antenna elements, depending on the relative position between transmitter and receiver. The losses generated by the polarization mismatch from dipole would have been compensated by the high gain and the total far-field, combining the separate far-field, would have a hemispherical character. However, a proper isolation between the two antenna elements would have been hard to achieve and interference and unwanted coupling between the antenna elements would have been likely to occur which presumably had resulted in distorted fields and unwanted reflections in the system. Also, the electric switch required to perform these operations would have to be designed as well, which would be outside the project scope, and the idea was eventually abandoned.
Figure 10: Visualization of a quadrifilar helix antenna enclosing a dipole antenna.

Another idea for the short antenna design, was to use two quadrifilar helix antenna structures; one larger antenna structure for the Legacy UL and DL which would enclose a smaller structure which would resonate solely on the MUOS DL as a parasitic element, see Figure 11. However, a design as such would also have uncontrollable coupling between the radiating elements and the idea was rejected, and it was found that the most plausible way to satisfy set requirement was to optimize the single quadrifilar helix design.

Figure 11: Visualization of a quadrifilar helix antenna enclosing another quadrifilar helix antenna.
4 General Theory

This section will describe the physical properties of the chosen antenna design, the quadrifilar helix antenna, and the relevant parameters used to describe the antenna system. Initially, the mathematical expressions for a helical structure will be presented. This is followed by a brief explanation of the two most common modes of operation for the monofilar helix antenna, with associated field properties and the relation between antenna dimensions and operating frequency. Thereafter, the quadrifilar helix antenna will have an in-depth explanation where the relation between antenna dimensions, operating frequency, and radiating far-field will be described in detail. The subsequent subsection will explain impedance matching and how to optimize transmission for the desired frequencies. Thereafter, the general far-field properties of an antenna will be described. Finally, the definition of the state of polarization and its relation to the electric field components of the far-field will be demonstrated. Throughout this section, the antenna will mostly be treated, both via text and figures, as a transmitting object. However, due to the reciprocal properties of the antenna, an identical behavior will arise when the antenna is on the receiving end of the system.

4.1 Monofilar Helix

A helix antenna is a wire antenna where the conductor is wrapped around its axis ascendingly, creating a spiral and, therefore, enclosing a cylindrical volume, see Figure 32. The helix geometry can be described using following parameters

\[ r = \text{radius of helix} \]
\[ s = \text{spacing between turns} \]
\[ \alpha = \text{pitch angle} = \arctan \left( \frac{s}{2\pi r} \right) \]
\[ N = \text{number of turns} \]
\[ h = \text{height} = Ns. \]

Using a cylindrical coordinate system with parameterization variable \( p \), the equations for a left hand winded helix vector, \( \ell \), can be expressed as

\[ \ell(p) = r \cos \left( \frac{2\pi Np}{s} \right) \hat{x} - r \sin \left( \frac{2\pi Np}{s} \right) \hat{y} + p \hat{z}, \quad (2) \]

where \( 0 \leq p \leq s \) and the line integral for the helix is

\[ \ell = \int_{2\pi r} ds = \int_{0}^{s} dp \sqrt{\left( \frac{dx}{dp} \right)^2 + \left( \frac{dy}{dp} \right)^2 + \left( \frac{dz}{dp} \right)^2} = \sqrt{s^2 + (2\pi Nr)^2} \quad (3) \]

i.e. an unfolded helix arm will create a right triangle with the circumference and height, and the helix arm length can be expressed using the Pythagorean identity. The radiation characteristics of the monofilar helix depends on the relation between geometrical dimensions and operating wavelength. Below follows a presentation of the two most common monofilar helix antennas, which both have electrical properties which are reminiscent of the later used quadrifilar helix antenna.
4.1.1 Normal Mode Helix Antenna

The first of the two most common monofilar helix antenna types, the normal mode helix antenna (NMHA) radiates along the normal direction relative to the lateral surface of the enclosing cylindrical volume if the helix antenna dimensions are small compared to operating wavelength, $s, r < \lambda$. When these requirements are satisfied, the helix current can be assumed to have a constant value in both phase and amplitude, and the current will have both circumferential and axial components [44]. With these assumptions, the helix can be seen as a resonator and the radiating fields can be approximated to be a sum of electric dipoles and magnetic loops using the superposition principle; each dipole length is equivalent to the spacing distance $s$ and each loop is equivalent to the circumference $2\pi r$, for each turn. [48]. If the elevation angle $\alpha \rightarrow 0^\circ$, the antenna will act as a magnetic loop, and if the conductor width is equal to the diameter of the structure, the antenna will instead act as an electric dipole. Using these assumptions, the electric fields can be expressed as

$$E_{\varphi-\text{NMHA}} = \frac{\omega \mu_0 I_L k \pi r^2}{4\pi d} \sin(\theta) e^{-jkd}$$  (4)

$$E_{\theta-\text{NMHA}} = j\frac{\omega \mu_0 I_D s}{4\pi d} \sin(\theta) e^{-jkd}$$  (5)

where $I_L$ and $I_D$ is the respective dipole and loop current, $d$ is the traveling distance, $k$ is the wave number, $\omega$ is the angular frequency, $\mu_0$ is the magnetic permeability, and $j$ denotes the imaginary complex unit. $E_{\varphi-\text{NMHA}}$ and $E_{\theta-\text{NMHA}}$ indicates azimuth and altitude components of the electric field, respectively. When $I_L = I_D$, the conditions are satisfied and the structure will then radiate beams along the normal direction of the antenna [49]. A cross-section of the far-field radiation pattern for the NMHA is shown in Figure 13.
CP for an antenna is achieved if the azimuth and altitude components of the electric field, $E_\theta$ and $E_\varphi$, has a phase quadrature relation, i.e.

$$\frac{E_\varphi}{E_\theta} = \pm j \Rightarrow \frac{|E_{\varphi-NMHA}|}{|E_{\theta-NMHA}|} = 1 \Rightarrow 2\pi r = \sqrt{2}\lambda s. \quad (6)$$

For the NMHA, as for all helical antenna structures, the winding direction determines the polarization state; for monofilar helixes, clockwise winding generates right hand circular polarization (RHCP) while counter clockwise winding generates left hand circular polarization (LHCP) [50]. The states of polarization will be explained more thoroughly in Section 4.5.

### 4.1.2 Axial Mode Helix

The second common monofilar helix antenna is called the axial mode helix antenna (AMHA). If the helix antenna is designed to have a circumference close to the approximate length of the operating wavelength, i.e. $2\pi r \approx \lambda$, the induced current will generate fields which radiate as the current rotates and progresses along the conductor. The guiding structure will, therefore, be the main radiating mechanism and it will act as a traveling wave antenna rather than a resonator antenna, which is the case for the NMHA [44]. The AMHA has the main radiation lobe aligned with the antenna axis with minor side-lobes on the sides of the structure. The radiation pattern has a relatively narrow beam width (BW) and instead has a high maximum gain. An example of a typical far-field radiation pattern for the AMHA is demonstrated in Figure 14.
Figure 14: Radiation pattern for an AMHA, radiating along the axial direction of the antenna.

For an AMHA, the magnetic vector potential, $A_{\text{AMHA}}$, can be expressed as

$$\mathbf{A}_{\text{AMHA}}(d) = \frac{\mu I_0 e^{-jkd}}{4\pi d} \int_0^{\varphi'} e^{j(kr\cos(\theta)\tan(\alpha))} e^{j(kr\cos(\theta)\tan(\alpha) - \omega h/v\varphi')} \hat{I} \, d\varphi'$$

where $I_0$ is the current magnitude, $v$ is phase velocity, and $\varphi'$ is an arbitrary azimuth point. The unit vector for the induced current along the helix wire, $\hat{I}$, can be expressed as

$$\hat{I} = -\hat{x}\sin(\varphi') + \hat{y}\cos(\varphi') + \hat{z}\sin(\alpha)$$

where $\hat{x}$, $\hat{y}$, and $\hat{z}$ are the Cartesian unit vectors. Neglecting the electric potential, the electric field components of an AMHA, $E_{\text{AMHA}}$, can be expressed as

$$E_{\text{AMHA}} = -\frac{dA_{\text{AMHA}}}{dt}$$

where $t$ is the time variable. If the time variation of the fields is assumed to be periodic sinusoidal and thus non-decaying, the time derivative, $\frac{d}{dt}$, can be assumed to have following representation in the frequency domain

$$\frac{d}{dt} \to j\omega,$$

since the information signal can be assumed to vary much slower than the oscillation. The spherical components of the magnetic vector potential can, therefore, be expressed as

$$E_{\phi,\text{AMHA}} = -j\omega((A_x \cos(\varphi) + A_y \sin(\varphi)) \cos(\theta) - A_z \sin(\theta))$$

$$E_{\varphi,\text{AMHA}} = -j\omega((A_y \cos(\varphi) + A_z \sin(\varphi)) \cos(\theta))$$

where $A_x$, $A_y$, and $A_z$ are the cartesian scalar components of the magnetic vector potential [51]. Due to the properties of a traveling wave antenna, the
degree of polarization for the AMHA depends largely on the number of turns in the antenna structure, rather than relations between geometries;

\[
\frac{|E_{\phi-AMHA}|}{|E_{\theta-AMHA}|} \propto \frac{2N + 1}{2N} \quad (13)
\]

i.e. with a higher number of turns, a higher rate of polarization purity is achieved [45]. The relations for the normal and axial modes, antenna dimensions, pitch angle, and operating wavelength are being shown in Figure 15 with the normalized axes \( r_\lambda = \frac{r}{\lambda} \) and \( s_\lambda = \frac{s}{\lambda} \).

![Figure 15: NMHA and AMHA dimensions, cf. [50].](image)

Observing Figure 15 and comparing to the technical requirements in Section 1.2, none of the presented monofilar helical structures have the exact properties sought after, neither in geometrical properties nor in radiation characteristics. The AMHA has a high gain beam aligned with the axis but the BW is too narrow to satisfy set requirements of the antenna design. Also, as underlined in Section 5, the monofilar structures needs a ground plane. However, by connecting several monofilar helices in an azimuthal array and using suitable dimensions, the maximum gain of the antenna will be decreased but the BW is instead increased. This model is called the quadrifilar helix antenna and is described next.
4.2 Quadrifilar Helix

The quadrifilar helix antenna (QHA) consists of an arrays of four monofilar helices with azimuth separation of by a quarter turn, i.e. the conductors have a relative separation of $\varphi = 0^\circ$, $\varphi = 90^\circ$, $\varphi = 180^\circ$, and $\varphi = 270^\circ$. As in the case of the monofilar helix structure, the total enclosed volume has a cylindrical shape and a quadrifilar structure will occupy equal amount of space as a monofilar structure of same dimensions. The four helix vectors can be expressed as

$$\ell_1(t) = r \cos \left(\frac{2\pi Np}{s}\right) \hat{x} - r \sin \left(\frac{2\pi Np}{s}\right) \hat{y} + p\hat{z} \quad (14)$$

$$\ell_2(t) = r \sin \left(\frac{2\pi Np}{s}\right) \hat{x} + r \cos \left(\frac{2\pi Np}{s}\right) \hat{y} + p\hat{z} \quad (15)$$

$$\ell_3(t) = -r \cos \left(\frac{2\pi Np}{s}\right) \hat{x} + r \sin \left(\frac{2\pi Np}{s}\right) \hat{y} + p\hat{z} \quad (16)$$

$$\ell_4(t) = -r \sin \left(\frac{2\pi Np}{s}\right) \hat{x} - r \cos \left(\frac{2\pi Np}{s}\right) \hat{y} + p\hat{z} \quad (17)$$

for left hand winding, and each arm is included in Figure 16.

![Quadrifilar helix diagram](image)

Figure 16: Quadrifilar helix enclosing a cylindrical volume. Start and endpoint for each helix arm as well as spherical coordinate system axis are included.

