ANTENNA MEASUREMENT AND DESIGN FOR THE CANX-7 NANOSATELLITE AND THE DEVELOPMENT OF A GLOBAL NAVIGATION SATELLITE SYSTEM BASED ATTITUDE DETERMINATION SYSTEM

by

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A thesis submitted in conformity with the requirements for the degree of Master of Applied Science
Graduate Department of Aerospace Science and Engineering
University of Toronto

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Abstract

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This thesis describes and presents solutions to various challenges small satellites may encounter during design and operation, particularly in the areas of communications and attitude determination. The first section of this thesis presents simulation and measurement of communications antennas on a nanosatellite to verify that the antennas have sufficient gain and polarization to enable near-omnidirectional operation. Near-omnidirectional antennas are essential to ensure reliable communication with the spacecraft regardless of its attitude, especially when fine pointing ability is unavailable or inadequate. Next, the following section covers the design of a circularly polarized patch antenna for use on an aircraft tracking payload. Lastly, the final section of this thesis presents the development and analysis of a technique for augmenting a single GPS antenna on a spacecraft to estimate attitude. It is possible for GPS measurements to partially supplement an existing attitude sensor that has been denied operation.
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Chapter 1

Introduction

In recent years, small satellites have seen increasing proliferation due to rapidly expanding technological capabilities. This term commonly encompasses some of the smallest classes of satellites such as microsatellites and nanosatellites. Such satellites are characterized by mass with microsatellites possessing a wet mass between 10 and 100 kg and nanosatellites possessing a mass between 1 and 10 kg [1].

Small satellites may encounter additional challenges during design and operations. For instance, some small satellites may possess limited attitude control and sensing capabilities. As such, to successfully complete their mission they may require the ability to establish stable communication links with a ground station and operate radio payloads independent of attitude. In addition, the ability to retain attitude estimation accuracy in unfavorable environments where the operation of the attitude sensors would be limited may also be required. This thesis seeks to address some of these communication and attitude estimation challenges encountered by small satellites. The implemented and theoretical solutions will also be discussed.

1.1 The University of Toronto Space Flight Laboratory

The Space Flight Laboratory (SFL) was founded in 1998 at the University of Toronto Institute for Aerospace Studies (UTIAS) to specialize in the development of small satellite technology [2]. Utilizing the microspace philosophy and a small team approach, SFL has been responsible for the successful development and launch of 13 microsatellites and nanosatellites and has become an international leader in small satellite development. These missions have continuously pushed the small satellite technological envelope, achieving the development and operation of some of the most advanced spacecraft in the world at the fraction of the costs of similar spacecraft developed elsewhere.

1.1.1 Design Philosophy

The Microspace Philosophy is the driving force behind the design methodology of all SFL spacecraft. This philosophy is centered about the axiom that smaller and cheaper missions lead to more accomplishments in space [3] and emphasizes the following points:

1. Development of smaller multiple low-cost missions favoring robustness rather than redundancy through design simplicity.
2. A small, tightly integrated, multidisciplinary team with aggressive schedules and minimal documentation.

3. Use of commercial off-the-shelf (COTS) technologies to reduce spacecraft size and cost while maintaining required capabilities and functionality.

The result of such a strategy is that spacecraft can be developed on rapid schedules while saving up to 90% in costs and scheduling without sacrificing functionality and quality [4]. Additionally, small inexpensive missions also allow for the validation of new technologies that may initially represent unacceptable risk to larger missions. Aggressive use of this strategy allows for the validation of new and COTS technologies, increasing confidence as these technologies gain space heritage and allowing for more missions which would not normally be feasible under traditional mission development.

Another fundamental aspect of SFL design methodology is the avoidance of death modes. A death mode is defined as: “Any condition the spacecraft can enter that would render it inoperable, unrecoverable, or limit its subsequent functionality to the point of making the fulfillment of mission objectives impossible” [5]. Death modes can be entered through self-command, self-action, or ground-based command and differ from failure modes in which they are not caused by equipment failure but rather by poor design.

1.2 Overview of Space Flight Laboratory Missions

Current development at SFL is divided between two main categories of mission. Research and operational spacecraft are primarily used to demonstrate advanced technology for nanosatellites and aggressively demonstrate performance capabilities with low-cost missions. The missions and spacecraft making up the Canadian Advanced Nanospace eXperiment (CanX) program fall into this category. The second category, commercial spacecraft, are missions in which SFL has been contracted by external end users. These end users seek smaller satellites with developmental and flight heritage which retain the benefits of reduced cost and risk associated with larger spacecraft. SFL missions typically utilize common form factors and technology and are designed to be adaptable between each to reduce the work necessary to incorporate new hardware into future missions. The most common form factor is the 20 × 20 × 20 cm Generic Nanosatellite Bus (GNB) [6]. The GNB form factor has also been expanded for missions requiring larger buses such as the NEMO-AM aerosol monitoring satellite which uses a 20 × 20 × 40 cm bus [7]. Additionally, SFL has also utilized the “3U” (3-unit) triple CubeSat bus in some of its missions [8]. The work presented in this thesis focuses primarily on two SFL nanosatellite missions: CanX-7 and CanX-4&5.

(a) CanX-2 - Triple CubeSat
(b) CanX-5 - GNB
(c) NEMO-AM - NEMO

Figure 1: Various SFL Spacecraft Buses
1.2. Overview of Space Flight Laboratory Missions

1.2.1 CanX-7

As with other spacecraft in the CanX series, the CanX-7 nanosatellite is designed to provide rapid and low-cost access to space for scientific experiments and technology demonstrations. This spacecraft was developed to fulfill three mission objectives [9]:

1. Demonstrate passive end-of-life deorbiting from low Earth orbit (LEO) through the deployment of modular, adaptable, and redundant drag sails.

2. Validate the use of space based ADS-B (Automatic Dependent Surveillance - Broadcast) receivers on nanosatellites for the use of aircraft tracking.

3. Visually confirm drag sail deployment and performance through the use of a SFL developed miniature VGA Inspection Camera (mVIC).

The CanX-7 payloads are exhibited in Figure 2 and reflect and support its mission objectives. To achieve its primary objective, CanX-7 carries four triangular drag sail modules supported by copper beryllium tape springs. Each module is made of Windform XT 2.0 and deploys a 1 m$^2$ upilex triangular sail. Collectively, the sails form a 4 m$^2$ drag surface capable of deorbiting a 15 kg satellite from an 800 km altitude [10]. The development of this payload is in response to growing concerns for low Earth orbiting space debris and the suggested Inter-Agency Space Debris Coordination Committee (IADC) de-orbit guidelines for future space missions [11]. The 60 g ADS-B payload was originally developed by COM DEV and RMC (Royal Military College) and consists of an L-band patch antenna and a based 1090 Mode S receiver [12]. This payload is to be operated for 6 months prior to sail deployment and will demonstrate the feasibility of tracking aircraft in remote areas or over the ocean where they would be out of range of line of sight tracking systems, terrestrial surveillance radar, and ADS-B ground stations. Such technology will expand surveillance coverage, allowing aircraft to take more direct routes to their destinations saving time and fuel [13]. Last of all, the mVIC payload consists of three miniature COTS cameras mounted within a deployable boom. These cameras serve to inspect and validate the performance of the drag sails.

The design of CanX-7 is based on a triple-cube (3U) nanosatellite measuring 10 x 10 x 34.05 cm with a mass of 4 kg (Figure 2). Deployables on the spacecraft entail a UHF (Ultra High Frequency) turnstile and a deployable boom containing the tertiary payload and a magnetometer. Along with the magnetometer, CanX-7 possesses magnetorquers to provide 2-axis attitude determination and control capability with ±3° attitude tracking [14]. Commands from Earth are received via an onboard UHF receiver connected to a four monopole quasi-turnstile while an S-band transmitter enables the spacecraft to transmit telemetry and payload data back to earth through the use of two patch antennas [9].

1.2.2 CanX-4 and CanX-5

CanX-4 and CanX-5 are an identical pair of 6 kg, 20 x 20 x 20 cm GNB based nanosatellites developed at SFL (Figure 3). These satellites were launched on 30 June 2014 with the primary mission objective of demonstrating precise, autonomous formation flight [15]. The formation flight portion of this mission involves a 1000 and 500 m along-track orbit (ATO) and a 100 to 50 m projected circular orbit (PCO). The spacecraft completed their primary formation flying mission in under 5 months and are now ready to perform additional experiments in the extended mission phase.
To accomplish their primary formation flying objectives, each nanosatellite is equipped with the following payloads [16]:

1. **Canadian Nanosatellite Advanced Propulsion System (CNAPS)**: A four nozzle sulphur hexafluoride cold gas thruster with a specific impulse of 40 s and a delta-v of 16 m/s. This thruster enables orbital control for drift recovery, station keeping, formation control, and orbit reconfiguration.

2. **Global Positioning System (GPS) Receiver**: A GPS receiver and antenna is used to collect orbit position information with sub-metre precision.

3. **Inter-Satellite Link (ISL)**: Allows for CanX-4 and CanX-5 to exchange navigation data to enable successful formation flight. The ISL is designed to operate in S-band with near-omnidirectional coverage at a 5 km maximum range.

Along with the payload instruments, CanX-4 and CanX-5 each contain a suite of attitude sensors and actuators for full three-axis attitude determination and control. Attitude sensors include six fine sun sensors, a three-axis rate sensor, and a three-axis magnetometer mounted inside a pre-deployed boom. Additionally attitude control actuators comprise of three orthogonally mounted magnetorquers and three reaction wheels.

### 1.3 Thesis Scope and Motivation

The thesis presents work completed on validating the communication antenna arrays on CanX-7, the design of a receiving antenna for the CanX-7 ADS-B Payload, and augmentation of a GPS antenna on CanX-5 to provide attitude measurements using the antenna’s radiation pattern. The following subsections give an overview on the scope and motivations behind each subject area contained in this thesis as well as their impact on small satellite technology within and beyond SFL.
1.3. Thesis Scope and Motivation

1.3.1 Performance Analysis and Validation of Small Satellite Antennas

Spacecraft communications is an essential element of all space missions. For spacecraft to be useful and fulfill their mission requirements, they must possess the ability to exchange information with a ground station or in some cases, other satellites. All SFL spacecraft are designed with radio links to exchange information with one or multiple ground stations. A typical SFL communications setup consists of a UHF band receiver and an S-band transmitter. Furthermore, it is essential that SFL communication systems are designed to be omnidirectional so a communications link can be established regardless of attitude. Omnidirectionality is important as small satellites do not always possess the pointing capability to use directional antennas. More importantly, this property negates a potential death mode where a satellite may fall into an attitude where it is unable to receive commands. Since spacecraft geometry can indirectly alter antenna performance, antenna specifications must be validated for each mission prior to launch.