The quadrifilar helix structure can also be seen as two bifilar helix antennas (BHA), i.e. the two opposing arms correspond to the first antenna structure.
and the orthogonal, opposing arms correspond to the second antenna structure. Relative to Figure 16 and Equations (14)-(17), \( \ell_1 \) and \( \ell_3 \) would be one BHA structure and \( \ell_2 \) and \( \ell_4 \) would be the other BHA. Assuming a sinusoidal current distribution with the phase velocity along the wires, the radiating fields \( E_{\theta \text{-BHA}} \) and \( E_{\varphi \text{-BHA}} \) for one BHA can, as in the case of the NHMA, be assumed to be electric and magnetic dipoles in superposition [52]. The fields can, therefore, be expressed as

\[
E_{\theta \text{-BHA}} = j Z_0 \frac{\ell_1}{d} \sin(\theta) \left[ \frac{s}{\lambda} e^{-jkd} \right] \\
E_{\varphi \text{-BHA}} = Z_0 \frac{\ell_2}{d} \sin(\theta) \left[ \frac{\pi r^2}{\lambda^2} e^{-jkd} \right]
\]

(18)

(19)

where \( \ell_1 \) and \( \ell_2 \) are the respective dipole and loop current, visualized in Figure 17, and \( Z_0 \) is the impedance of free space. As in the case of the NMHA, when \( I_L = I_D = I_0 \), the antenna will radiate circularly polarized waves [53].

\[
\approx I_L + j \ell_1
\]

Figure 17: Illustration of the superposition principle of the helix current for one turn BHA.

Following the assumption with summation of loop and dipole currents for each BHA, the fields for the full QHA structure, with another orthogonal BHA connected, can be expressed as

\[
E_{\theta \text{-QHA-1}} = \frac{\omega \mu I_0}{4\pi d} \cos(\theta) \sin(\theta) e^{-jkd} \\
E_{\varphi \text{-QHA-1}} = \frac{\omega \mu I_0}{4\pi d} \cos(\theta) e^{-jkd} \\
E_{\theta \text{-QHA-2}} = \frac{\omega \mu I_0}{4\pi d} \cos(\theta) \sin(\phi) e^{-jkd} \\
E_{\varphi \text{-QHA-2}} = \frac{\omega \mu I_0}{4\pi d} \cos(\theta) \cos(\phi) + j \sin(\phi) e^{-jkd}
\]

(20)

(21)

(22)

(23)

where \( E_{\theta \text{-QHA-1}} \) and \( E_{\varphi \text{-QHA-1}} \) are the electric angular and azimuth fields of the first BHA and \( E_{\theta \text{-QHA-1}} \) and \( E_{\varphi \text{-QHA-1}} \) are the electric angular and azimuth fields of the second BHA. If the separate BHAs have equal amplitude and are in a phase quadrature relation, \( i.e. \phi = 90^\circ \), the total azimuth and elevation
electric fields, $E_{\varphi\text{-QHA}}$ and $E_{\theta\text{-QHA}}$ respectively, can be expressed as

$$E_{\theta\text{-QHA}} = |E_{\theta\text{-QHA-1}} + E_{\theta\text{-QHA-2}}| = \frac{\omega \mu I_0 r \cos(kr)}{4\pi d} \sqrt{\sin^2(\theta) + \sin^2(\varphi) + \cos^2(\theta) \cos^2(\varphi) + 2 \sin(\theta) \sin(\varphi)}$$

$$\exp\left(j \arctan\left(\frac{\cos(\theta) \cos(\varphi)}{\sin(\theta) + \sin(\varphi)}\right)\right),$$

$$E_{\varphi\text{-QHA}} = |E_{\varphi\text{-QHA-1}} + E_{\varphi\text{-QHA-2}}| = \frac{\omega \mu I_0 r \cos(kr)}{4\pi d} \sqrt{\sin^2(\theta) + \sin^2(\varphi) + \cos^2(\theta) \cos^2(\varphi) + 2 \sin(\theta) \sin(\varphi)}$$

$$\left(-j \arctan\left(\frac{\sin(\theta) + \sin(\varphi)}{\cos(\theta) \cos(\varphi)}\right)\right),$$

and if these requirements are satisfied, the antenna structure will radiate a cardioid pattern with high circular polarization purity along the antenna axis [54]. It was shown in [52] that this approximation, with an electric dipole and magnetic loop superposition, was valid when increasing and decreasing the fractional number of turns. In [55], the geometrical properties, such as radius and spacing, was mapped in relation to electrical properties, such as wavelength and radiation pattern. Following parameters can be used for describing the QHA dimensions and far-field:

$$r_\lambda = \frac{r}{\lambda} = \text{radius in wavelengths}$$

$$s_\lambda = \frac{s}{\lambda} = \text{spacing in wavelengths}$$

$$\kappa = \frac{r_\lambda}{s_\lambda} = \frac{r}{s} = \text{radius/spacing coefficient}$$

$$\theta_p = \text{angle between the antenna axis and maximum lobe}$$

$$\theta_{BW} = \text{angle from antenna axis to where maximum lobe has dropped 3 dB}$$

$$\gamma = \text{depth of pattern minimum at } \theta = 0^\circ \text{ relative to pattern maximum}$$

where the first three parameters are the normalized dimensions with respect to the operating frequency and latter three parameters indicate the shape of the radiation pattern. It was found that if the normalized antenna dimensions is positioned inside the marked region in Figure 18, the radiation pattern of the QHA will have a hemispherical shape. However, it was found that the exact shape of the radiation pattern is highly dependent on the operating wavelength. In Figure 19 and Figure 20, the mapped data is shown for $N = 1$ QHA and $N = 2$ QHA, respectively, the two relevant graphs for this project. In the figures, the shape of the radiation pattern is expressed using the parameters $\theta_{BW}$, $\theta_p$, and $\gamma$ along set $\kappa$ values. When the normalized antenna dimensions are set, the behavior of the radiation pattern for the antenna design can be predicted using this data.
Figure 18: Region of shaped conical beam performance for QHA, cf. [55].

Figure 19: Radiation pattern characteristics for QHA when $N = 1$, cf. [55].
Figure 20: Radiation pattern characteristics for QHA when \( N = 2 \), cf. [55].

Figure 21 below shows two examples of radiation patterns of the QHA and how it correlates to the mapped data in Figure 19 and Figure 20. In Figure 21(a), the polar plot shows an example of when the radiation pattern of the QHA has a main beam aligned with the antenna axis, radiating in a typical hemispherical manner. In those cases, \( \gamma = 0 \, \text{dB} \) and \( \theta_p = 0^\circ \), corresponding to the behavior when the antenna operates on lower frequencies in Figure 19 and Figure 20. In Figure 21(b), the main antenna lobe is offset and \( \gamma \), as well as \( \theta_p \), will, therefore, have non-zero values, corresponding to when the QHA operates on higher frequencies on said figures. Using this information, it can be predicted that the antennas will radiate perfectly hemispherically on the lower frequencies and have a main beam offset on the higher operating frequencies.
(a) Radiation pattern when the main lobe is aligned with the antenna axis.

(b) Radiation pattern when the main lobe is offset from the antenna axis.

**Figure 21:** Radiation pattern for QHA, cf. [55].

The generated wave will be reflected at the bottom of the antenna and will thereafter radiate toward the feed, in a backward radiating mode. Therefore, the feed arrangement needs to be on the top of the antenna structure and, in order to achieve RHCP, the QHA needs counter-clockwise turning on the conductors, the opposite of NMHA and AMHA. An additional condition that needs to be satisfied is that the total length of the helix arm, $\ell_{\text{helix}}$, consisting of the rotating arm and the two connecting radius’s, must be in the order of the intended operating wavelength i.e.

$$\ell_{\text{helix}} = \ell + 2r = \sqrt{(2\pi r)^2 + h^2} + 2r \approx \lambda$$

which also sets constraints on the antenna design [56].
According to the figures, the normalized dimensions needed to design a QHA is varying between $r_\lambda = [0.01-0.13]$, and $s_\lambda = [0.3-0.7]$. Comparing these dimensions to the normalized dimensions of the two antenna designs presented in Section 3.1, this antenna design is expected to be able to satisfy set size constrains and, due to the properties of the QHA, radiating with wanted characteristics. As stated in the antenna review in Section 3, the QHA does not need a ground plane due to the backward generated wave, a great advantage when volume reduction is wanted. Also, the QHA is of low weight and is easy to produce. The QHA does, however, have a fairly low achievable bandwidth, and impedance matching is needed to be designed in order to potentially meet the requirements regarding bandwidth.

4.2.1 Feed Arrangement

There are mainly two different approaches to set up a feed arrangement for the QHA; externally phased and self-phased feed arrangement [57]. Both will be described below.

The first feed structure, externally phased arrangement, is constructed by using external hybrid phase shift components in order to shift the phase of current running through the individual helix arms relative to each other. The hybrid phase shift component is a four-port device which splits the input signal into two outgoing signals where one of the signals will remain unchanged regarding the value of phase and the other signal will have a relative phase rotation, which is specified by the component. The non-used port is grounded. Two different designs can be used to achieve a phase quadrature relation when using externally phasing; the first approach is to use a setup where each of the monofilar helix arms are fed with a phase shift of $\phi = 90^\circ$ relative to the next arm circumferentially, i.e. the individual conductors will have the relative phases $\phi = 0^\circ$, $\phi = 90^\circ$, $\phi = 180^\circ$, and $\phi = 270^\circ$. The opposing conductor arms will be electrically shortened on the opposite side of the feed arrangement, i.e. conductor 1 and 3 will be connected as well as conductor 2 and 4, as defined in Figure 16 and the connected, perpendicular, bifilar structures will, therefore, be in quadrature phase. This quadrature phasing relation between the bifilar helices is necessary for achieving CP RF waves. To achieve this phase relations, two $\phi = 90^\circ$ phase hybrids and one $\phi = 180^\circ$ phase hybrids need to be used in the circuit feed arrangement, as demonstrated in Figure 22(a). Another way to achieve these phase relations for the conductors is to use four $\phi = 90^\circ$ phase hybrids.

The other way to achieve phase quadrature relation between the radiating fields using an externally phased arrangement is to feed the arms in pairs of BHAs and phase shift the current in one of the arms with $\phi = 90^\circ$. With this approach, only one $\phi = 90^\circ$ hybrid need to be used to phase shift one of the perpendicular bifilar helices, see Figure 22(b). The scattering parameters of the phase hybrid components, i.e. the small-signal representation of the relation between the input and output voltages at the ports, are

$$ [S]_{\phi=90^\circ} = -\frac{1}{\sqrt{2}} \begin{pmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{pmatrix} $$  \hspace{1cm} (27)
\[
[S]_{\phi=180^\circ} = \frac{1}{\sqrt{2}} \begin{pmatrix}
0 & 1 & 1 & 0 \\
1 & 0 & 0 & -1 \\
1 & 0 & 0 & 1 \\
0 & -1 & 1 & 0
\end{pmatrix}
\] (28)

and observing the expressions, the power is divided between the two quadrature-phased out-ports.

The second way to design a feed arrangement is to use a self-phased arrangement. A self-phased quadrifilar helix antenna (SP QHA) is composed of two mutually orthogonal BHAs where one of the antenna structures is larger in regards to height and radius than the other structure, i.e.

\[
\begin{align*}
\ell_{1\text{maj}}(p) &= r_{\text{maj}} \cos \left( \frac{2\pi Np}{s_{\text{maj}}} \right) \hat{x} + r_{\text{maj}} \sin \left( \frac{2\pi Np}{s_{\text{maj}}} \right) \hat{y} + p\hat{z} \\
\ell_{2\text{maj}}(p) &= -r_{\text{maj}} \sin \left( \frac{2\pi Np}{s_{\text{maj}}} \right) \hat{x} - r_{\text{maj}} \cos \left( \frac{2\pi Np}{s_{\text{maj}}} \right) \hat{y} + p\hat{z} \\
\ell_{1\text{min}}(p) &= r_{\text{min}} \cos \left( \frac{2\pi Np}{s_{\text{min}}} \right) \hat{x} - r_{\text{min}} \sin \left( \frac{2\pi Np}{s_{\text{min}}} \right) \hat{y} + p\hat{z} \\
\ell_{2\text{min}}(p) &= -r_{\text{min}} \sin \left( \frac{2\pi Np}{s_{\text{min}}} \right) \hat{x} + r_{\text{min}} \cos \left( \frac{2\pi Np}{s_{\text{min}}} \right) \hat{y} + p\hat{z}
\end{align*}
\] (29-32)

for left hand winding, where \(r_{\text{maj}} > r_{\text{min}}\) and \(s_{\text{maj}} > s_{\text{min}}\), see Figure 23. This setup removes the need for phase hybrid components, instead the quadrature phasing relation is generated by the different resonant lengths of the structure. Adjusting the ratios accordingly

\[
\frac{r_{\text{maj}}}{r_{\text{min}}} = \frac{s_{\text{maj}}}{s_{\text{min}}},
\] (33)
the induced current on the major BHA will have a phase lead of $\phi = 45^\circ$ and the minor BHA will instead have a current with a phase lag of $\phi = -45^\circ$, and when fed in parallel, the total field will have a phase quadrature relation. The induced currents are summed due to the superposition principle and the antenna impedance will, theoretically, have no reactive component for the operating frequency and due to the two resonant frequencies of the two separate BHAs, the achieved bandwidth of the SP QHA will be wider than for a QHA using an externally phased arrangement [58].