1.3.2 Antenna Design for an Aircraft Tracking Payload

ADS-B is a new form of co-operative aircraft tracking used in some regions and is expected to be adopted as a global standard of aircraft surveillance by 2020 [17]. However, this system relies on ground stations which are location limited by costs and complexity. One way to circumvent this issue is to install ADS-B receivers onto satellite constellations. The expanded coverage would allow aircraft to take more direct routes to their destinations, saving time and fuel. CanX-7 carries an ADS-B payload to be operated six months prior to sail deployment to demonstrate the feasibility of orbital ADS-B receivers. While this payload was originally externally designed, SFL inherited the development process in early 2015. The author’s main role during this phase was the continuing development of the antenna for this payload. The design of this antenna was subject to constraints to ensure compatibility with the payload structure and spacecraft.
1.3.3 Single GPS Antenna Based Attitude Determination

Small satellites are often constrained in terms of available mass and volume, limiting the quantity and type of attitude sensors available for use. Moreover, some of these attitude sensors may be limited in operation due to unfavorable environmental characteristics or operating scenarios. To counter this, a single GPS antenna on the spacecraft can be augmented to function as a coarse attitude sensor, providing additional measurements to supplement missing sensor data without adding mass, volume, or power requirements to the spacecraft. Attitude sensing with a single GPS antenna is also a novel practice, as typical means of GPS attitude determination utilize multiple antennas. Such a technique is significant for current and future GPS enabled SFL spacecraft since an additional means of attitude determination enables greater safety and flexibility in mission planning and operations.

1.4 Thesis Organization

The thesis is organized into three modules, each covering a main area of work. Each chapter is meant to be relatively self-contained although some background information may be shared between them.

Chapter 2: CanX-7 Antenna Modeling and Measurement

This chapter presents the performance validation of the communications antennas installed on the CanX-7 nanosatellite. Initially, antenna performance is simulated using a finite element solver which is inexpensive and allows the establishment of performance baselines. After the design has been verified, these baselines are then validated through real world measurement of antenna performance. The methodologies and procedures used in communications link design and validation are also disclosed.

Chapter 3: CanX-7 ADS-B Antenna Design

The design of an L-band patch antenna for CanX-7’s ADS-B payload is presented in this chapter. Initially, an analytical design for a linearly polarized antenna is performed. Circular polarization is then induced in this antenna using a cross slot cutout. The dimensions of the cutout are determined using a finite element solver to conduct parametric studies. This solver is then used to optimize and fine tune the antenna dimensions for optimal performance. The antenna design undergoes multiple iterations to address issues related to material availability, antenna gain, and the effect of the spacecraft bus itself on the antenna’s performance.

Chapter 4: Global Navigation Satellite System Based Attitude Determination

A single GPS antenna on the CanX-5 nanosatellite is augmented to function as a coarse attitude sensor. Carrier to noise density ratio measurements from the antenna along with sun-sensor, magnetometer, and rate measurements are incorporated into an extended Kalman filter. The ability of the GPS measurements to supplement the loss of another attitude sensor is determined through the assessment of the filter using simulated or flight datasets containing either naturally occurring or artificially created periods of eclipse, where the spacecraft’s sun-sensors are denied functionality.
Chapter 2

CanX-7 Antenna Modeling and Measurement

To ensure mission success, all spacecraft must be able to communicate with Earth. This can take the form of telemetry, tracking, and control (TT&C) in support of spacecraft operations, transmission of observation and scientific payload data, or as itself a primary mission objective. Most spacecraft accomplish this through the transmission of electromagnetic signals at radio frequencies (RF) [18].

A communications link consists of either a radio or optical modulated carrier between transmitting and receiving equipment. In general, there are three categories of links associated with satellite communications [19]:

- **Uplinks**: Communication links from Earth stations to satellites.

- **Downlinks**: Communication links from satellites to Earth stations.

- **Intersatellite Links**: Communication links between satellites.

Typically, SFL spacecraft are designed to communicate with a number of ground stations. However, the CanX-4&5 formation flying mission has required the development and use of an intersatellite link for their mission.

For spacecraft to send and receive data in orbit, antennas are used to wirelessly transmit radio waves modulated with data or to intercept transmitted radio waves for demodulation. The antennas installed on SFL spacecraft are designed to operate with low directivity and radiate in all directions as equally as possible. This is useful as it allows the spacecraft to maintain a communications link regardless of its attitude which may be critical to mission success and compliant with SFL’s design philosophy of avoiding death modes.

The communications subsystem on CanX-7 are based on heritage designs used on previously launched missions where the communication subsystem designs have been proven to work. Regardless of heritage, CanX-7’s communications subsystem must be empirically proven to function within bounds set by mission, system, and subsystem level requirements. This is achieved through the use of finite element modeling and real world antenna measurement.
2.1 Chapter Overview

This chapter summarizes the work performed on the modeling and measurement of the uplink and downlink antennas on the CanX-7 spacecraft. As evidenced, most of the focus is placed on performance related parameters of the antenna arrays themselves. The chapter is arranged as follows:

1. **Driving Requirements**: SFL requirements which govern the design and performance requirements of the uplink and downlink antenna arrays are identified.

2. **Overview of Antenna Performance & Radio Communication Links**: Crucial performance related parameters related to antenna measurement are defined. A brief background on communications link design is presented as it relates to the requirements and antenna measurement.

3. **Overview of CanX-7 Communications**: Overview of the workings and design uplink and downlink communications subsystem on CanX-7 and how they relate to the driving requirements.

4. **Antenna Modeling**: Formation of CanX-7 antenna models in a finite element solver for the uplink and downlink antennas in both sail deployed and stowed configurations.

5. **Antenna Measurement**: Explanation of the procedures and experimental setup used to measure the CanX-7 antenna arrays.

6. **Results**: Both simulation and measurement results are presented with calculated coverages.

2.2 Driving Requirements

Table 1 outlines relevant SFL requirements where “shall” implies a strict requirement and “should” implies a desire. Requirements CX7-SYS-42 and CX7-SYS-19 justify the reasoning behind the design and setup of the communications subsystem. Likewise, requirements CX7-SYS-45 and CX7-SYS-12 enshrine the antenna properties that antenna pattern modeling and measurement seek to verify; that the spacecraft can maintain communications with a ground station from an orbital environment regardless of attitude and at a required elevation angle with respect to the horizon. It should be noted that while Requirement CX7-SYS-12 was verified in modeling, this requirement was dropped prior to antenna measurement. This was because of the expenses and difficulty associated with sail transport and setup as well as the fact that the design and material of the sails were locked, limiting the utility of the tests. Furthermore, it was determined that post-deployment communications was not a strict requirement. Consequently, Requirement CX7-SYS-12 is presented as “non-compliant”.

2.3 Overview of Antenna Performance and Radio Communication Links

Figure 4 shows the components typically involved in a simple radio communications link. Such a link is comprised of a transmitter element that modulates the data to be transmitted in an RF carrier frequency. The resultant RF signal is then sent to an antenna for transmission in the form of electromagnetic (EM) waves which are then intercepted by a receiving antenna, producing a small voltage. The resultant voltage can then be amplified and demodulated by the receiver to recover the transmitted data.
Table 1: CanX-7 Communications Driving Requirements [20]

<table>
<thead>
<tr>
<th>Requirement</th>
<th>Comments/Rationale</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>CX7-SYS-42:</strong> CanX-7 shall have separate communication frequency bands for telemetry and command.</td>
<td>Minimizes risk by eliminating the need to switch between transmit and receive when communicating with the satellite. (Full duplex operation)</td>
</tr>
<tr>
<td><strong>CX7-SYS-19:</strong> The CanX-7 TT&amp;C subsystem should use equipment sourced from previous SFL missions to the extent practical.</td>
<td>Heritage proven designs; reusing tested and proven equipment and designs is economical and poses less risk.</td>
</tr>
<tr>
<td><strong>CX7-SYS-45:</strong> Prior to sail deployment, link margins (uplink and downlink) shall be no less than 6 dB with a minimum coverage of 75% about the spacecraft, at a minimum elevation angle of 5°.</td>
<td>Spacecraft may be prone to tumbling; near-omnidirectional communications needed to maintain command and control regardless of spacecraft attitude. (800 km operational altitude)</td>
</tr>
<tr>
<td><strong>CX7-SYS-12:</strong> Following sail deployment, link margins (uplink and downlink) shall be no less than 3 dB with a minimum coverage of 75% about the spacecraft, at a minimum spacecraft elevation angle of 30°.</td>
<td>Ensures that spacecraft-ground communications are still possible following sail deployment. This requirement was dropped for antenna measurement</td>
</tr>
</tbody>
</table>

Figure 4: Radio Communications Link Anatomy

The physical properties (amplification, gain, and attenuation) and quantities (i.e. power) encountered in communications engineering often change by many orders of magnitude and are usually multiplicative. As such, it is general convention to express these properties in terms of decibels (dB). Doing so turns multiplication into addition and hence writing and managing budgets are easier. A decibel in terms of power (W) is taken as a base ten logarithmic unit with the unit attached (dBW). For instance the
decibel value of $P$ is computed as:

$$P(dBW) = 10 \log_{10} \left( \frac{P}{P_{ref}} \right)$$

(2.1)

Where the decibel expressed quantity is referenced to $P_{ref}$. Ratios without units are expressed in dB.

As the main focus of this chapter is on antenna measurement, this section will concentrate on the role and properties of the antennas in a communications link. However, a high level overview will also be given to the other components and link design in general. This is done to provide insight into the origin of the coverage requirements for this spacecraft and come context into the antenna measurement procedures. Nonetheless, this should not be taken as a comprehensive guide on the topic.