![Figure 23: Geometry of a self-phased QHA consisting of two orthogonal bifilar helices. The major and minor helices are visualized with a blue and a green line, respectively.](image)

### 4.2.2 Conductor Geometry

When a conductor is fed with a current, the current will run on the surface of the conductor. If the skin depth of the RF wave approaches or exceeds the thickness of the conductor, great ohmic losses due to surface resistance will occur in the transmission lines. The skin depth for a good conductor can be expressed as,

$$
\delta = \sqrt{\frac{2\rho}{\omega \mu}} \sqrt{1 + \left(\omega \varepsilon \rho\right)^2} + \omega \varepsilon \rho \approx \{1 \gg \omega \varepsilon \rho\} \approx \sqrt{\frac{2\rho}{\omega \mu}}
$$

(34)

where $\rho$ and $\varepsilon$ is the resistivity and permittivity of the conductor, respectively. It can be observed that the depth decreases with higher frequency. The skin depth can also be expressed as

$$
\delta = -\frac{1}{k''}, \quad k'' = \text{Im}\{k\}
$$

(35)
i.e. the complex component of the wave number [59]. The ohmic power losses, \( P_s \), are caused by the surface resistance

\[
P_s = I^2 R_s
\]

(36)

where \( I \) is the current running through the conductor and \( R_s \) is the surface resistance,

\[
R_s = \sqrt{\frac{\omega \mu_0 \rho}{2}}.
\]

(37)

The ohmic losses of the system can be neglected if the antenna conductor depth, \( D \), is much greater than the skin depth at relevant frequencies, i.e. \( D \gg \delta \), which needs to be considered in the conductor geometry design.

4.3 Impedance Matching

An antenna system can be simplified and translated to a circuit model consisting of a voltage source, \( V_S \), a source impedance, \( Z_S \), and the antenna load is represented with an impedance, \( Z_L \), as demonstrated in Figure 24. To increase the bandwidth of the system, additional components can be added to the circuit, connected between the source impedance and antenna load impedance, to optimize the power transfer of the process at the antenna load. In other words, the aim is to maximize the transmission and minimize the reflection at the antenna load.

\[
\begin{align*}
\text{Figure 24: Circuit representation of the antenna system.}
\end{align*}
\]

The incident electric fields towards the antenna impedance, \( E_i \), will be divided into two waves when reaching the load; one part reflected fields, \( E_r \), and one part transmitted fields, \( E_t \). The rate of reflected energy in the system can be represented by voltage signals using the components in Figure 24, which is proportional to the intensity of the electric fields;

\[
\frac{E_r}{E_i} = \frac{Z_L - Z_S}{Z_L + Z_S} = \Gamma
\]

(38)

and the transmission of the system will, according to the theory of electric waves at boundaries, be

\[
E_t = E_i + E_r
\]

(39)
and, therefore,
\[ \frac{E_t}{E_i} = \frac{2Z_L}{Z_L + Z_S} = \sigma \]  
(40)

where \( \Gamma \) is called the reflection coefficient and \( \sigma \) is called the transmission coefficient of the system. For a lossless network, which the system in this case is assumed to be, it can be assumed that the input power, \( P_i \), is divided into a reflected part \( P_r \), and a transmitted part, \( P_t \), and the law of conservation of energy says that over the load

\[ P_i = P_t + P_r \Rightarrow E^2_t = E^2_r + E^2_i \]  
(41)

which gives

\[ 1 = \Gamma^2 + \sigma^2 \]  
(42)

and the aim is, therefore, to minimize \( \Gamma \) in order to have a maximized \( \sigma \) over the antenna load.

If the source impedance is modified to have the value of the complex conjugate of the load, \( Z_S = Z_L^* \), Equation 42 will give \( \sigma = 1 \), and the system will, theoretically, transfer all the power from the incident wave to transmission power over the load and the system will be perfectly matched [60]. The antenna load, \( Z_L \), is usually frequency dependent, thus when matching on a certain frequency affects the whole system [44].

Voltage Standing Wave Ratio (VSWR) is a function of the reflection coefficient and indicates the impedance mismatch in the system, where

\[ \text{VSWR} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \]  
(43)

and the aim is to have as low VSWR on the desired frequencies as possible.

On frequencies where the wavelength of the RF waves are of a comparable length to the transmission line between the source load an antenna load, the position of motion for the wave may cause great losses in the system if the reflection at the load end up in counter phase compared to the incoming wave and this relation must, therefore, be taken into account in the calculations, and circuit theory is then abandoned in favor of transmission line theory [61]. The different approaches prevail when the traveling distance of the waves, \( d \), relates to the wavelength accordingly:

\[ d \ll \lambda \quad \Rightarrow \quad \text{Circuit theory} \]  
(44)

\[ d \approx \lambda \quad \Rightarrow \quad \text{Transmission line theory} \]  
(45)

and needs to be taken into account when performing design and calculations.
4.4 Gain and Radiation Pattern

The radiating pattern of an antenna is the body characteristics of the different regions of transmitting, or receiving, compatibility from the antenna structure. As stated in 2.2, the design options regarding antenna design and radiation pattern usually is to either have a narrow, high-gain radiation pattern, or broad, low-gain radiation pattern, and in this thesis, the antenna designs are aimed to have characteristics as the latter.

The antennas will be observed in the far-field region, which is defined to be when the traveling distance, $d$, between the antenna and the nearest interfering object, in relation to the antennas electrical properties, satisfies following conditions:

$$d > \frac{2\ell^2}{\lambda}$$  \hspace{1cm} \text{(46)}

$$d \gg \lambda$$  \hspace{1cm} \text{(47)}

$$d \gg \ell$$  \hspace{1cm} \text{(48)}

where $\ell$ is the largest dimension of the antenna structure, in this case, the helix element length. When these conditions are satisfied, the antenna is assumed to be isolated and free from interference. If interfering objects are placed within these distances to the antenna, either in the reactive field region or the near-field region, see Figure 25, the antenna may couple with the object and the radiation pattern will, in that case, be distorted.

![Field regions of an antenna.](image)

Figure 25: Field regions of an antenna.

The RF waves are propagating in spherical waves from the source, i.e. the antenna. However, when in the far-field region, the curvature angle at a given point negligible and the waves are approximated as planar and the radiation pattern does not change shape with distance [30]. With this, the electromagnetic...
fields do not have components altering in the direction of propagation and the wave satisfies the definition of being in transverse electromagnetic (TEM) mode. A TEM wave propagating along the traveling distance axis, \( \hat{d} \), in free-space with the electric field \( E = E_\phi \hat{\phi} + E_\theta \hat{\theta} \) with alternating components \( E_\phi \) and \( E_\theta \), will have a corresponding magnetic field \( H = \hat{d} \times E / Z_0 \). The power density vector of the wave, Poyntings vector, \( W \), will have following appearance:

\[
W = E \times H^* = \begin{pmatrix} 0 \\ E_\phi \\ E_\theta \end{pmatrix} \times \begin{pmatrix} 1 \\ 0 \\ 0 \end{pmatrix} \times \begin{pmatrix} 0 \\ E_\phi / Z_0 \\ E_\theta / Z_0 \end{pmatrix} = \begin{pmatrix} 0 \\ -E_\theta^* / Z_0 \\ E_\phi^* / Z_0 \end{pmatrix}
\]

\[
= \left( \frac{|E_\phi|^2}{Z_0} + \frac{|E_\theta|^2}{Z_0} \right) \hat{d} = \frac{|E|^2}{Z_0} \hat{d} = W \hat{d}
\]

(49)

i.e. in the far-field region, all the power of the radiated wave will be in the direction of propagation and only the values of the electric field is needed to be identified in order to determine the full characteristics of the wave. The antenna gain, \( G \), relates to the power density via

\[
G(\theta, \phi) = \eta \frac{1}{4\pi} \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} W(\theta, \phi) \sin(\theta) \, d\theta \, d\phi
\]

(50)

where \( \eta \) is the efficiency factor, \( 0 < \eta < 1 \), and the fraction in the expression is the directivity of the antenna.

4.5 Polarization

The polarization of an electromagnetic wave is defined by the orientation of the electric field vector as it varies in time [38]. Assuming a time-harmonic TEM wave propagating in the direction of the \( \hat{d} \)-component, i.e. \( E_d = 0 \), the electric field can be expressed as

\[
E(d, t) = E_\phi e^{i(kd - \omega t + \phi_\phi)} \hat{\phi} + E_\theta e^{i(kd - \omega t + \phi_\theta)} \hat{\theta}
\]

(51)

where the phasor components of the electric fields are included, \( \phi_\phi \) and \( \phi_\theta \) are the initial phases of the wave components, respectively. The \( \hat{\phi} \)-axis is defined to align with the horizontal plane and the \( \hat{\theta} \)-axis is defined to align with the vertical plane. A wave alternating solely along the \( \hat{\phi} \)-axis, i.e. \( E_\theta = 0 \) will have a state of horizontal polarization (HP) while a wave alternating only along the \( \hat{\theta} \)-axis, i.e. \( E_\phi = 0 \) will have vertical polarization (VP). Both of the waves will, seen from a transmitter or receiver, alternate along a line and both states are therefore defined as LP. If the electric field of a radiating object have equal amplitudes on the orthogonal field components and equal initial phase shift, i.e. \( E_\phi = E_\theta \) and \( \phi = |\phi_\phi - \phi_\theta| = 0^\circ \), the resulting wave will, according to the superposition principle, be a summation of the two separate waves and seen from a transmitter or receiver, the electric field will alternate along a line, see Figure 26. This state is also defined as LP, with equal sized horizontal and vertical fields components. However, if the orthogonal component amplitudes are kept equal but there is a phase quadrature relation between the components, i.e. \( \phi = |\phi_\phi - \phi_\theta| = 90^\circ \), the resulting vector will instead rotate circularly along the axis of propagation, see Figure 27. Depending on the phase relation between the
horizontal and vertical components, the total wave will either have an RHCP or an LHCP state, where

\[
E_{\text{RHCP}} = \frac{1}{\sqrt{2}}(E_{\phi} - jE_\theta) \tag{52}
\]
\[
E_{\text{LHCP}} = \frac{1}{\sqrt{2}}(E_{\phi} + jE_\theta) \tag{53}
\]
i.e. circular polarization is the superposition of two orthogonal linear polarized waves in phase quadrature relation. The electric field components can also be expressed as a complex unit with real and imaginary components, i.e.

\[
E_\phi = E_{HP} \cos(\phi_{HP}) + jE_{VP} \sin(\phi_{VP}) \tag{54}
\]
\[
E_\theta = E_{VP} \cos(\phi_{VP}) + jE_{HP} \sin(\phi_{HP}) \tag{55}
\]
where \( E_{HP}, E_{VP} \) is the horizontal and vertical amplitude, respectively. \( \phi_{HP}, \phi_{VP} \) are the associated phase components. By rearranging the electric field components, following expressions for the CP field components can be found

\[
E_{\text{RHCP}} = \frac{1}{\sqrt{2}} \left\{ \left[ E_{HP} \cos(\phi_{HP}) - E_{VP} \sin(\phi_{VP}) \right] - j \left[ E_{HP} \sin(\phi_{HP}) + E_{VP} \cos(\phi_{VP}) \right] \right\} \tag{56}
\]
\[
E_{\text{LHCP}} = \frac{1}{\sqrt{2}} \left\{ \left[ E_{HP} \cos(\phi_{HP}) + E_{VP} \sin(\phi_{VP}) \right] + j \left[ E_{HP} \sin(\phi_{HP}) + E_{VP} \cos(\phi_{VP}) \right] \right\} \tag{57}
\]
and it can be concluded that the CP properties of an electric field can be found if the amplitude and phase of the HP and VP fields are known. The polarization
(a) Electric field with two perpendicular components in phase quadrature.  