### 2.3.1 Antenna Gain and Radiation Pattern

Antenna gain is a performance figure combining an antenna’s directivity and radiation efficiency. The radiation efficiency of an antenna is defined as the ratio of radiated power to the input power [19].

$$\epsilon_{R} = \frac{P_{radiated}}{P_{input}}$$

(2.2)

Subsequently, the directivity of an antenna is the measurement of the power density of its direction of strongest emission with respect to the radiated power density from a lossless isotropic emitter. That is, a hypothetical emitter that is completely omnidirectional and efficient. This property is directionally dependent and is often expressed in spherical coordinates. Antenna gain ($G$) is thus the product of the antenna’s directivity ($D(\theta, \phi)$) and radiation efficiency.

$$G(\theta, \phi) = \epsilon_{R}D(\theta, \phi)$$

(2.3)

Gain and directivity are represented in terms of decibels with respect to an isotropic ($G = 0$ dB) or dipole antenna using the unit notations dBi and dBd respectively. Additionally dBic and less commonly dBiL, are used to represent circularly and linearly polarized reference antennas. A very popular manner to express an antenna gain or directivity pattern is by plotting its radiation pattern.

A radiation pattern [21] is a graphical or mathematical representation of the directivity or gain of an antenna. If the antenna in question is in anyway directive the pattern will exhibit a number of lobes. The major lobe of the pattern consists of the region in the direction of maximum radiation. Sidelobes are smaller beams radiating away from the main beam which are caused by spurious emissions of radiation in undesirable directions. Most of the time, sidelobes and their effects cannot be fully eliminated.

The half power beamwidth is defined as the angular separation between the points of the major lobe where the power is halved (-3 dB) from the peak of the main lobe. Conversely, null to null beamwidth is defined as the angular separation where the radiation pattern’s magnitude decreases to zero.

Knowledge of an antenna’s radiation pattern is essential in antenna design to ensure that a selected antenna will perform as desired. At SFL, emphasis is placed on spacecraft antenna arrays which permit communications with the spacecraft regardless of spacecraft attitude. Since a true isotropic antenna does not exist, emphasis is placed on designing arrays with adequate signal strength across most of their radiation pattern. The percentage of the array’s radiation pattern with enough gain to permit communications is referred to as antenna coverage.
2.3. Overview of Antenna Performance and Radio Communication Links

It should also be noted that since all antennas have a degree of directionality, the transmitter and receiver antennas may operate in orientations where both antennas do not experience maximum gain. This reduction in antenna gain due to misalignment is known as pointing loss. Pointing loss can be estimated and accounted for by determining the gain associated with the off-boresight angle corresponding to the maximum pointing error in both the receiving and transmitting antennas.

2.3.2 Antenna Polarization

Antenna polarization is another fundamental element which must be taken into account during antenna design. Polarization refers to the figure traced out by the electric “E-”field of the radiated radio wave with respect to a plane orthogonal to the direction of propagation [22]. Three modes of polarization exist and are discussed below.

Linear

Linear polarization occurs when the E-field oscillates linearly along a single plane (Figure 5). Linearly polarized antennas typically exist in vertical or horizontal polarization where their electric field is either perpendicular or parallel to the Earth’s surface respectively. To ensure minimal loss, it is essential for transmitting and receiving antenna pairs to possess the same polarization and respective orientation. If two linearly polarized antennas are rotated from each other by angle $\theta$, the signal power is degraded by a polarization loss factor ($L_{Pol}$).

$$L_{Pol} = \cos^2(\theta)$$

(2.4)

For instance, if both antennas are linearly polarized, a misalignment of 45° can result in signal degradation of 3 dB whereas 90° can result in more than 20 dB of loss.

![Figure 5: Linear Polarization](image)

Circular

Circular polarization occurs when the E-field of an EM wave contains both horizontally and vertically polarized components. This results in a constant magnitude E-vector that changes direction in a rotary manner tracing out a circular on a plane orthogonal to the direction of propagation. Circular polarization sense is defined as the direction of rotation of the E-vector as seen by an observer looking in the direction of propagation. A clockwise rotation is referred as right-handed circular polarization (RHCP) and a counter-clockwise rotation as left-handed circular polarization (LHCP) (Figure 6). A circularly polarized transmit and receive antenna pair must use the same sense to enable successful signal transmission and
Chapter 2. CanX-7 Antenna Modeling and Measurement

reception. Alternatively, a circular polarized antenna can be paired with a linear polarized antenna to obtain a fixed polarization loss of 3 dB resulting from a constant 45° polarization misalignment.

Figure 6: Right and Left Handed Circular Polarization [23]

Since a spacecraft may find itself in numerous orientations with respect to a ground station or other spacecraft, circular/circular or circular/linear links are preferred in astronautical applications. If a spacecraft and ground station were both to use linearly polarized antennas, constantly fluctuating polarization losses would occur, resulting in frequent communications cutoffs. The uplink and downlink of most SFL spacecraft are designed to operate in circular polarization.

Elliptical

An EM wave with fluctuating vertical and horizontal E-field components results in a rotating E-vector with changing magnitude that traces out an ellipse orthogonal to the direction of propagation. This is known as elliptical polarization. In reality, systems designed for circular polarization are slightly elliptical due to naturally occurring amplitude and phase fluctuations of the E-field.

Polarization Ellipse and Axial Ratio

Polarization can be described in terms of the polarization ellipse shown in Figure 7. This ellipse is a representation of the E-vector’s rotation orthogonal to the direction of EM propagation. The sense of elliptical and circular polarization is determined by the E-vector’s clockwise or counterclockwise direction of travel (\(\omega\)). Polarization quality can be quantified by the axial ratio (\(p_{AR}\)) and tilt angle (\(\tau\)). The axial ratio is defined as the ratio between the horizontal and vertical E-field components in the polarization ellipse [24].

\[
p_{AR}(dB) = 20 \log_{10} \left( \frac{a(E_{max})}{b(E_{min})} \right)
\]

Where \(a(E_{max})\) and \(b(E_{min})\) are the maximum and minimum magnitudes (V) of the E-vector encountered along the semi-major (2a) and semi-minor (2b) axes of the polarization ellipse. Since the axial ratio is a voltage ratio, conversions to decibels utilize a factor of 20 rather than 10.

The axial ratio for a circular polarized electromagnetic wave is equal to 1 (0 dB) since the magnitude of the horizontal and vertical E-field components are equal. Likewise, the axial ratios for linear and elliptical polarization are \(\infty\) and \(> 1\) (> 0 dB) respectively. From this, it can be seen that circular and linear polarization are special cases of elliptical polarization.
2.3.3 Two Port Network S-Parameters

A transmitting and receiving antenna can be modeled as the two port network in Figure 8. In this figure, the two coaxial ports in this network are intuitively labeled 1 and 2 [25]. Additionally, \( a \) and \( b \) correspond to incoming and outgoing traveling waves at each port:

\[
a_1 = V_1^+, \quad a_2 = V_2^+, \quad b_1 = V_1^-, \quad b_2 = V_2^-
\]  

(2.6)

Where + and − refer to the forward or reverse voltage wave directions. In this case, the input-output relationship can be described by the S(Scattering)-Parameter matrix.

\[
\begin{bmatrix}
  b_1 \\
  b_2 
\end{bmatrix} = \begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix} \begin{bmatrix}
  a_1 \\
  a_2
\end{bmatrix}
\]

(2.7)

\( S_{21} \) represents the power transferred from port 1 to 2 and is known as the forward voltage gain while \( S_{12} \) is the power transferred in the opposite direction and is defined as the reverse voltage gain. \( S_{11} \) and \( S_{22} \) are the reflected power at each port and are known as the input and output port voltage reflection.
coefficients respectively. Expanding the S-parameter matrix yields:

\[ b_1 = S_{11}a_1 + S_{12}a_2 \]  \hspace{1cm} (2.8)

and

\[ b_2 = S_{21}a_1 + S_{22}a_2 \]  \hspace{1cm} (2.9)

Where the reflection coefficients are:

\[ S_{11} = \frac{b_1}{a_1} = \frac{V_1^-}{V_1^+} \quad \text{and} \quad S_{22} = \frac{b_2}{a_2} = \frac{V_2^-}{V_2^+} \]  \hspace{1cm} (2.10)

and the voltage gains are:

\[ S_{12} = \frac{b_1}{a_2} = \frac{V_1^-}{V_2^+} \quad \text{and} \quad S_{21} = \frac{b_2}{a_1} = \frac{V_2^-}{V_1^+} \]  \hspace{1cm} (2.11)

S-Parameters can be measured directly through the use of a vector network analyzer (VNA) and are important for both antenna design and antenna measurement. The \( S_{21} \) is the property measured during radiation pattern measurement. Likewise, the \( S_{11} \) parameter is commonly defined as the reflection coefficient (\( \Gamma \)) or return loss and is used to impedance match an antenna and determine an antenna’s voltage standing wave ratio (VSWR); more of which will be discussed next chapter. Since the S-parameters are voltage ratios, as with axial ratio, conversions to decibels also utilize a factor of 20 \((i.e. V(dB) = 20 \log_{10} \left( \frac{V}{V_{ref}} \right) )\).

### 2.3.4 Transmitter Characteristics

The transmitter is an electronic device that when coupled with an antenna generates radio waves. In a communications link, the transmitter modulates data in the form of an information signal with a carrier frequency to produce an alternating current at a radio frequency. When this current excites the transmitting antenna, EM waves are emitted. Transmitters also contain a power amplifier for increasing the power of the RF signal to drive the transmitting antenna. Conventionally when expressed in decibels, transmitter power is expressed with respect to 1 mW in decibel-milliwatts (dBm). Given the increased distances encountered in space-based communications links, transmitter power for astronautical applications are often expressed with respect to 1 W or in decibel-watts (dBW).

The effective isotropic radiated power (EIRP) \cite{19} is used as a metric for the overall performance of the transmitting equipment. EIRP is defined as the power fed to the transmitter antenna multiplied by the antenna gain in the desired transmit direction.

\[ P_{EIRP}(dBW) = P_{Tx}(dBW) + G_T(dBi) - L_{Tx}(dB) \]  \hspace{1cm} (2.12)

Where \( P_{Tx} \) is the power of the RF signal exiting the transmitter’s power amplifier, \( L_{Tx} \) are the losses between the transmitter and the antenna (mostly from the transmission line, \( L_{FTx} \)), and \( G_T \) is the gain of the transmitting antenna.