(b) Resulting wave due to the superposition principle.  

(c) The total electric field alternating in a circular motion, seen from the receivers end of the wave.

Figure 27: Visualization of a circularly polarized wave.

Powers density for each CP state is defined as

\[ W_{\text{RHCP}} = 10 \log_{10} \left( \frac{E_{\text{RHCP}}^2}{Z_0} \right) \]  \hspace{1cm} (58)  

\[ W_{\text{LHCP}} = 10 \log_{10} \left( \frac{E_{\text{LHCP}}^2}{Z_0} \right) \]  \hspace{1cm} (59)  

And the axial ratio (AR) can be expressed as

\[ \text{AR} = 20 \log_{10} \left( \frac{1 + 10^{-|W_{\text{RHCP}} - W_{\text{LHCP}}|/20}}{1 - 10^{-|W_{\text{RHCP}} - W_{\text{LHCP}}|/20}} \right) \equiv 20 \log_{10} \left( \frac{E_{\text{maj}}}{E_{\text{min}}} \right) \]  \hspace{1cm} (60)  

Where \( E_{\text{maj}} \) and \( E_{\text{min}} \) denotes the major and minor axis of the polarization, respectively, and \( |W_{\text{RHCP}} - W_{\text{LHCP}}| \) indicates the level of cross-polarization. AR indicates the polarization purity for the antenna system and, as stated in 2.2, if there are different polarization states between transmitter and receiver, significant losses can occur. AR varies depending on frequency and radiation angle, and in the majority of cases, the AR cross-section will have a degree of elliptically polarization state with a major and minor axis of the electric field components, see Figure 28. AR follows the behavior of the gain pattern; along the main beam of the antenna, AR will have optimal value, assuming the antenna is correctly designed.
Figure 28: Elliptically polarized electric field with major and minor axis.
5 Design Approach

This sections will describe the antenna design implementation of the project. The design process of the project is visualized with a flowchart in Figure 29 and is described in general terms next.

Initially, with given technical scope, suitable design choice could be identified and theory and data describing the behavior of the antenna structure were identified and mapped. These operations were performed in previous sections, the yellow boxes. When entering the design procedure of the project, visualized with the blue boxes, the presented data, and theory from the previous section was used as the basis when calculating optimal antenna conductor lengths for the two antenna designs. After the dimensions of the antenna conductors were set, simulation models were generated using a full domain time solver software. In the simulation software, different feed arrangements, matching circuits, and conductor geometries were evaluated in an iterative process in order to find the optimal outcome in regards to performance; VSWR, far-field gain characteristics, and AR. When the simulations were performed, relevant data were extracted and saved, and prototypes were thereafter constructed. The antenna prototypes were based on the parameter values of the simulation models, both in regards to electrical and geometrical values. After the construction was done, measurements on the antenna prototypes were performed. With the measurements performed and clear results were obtained, analyses and comparisons of the simulated antenna models and the antenna prototypes could be performed, and conclusions could thereafter be drawn whether the antenna designs could satisfy set requirements. This actions will be presented in Section 8 and Section 11, the green boxes.
5.1 Dimensions for Antenna Elements

There were two defining relations between electrical and geometrical parameters for the QHA which had to be satisfied for the antenna to radiate with desired characteristics on desired frequencies, while still fit inside the allocated spaces. This gave three main conditions:

- Helix radius $r$ and total height $h$ were set in regards to maximum allowed value.

- The normalized dimensions $s_\lambda$ and $r_\lambda$ had to be adjusted correctly so that the antenna dimensions were placed optimally in relation to the shaped conical beam region.
The total helix arm length had to be approximately equal to the desired operating wavelength \( i.e. \ell_{\text{helix}} \approx \sqrt{(2\pi Nr)^2 + h^2 + 2r} \approx \lambda \).

These three set conditions gave a small degree of freedom in regards to the element designs. The placement of the antenna dimensions in relation to the shaped conical beam region, Figure 18, was found by normalizing the radius and pitch distance of the helix arm with the wavelengths of the highest and lowest operating frequency, \( i.e. \lambda_{\text{min}} \) and \( \lambda_{\text{min}} \). By performing this operation, the start- and end point of the coefficient line \( \kappa \) when the QHA varied in frequency could be determined and it could be observed if the normalized dimensions would be placed inside the shaped conical beam region. This line placement could be shifted by altering the radius and height of the helix arm but also by changing number of turns, \( s = h N \).

When the most optimal combination of \( r, s, N, \) and \( \ell_{\text{helix}} \) for the helices were found, the start- and end point of the coefficient lines were compared to the mapped data of the QHA in Figure 19 and Figure 20. Doing so and comparing the line placement to the mapped data, the outcome regarding far-field radiation pattern when the frequency is altering could be predicted.

### 5.1.1 Short Antenna Helix Length

When using a design where the QHA had one turn, \( N = 1 \), the start- and endpoint of the coefficient line \( \kappa \) for the short antenna design got placed inside the shaped conical beam region, see Figure 30. This was desired and a QHA design using those dimensions would, therefore, radiate a hemispherical far-field pattern. By comparing the line placement of the antenna to the mapped data, Figure 31, it could be observed that the antenna dimension coefficient had a similar alignment as \( \kappa = 0.164 \), and a comparable outcome was, therefore, to be expected for the short antenna. In Figure 31, the regions of the MUOS/Legacy wavelengths from Figure 4 are also included.

Given this information, it could be predicted that on the Legacy DL frequencies, the main beam would align with the antenna axis with a broad beam and have a BW of around \( \theta_{\text{BW}} \approx 50^\circ \), and on the Legacy DL/MUOS DL frequencies, the main beam would be kept aligned with the antenna axis and the BW would increase slightly. On the highest frequencies, the MUOS UL, the main beam was expected to be offset from the antenna axis with approximately \( \theta_p \approx 45^\circ \) and there would be a relative gain loss of \( \gamma = 1 \) dB at \( \theta = 0^\circ \) compared to the maximum main beam level. Using these dimensions, the antenna had a resonance frequency at around \( f = 300 \) MHz, \( i.e. \) in the middle of the frequency band, providing advantageous conditions for frequency matching. This was considered the most optimal antenna conductor length and was therefore chosen as the design for the short QHA. The four helices had a left-hand winding in order to generate RHCP.

### 5.1.2 Long Antenna Helix Length

Using one turn, \( N = 1 \), for the long QHA, the antenna structure would alternate within the wanted frequency band. However, the coefficient line would be placed too far to the right in Figure 30 and the far-field radiation pattern would not have a cardioid shape, but instead, have multiple lobes. By altering the geometry and increasing the number of turns to two, \( N = 2 \), the line coefficient
line would have a better placement in the shaped conical beam region. The antenna would alternate within the frequency band, although on a lower frequency compared to the one-turn structure. The coefficient line for the short antenna is placed between $\kappa = 0.83$ and $\kappa = 0.164$, and by comparing the line alignment to the mapped far-field data, the characteristics of the far-field radiation pattern could be predicted. The radiation pattern would have a cardioid shape aligned with the antenna axis at the lower, Legacy DL, frequencies. However, the beam would be narrower compared to the short antenna and BW would be around $\theta_{BW} = 35^\circ$, which also can be predicted by observing the coefficient line placement for those frequencies in Figure 30. In the middle of the frequency band, on the MUOS UL/Legacy UL frequencies, the main beam would still be aligned with the antenna axis and there would be a slight increment in BW. On the highest operating frequencies, MUOS DL, a significant main lobe offset is to be expected with $\gamma = 2$ dB and the offset would be in the order of $\theta_p = 45^\circ$. With these dimensions, the resonance frequency was located at around $f = 240$ MHz which was not optimal from a frequency matching point of view. However, no other options were available in order to make the antenna dimensions fit inside the conical beam zone and this was set as the final dimensions for the long QHA. Same as for the short antenna, the winding direction was left-hand winded.

An antenna design with $N = 1.5$ turns would shift the start- and end point of $\kappa$ to be positioned inside the shaped conical beam region, and the resonance frequency would be at $f = 270$ MHz making the initial conditions more promising than with a QHA design with $N = 2$. However, simulations showed that the beam offset for such design would start at a much lower frequency than for a QHA with $N = 2$ and the level of the gain drop and the displacement position of the main lobe from the antenna axis would have made it difficult to establish a communication link from frequencies $f = 285$ MHz and higher. No mapped measurement data were available for a QHA with $N = 1.5$ and this design option was abandoned.
Figure 30: Antenna dimensions in regards to shaped conical beam zone.

Figure 31: Radiation pattern characteristics with antenna dimensions included.
5.1.3 Antenna Conductor Design

The aim when designing the antennas in the simulation environment was to minimize the differences between the theoretical and practical models in order to make accurate comparisons of the final results later on. In the prototype construction, copper tape was used. The reason for using copper tape for the prototypes was to facilitate the construction phase. Using massive copper pipes as conductors would have resulted in more robust and power resistant antennas but it would have demanded crafts and tools outside the scope of the project, and would also have been significantly more expensive to produce. The cross-section of the conductors, \( w \times D = 2.54 \, \text{cm} \times 25.4 \, \text{µm} \), were therefore set to be \( w \times D = 2.54 \, \text{cm} \times 25.4 \, \text{µm} \) in the simulations since these dimensions were the same as the cross-section for the copper tape later used. Using Equation (34) with value for the copper resistivity set to \( \rho = 16.8 \, \text{nΩm} \) in room temperature, and with the applicable frequencies inserted, it was found that the skin depth would vary in the span of \( 3.34 \, \text{µm} \leq \delta \leq 4.21 \, \text{µm} \) and it could be concluded that \( D > \delta \) for all relevant frequencies. Also, using Equation (37), the surface resistance varied between \( R_s = [4.0-5.0] \, \text{mΩ} \) over relevant frequencies and the current would, therefore, flow along the surface of the conductors and the ohmic losses of the conductors could be neglected.

With all the dimensions and structures set for the conductors, simulation software models of the antennas could be constructed. Initially, a helix line was generated using a parameterization function and used the expressions from Equation (2) for the \( x \), \( y \) and \( z \) axes. Thereafter, a cross-section of the conductor was defined and by using a sweep function, the helix conductor was generated, see Figure 32. As stated previously, the helix conductors were to mimic copper tape and it was, therefore, desired to have a rectangular cross-section.

![Helix design in the simulation software.](image)

Figure 32: Helix design in the simulation software.

When the first helix conductor was completed, the shape was copied and rotated a quarter turn in the azimuth direction for each new helix arm. Identical approaches were used for both QHAs. The conductors for both QHA models were defined to consist of lossy copper, which was available as a predefined
material in the software. Figure 33 shows the final design of the two antenna simulation models. In the software, the antennas were defined to be in a free-space environment.

(a) Simulation model of the short QHA. (b) Simulation model of the long QHA.

Figure 33: Final antenna element design models in the simulation software. The antenna models have different scaling in the figure.

5.2 Feed Design

When evaluating the feed design for the antennas, all of the arrangement techniques presented in Section 4.2.1, i.e. self-phased arrangement, externally 90° phased monofilar feed arrangement, and externally 90° phased bifilar feed arrangement were evaluated in the simulation software and the arrangement delivering the best results was chosen as the final type of feed design.

The self-phased arrangement was rejected to be the final feed design partly due to presumptive construction complications; simulations on the SP QHA model indicated that the optimal performance was when the ratio between the major and minor BHA dimensions was \( \frac{r_{\text{maj}}}{r_{\text{min}}} = \frac{s_{\text{maj}}}{s_{\text{min}}} \approx 1.15 \) and a construction which would suspend the antenna conductors with such small differences in dimensions would have been difficult to build. Since the two BHAs would have needed two separate cylindrical suspensions, the antennas would have a fragile and unstable construction where the separate BHAs could eas-
ily have shortened each other. Also, simulations showed that VSWR and AR results for SP QHA, in general, were worse compared to QHAs using externally phase arrangement. The BW of the SP QHA was narrower and the cross polarization was higher, making the axial ratio worse. The externally phased arrangement showed to be superior regarding frequency range, BW, and polarization purity. The differences in performance between the two externally phased feed alternatives were found to be insignificant, and the 90° phased bifilar feed arrangement was chosen as feed design due to lower complexity; for the 90° phased bifilar feed arrangement, one phase shifter and two matching networks were needed instead of three phase shifters and four matching networks which would have been the case if using a 90° phased monofilar feed arrangement. Using more components will generate more losses, which is to be avoided.

In the simulation model, the two bifilar structures were separated by constructing bridged connections on one of the BHAs where the two structures would have shorted otherwise, see Figure 34. On the top of the BHAs, two feed ports were connected to each bifilar structure. The phase shift on one of the ports was performed in the next step.

![Figure 34: Top and bottom of the short QHA, consisting of two perpendicular, separate BHAs with individual feeding ports.](image)

5.3 Matching Network

For the matching network, the complexity and possibility for practical implementation were taken into account when designing the circuits; the approach was to design a basic network that also would be easy to realize in the construction phase of the project. The operating wavelength was in the order of $\lambda = 1 \text{ m}$, compared to the length of the transmission lines which would be in the order of $d = 2 \text{ cm}$, and Equation 45 in Section 4.3 was satisfied, and circuit theory model was used, i.e. the length of the transmission lines needed not be
considered and possible effects due to these relations were neglected.

To increase the bandwidth of the QHAs, matching networks were connected to each feed port. The voltage source was defined to have an internal impedance of $Z_S = 50 \, \Omega$. L-filters, i.e. components positioned in an L shaped configuration, consisting of two reactive elements were connected between the source and antenna load to generate a complex conjugate matching. The initial values of the components were generated by calculations from a software built-in macro. One of the antenna ports were phase delayed with $\phi = 90^\circ$ to generate the desired circular polarization. By defining the values of the matching components as variables, i.e. capacitance $C_1$ and $C_2$, and inductance $L_1$ and $L_2$, frequency fine-tuning of the network could be performed in the software. Where altering component values, a real-time update of the VSWR curve was generated and could be observed, and by using this hands-on approach, optimization of the matching network was performed. Different filter configuration, as well as cascaded L-filters, were also evaluated briefly. However, no improvement was made regarding the bandwidth performance using other filter configurations, and single L-filters topology was chosen as matching network. Identical approaches were used for both the short and long antenna, resulting in similar circuit models. The matching network circuits are shown in Figure 35 where the BHAs are visualized as the loads $Z_{\text{BHA-1}}$ and $Z_{\text{BHA-2}}$, respectively. Optimal matching for the short antenna was found when $L_1 = 40.0 \, \text{nH}$, $L_2 = 61.5 \, \text{nH}$, $C_1 = 5.48 \, \text{pF}$, and $C_2 = 3.74 \, \text{pF}$. For the long antenna, optimal matching was found when $L_1 = 26.1 \, \text{nH}$, $L_2 = 55.0 \, \text{nH}$, $C_1 = 5.65 \, \text{pF}$, and $C_2 = 1.56 \, \text{pF}$. The bridge connection will make one of the bifilar arms slightly longer, resulting in different impedance values for the matching. The bifilar arms will have equal lengths for the prototype since as bridge connection will be used. Therefore, i.e. $L_1 = L_2$ and $C_1 = C_2$ for the prototypes.
(a) Matching circuit arrangement representation for the short antenna model.

(b) Matching circuit arrangement representation for the long antenna model.

Figure 35: Matching circuit arrangements.
6 Construction of Prototypes

The construction of the antenna prototypes was based on the geometrical dimensions calculated from presented theory, and the matching network, with associated component values, were based on the results produced in the simulations. The antenna simulation models were designed to facilitate the prototype construction phase, as underlined previously in this report. Cylindrical cardboard tubes were used as suspension frames for the antenna prototypes. Copper tape, with the dimensions $w \times D = 2.54 \text{ cm} \times 25.4 \text{ µm}$, was glued on to the cardboard tubes in an ascending spiral shape with a 45° degree separation in the azimuth direction between the conductors, creating the four helical arms of the antennas, see Figure 36.

\begin{figure}[h]
\centering
\includegraphics[width=0.5\textwidth]{antenna_prototypes_construction.jpg}
\caption{Antenna prototypes under construction.}
\end{figure}

On the bottom of the antennas, the opposing helices were connected using isolated conductors, and the connection between the helices were soldered together. The connectors could overlap without the bifilar structures shorting out by using isolated conductors and the need to use the bridged connection, as used in the simulations, was removed, see Figure 37. Figure 38 shows the top of the long antenna, with the hybrid phase shifter connected and the L-filter matching network visible. The input signal was split into two components with a relative phase difference of $\phi = 90^\circ$, feeding each of the bifilar arms. Tunable capacitors were used and the inductors were air winded on a plastic frame. By compressing or expanding the pitch distances of the coil windings, the induction could be changed, making all the components tunable.
Figure 37: Bottom of the short QHA where the isolated conductors connects the opposing helix arms.

Figure 38: Feed arrangement of the long antenna with L-filters connected to each bifilar antenna arm and a 90° hybrid phase shifter connected.

With this setup, the difference in length between the two bifilar structures were assumed to be negligible and therefore $L_1 = L_2 = L$ and $C_1 = C_2 = C$ were set when constructing the prototypes. The initial components values used were $L = 52$ nH, $C = 5$ pF for the short antenna, and $L = 45$ nH, $C = 4.5$ pF for the long antenna.
7 Measurements Setup

In this section, the general setups and procedures when performing measurements on the antenna prototypes will be presented. The schematics of the systems, as well as the method of operation for the measurements, will be presented.

7.1 Frequency Response Measurements Setup

For system coefficients and VSWR measurements, a Network Analyzer was used to measure the frequency response of the antenna systems. The setup of the Network Analyzer is demonstrated in Figure 39 where the antenna under test (AUT) is connected to the system and is seen as an impedance.

![Figure 39: Simplified block circuit of a network analyzer setup. With current settings, reflection coefficient is measured.](image)

With a frequency sweep source acting over frequencies of interest, phase and magnitude of both reflected and transmitted antenna signals can be obtained. The directional couplers will register the measured levels in a one-way direction where the reference channel will obtain the value of the waves going into the circuit and the reflection channel and transmission channel will register the data after the waves have been reflected and transmitted from the load, respectively. Using these channel inputs, the network analyzer calculates the reflection coefficient, \( \Gamma \), and transmission coefficient, \( \sigma \), by comparing the data reflection measurement channel and transmission measurement channel from the load to the reference channel data.
Before the coefficient measurements were performed, the network analyzer was measured with an open load and short load in order to ensure that system was correctly calibrated. During the measurement session, the antennas and analyzer were placed where no sources of electrically interfering objects were present, and measurements were thereafter performed over the relevant frequency band for both reflection and transmission coefficients. Further analyses and calculations based on the measurement data were later performed using the equations presented in Section 4.3.

7.2 Far-Field Measurements Setup

The far field of the antennas was measured on two different occasions using two different setups, called slanting field measurements and direct illumination measurements. Below follows the setup and schematics of these two measurement types.

7.2.1 Slanting Field Measurements

On the first measurement session, a technique called slanting field measurements was used. Slanting field range measurements uses the ground reflection and adjust the relative height relation between sender and receiver to generate a positive interference between the direct and reflected waves. This in order to obtain a uniform amplitude and phase distribution over the AUT. If the AUT is mounted on the receiving end and is illuminated with RF waves from a transmitting antenna, the power received by the AUT, $G_{AUT}$ [dBm], can be measured. By thereafter replacing the AUT with a standard gain antenna (SGA) with known antenna gain, $G_{SGA}$ [dBi], at the receiving end of the system and repeat the procedure, power received by the SGA, $G_{SGA}$ [dBm], will be found. The sought gain for the AUT, $G_{AUT}$ [dBi], can thereafter be calculated by using gain comparison,

$$G_{AUT} [\text{dBi}] = G_{AUT} [\text{dBm}] - G_{SGA} [\text{dBm}] + G_{SGA} [\text{dBi}]$$

(61)

where dBi is decibels relative to an isotropic antenna and states the gain of the AUT compared to a theoretical source which radiates perfectly omnidirectional, and dBm is power ratio in decibels of the measured power referenced to 1 mW [62]. The measurement facility and block schematics of the measurement system is shown in Figure 40 and Figure 41, respectively.
Figure 40: Measurement facility of the first far-field measurement occasion. Photo courtesy of Combitech AB.

Figure 41: Setup and block circuit of the slanting field measurements.

During the measurements, the AUTs and SGA were mounted on a rotatable mast with the antenna axis along the horizon. The antennas were illuminated with RF waves from the transmitter antennas and the mast was rotated 360° during a test cycle, and the received power levels were registered for the azimuth angles. The process was repeated with different states of polarization radiating from the transmitting antenna. The AUTs and SGA were illuminated with VP, HP, LHCP, and RHCP polarized waves, and the full turn far-field characteristics could, therefore, be obtained from the measurements. The total gain for the AUT could thereafter be calculated by adding the results from the HP
illumination, $G_{AUT-VP}$, and from the VP illumination, $G_{AUT-VP}$,

$$G_{AUT} = G_{AUT-HP} + G_{AUT-VP},$$

and using the data from the measurements CP levels and AR could be calculated using the expressions in Section 4.5.

### 7.2.2 Direct Illumination Measurements

The second far-field measurements were performed using direct illumination. The AUT was mounted with the antenna axis facing the axis of the illuminating antenna, an AMHA radiating RHCP waves. The system was however tilted; the AMHA was tilted upwards and was positioned at a lower altitude compared to the and the AUT which was tilted downwards from a high altitude. Both suspensions had equal tilt angles $\theta_m$ in order to maintain said axis alignment of the AUT and AMHA, see Figure 42. This setup was used in order to have a region of free air behind the AUTs, and therefore avoid unwanted reflections from nearby objects.

![Figure 42: Measurement setup of the second far-field measurement occasion using direct illumination.](image-url)

The AUT was illuminated with RHCP waves from the AMHA. The antenna suspension could rotate a full turn along the relative altitude angle and full far-field characteristics of the AUT could, therefore, be measured. The measurement facility setup is shown in Figure 43. Only far-field gain measurements were performed during this session.
The reason for performing two different far-field measurements was due to the fact that the results from the first measurement session, where slanting field measurements was used, indicated an offset in regards to gain. The second measurement, direct illumination measurements, was performed after the prototypes had been surveyed and all the connections and crossings on the had been re-soldered as a troubleshooting measure. The direct illumination measurements was a scaled-down measurement where only one polarization state was measured in order to verify that the deviations from the first session were caused due to construction errors.
8 Results and Analysis

This section will present the results gathered from this project, both from the simulations and from the antenna preliminary measurements. Initially, the simulation results will be analyzed and it will be determined whether the antennas satisfy the requirements set. Thereafter, the results from both simulations and prototype measurements will be combined when possible in order to compare and analyze possible similarities and deviations from the different outcomes. In the end of the analysis of the antenna designs, a summary of the results will be presented. The results of the short antenna will initially be presented, followed by the results from the long antenna. Since both antennas have the same design, the same setup, and similar physical rules apply, the analysis of the long antenna results will, in general, be kept slightly shorter. Each measured performance will be presented chronologically.