### 2.3.5 Propagation Losses

While in transit between the transmitting antenna and the receiving antenna(s), the emitted radio wave is subject to multiple sources of propagation losses. These include free space loss, atmospheric
loss, and precipitation losses. Atmospheric and precipitation losses are the result of absorption of the radio wave by atmospheric gasses or precipitation encountered and may vary based on environment and altitude. These losses can estimated using geographic databases and precipitation models published by the International Telecommunication Union (ITU) [26]. However, these losses are often neglected from space-based communications design since they have very little impact on highly powered and directive space links.

Free-space path loss (FSPL) accounts for the majority of losses experienced in a radio communications link and is defined as: “The loss between two isotropic radiators in free space, expressed as a power ratio” by IEEE Std 145-1983 Standard Definitions of Terms for Antennas [27]. It is a common misconception that FSPL is caused by the direct attenuation of radio waves by free space since there is no physical mechanism that can cause this. In reality, FSPL encapsulates two factors. First, radiated EM waves have a tendency to spread out, reducing the power spectral density (\( S \) (W/m\(^2\))) as a function of range, \( d \) in meters. This phenomena can be described by the inverse square law:

\[
S = \frac{P_{EIRP}}{4\pi d^2} \tag{2.13}
\]

The second contributing factor is the frequency dependency of the receiving antenna’s effective area. The effective area of an antenna (\( A_{Eff} \)) determines how well an antenna can pick up power from an intercepted radio wave and is related to the receiving antenna gain (\( G_R \)) and EM wavelength (\( \lambda \)) by:

\[
A_{Eff} = G_R \frac{\lambda^2}{4\pi} \tag{2.14}
\]

As FSPL is referenced with respect to an isotropic antenna (\( G_R = 0 \) dBi), the received power (\( P_R \)) is:

\[
P_R = S \frac{\lambda^2}{4\pi} \tag{2.15}
\]

This characteristic is thus dependent on the operating frequency and FSPL (\( L_{FSPL} \)) can then be calculated as:

\[
L_{FSPL} = \left( \frac{4\pi d}{\lambda} \right)^2 = \left( \frac{4\pi df}{c} \right)^2 \tag{2.16}
\]

Where, \( f \) is the EM frequency (Hz) and \( c \) is the speed of light (2.99792458 \times 10^8 \) m/s). In decibels, FSPL is then expressed as:

\[
L_{FSPL}(dB) = 20 \log_{10}(R) + 20 \log_{10}(f(MHz)) + 32.45 \tag{2.17}
\]

Where \( R \) is the distance expressed in kilometers (\( R = d/1000 \)). The sum of the FSPL, precipitation, atmospheric losses, pointing, polarization losses form the propagation losses \( L \).

### 2.3.6 Noise

The purpose of the receiver in a communications link is to amplify and demodulate the electrical signal from the receiving antenna to extract the transmitted data. However, the receiver also must contend with unwanted disturbances and signals (Noise) which will degrade and obscure the desired signal. In a communications link, there exists external and internal sources of noise. External sources of noise include man-made sources such as other communication links, radars, radio telescopes, and power lines.
and naturally occurring sources such as celestial bodies (including the Earth), and background cosmic noise. Noise generated from external sources enter the system through the antenna and are quantified in terms of an antenna noise temperature $T_A$ expressed in kelvins. The noise temperature is equivalent to the temperature of a theoretical resistor outputting the same amount of thermal (Johnson) noise output as the receiver system \cite{27}. The noise temperature of a celestial body is its brightness temperature \cite{28}:

$$T_{\text{body}} = (1 - |\Gamma_s|^2)T_p$$  \hspace{1cm} (2.18)

Where $\Gamma_s$ is the albedo of the body and $T_p$ is the maximum surface temperature of the body. For low earth orbiting satellites, major bodies which must be taken into consideration are the Earth, its Moon, and the Sun.

Receivers also possess a number of active devices such as amplifiers which also contribute noise along as passive devices such as filters and feeds which will attenuate the received signal. The properties of this signal chain can be approximated as a single system noise temperature, $T$ \cite{18}.

$$T = T_A(L_{FRx}) + (1 - L_{FRx})T_{LFRx} + T_{Rx}$$  \hspace{1cm} (2.19)

Where $T_A$ is the noise temperature from the antenna, $T_{L_{FRx}}$ is the effective temperature of the receiving feed, $L_{FRx}$ is the loss through the feed, and $T_{Rx}$ is the noise temperature from the receiver including its active and passive components. The receiver’s noise temperature can also be given in terms of a noise figure ($N_F$) where:

$$T_{Rx} = T_0(N_{NF_{Rx}} - 1)$$  \hspace{1cm} (2.20)

Noise figures are always quoted at a reference temperature of $T_0 = 290$ K. The resulting system noise temperature thus allows the noise power spectral density, $N_0$ to be calculated in W/Hz as:

$$N_0 = kT$$  \hspace{1cm} (2.21)

or in decibel form:

$$N_0(dB/Hz) = -228.6 + T$$  \hspace{1cm} (2.22)

Where $k$ is the Boltzmann constant ($1.3806488 \times 10^{-23}$ J/K or -228.6 dBJ/K)

### 2.3.7 Receiver Characteristics and Link Budgets

As with the transmitter, the performance of the receiving components is often characterized by its figure of merit $G/T$ \cite{19}:

$$G/T(dB/K) = G_R(dBi) - T(dBK)$$  \hspace{1cm} (2.23)

The total power received at the receiver, $C$ can then be expressed in decibels:

$$C(dB) = P_{EIRP} + G_R - L - L_{FRx}$$  \hspace{1cm} (2.24)

A bandwidth independent carrier-to-noise power spectral density ratio, $C/N_0$ can then be obtained.

$$C/N_0 = P_{EIRP} + G/T - L - L_{FRx} + 228.6$$  \hspace{1cm} (2.25)
Finally, the energy per bit to noise power spectral density ratio \( \frac{E_b}{N_0} \) is calculated by dividing \( C/N_0 \) by the data rate \( r_b(dB-Hz) = 10\log_{10}(r_b(bps)) \).

\[
\frac{E_b}{N_0} = \frac{C}{N_0} - r_b
\]  

(2.26)

\( E_b/N_0 \) is compared to the bit error rate (BER) performance of the receiver. BER refers to the ratio of bit errors to total number of transferred bits in a given time interval, where bit errors are the number of received bits in a data stream that have been altered due to interference, noise, distortion, or synchronization errors. In general, communication links are designed with an intended maximum BER which is related to a required \( (E_b/N_0) \) or receiver sensitivity that is dependent on the modulation scheme and coding rate used. The link margin \( (M_L) \) is thus:

\[
M_L = \frac{E_b}{N_0} - (E_b/N_0)_{\text{Required}}
\]  

(2.27)

Where \( (E_b/N_0)_{\text{Required}} \) is the \( E_b/N_0 \) required to achieve a certain BER for the given receiver. If \( M_L \geq 0 \) dB then the communications link is closed and data can be transferred. Typically communications links are designed with a link margin greater than zero to account for variations in component performance and other unforeseen circumstances. It should also be noted that in some cases the sensitivity of a certain component in the receiver (i.e. a discriminator) may be the limiting factor of the communications link rather than the sensitivity of the receiver itself [29]. In these cases, an additional margin is calculated with respect to that component. In communications engineering, the link components are expressed in a tabulated form known as a link budget. This allows the effects of variations in the link components on the performance of the communications link to be determined easily. In this context, the antenna coverage as expressed in the requirements is the percentage of the 3D radiation pattern of each antenna array on CanX-7 possessing sufficient gain for their respective link budgets to close with the required link margin. Conversely, areas of a radiation pattern with insufficient gain to achieve the desired link margin are known as nulls.

2.4 Overview of CanX-7 Communications

CanX-7’s RF architecture consists of an UHF (Ultra High Frequency) uplink, an S-band downlink, and an L-band ADS-B payload receiver (Figure 9) [30]. The following section details the properties of the uplink and downlink on CanX-7. The ADS-B payload along with more detailed workings of patch antennas will be discussed in the following chapter.

2.4.1 Uplink Receiver and Turnstile

To receive uplinked commands from a ground station, CanX-7 contains a 4-port UHF receiver (Figure 10a) based on the radio designs for CanX-2 and the GNB satellites. Uplink signals use a frequency-shift keying (FSK) modulation scheme. Due to the modulation scheme, this receiver contains a discriminator which is the limiting factor for receiver sensitivity. Also, signals entering each input port is phased shift 90° with respect to each port.

Reception of uplink signals from the ground station is performed using four deployable monopoles (Figure 10b). These are based on designs used on other SFL spacecraft, most notably CanX-2 [31]. Each
monopole element consists of a straight rod shaped piece of steel approximately $1/4$ of the operating wavelength which acts as a resonator. That is, radio waves are generated from oscillating standing waves of voltage and current throughout each element’s length. A single monopole emits linearly polarized radio waves along its length with a relatively non-directive radiation pattern.

By combining the four monopole elements and feeding each element with signals each with a phase shift of $90^\circ$, a quasi-turnstile array can be created which emits circular polarized radio waves. It should be noted that since the elements protrude from the surface of the bus, the effective area (or aperture) of the resultant turnstile takes the form of a plane approximately orthogonal to the spacecraft’s Y-axis. This means that emitted radio waves in the -Y direction are RHCP and those in the +Y direction are LHCP (Figure 11). Such a radiation pattern theoretically allows near-omnidirectional coverage of the spacecraft providing that the ground station is capable of switching polarization.
2.4.2 Downlink Transmitter and Patch Antennas

As with the uplink receiver, CanX-7’s downlink S-band transmitter using phase-shift keying (PSK) modulation is also borrowed from CanX-2 and GNB radio designs (Figure 12a). Unlike the UHF receiver, the two output ports are kept in phase with each other. Additionally, the radio components and amplifier components are kept on separate boards which are folded together for installation.

Another heritage design from CanX-2, CanX-7 is equipped with two patch antennas located on opposing faces (+X and -X in the reference frame defined in Figure 14) of the spacecraft bus (Figure 12b). Each patch antenna consists of a $33.7 \times 33.7$ mm metallic patch bonded onto a $55 \times 55$ mm laminate substrate approximately 1.5 mm thick. Circular polarization is induced by slotting the patch and repositioning the feed to cause certain parts of the antenna to radiate with a 90° phase shift. Unlike the UHF turnstile, each patch antenna functions as an individual stand-alone antenna, emitting a RHCP lobe in the antenna plane’s normal direction. Antennas are mounted on opposing faces of spacecraft to maximize coverage and to reduce possibility of interference between antennas.