8.1 Short Antenna

Firstly, the results for the short antenna, i.e. the antenna design with cylindrical size constraints maximum radius $r = 10\, \text{cm}$ and maximum height $h = 50\, \text{cm}$ is presented. In Table 1, the values of the geometrical and electrical parameters used for this design in the simulations, and later in the construction, are shown. Figure 44 shows an image from the simulation software when the antenna structure is radiating on the frequency $f = 300\, \text{MHz}$ in free space. The relative radiated effect from the antenna in the simulation software is visualized with a scale according to the light spectrum where the maximum and minimum intensity is shown with a red and blue color, respectively. Below follows the results and analysis for VSWR, far-field gain, and AR.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radius</td>
<td>$r$</td>
<td>10 cm</td>
</tr>
<tr>
<td>Height</td>
<td>$h$</td>
<td>50 cm</td>
</tr>
<tr>
<td>Number of turns</td>
<td>$N$</td>
<td>1</td>
</tr>
<tr>
<td>Capacitance</td>
<td>$C_1$</td>
<td>5.48 pF</td>
</tr>
<tr>
<td>Capacitance</td>
<td>$C_2$</td>
<td>3.74 pF</td>
</tr>
<tr>
<td>Inductance 1</td>
<td>$L_1$</td>
<td>40.0 nH</td>
</tr>
<tr>
<td>Inductance 2</td>
<td>$L_2$</td>
<td>61.5 nH</td>
</tr>
<tr>
<td>Conductor width</td>
<td>$w$</td>
<td>2.54 cm</td>
</tr>
<tr>
<td>Conductor depth</td>
<td>$D$</td>
<td>25 µm</td>
</tr>
</tbody>
</table>

Table 1: Parameter list for short QHA.
8.1.1 VSWR

The results for the reflection in the antenna system for the short antenna is presented in Figure 45 below. By analyzing the graphs, we can conclude that the VSWR curve for the short antenna is significantly improved with the matching network, according to the simulated results. The resonance frequency for the antenna is around $f = 310$ MHz and when performing complex conjugate matching and fine-tune the system, $\text{VSWR} \leq 2$ on the desired frequencies and meets the requirements.

![VSWR results for the short antenna; matched, unmatched and measured system.](image)

**Figure 45:** VSWR results for the short antenna; matched, unmatched and measured system.

By analyzing the measured result of the reflections, we can see that VSWR behaves in a similar manner to the simulations and $\text{VSWR} \leq 2$ for the whole relevant frequency band. However, when measuring the reflection and transmission coefficients of the system, demonstrated in Figure 46, we observe that the transmission coefficient has a lower value than expected and that the insertion losses of the system are significant over most relevant frequencies. These losses might have arisen from insufficient soldering between the antenna conductors or leakage in the matching network.
8.1.2 Far-Field Gain

Figure 47 shows the far-field gain of the short antenna as a function of frequency and elevation angle for the simulated antenna model, seen from two $\varphi = 90^\circ$ azimuthally separated angles. By observing Figure 47, it can be seen that the far-field behaves according to the theory presented Section 4.2; on the lower-to-mid frequencies, there is a broad beam radiation with the main lobe aligned with the antenna axis and on the highest operating frequencies, the main lobe will be slightly offset.

Observing the far-field radiation in polar plot format on set frequencies will facilitate the process of extracting the numerical values of the results. Figure 48 shows the far-field on set frequencies $f = [240, 280, 320, 370]$ MHz. By analyzing the sub figures, it can be seen that the antenna has a broad beam and the BW is around $\theta_{BW} = 50^\circ$ on the low-to-mid operating frequency, and the main beam gain is around $G_i(\theta = 0^\circ, \varphi) = 4$ dBi. When the frequency is increased, the gain offset will be more palpable and the gain level at $\theta = 0^\circ$ will decrease. On MUOS UL frequencies, the gain offset will be around $\theta_p = 50^\circ$ and $\gamma = 2$ dB. On the highest operating frequency, $f = 380$ MHz, the gain is $G_i(\theta = 0^\circ, \varphi) = 1.8$ dBi and the antenna design, therefore, meets the gain requirement for $\theta = 0^\circ$. As seen in Figure 44, the antenna structure will generate nulls where the signal is almost canceled out completely. These nulls occur at around $\theta \approx 90^\circ$.

Comparing the results from the two cut section angles, $\varphi = 0^\circ$ and $\varphi = 90^\circ$, it can be observed that if a null occurs, the gain level decays rapidly close to $\theta = 90^\circ$. However, when nulls do not occur, $G_i(\theta = 65^\circ, \varphi) \geq 0$ dBi for all the
frequencies and thus the antenna meets the requirements.

Figure 47: Simulated 3D results for the short antenna.
Figure 48: Simulated polar plot results of the far-field radiation for the short antenna.
The results from slanting field measurements are shown in Figure 49. The graph shows a radiation pattern similar to the simulations; a broad beam where maximum gain is aligned with the antenna on the lower frequency and with a gain offset on the higher frequencies. The simulated and measured results for the far-field behave in a similar manner, overall, but differ significantly in regards to gain level. By observing the far-field polar plot of the radiation field, Figure 50, demonstrating both the simulated and measured results on set frequencies \( f = [240, 280, 320, 370] \) MHz, we can see that the antenna, in general, has a noticeably lower gain value in the order of 10 to 15 dBi. These deviations might have occurred due to the losses in the matching network or due to a sub-optimal construction of the prototype. Also, the maximum gain level for the field itself differs 10 dBi, and the radiation patterns at \( f = 240 \) MHz and \( f = 370 \) MHz indicates the presence of measurement errors.

Figure 49: Measured far-field 3D plot for the short antenna.
\( G_{AUT}(\theta, \phi) \) measured
\( G_i(\theta, \phi = 0^\circ) \) simulated
\( G_i(\theta, \phi = 90^\circ) \) simulated

\( \theta / \text{Degree} \)
\( \text{Gain/dBi} \)

(a) \( f = 240 \text{ MHz} \)

(b) \( f = 280 \text{ MHz} \)
Figure 50: Far-field gain radiation for the short antenna, simulated and measured results combined.
New far-field measurements were performed using the second setup, direct illumination, displayed in Figure 42 in Section 7.2, where the antenna is directly illuminated with RHCP. Before measurements were performed, the connections between the conductors were re-soldered in order to ensure a good connection. The normalized 3D results are displayed in Figure 51 and on set frequencies $f = [243, 284, 311, 366]$ MHz in Figure 52. The new results show an outcome which resembles the simulated values much more closely, both in regards to shape of the far-field radiation pattern and relative gain level over the frequencies. The new measured indicates an even level for the far-field radiation, and a correct representation of the antenna radiation compared to the simulated values, once again indicating measurement errors from the slanting field measurement.

Figure 51: New antenna prototype measurements for the short antenna.
Figure 52: Polar plot results for the new antenna prototype measurements of the short antenna.
8.1.3 Polarization

In the polarization results, the simulated and measured data are combined and presented on frequencies $f = [240, 280, 320, 370]$ MHz, displayed in Figure 53 below. As in the case of far-field gain results, the simulated results have data from two azimuth angles with a relative $90^\circ$ separation and the measurements from the antenna prototype only have data from one azimuth angle, and all the results have a full $360^\circ$ sweep in altitude angle. By observing the figures, it can be seen that for both simulated results, the dotted graphs, the antenna have a high polarization purity with RHCP signal over a broad angle and a corresponding low AR. Comparing the polarization plots to the far-field polar plots, Figure 48, it can be seen that the spikes in the AR graphs occur whenever passing a null, otherwise, the AR of the antenna is kept low. When not passing a null, $\text{AR}(\theta, \phi) < 5 \text{ dB}$ for $-65^\circ < \theta < 65^\circ$ which was the requirement set. The measurement data was gathered from the slanting field measurement and corresponds to Figure 49 and Figure 50 from the previous subsection. From the measurements, it can be concluded that the AR and polarization for the antenna prototype deviate from the simulated results in regards to purity. However, the measurements indicate that there is a dominant RHCP component radiating from the antenna and the general tendencies of the measurements resemble the simulated values. Comparing the sub figures in 53 to 50, it appears that the far-field measurements which resembled the simulated values most closely, i.e. on frequencies $f = 280$ MHz and $f = 320$ MHz, also resembles the polarization curves more accurately, once again indicating that measurement errors are the causes of this deviations.
\begin{align*}
\text{AR}(\theta, \phi) \text{ measured} & \quad \text{AR}(\theta, \phi = 0^\circ) \text{ simulated} \\
\text{RHCP}(\theta, \phi) \text{ measured} & \quad \text{RHCP}(\theta, \phi = 0^\circ) \text{ simulated} \\
\text{LHCP}(\theta, \phi) \text{ measured} & \quad \text{LHCP}(\theta, \phi = 0^\circ) \text{ simulated} \\
\text{AR}(\theta, \phi) \text{ measured} & \quad \text{AR}(\theta, \phi = 90^\circ) \text{ simulated} \\
\text{RHCP}(\theta, \phi) \text{ measured} & \quad \text{RHCP}(\theta, \phi = 90^\circ) \text{ simulated} \\
\text{LHCP}(\theta, \phi) \text{ measured} & \quad \text{LHCP}(\theta, \phi = 90^\circ) \text{ simulated} 
\end{align*}

(a) $f = 240$ MHz

(b) $f = 280$ MHz
Figure 53: Axial ratio, co and cross polarization for the short antenna on set frequencies.
8.1.4 Summary of Results for the Short Antenna

The results for the short antenna is compiled in Table 2 below. The results from the simulations are consistent with the theory presented in Section 4 and the simulation model antenna can meet all the requirements set.

<table>
<thead>
<tr>
<th>Requirements</th>
<th>Goal met?</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency ranges</td>
<td></td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 244-270 MHz</td>
<td>Yes</td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 292-320 MHz</td>
<td>Yes</td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 360-380 MHz</td>
<td>Yes</td>
</tr>
<tr>
<td>Far-field gain within the frequency ranges</td>
<td></td>
</tr>
<tr>
<td>( G_i(\theta, \varphi) \geq 1.5 \text{ dBi at } \theta = 0^\circ )</td>
<td>Yes</td>
</tr>
<tr>
<td>( G_i(\theta, \varphi) \geq 0 \text{ dBi at } \theta = \pm 65^\circ )</td>
<td>Yes</td>
</tr>
<tr>
<td>RHCP within the frequency ranges</td>
<td></td>
</tr>
<tr>
<td>( \text{AR}(\theta, \varphi) &lt; 5 \text{ dB when } -65^\circ &lt; \theta &lt; 65^\circ )</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 2: Results from the simulations for the short antenna design.

Due to the results from antenna prototype measurements and its similarities to the simulation results, there are enough reasons to conclude that the simulation results are accurate, and the presented theory and design procedures were correctly handled. Improvements on the prototypes can be performed in order to optimize the antenna performance. This is, however, not within the scope of this thesis.
8.2 Long Antenna

This subsection will present the relevant results for the long antenna, i.e. the antenna with the size constrains \( r = 5 \text{ cm} \) and \( h = 90 \text{ cm} \). The parameters used in the simulations and for the antenna prototype are presented in Table 1. Figure 54 shows an image of the antenna radiating in a free-space environment at \( f = 280 \text{ MHz} \) in the simulation software.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radius</td>
<td>( r )</td>
<td>5 cm</td>
</tr>
<tr>
<td>Height</td>
<td>( h )</td>
<td>90 cm</td>
</tr>
<tr>
<td>Number of turns</td>
<td>( N )</td>
<td>2</td>
</tr>
<tr>
<td>Capacitance 1</td>
<td>( C_1 )</td>
<td>1.56 pF</td>
</tr>
<tr>
<td>Capacitance 2</td>
<td>( C_2 )</td>
<td>5.65 pF</td>
</tr>
<tr>
<td>Inductance 1</td>
<td>( L_1 )</td>
<td>26.1 nH</td>
</tr>
<tr>
<td>Inductance 2</td>
<td>( L_2 )</td>
<td>55.0 nH</td>
</tr>
<tr>
<td>Conductor width</td>
<td>( w )</td>
<td>2.54 cm</td>
</tr>
<tr>
<td>Conductor depth</td>
<td>( D )</td>
<td>25.4 µm</td>
</tr>
</tbody>
</table>

Table 3: Parameter list for short long.

8.2.1 VSWR

The VSWR results is visualized in Figure 55. Observing the figure, we can see that the VSWR is significantly improved by the matching network but does not go below the desired value of \( \text{VSWR} \leq 2 \) for all relevant frequencies, which was to be expected from the design approach analysis in Section 5. It can be observed that the resonating frequency for the antenna is around \( f = 240 \text{ MHz} \) and this made it hard to suppress reflections on the higher frequencies without generating large reflections on the lower frequencies. When tuning the matching network, the focus was to achieve a good VSWR for the MUOS frequencies while Legacy had a lower priority. This was based on the higher transmission capacity the MUOS system provides compared to the Legacy system. For the MUOS uplink frequencies, \( \text{VSWR} \approx 3 \), which means that half of the delivered power is lost in matching losses. For the DL frequencies, \( \text{VSWR} \leq 2 \) and the requirement is locally met.

Figure 54: Long QHA radiating at \( f = 280 \text{ MHz} \)
For the measured results, the VSWR curve is performing better on the higher frequencies, as in the simulations. The plot of the reflection and transmission coefficients, Figure 56, shows that there are significant insertion losses in general for the antenna. The matching network, which was used for the slanting field measurements, was broken during transportation to direct illumination measurements session and a new matching network was, therefore, constructed. The coefficients of the new matching network are presented in Figure 57 and it can be concluded that the new networks indicated on a superior performance and that the underwhelming performance of the previous matching networks was due to erroneous construction and not due to the design itself.
Figure 56: Reflection and transmission coefficient of the measured long antenna prototype.

Figure 57: Results of the reflection and transmission coefficient of the new matching network for long antenna system used for the direct illumination measurements.
8.2.2 Far-Field Gain

The simulation setup for the long antenna is identical to the setup used for the short antenna. For the long antenna far-field gain results, the simulated results align with the presented theory; the main beam is along the antenna axis on the lower frequencies and the beam gets offset on higher frequencies, see Figure 58. The offset beam is more prominent for higher frequencies compared to the short antenna, which the theory presented in Section 4 predicted. The polar plot representation of the results is presented in Figure 59. Comparing to the results for the short antenna, Figure 48, we can see that the main lobe for the long antenna is more narrow and the back-lobe beam is more apparent compared to the results the short antenna. This results in an overall lower gain level for the long antenna. On the lower Legacy DL frequencies, $\theta_{BW} \approx 40^\circ$ and on the MOUS UL frequencies $\theta_{BW} \approx 65^\circ$, as predicted from Figure 31. On the MOUS DL frequencies, $\theta_p \approx 70^\circ$ and $\gamma \approx 3$ dB. In dBi values, on the Legacy DL frequencies $G_i(\theta, \varphi = 0^\circ) \approx 3.5$ dBi and on the MUOS UL frequencies $G_i(\theta, \varphi = 0^\circ) \approx 1.6$ dBi and the requirement $G_i(\theta, \varphi = 0^\circ) \geq \gamma$ dB is satisfied for all frequencies. As previously stated, the beam is, in general, more narrow shaped compared to the short antenna and the $G_i(\theta, \varphi = 65^\circ)$ requirement cannot be satisfied for all frequencies. However, the MUOS UL and DL frequencies satisfies set requirements.

The outcome from the slanting field measurements on the antenna prototype shows similar radiation pattern characteristics as in the simulated results, see Figure 60. However, the gain level is significantly lower compared to the simulations. Observing the polar plot representation, Figure 59, we can see that there is a relative difference in gain in the order of 20 dBi difference between simulated and measured results across all frequencies. This might partially be caused by the losses in the matching network but also by e.g. bad connections between the antenna conductors. The results on the $f = 370$ MHz show no expected outcome, especially compared to the results on the other frequency ports, indicating some sort of measurement error.

The results from the measurements on the long antenna using the second measurement setup, direct illumination, is presented as 3D plot in Figure 62 and as polar plots in Figure 63 on the set frequencies $f = [256, 272, 311, 366]$ MHz. As in the case of the short antenna, the connections between the conductors were re-soldered. Also, the new matching network with the results presented in Figure 57 was used. Observing the figures, it can be seen that the antenna shows a much more consistent behavior over all the relevant frequencies and the results resembles the simulated values much more closely compared to the results from the slanting field measurements, indicating that bad construction was one of the main reasons for the deviating results from the first measurement session.
Figure 58: Simulated 3D results for the long antenna.
Figure 59: Simulated polar plot results of the far-field radiation for the long antenna.
Figure 60: Measured far-field 3D plot for the long antenna.
\( \theta \) / Degree
Gain/dBi
\( \text{GAUT}(\theta, \phi) \) measured
\( \text{Gi}(\theta, \phi = 0^\circ) \) simulated
\( \text{Gi}(\theta, \phi = 90^\circ) \) simulated
0° ±180° 90° −90° −45° −135° 135° −105° 105° 165° −165° 15° −75° 75° −30° −120° 60° −60° 120° −150° 150° −50 −30 −10 10

(a) \( f = 240 \text{ MHz} \)

(b) \( f = 280 \text{ MHz} \)
Figure 61: Far-field gain radiation for the long antenna, simulated and measured results.
Figure 62: New antenna prototype measurements results for the long antenna.

Figure 63: Polar plot results for the new antenna prototype measurements of the long antenna.
8.2.3 Polarization

For the long antenna, an identical setup was used as in for the short antenna results. The measured results are from the slanting field measurements. The polarization for the antenna is dominated by the RHCP component and the purity of the AR, therefore, achieves a high purity across a wide span and for all frequencies along the main beam, see Figure 64. \( \text{AR}(\theta, \varphi) < 5 \, \text{dB} \) for \(-65^\circ \leq \theta \leq 65^\circ\) is kept and the requirements set are, therefore, satisfied. The results from the antenna prototype measurements correlate with the simulations where there also is a dominant RHCP component, but with a relatively worse polarization purity compared to the simulated case. As previously stated, construction errors might be the cause of these deviations.
AR(θ, ϕ)/dB
RHCP(θ, ϕ)/dBi
LHCP(θ, ϕ)/dBi

(a) f = 240 MHz

(b) f = 280 MHz
Figure 64: Axial ratio, co- and cross polarization for the long antenna on set frequencies.
8.2.4 Summary of Results for the Long Antenna

The simulated results show that there will be difficult to establish a good communication link on the lower frequencies, i.e. Legacy UL and DL, due to the losses in VSWR when using the long antenna design. The radiation pattern is in general narrower compared to the short antenna but satisfied the requirements on the MUOS links. This antenna does not satisfy all of the requirements set. However, the antenna design can still be useful if the intention is to solely communicate via the MUOS frequencies. On the MUOS frequencies, the losses from VSWR will be tolerable and the far-field will have a broader beam and therefore a better AR and a decent communication link can, therefore, be established. The results from the simulations are summarized in Table 4 below.

<table>
<thead>
<tr>
<th>Requirements</th>
<th>Goal met?</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency ranges</td>
<td></td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 244-270 MHz</td>
<td>No</td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 292-320 MHz</td>
<td>No</td>
</tr>
<tr>
<td>VSWR ≤ 2 for f = 360-380 MHz</td>
<td>Yes</td>
</tr>
<tr>
<td>Far-field gain within the frequency ranges</td>
<td></td>
</tr>
<tr>
<td>$G_i(\theta,\varphi) \geq 1.5$ dBi at $\theta = 0^\circ$</td>
<td>Yes</td>
</tr>
<tr>
<td>$G_i(\theta,\varphi) \geq 0$ dBi at $\theta = \pm 65^\circ$</td>
<td>Partially</td>
</tr>
<tr>
<td>RHCP within the frequency ranges</td>
<td></td>
</tr>
<tr>
<td>$\text{AR}(\theta,\varphi) &lt; 5$ dB when $-65^\circ &lt; \theta &lt; 65^\circ$</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 4: Results from the simulations for the long antenna design.

The antenna prototype indicates that there were significant insertion losses caused by the matching network. However, with a new matching network and improved connections between the conductors, it was showed that the results correspond well to the simulated values.
9 Setup for the Submarine Simulations

Simulations in a scenario where the designed antennas were mounted on a submarine in a surface mode were performed. This in order to see how the antenna systems would behave in a more real-life situation, compared to the case where the antennas were simulated in the free-space environment. The parameter values for the antennas, geometrical and electrical, were unchanged from when performing the free-space simulations. The dimensions of the ship were based on an existing submarine design where a foldable antenna mast structure could be a conceivable alternative for the V/UHF SatCom system. The submarine hull had a total length of 10.45 m and the main radius of the ship was set to 1.7 m. The hull was defined to be hollow with a thickness set to be 3 cm and was positioned in a cuboid of water, submerged 5 cm below the surface, as if the submarine was in a communication mode. The radomes covering the antennas consisted of epoxy and were defined to have a thickness of 1.0 cm and there was a 1.5 mm air gap separating the antennas from the radomes. The antenna mast was 1 m long for the short antenna setup and 25 cm long for the long antenna setup. This setup is visualized in Figure 65. All the materials used for the simulation model, with associated electrical and magnetic properties, were predefined in the software material library.

![Short antenna.](a)

![Long antenna.](b)

**Figure 65:** Antennas highlighted inside the radomes.

The boundaries in the simulation environment were defined as touching the sides and bottom of the water cuboid, and to have open space boundaries above the surface. By having these settings, the simulation software would interpret the scenario as if there was a semi-infinite region of water downwards while there would be perceived to be free-space environment upwardly. These boundaries were set in order to imitate a situation where the submarine was placed in an open position on the sea. The full simulation environments for the antennas is demonstrated in Figure 66, where the boundary boxes of the simulations are included.
(a) Short antenna simulation environment.

(b) Long antenna simulation environment.

Figure 66: Submarine simulation environment with boundary box included.
10 Results and Analysis of Submarine System Simulations

This section will present the results of the antenna performance when the antennas were mounted inside the radome and simulated in an environment which mimics a submarine in communication mode on an open sea. Comparisons between the outcome from the simulations and the ideal free-space model results will be drawn. Similarities or deviations will be discussed and briefly analyzed.

10.1 Short Antenna

The submarine setup when the antenna is placed inside the radome, operating on the frequency \( f = 320 \text{ MHz} \) is visualized in Figure 67. As seen, the far-field still has the wanted hemispherical characteristic, as in the free-space situation. Below follows the results and analysis of VSWR and far-field gain from the simulations for the short antenna.

![Figure 67: Short antenna radiating at \( f = 320 \text{ MHz} \).](image)

10.1.1 VSWR

The results for VSWR from the full submarine simulation is presented in Figure 68. Observing the figure, we can see that the antenna behaves in a similar manner in the submarine environment simulation as when it is simulated in free-space but with a frequency lag relation compared to the ideal case. This frequency shift probably arises due to coupling with the covering radome and the antenna mast.
10.1.2 Far-Field Gain

The results of the submarine environment simulations indicate that the far-fields gain pattern of the short antenna has similar characteristics compared to when the antenna is placed in a free-space environment, which can be observed by comparing Figure 69 to Figure 47. The main radiation beam is aligned with the antenna axis on the lower frequencies and has an offset from the antenna axis of about $\theta_p \approx 40^\circ$ on the MUOS UL frequencies in both cases. The frequency monitor polar plots, Figure 70, show that the gain, in general, is in the same order of magnitude as the free-space situation, Figure 48, but with some fringing in the fields occurs. This is presumably due to reflections and near-field effects in form of coupling with the radome and mast. The orientation of the submarine has a notable effect on the far-field on the highest frequencies, which can be observed by comparing the two sub-figures in Figure 69. On the free-space simulations, the results from the two azimuth angles had near identical far-field characteristics. In this case, however, when observing the fields parallel to the ship’s alignment, $\varphi = 90^\circ$, the beam has a more omnidirectional coverage compared to the fields perpendicular to the ship orientation, $\varphi = 0^\circ$, where the gain offset and angle between the antenna axis and maximum lobe is more palpable.
Figure 69: Far-field 3D plot for the short antenna.