2.4.3 Relation to CanX-7 Communication Requirements

To reiterate, a communications link can only be “closed” and data transferred when there exists a link margin greater than or equal to zero. At SFL, the required BER for communications links is $10^{-5}$ with the required link margin during design to be 12 dB and 6 dB after a mission’s critical design review (CDR). From CanX-7’s link budgets, the communications links close with uplink and downlink margins of 6 dB if the S-band and UHF antennas possess a minimum gain of -10 dBic and -11.36 dBic respectively for pre-sail deployment. For post-sail deployment, the minimum gains required are -19.5 dBic for S-band and -20.4 dBic for UHF. To meet requirements, these antennas must achieve these gains for over 75% of their 3D radiation patterns. Since the S-band antenna only operates in RHCP, only coverage in that polarization sense is permissible. Conversely, the UHF array meets coverage requirements if the individual coverage of at least one of its lobes or the combined coverage of both RHCP and LHCP lobes exceed 75%. Permitted elevation angle also plays a significant role in the minimum gain requirements.
as well. A lower elevation angle with respect to the horizon means that the maximum design distance between the spacecraft and the ground station is increased.

2.5 Antenna Modeling

Prior to performing any real world antenna verification, the antenna patterns on the spacecraft were simulated using ANSYS HFSS (High Frequency Structural Simulator), a finite element solver for electromagnetic structures and is popular for antenna and RF circuit design. Simulations provide an efficient and cost effective means to verify design functionality and provide baseline expectations of performance.

The methodology of antenna modeling was to analyze the uplink and downlink antennas separately for each drag sail deployment case. This meant that two models would be produced for each of the drag sail pre-deployment (stowed) and post-deployment cases. Components such as conductive tape and solar panels were not included as previous modeling in other missions had shown that educated simplifications of less critical design aspects would still yield acceptable results that can be taken to test.

2.5.1 Old CanX-7 Model

The original CanX-7 layout (Figure 13) was similar to the CanX-2 nanosatellite with a magnesium spacecraft bus and deorbit sail modules. At the time, the secondary payload consisted of a femtosatellite from Ryerson University mounted on a deployable boom [30]. The antennas in this model were each fed directly using a lumped port.

2.5.2 Revised CanX-7 Model

The design of CanX-7 had undergone drastic changes since its preliminary design review (PDR). The 3U cube satellite bus is still being used, however the secondary payload was replaced by the ADS-B
receiver which required the S-band antennas to be shifted. Furthermore, a camera-based tertiary sail deployment verification payload was added to the deployable boom replacing the femtosatellite. As a result, the boom’s location was shifted and the boom was angled to provide a better view of the sails. The HFSS antenna model was updated in Fall 2013 to reflect the following design changes:

1. The deployable boom was repositioned and its dimensions and angle were altered to accommodate the tertiary payload.

2. The ADS-B secondary payload antenna was modeled as a generic $75 \times 75$ mm patch antenna.

3. The S-band antennas were shifted to their present centered locations 0.5 mm from the -Y edge of the $\pm X$ panels.

4. The dimensions of the deorbit modules and sails were updated and aluminum spacer and end plates were inserted between the modules.

5. The body was confirmed to be constructed of aluminum and the sail modules from Windform XT 2.0; a carbon filled, polyamide based composite.

6. The reference frame was revised to reflect system level conventions used in solid modeling.

The updated models are presented in Figure 14.

2.6 Antenna Measurement

While simulations are cost efficient and useful to initially validate antenna performance and set baseline assumptions, real world testing is required to confirm that the actual antenna will perform as designed. The field of antenna measurement encompasses the testing and characterization of an antenna’s performance parameters. These parameters can include, gain, radiation pattern, polarization, beamwidth, and impedance. In the context of CanX-7 which uses near-omnidirectional arrays, the emphasis is on verifying antenna radiation patterns and circular polarization.
Figure 14: Revised CanX-7 Antenna Model
2.6. Antenna Measurement

2.6.1 Experimental Setup

Antenna measurement was performed in March 2014 on an assembled mechanical model of CanX-7. This mechanical model consisted of the qualification structure with the S-band and UHF antenna arrays installed. These arrays were connected to splitters which allowed them to be connected to test equipment. The mechanical model can be seen in Figures 10b and 12b while the splitters can be found in Figure 15.

![CanX-7 Mechanical Model with UHF (Blue) and S-band (Red) Splitters](image)

General antenna measurement techniques and principles are described in the *IEEE Standard Test Procedures for Antennas* [32]. Additionally, SFL has used similar antenna measurement techniques on past spacecraft such as the NEMO-HD satellite. While, radiation patterns are three-dimensional in nature, measuring the pattern at an infinite amount of spherical points is impossible. Instead, it is more efficient and cost-effective to measure a number of two-dimensional *pattern cuts*. Pattern cuts are measured by placing the spacecraft on a platform which rotates azimuthally in discrete steps (Figure 16) while being illuminated by a source antenna possessing a known radiation pattern. The source antenna frequency is also swept over a S-band and UHF frequency ranges during measurement. Although only S-band results at the operating frequency are used, the full UHF sweep is evaluated to permit optimization of monopole length.

**Antenna Calibration**

Calibration is performed to characterize the radiation pattern of the source antenna and to account for unknown losses in the test equipment. This is especially important since the source antenna will never be truly isotropic and inconsistencies in its radiation pattern may affect the validity of the results. The calibration procedure is performed through two measurements. First, a line consisting of a coaxial cable and a 30 dB attenuator with a known loss is connected to the source antenna port and the receiving port
on the turnstile and the loss of the cable is measured with the network analyzer. After this, the source antenna is connected to the source port while a duplicate source antenna is placed on the turnstile and a pattern cut taken. After calibration, the true path loss is calculated [33]:

\[
L_{\text{path}} = A_{2\text{Antennas}} + L_{\text{Cable+Attenuator}} - A_{\text{Cable+Attenuator}}
\]  

(2.28)

Where \(L_{\text{Cable+Attenuator}}\) is the known loss of the cable and attenuator line, \(A_{\text{Cable+Attenuator}}\) is the signal amplitude of the cable-attenuator measured by the test apparatus and \(A_{2\text{Antennas}}\) is boresight signal amplitude measured with the source antenna pair. In this case, only the boresight magnitude value of the source antenna is needed since the source antenna remains stationary during measurement and both antennas are linearly polarized and aligned. The gain of the source antenna at boresight (\(G_{\text{Source}}\)) can then be calculated by:

\[
G_{\text{Source}} = \frac{L_{\text{path}} - L_{\text{FSPL}}}{2}
\]  

(2.29)

Where \(L_{\text{FSPL}}\) is the free space path loss between the two antennas calculated from the known distance and frequency. A correction factor (\(q_{CF}\)) can then be calculated which is factored into the antenna measurement results.

\[
q_{CF} = G_{\text{Source}} - A_{2\text{Antennas}} - L_{\text{SatCables}} - L_{\text{Splitter}}
\]  

(2.30)

Where \(L_{\text{SatCables}}\) are the losses of the cables installed in the CanX-7 bus and \(L_{\text{Splitter}}\) is the loss from the S-band or UHF splitter (Figure 15). It is generally good practice to roughly calculate the boresight gain of the source antenna during calibration as it allows test personnel to verify proper functionality of
2.6. Antenna Measurement

the equipment.

Test Orientations

The source antennas used in this case were linearly polarized and two pattern cuts were obtained for each test orientation; one with the source antenna in horizontal polarization and another in vertical. This permits measurement of the horizontal and vertical E-field components allowing determination of LHCP and RHCP characteristics [22]. After measurement, the model was rotated 45° to a new orientation and another pair of pattern cuts were taken for four orientations in total. Four pattern cuts were chosen as a cost effective option since the arrays were designed to operate omnidirectionally and expected to possess a symmetrical patterns based on pattern simulations. In addition, this number of cuts has proven to provide accurate enough coverage estimates for previous SFL spacecraft.

The test orientations for CanX-7’s UHF and S-Band antennas are shown below in Figures 17 and 18 resulting in the pattern cuts shown in Figures 19a and 19b. Note that the black dots in Figure 19 indicate the portion that is facing the source antenna’s boresight; designated as 0° azimuth for each pattern cut.

Isolation

An isolation measurement was conducted parallel to antenna pattern measurement with the objective of determining the degree of isolation, the attenuation to undesired coupling (transfer of energy), between the transmitter and receiver antennas. If isolation was too low between the arrays, transmitting antennas may interfere with receiving antennas, decreasing operational efficiency. SFL does not mandate an
isolation requirement and this property was characterized as a cautionary measure. Also, most SFL spacecraft rely on frequency diversity by using widely spaced uplink and downlink bands. General isolation measurement procedures can be found in land-based telecommunications literature [34] and instrument setup was identical to antenna measurement. The difference in this setup was that S-band transmitter antennas on CanX-7 were used as the source antenna (Figure 20). Switching antenna feeds was predicted to have little impact on results, as the network was reciprocal; $S_{21} = S_{12}$ [35]. Isolation was then determined by measuring the forward voltage gain ($S_{21}$) between the S-band and UHF arrays while sweeping the test frequency over S-band and UHF frequency ranges.

![Figure 19: CanX-7 Pattern Cuts](image1)

![Figure 20: CanX-7 Isolation Measurement Setup](image2)
2.7 Results

Full 3D antenna patterns are obtained from the HFSS (High Frequency Structural Simulator) simulations while antenna measurement produces four 2D pattern cuts. Antenna coverage can be calculated directly from the 3D HFSS patterns and estimated by interpolating between pattern cuts. Tables 2 and 3 show the calculated coverages and the following subsections cover further details regarding the simulated patterns, measured pattern cuts, and coverage calculations. The simulated sail deployed coverages in Table 3 are increased due to relaxed link margin and contact elevation requirements. These relaxed requirements result in lower gain design points. To reiterate, the sail deployed case was not measured, due to the impracticality of transportation and setup and the non-necessity of post-deployment communications; requirement CX7-SYS-12 in Table 1 was declared “non-compliant”.