(a) $\varphi = 0^\circ$

(b) $\varphi = 90^\circ$
Figure 70: Simulated polar plot results for the far-field radiation on set frequency monitors from perpendicular azimuth angles for the short antenna.
10.2 Long Antenna

Figure 71 shows the far field pattern when the antenna is simulated in the submarine environment, radiating at \( f = 280 \) MHz. As in the case of the short antenna, the far-field has a hemispherical characteristic. Below follows the results and analysis of the VSWR and far-field gain from the simulations for the long antenna in the submarine environment.

![Figure 71: Long antenna radiating at \( f = 280 \) MHz.](image)

10.2.1 VSWR

The VSWR is presented in Figure 72 together with the results from the matched results from the free-space simulations. Comparing the results, we can see that similar results are obtained. As with the short antenna, a frequency shift occurs compared to the free space model. The frequency shift has, however, a lead relation compared to the free-space model, i.e. the long and short antenna have different frequency shift relations compared to each other. These differences in phase relation might arise due to the differences in proximity to the hull which causes different coupling between the cases.
10.2.2 Far-Field Gain

The results from the far-field simulations, Figure 58, shows a clear resemblance to the free-space simulations, Figure 73, with the main beam centered around the axis on the lower frequencies and gain offset at the highest operating frequencies. The gain offset has a more significant drop compared to the free-space results where the MUOS UL frequencies have a gain drop of $\gamma = 4$ dB and the offset angle $\theta_p = 55^\circ$. In general, however, the levels are kept in the same order of magnitude as in the ideal case. As in the case of the short antenna, the hull alignment seems to affect the far-field pattern at the higher frequencies which can be seen by observing the polar plots of the results, displayed in Figure 74.
Figure 73: Far-field 3D plot for the long antenna.
Figure 74: Simulated polar plot results for the far-field radiation on set frequency monitors from perpendicular azimuth angles for the long antenna.
10.3 Summary of Results

Both antennas show a similar behavior when placed in the submarine environment as when in free space. Minor changes such as fringing fields, frequency drifts, and local gain drops can be observed from the outcome. This might have been caused by propagation near-field effects and conductive coupling to the surroundings. However, the general gain levels and patterns clearly coincide with the results from the free-space simulations presented in Section 8. AR results produced showed anomalous values and were omitted. This will be discussed in the next section.
11 Discussion

This section will discuss possible sources of error, the credibility of the produced results, and other ambiguities which have arisen during this project execution. Initially, the results produced from the simulation will be discussed and thereafter, the measurement procedures and the deviating results from the prototypes will be discussed.

11.1 Simulations

As stated throughout the report, efforts were made to have a simulation model which would resemble the antenna prototypes as close as possible. The order of magnitudes for the antenna dimensions differed significantly. The largest dimension was in the order of 1 m and the smallest dimension was in the order of 30 µm. This relative large differences in order of magnitude showed to be difficulties in the modeling process when the simulation program were to interpret the geometries in the simulation environment. This, in return, caused longer simulation time. When the simulation model behaved properly, the final simulation resolution was set to have a high accuracy in order to produce credible results.

In Section 5.1.2, using $N = 1.5$ turns for the long antenna would have placed the antenna dimension line on a better position on Figure 30 and the resonance frequency would have arisen towards the middle of the frequency band instead of on the lower boundary. This would have improved the initial conditions for a more optimal outcome regarding frequency matching on wanted frequencies. However, the results of the simulation with a $N = 1.5$ turn QHA, Figure 75, showed an unexpected outcome. The gain offset was significant over most frequencies and the antenna would not have been able to operate on the mid to higher frequencies. A similar outcome arose when using $N = 1.25$ turns. There are design where fractional turns have been used in QHA; in [63], $N = 0.25$ turn and $N = 0.5$ turn are used. No data was available for the expected behavior when using e.g. $N = 1.25$ or $N = 1.5$ turns in the construction and no additional information was found, why this design was omitted.

When simulating the full submarine environment, the far-field gain plots and VSWR showed credible values. However, the polarization plots deviated from expected behavior and the simulation environment seemed to affect the AR in a negative way where the polarization purity was irregular and alternated between RHCP and LHCP levels in a non-cohesive manner, which deviated from the free-space simulations. The full submarine simulation increased the volume of the simulation environment greatly, thus the mesh size was chosen according to the available computational resources. There had to be a balance between having a mesh fine high enough for the software to be able to reasonably represent the model but also low enough to be possible to compute with resources available. The relatively low mesh count settings were suspected to have affected the AR calculations in the program. A scaled-down environment where the hull was removed, and instead the environment consisted of the antenna, radome, mast, and a small cuboid of water, which the simulation boundaries where set around, was simulated to troubleshoot if the relatively low numbers of meshes in the
Figure 75: Long QHA when N=1.5.
simulation were the root cause. The outcome from the simulations resembled the results obtained from the full submarine environment simulations regarding VSWR and far-field pattern. Also, the outcome for the AR showed a behavior which resembled the free-space simulations which indicate that the low mesh fine indeed was the source of error. This problem might have been avoided by decreasing the number of details, e.g. on the radome but mostly on the hull. The hull could have been replaced by a cylinder or rectangle, as the structure. This since the details of the submarine vessel design gives a minimal impact on the electric parameter and by simplify the design, the number of mesh needed in order to perform the simulation could have been increased.

Overall, the simulated results are expected and follow the theory presented in Section 4. The results also seems to behave similarly in a more real-life simulation environment which strengthens the reliability of the simulation results and the project itself.
11.2 Measurements

Comparing the first far-field measurements on the antenna prototypes to the simulated results, we see a clear difference in gain level and pattern shape. The simulation results show a best case scenario and are expected that the simulated values differ from the preliminary measurements. In real life, external factors such as losses in cables, components, equipment, surrounding, and calibrations can make themselves apparent and affect the outcome. However, the differences are too apparent to rule out that construction and measurement errors not took place.

For the first far-field measurements, slanting field measurements, the available SGA had a frequency range of $f = [200-350]$ MHz and it was expected for the results to have some gain losses for the higher frequency measurements since a mismatch between transmitter and receiver would occur, and a significant drop is evident at the higher frequencies for both antennas, which can be seen by observing Figure 50d and Figure 61d. However, the results on the lower frequencies deviate from the expected appearance while the center frequencies show higher values and mimic the pattern of the simulated results more closely. This behavior appears on both antennas, in various degrees, and it is unlikely that a local measurement error occurred multiple times. Instead, there might be something systematically incorrect, e.g, the SGA is faulty and shows the wrong levels. At the time of measurements, a plane was mounted on the rig and the antennas were elevated 6 m above the plane. This might have affected the measurements since this distance does not satisfy the far-field requirements interference from the fuselage probably affected the end results.

The prototypes might also have been built incorrectly. Potential gap between conductors, imperfect cylindrical shapes on the upholding paper tubes, and the tolerance in the components might also be sources of error. Also, the component values calculated, and later used, had very low values and were, therefore, sensitive to interference. As shown when the matching network for the long antenna was reconstructed, the performance was greatly improved in regards to antenna gain. Also, with the results from the direct illumination measurements, where the antennas had been inspected and connections had been re-soldered, the antennas shows a consistent behavior over all frequencies and mimics the results from the simulations much more closely compared to the slanting field measurements. As stated previously in the report, the main objective of this thesis was not to optimize the performance of a prototype, but rather to suggest a suitable antenna design for this specific application based on simulation results. The antenna prototypes was used as an initial tool of verification, to verify the validity of produced simulation results. Complete measurement will be obtained with full-scale measurement. However, given the technical limitations and temporal constraints, this projects has shown that the design chosen seems to be a viable approach to meet the technical requirements of the antenna system. When the measurements reproduce results that correlate strongly with the simulated results, both in regards to far-field gain and polarization, it is reasonable to assume that the model, the simulation and the design and construction of the antennae is satisfactory and the results are reliable.
12 Conclusions and Future Work

This section will summarize this thesis and draw conclusions based on the gathered results. Suggestions on optimization and continued work based on this thesis which there was no time or budget for to perform will lastly be presented.

12.1 Summary and Conclusion

In this thesis, two antenna elements were designed and evaluated with the main focus to be used towards naval applications. The first design had a maximum allowed radius of 10 cm and a maximum allowed height of 50 cm, referred throughout the report as the short antenna, and the other design had a maximum allowed radius of 5 cm and a maximum allowed height of 90 cm, referred throughout the report as long antenna. The antennas were required to communicate with the satellite system MUOS/Legacy and would, therefore, have to operate on the higher VHF band to the lower UHF band and radiate a hemispherical pattern with high purity of right-hand circular polarization.

After evaluating existing antenna designs and comparing the electrical properties and size constraints with the set technical scope of the project, it was found that the most appropriate design to satisfy all the requirements set was to use a quadrifilar helix antenna structure, on both antenna element designs.

When optimal dimensions for both designs were found, simulation models of the antenna elements were constructed and analyzed using a time domain full-wave solver software. Using the program, it was found that the most well-performing matching network that could be obtained consisted of a $\phi = 90^\circ$ hybrid phase shifter and L-filters consisting of reactive components.

In order to verify the simulated results, antenna prototypes based on the produced values were built and measured. Although the simulated and measured results showed small deviations, mainly due to minor construction and measurement errors, the measurements indicate that the simulations showed realistic values and that the antenna designs were correctly performed.

Finally, simulations in a full-scale submarine environment were performed in order to observe how the antennas behaved in a real-life situation. The results indicated that the antennas performed in a similar manner as in the free-space simulations.

Final conclusions:

- The results from both simulation and prototype verification shows that the antennas perform in accordance to the theory presented
- The short antenna could meet all the technical requirements set
- The long antenna could not meet local requirements regarding VSWR and gain. However, an acceptable communication link can still be established at the higher frequencies if no other design option is possible
- In a choice between the two alternatives, the short antenna is the superior design
12.2 Future Work

A more technically advanced matching network for the long antenna might solve the unwanted reflections in the system and the antenna would perhaps thereafter meet the VSWR requirement. Some variants were tested but nothing was found that improved the final outcome. Due to the given time constrain, a basic filter topology was implemented for the matching network. However, with an expanded time budget, this concept could be expanded and improved. By constructing a better matching network with a lower VSWR on the lower frequencies, the long antenna could be used to communicate with the whole MUOS/Legacy system.

As seen from the measurement results, there is optimization left to be done on the antenna prototypes, e.g. more accurate soldering and a more rigorous matching network which presumably would make the measured results resembles the simulated values more closely. In the case of a sharp product, microstrip transmission lines would probably be used instead of coils and inductors, and the matching network would, therefore, need to be re-designed. The preliminary measurements were only performed from one azimuth angle. As each measurement setup was time-consuming to set up and implement and three antennas were measured, long antenna, short antenna, and SGA, with four different polarization states, there was no time for additional measurements. Measuring a second azimuth angle for the AUT would double the number of measurements and there was no time to complete a new set of measurements. With more time given, these additional measurements with the AUT rotated 90° in azimuth angle relative to the first measurement, as with the simulation, could be performed and two cut sections of the far-field could, therefore, be evaluated instead of only one angle, providing more data to observe potential anomalies.

On the second far-field measurement session, only far-field gain measurements were performed. If more time was given, performing polarization measurements using this measurement setup would have helped determine if the correlation between the main lobe and polarization purity would have had similar values as the simulations indicate.

Since the antennas will be applied to naval use, it is advisable to evaluate the robustness of the construction and if the performance remains the same or change in the case of vibrations and other acting forces which might occur out on the sea. With a radome included, it would be interesting to evaluate the performance of the communication system with a soak test and observe how the system and radiation pattern behaves when the radome is wet. Near-field effects due to the covering radome would probably also need to be analyzed on a much more detailed level than what was performed in this thesis. Also relevant would be to perform a temperature drift test in order to see how much the frequencies shifts when the temperature is increased and decreased.

If the concept is to be expanded and more space was allocated, the possibility to include additional communication options such as LP LoS on the $f = [200-500]$ MHz band and also L-band coverage, $f = [1-2]$ GHz, which are standard bands in submarine communication, could be evaluated. With such concept, there is a need to evaluate eventual telecommunications conflict and
interference signals internally in the system.
References