<table>
<thead>
<tr>
<th>Band</th>
<th>Required Coverage</th>
<th>Required Gain</th>
<th>Polarization</th>
<th>Simulated Coverage</th>
<th>Measured Coverage</th>
</tr>
</thead>
<tbody>
<tr>
<td>S-Band</td>
<td>75%</td>
<td>-10 dBi</td>
<td>RHCP</td>
<td>89.9%</td>
<td>89.8%</td>
</tr>
<tr>
<td>UHF</td>
<td>75% (Combined)</td>
<td>-11.36 dBi</td>
<td>RHCP</td>
<td>88.2%</td>
<td>84.0%</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>LHCP</td>
<td>83.0%</td>
<td>69.5%</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Band</th>
<th>Required Coverage</th>
<th>Required Gain</th>
<th>Polarization</th>
<th>Simulated Coverage</th>
</tr>
</thead>
<tbody>
<tr>
<td>S-Band</td>
<td>75%</td>
<td>-19.5 dBi</td>
<td>RHCP</td>
<td>92.6%</td>
</tr>
<tr>
<td>UHF</td>
<td>75% (Combined)</td>
<td>-20.4 dBi</td>
<td>RHCP</td>
<td>89.0%</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>LHCP</td>
<td>97.4%</td>
</tr>
</tbody>
</table>

2.7.1 Simulated 3D Radiation Patterns and Coverage Calculation

The 3D HFSS patterns are shown below in Figures 21, 22, and 23 with respect to the reference frame defined in Figure 14 (Green: +X, Blue: +Y, Red: +Z).

These HFSS simulations reveal symmetrical radiation patterns prior to sail deployment. The sail stowed S-band pattern is characterized by two individual main lobes from each antenna which meet along the Y-axis to form a series of nulls along the YZ plane (Figure 21a). Additionally, the sail stowed UHF patterns (Figures 22a and 23a) confirm the near-omnidirectional radiation pattern described in Figure 11. When the sails are deployed near the +Y side of the spacecraft (Figure 14b) they interfere with the radiation patterns. This interference is shown in the sail deployed cases where the S-band and UHF LHCP radiation patterns (Figures 21b and 23b) are bisected by the sails and experience scattering and reflection in the -Y direction. Reflection is more pronounced in the UHF RHCP pattern (Figure 22b) as the lobe is already facing away from the sail towards the -Y direction. Antenna coverage can be calculated from these patterns by first defining a unit sphere using the solid angle equation [25]:

\[ d\omega = \sin \theta d\theta d\phi \]  

(2.31)
Where $\theta$ and $\phi$ are angular spherical coordinates with respect to the ZX and XY planes respectively (Figure 14). HFSS outputs antenna gain values at discrete angular increments of $d\Theta = d\Phi = 2.5^\circ$. From this, the ideal coverage area can be obtained by taking the sum of all solid angle segments $d\omega$ at each angular increment. Effective coverage area can then be calculated by taking the sum of the solid angle segments that have corresponding gains higher than the required minimum gain. The percent coverage can then be calculated from the ratio of the effective coverage area to the total area of a unit sphere.

Table 2 shows that in the sail stowed case, the UHF LHCP coverage is lower than that of the UHF RHCP lobe. This is likely attributed to the bus geometry as most of the structure sits within the LHCP Lobe. However this situation is reversed in the sail deployed case (Table 3) where UHF LHCP possess the most coverage out of all setups while the UHF RHCP possesses the least. The increase in coverage is due to lower gain design points resulting from a relaxed link margin and elevation requirements for the deployed case. However, the reversal in coverage can also be attributed to the electromagnetic properties...
of the sails. The result of the reflective effect of the sails is the concentrates most of the effective area of
the UHF RHCP sphere in the -Y direction, increasing directivity and narrowing the lobe (Figure 22b). Conversely, the sails intercept the UHF LHCP and S-band lobes, causing scattering and an increase in
the areas of the pattern possessing sufficient gain (Figure 21b and 23b). Regardless, simulations have
shown that CanX-7 is capable of meeting both pre- and post-sail deployment coverage requirements.

2.7.2 Antenna Measurement Pattern Cuts

Measurements of the signal amplitudes ($A_H$, $A_V$) and phase ($P_H$, $P_V$) for both a horizontally and
vertically polarized source antenna were output by the network analyzer for each 1° azimuth increment.
As antenna gain is the ratio of the power produced by an antenna to that of an ideal isotropic emitter,
the horizontal and vertical gains ($G_H$, $G_V$) in each pattern cut would be directly equal to the measured
signal amplitudes if the source antenna was isotropic. Isotropic emitters do not exist under real-world
circumstances, thus the correction factors ($q_{CF}$) calculated in Equation 2.30 were added to the measured
CanX-7 signal amplitudes to account for the gain of the source antenna [36].

$$G_{HorV}(dB) = A_{HorV} + q_{CF} \quad (2.32)$$

The LHCP and RHCP antenna gains in decibels can then be calculated from the measured phases and
resultant horizontal and vertical antenna gains [22].

$$G_{RHCP} = 10 \log_{10} \left( \left[ G_H \cos(P_H) - G_V \sin(P_V) \right]^2 + \left[ G_H \cos(P_H) + G_V \sin(P_V) \right]^2 \right) - 3 \quad (2.33)$$

$$G_{LHCP} = 10 \log_{10} \left( \left[ G_H \cos(P_H) + G_V \sin(P_V) \right]^2 + \left[ G_H \cos(P_H) - G_V \sin(P_V) \right]^2 \right) - 3 \quad (2.34)$$

The axial ratio along the pattern cuts can also be calculated:
\[ p_{AR} = 20 \log_{10} \left( \frac{10 \frac{G_{RHCP}}{20}}{10 \frac{G_{LHCP}}{20} + 10 \frac{G_{LHCP}}{20}} \right) \] (2.35)

By plotting the calculated RHCP and LHCP antenna gains at each azimuth point, a pattern cut can be constructed. The resultant S-band and UHF pattern cuts are shown in Figure 24, 25, and 26. These are compared to pattern cuts extracted from the simulated 3D patterns.

Figure 24: S-Band RHCP Pattern Cuts - [Blue: Measured, Red: HFSS, Green: Required(-10 dBic)]

The geometry of the S-band pattern cuts (Figure 24) were similar to the HFSS simulations; consisting of two symmetrical main lobes originating from the location of the antennas. Additionally, the gains along the null plane in the measured cut appear to be higher in magnitude and rotationally offset by
2.7. Results

Figure 25: UHF RHCP Pattern Cuts - [Blue: Measured, Red: HFSS, Green: Required(-11.36 dBic)]

10-15° (Figure 24a). This effect was likely a result of the accuracy of the measurement instrumentation and pattern cut calculations. For instance, the measured pattern cuts were taken with respect to an origin centered between the S-band antennas whereas the HFSS simulations were taken with respect to the geometric center of the spacecraft bus (Figure 21a). There were also differences in measured antenna gain especially in the 45° and 135° orientations (Figures 24b and 24d) where some areas of the measured pattern cut were lower than the simulated cuts. Regardless, these reduction in gain at the nulls were not low enough to expand the existing nulls.

The UHF lobes from the pattern cuts (Figures 25 and 26) were also geometrically similar to their HFSS counterparts. The peak gains were 0-3 dBic lower in the measurements than in the simulations
for all orientations and polarization senses. Another issue was that in all of the orientations except for 90° cut, the null in the -Y UHF LHCP lobe increased in width by 45°. This widened null resulted in a significant decrease LHCP coverage. Although widening was present in the UHF RHCP null, especially in the 135° cut (Figure 25d), the decrease in main lobe width and resultant loss of coverage was much less. The measured UHF RHCP nulls also appear to be more defined than those observed in the S-band patterns, although their minima also experience a rotational offset.

Three-dimensional spherical radiation patterns were calculated by interpolating between the four pattern cuts for the UHF and S-band pattern cuts. Coverage was calculated by taking the sum of all solid angle segments with sufficient gain with respect to unit sphere formed by the interpolated pattern.
2.7. Results

cuts. This effective area was then divided over the total solid angle of the sphere. To reiterate, the antenna arrays must possess sufficient signal power over 75% of their 3-D radiation pattern to close the uplink and downlink link budgets with their required margins.

The measured coverages were also tabulated in Table 2. The S-band coverage experienced almost no change between the measured and HFSS results. This stability led credence to the observation discussed above that width of the nulls in the measured patterns were similar to the simulations as well as to the fact that any decreases in the magnitude of the main lobes were not sufficient to create new nulls. The consistency between measured and simulated pattern cuts was likely due to the fact that the S-band antennas were more self-contained and would be less affected by losses originating from assembly errors. Likewise, the the test orientations used kept the antennas facing out the opposite sides of the test structure. Since at least one antenna would remain exposed to the source antenna at all times, the chances of interference from the test structure was minimal.

The UHF observations also appear to be supported by the calculated coverages. The UHF RHCP null which experienced a minor increase in width, only cost the UHF RHCP pattern a coverage loss of 4.2%. However, the much greater increase in the width of the UHF LHCP null decreased LHCP coverage to 69.5%. This was hypothesized to be caused by implementation losses or the occultation of the LHCP lobe by the foam support structure. Regardless, since the UHF array was designed to function in both polarization senses the combined coverage of both lobes exceeded the 75% requirement and no further testing was performed. Additionally, the results of the UHF frequency sweep was also used to calculate coverages for each frequency (Figure 27). The sweep was performed to determine if it would be possible to optimize performance at the centre frequency by adjusting the length of the UHF monopoles. This was not performed for CanX-7 since both curves did not possess any local maximas near that frequency.

![Figure 27: UHF Frequency Sweep](attachment:image.png)
Chapter 2. CanX-7 Antenna Modeling and Measurement

Isolation

From Figures 28a and 28b, the isolation measured at the UHF and S-band sweep center frequencies (437 and 2235 MHz) were 57.4 dB and 66.2 dB respectively. These were expected and satisfactory values for these arrays.

![Figure 28: CanX-7 Isolation](image)

2.8 Summary

The performance of the uplink and downlink antenna arrays on the CanX-7 nanosatellite have been successfully validated through simulation and measurement. Initially, the original CanX-2 derived HFSS model was updated to account for new materials and payloads. After which, 3D radiation patterns were simulated for both sail stowed and deployed cases. It was confirmed that the sail had noticeable effects on coverage due to a reflective or scattering effect on the radiation patterns. Nonetheless, it was found that the simulated antenna arrays possessed adequate coverage to meet requirements for both the stowed and deployed setups. The results of this simulation also provided increased confidence and baseline values for the subsequent real world antenna measurement.

Antenna measurement was only performed for the sail stowed case. The sail deployed case was dropped and the requirement declared “non-compliant” due to the increased logistics of transporting and testing deployed sails along with the limited utility of guaranteeing communications during deorbiting. Four pattern cuts each were measured for the S-band in RHCP and UHF in both polarization senses. Additionally, the isolation between the S-band and UHF antennas was also measured for characterization purposes. Four pattern cuts were used as this arrangement was cost effective and has provided reliable coverage estimates for previous SFL spacecraft. It was verified geometrically that the measured pattern cuts were similar to those simulated by HFSS. However, the width of the UHF LHCP null increased significantly causing its coverage to fall drastically. Despite this, both antenna arrays still successfully fulfilled requirements. Coverage was also calculated over a UHF frequency sweep but did not provide any clear optimization points for tuning the UHF array. While this test proved successful, it is recommended in future antenna measurement campaigns that the major drop in UHF LHCP coverage be investigated further if it reoccurs. Diagnostic tests may include increasing the amount of pattern cuts in a coverage calculation or adopting orientations and test structures that do not enter the turnstile area.
Chapter 3

Antenna Design for the CanX-7
ADS-B Payload

Air traffic services (ATS) play a fundamental role in both civil and military aviation by tracking and directing aircraft on the ground and in controlled airspace as well as providing advisory services to aircraft in non-controlled airspace [12]. These services are provided for the purposes of preventing collisions, organizing traffic, and providing information and support to pilots.

Primary and secondary surveillance radar systems have traditionally been used by ATS to track aircraft. While proven to be a useful and reliable means of aircraft tracking, radar installations can be costly and impractical to install. Consequently, there is very limited radar coverage over much of the planet especially in remote and oceanic regions. In areas outside of radar coverage, air traffic is managed by assigning aircraft to different flight attitudes with longitudinal spacing along flight paths. These margins between aircraft positions are mandated by the International Civil Aviation Organization (ICAO) and are large to account for uncertainties in instrumentation, position information, and flight path [37].

Automatic Dependent Surveillance - Broadcast (ADS-B) is a new cooperative surveillance technology for air traffic management (ATM). This method relies on a cooperative aircraft which determines its own position using GNSS (Global Navigation Satellite System) and automatically broadcasts it along with itinerary and identification information to other aircraft and ground stations within its vicinity. ADS-B is known as “dependent” surveillance as it requires cooperation and contributions from the tracked target to function properly [12]. As part of the Single European Sky ATM Research (SESAR) program and the U.S. Federal Aviation Administration (FAA) Next Generation of Air Transportation System (NextGen) plan, ADS-B is expected to be adopted as the global standard for ATM before 2020 [17]. ADS-B ground stations are currently deployed as an add-on to current radar sites or in areas where traditional radar sites are not cost effective allowing the extension of ATM services.

Despite the increased flexibility of ADS-B Ground Stations, the installation of stations for much of the world is still not cost effective and too complex. This is especially the case in very remote or oceanic areas. One such means to circumvent this issue is to utilize a constellation of satellite based ADS-B receivers. This constellation could provide global coverage would allowing aircraft to take more direct routes to their destinations without having to resort to inefficient and time consuming techniques to avoid separation which would save time and fuel.
As a secondary mission, CanX-7 is to demonstrate the feasibility of an orbital ADS-B receiver on a nanosatellite platform. The ADS-B payload will be operated for 6 months on-board the CanX-7 spacecraft in LEO. The payload was originally to be provided by COM DEV Ltd. and the Royal Military College (RMC). Due to resource availability and scheduling issues, SFL took over the design and implementation of the ADS-B payload in early 2015. Part of the tasks related to this hand off was the continuing design of the ADS-B antenna for this payload.

### 3.1 Chapter Overview

This chapter summarizes and documents the design of an ADS-B antenna for CanX-7’s secondary payload. The focus of this chapter is on the iterative design of this antenna. The arrangement of this chapter is as follows:

1. **ADS-B Overview**: This section covers the history and workings of ADS-B. The benefits of space based ADS-B receivers are also presented as well as the similarities with space based ship tracking systems. This section concludes with a synopsis of the CanX-7 secondary payload.

2. **Driving Requirements**: Relevant SFL subsystem level requirements pertaining to the ADS-B payload are described along with their effect on the antennas design. Additional, antenna design constraints and performance requirements that are not recorded as SFL requirements are also presented.

3. **Antenna Design Specifications**: Expanding on the antenna performance parameters described in Chapter 2 this section defines additional antenna parameters relating to antenna design.

4. **Patch Antenna Overview & Design Characteristics**: The anatomy of patch antennas, common designs, and configurations are presented. Electromagnetic transverse modes are also introduced to provide insight to how patch antennas radiate and how circular polarization can be induced.

5. **Antenna Design**: This section summarizes the iterative design process of the ADS-B antenna. Each design is performed through modeling and simulation. Previous iterations were discarded either due to out of spec materials, inadequate performance, or electromagnetic incomparability with the spacecraft bus.

### 3.2 ADS-B Overview

ADS-B is based upon a Mode S extended squitter operating at 1090 MHz. The definition of a squitter is a reply format transmission which is constantly being transmitted without being interrogated by a receiving station [38]. Mode S (Select) is a secondary surveillance radar (SSR) interrogation process that was developed to improve upon SSR methods which had reached their operational limits. To clarify, a SSR differs from a primary surveillance radar (PSR) in that it requests information from a cooperative target equipped with a transponder. The DF17 extended squitter transmitted from an aircraft based Mode S transmitter forms an integral and working part of ADS-B.

This DF17 extended squitter continuously broadcasts the following information packets: airborne position, surface position, extended squitter status, aircraft identity and category, and airborne velocity.
The airborne position packet contains GNSS longitude, latitude, barometric altitude, GNSS height, and surveillance status. Similarly, surface positional information includes longitude, latitude, movement, and heading of the aircraft. The extended squitter status report is comprised of surface squitter rate, altitude type and extended squitter status. Moreover, the aircraft identity and category field reports the ABS-B emitter category which ranges from “no reporting” to surface vehicle to space vehicle. Lastly, the airborne velocity reports east/west and north/south velocities, heading, vertical rate and attitude source, the difference between the aircraft’s barometric and GNSS altitude, IFR (Instrument Flight Rules) capability, and airspeed.

ADS-B is comprised of two operational components, \textit{ADS-B out} encompasses the flight parameter informational output from an aircraft transmitted with a refresh rate of 0.5 seconds to ground stations and other aircraft. \textit{ADS-B in} refers to the reception of transmitted ADS-B broadcasts by other aircraft \cite{39} allowing an aircraft with this component to determine nearby air traffic. As transmitting elements are located on aircraft, ADS-B ground components only need to consist of passive receiving components. This is contrasted to PSR which must continuously transmit an electromagnetic (EM) beam and SSR which has to actively interrogate its targets. As such, ADS-B ground stations are mechanically simpler, cost effective, and require less power than traditional radar installations. Consequently, this simplicity allows for greater flexibility in the installation of ground stations. Despite the increased flexibility, there remains limitations on where ADS-B ground stations can be installed such as at sea or highly remote areas. One solution to remedy this lack of coverage is to place ADS-B receivers on orbital platforms.

Receivers based on orbital platforms have already been used with the VHF (Very High Frequency) based AIS (Automatic Identification System) transponder system used for tracking maritime vessels. Two AIS receivers, NORAIS (Norwegian Automatic Identification System) and LUXAIS (Luxembourg Automatic Identification System) have been deployed by the European Space Agency (ESA) onboard the International Space Station (ISS) \cite{40}. SFL has also developed missions with AIS payloads such as NTS (Nanosatellite Tracking Ships) \cite{41}, the AISSat constellation \cite{42}, and M3MSat (Maritime Monitoring and Messaging Microsatellite) \cite{43}.

### 3.2.1 CanX-7 ADS-B Payload

![ADS-B Receiver Payload](image1.png) ![ADS-B Receiver + Payload Computer](image2.png)

Figure 29: ADS-B Receiver Payload (Left) and ADS-B Receiver + Payload Computer (Right) \cite{44}

Figure 29 shows the self-contained ADS-B payload for CanX-7 as well as the as well as the payload
Chapter 3. Antenna Design for the CanX-7 ADS-B Payload

electronics. The payload electronics are based on a Beaglebone computer. Reception and decoding of ADS-B messages is performed by a Radarcape receiver board developed by Planevision Systems. For the CanX-7 payload, the Radarcape was configured to operate without an attached GPS module (Grey cable in Figure 29. Furthermore the Ethernet, USB, and power ports on the payload computer were removed to reduce power consumption [44]. SFL took over the design and development of the ADS-B payload from RMC in early 2015. In order to finish the design, link analysis was completed and updates were made to the both the payload hardware and software. In particular, the L-band patch antenna had to be completely redesigned.

A circularly polarized antenna is required for this payload as ADS-B aircraft transmitters operate in vertical polarization. As such, a circularly polarized receiver antenna reduces and keeps constant polarization mismatch losses between CanX-7 and the transmitting aircraft. This is especially important as CanX-7 only possesses 2-axis attitude control through local magnetic field tracking [14]. Additionally, the receiver must receive signals from multiple moving targets scattered over a wide area making it extremely difficult for the spacecraft to maintain an optimal alignment for reception. It should also be noted that circular polarization will negate the need to compensate for Faraday rotation of the ADS-B signals [44].

3.3 Driving Requirements

As the ADS-B payload was originally developed externally, there exist few internal SFL requirements directly relating to the design of the antenna. Instead the antenna was required to fit more specific design criteria to enable proper functionality of the secondary payload:

1. **Operating Frequency:** The ADS-B antenna shall possess an operating frequency of 1090 MHz to receive ADS-B signals.

2. **Antenna Gain:** ADS-B antenna gain should be as high as possible without interfering with the other design requirements. A wide beamwidth is also desired to enable a wide, usable field of view.

3. **Form Factor:** The antenna is limited to a maximum area of 75\times75\text{ mm} and thickness of 12.2\text{ mm.} This enables the antenna to fit between the bus’ rails and for CanX-7 to fit within its deployment system.

4. **Circular Polarization:** A circularly polarized antenna is able to receive signals from linearly polarized transmitting antennas with a fixed polarization mismatch loss (3 dB) regardless the attitude of CanX-7 and the transmitting aircraft. A circularly polarized antenna also mitigates the effects of Faraday rotation. As a general rule, circular polarization is considered achieved with an axial ratio \leq 3\text{ dB}. High quality antennas will possess axial ratio minimas \leq 1\text{ dB.}

5. **Single Feed:** The secondary payload structure is designed for a single feed antenna and there also exists a desire to avoid the extra mass, volume, and complexity of adding a combiner required for a dual fed circularly polarized antennas.

6. **Thermal Stability:** Based on thermal analysis of the CanX-7 spacecraft, the externally mounted ADS-B antenna is expected to see temperatures ranging from -20 to 60\degree\text{ C}, therefore, the antenna must not be sensitive to changes in temperature. This applies to temperature dependent changes in the antenna substrate’s dielectric constant and the antenna’s thermal expansion.
7. Non-Magnetic: Reduces interference with other spacecraft components.

3.4 Antenna Design Specifications

This section expands upon the antenna performance overview section in Chapter 2 (Section 2.3). Here, additional terminology is introduced which is more relevant to the design and characterization of antennas.

3.4.1 Impedance

Impedance ($Z$) is known as the opposition a circuit provides to a current when voltage is applied. This property is often a complex property with electrical resistance $R$ as its real component and reactance $X$ as the complex component, both expressed in Ohms ($\Omega$):

$$Z = R + jX$$

(3.1)

Electrical resistance is the opposition of the passage of electric current through a conductor while reactance is the opposition to a change in either current (inductive) or voltage (capacitive) by a circuit element. As such, an ideal resistor possess zero reactance while ideal inductors and capacitors possess no resistance. Reactance can be either a positive or negative value. A reactance of $X = 0$ means a purely resistive impedance, $X > 0$ is inductive, while a reactance of $X < 0$ is capacitive. It should be noted that admittance is known as the inverse of impedance.

Impedance matching is a significant consideration in antenna design. Transmission lines typically possess a characteristic impedance $Z_0$ of 50 $\Omega$ (75 $\Omega$ is also common for some applications). If the input impedance $Z_{in}$ of the antenna is not equal to this value, some power will be reflected at this discontinuity and will not make it to the antenna or receiver. This is especially problematic for transmitting antennas as the reflected power may also cause damage to the transmitter and cable. Impedance matching is usually verified in terms of a pass/fail criteria and can by quantified by the following characteristics.

Return Loss

The quality of impedance matching is often expressed in terms of a return loss ($L_{\text{Return}}$); the loss of input power due to a transmission line discontinuity. The return loss can be expressed in terms of incident power $P_i$ and reflected power $P_r$.

$$L_{\text{Return}}(dB) = 10 \log_{10} \frac{P_i}{P_r}$$

(3.2)

Return loss can also be specified as the negative of the reflection coefficient ($\Gamma$) in dB.

$$L_{\text{Return}}(dB) = -20 \log_{10} |\Gamma|$$

(3.3)

The reflection coefficient (also known as the 2-port $S_{11}$ parameter) can be calculated from voltage or impedance.

$$\Gamma = \frac{V_r}{V_i} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

(3.4)
In practice, an $S_{11} \leq -10$ dB is considered good impedance matching.

**Voltage Wave Standing Ratio**

The Voltage Wave Standing Ratio (VSWR) is a common measure of how well an antenna is matched to a feed and is expressed as a real number equal or greater than 1. A VSWR of 1 indicates no mismatch loss while higher values indicate higher levels losses. VSWR is calculated as a product of the reflection coefficient.

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$

It should be emphasized that impedance matching parameters such as $S_{11}$ and VSWR alone are not reflective of the performance or functionality of the antenna but only of the power that is delivered to the antenna; the potential of the antenna to radiate.

### 3.4.2 Cross-Polarization

In practice, antennas are never fully polarized in one sense and will radiate in other polarizations. Cross-polarization is thus defined as the polarization orthogonal to the design polarization (the inverse being co-polarization). For example, a vertically polarized antenna will emit some EM waves in horizontal polarization and an right-handed circular polarized (RHCP) antenna will emit some EM waves in left-handed circular polarized (LHCP).

Cross-polarized radiation patterns may be presented together with their co-polarized counterparts. Additionally, cross-polarization is generally specified in terms of a negative power level indicating how many decibels below the co-polarized power level it is. The co-/cross-polarization ratio is the ratio of the desired polarization component to the undesired component and is a measure of polarization purity.

### 3.4.3 Bandwidth

The bandwidth covered by the antenna is often an important design parameter of an antenna. However, only the impedance (return loss) bandwidth of an antenna is usually quoted. Other bandwidth definitions exist and may also be important depending on design requirements. These include bandwidths defined on directivity/gain, efficiency, polarization, and axial ratio.

**Impedance Bandwidth**

This is defined as the frequency range for which the structure or network possess good impedance matching with respect to a reference impedance (Usually the characteristic impedance of the feed) and can be defined either in terms of the real impedance, VSWR, or return loss ($S_{11}$). This property is dependent on properties of the patch antenna itself such as the type and the positioning of feed used as well as the antenna’s Q(Quality)-factor. The Q-factor is a measure of the operating band of the antenna ($\Delta f$) relative to the center frequency ($f_r$) of the antenna:

$$Q = \frac{f_r}{\Delta f}$$

Antennas with a high Q are considered narrowband while those with a low Q are wideband. A narrowband antenna’s input impedance is more sensitive to minor changes in frequency.
3.4.4 Additional Bandwidth Definitions

Additional bandwidth definitions include:

- **Directivity/Gain Bandwidth**: Range of frequencies where the antenna meets a gain or directivity requirement.

- **Efficiency Bandwidth**: Range of frequencies where the antenna meets an efficiency requirement.

- **Polarization Bandwidth**: Range of frequencies where the antenna possesses an acceptable co-/cross-polarization ratio.

- **Axial Ratio Bandwidth**: Range of frequencies where the antenna possesses a desired type of polarization. Can be related to polarization bandwidth.

3.5 Patch Antenna Overview and Design Characteristics

A patch antenna (often known as a microstrip antenna) is a popular printed antenna often possessing a narrow bandwidth and wide beamwidth. The first patch antenna was invented by Howell in 1972 [45] and consisted of two metal sheets approximately one-half wavelength in length with a space between them forming a resonate piece of microstrip transmission line. A modern patch antenna consists of a radiating element etched on metal and bonded on to an insulated dielectric substance layered on a continuous metal backing. Patch antennas are very attractive for aerospace applications due to their slim profile, ease of manufacture, and cost. However, drawbacks include narrow impedance bandwidth, low efficiency, and suitability for only low power applications. Patch antennas are also very configurable and a designer can alter the antenna’s structure, shape, substrate, and feeding technique to meet design requirements.

Figure 30 displays the anatomy of a basic patch antenna, comprising of four main parts: the patch, dielectric substance, ground plane, and feed.

Figure 30: Components of a Basic Patch Antenna
3.5.1 Patch

The patch is mounted to one side of the substrate and is comprised of a thin metal sheet. While the most popular patch shapes are the rectangular and circular patch, patches may take other geometries to meet design requirements.

3.5.2 Dielectric Substrate

There are a wide variety of options for the dielectric material between the patch and ground plane. Typically, the two main properties which define the dielectric substrate are substrate height \( (0.003\lambda \leq h \leq 0.05\lambda) \) which is constrained by the operating wavelength \( \lambda \) and dielectric constant \( \epsilon_r \). An increase in dielectric constant leads to a smaller antenna profile. The dielectric substrate may also be selected based on other qualifying factors. For instance an antenna developed for space applications will be operating in a vacuum and may have to contend with cycling temperature extremes. As such, factors such as outgassing, thermal expansion, and the sensitivity of the dielectric constant to changes in ambient temperature would also be considered.

Increasing the thickness of the substrate increases the gain of the antenna, however doing this may lead to undesirable emissions which may decrease antenna efficiency and perturb the radiation pattern. Furthermore, a thicker substrate requires a longer feed probe which may increase the difficulty of impedance matching. In general, the distance between the edge of the patch and the edge of the substrate should be two to three times the thickness of the substrate to permit proper operation of the antenna.

3.5.3 Ground Plane

The ground plane is the metallic layer covering the side of the substrate opposite the patch. Antenna gain may also be increased by increasing the size of the ground plane. However, the effectiveness of this technique is limited past a certain ground plane size. Other perturbations such as inserting slots and shapes may also enhance certain antenna performance aspects.

3.5.4 Feed

Four popular techniques of feeding the patch in a patch antenna are described below. These techniques are: direct feeding, probe feeding, coupled feeding, and aperture feeding (Figure 31) [46].

![Figure 31: Patch Antenna Feeding Techniques (Feeds Indicated in Red)](image-url)
3.5. Patch Antenna Overview and Design Characteristics

Microstrip Feed

A microstrip feed consists of a microstrip transmission line which feeds directly to the edge of the patch. This method of feeding is simple to model, match, and fabricate as the feed and patch form a singular structure and can be connected to electronic components without the need for a multi-layer board. However, microstrip fed antennas possess narrow bandwidth and require a larger profile. They may also experience coupling between the feed line and patch which can cause spurious radiation emissions. Three popular techniques of microstrip feeding exist (Figure 32):

- **Inset Feed**: An inset feed allows the feeding point to be set within the patch and is popular for feeding microstrip arrays. This allows the input impedance to be tuned by modifying the thickness and depth of the air gap around the feed. It should be noted that the gap may cause distortion of the radiation pattern.

- **Direct Feed**: In a direct feed, the microstrip feed is attached to the edge of the patch. Due to the direct connection, a matching network is needed between the feed line and patch for proper impedance matching. This is often performed using a quarter wavelength transformer to compensate for the impedance differences between the patch and the feed line.

- **Gap-Coupled Feed**: In a gap-coupled feed antenna, the feed line does not directly contact the patch and feeding is done through coupling. Coupled feeds introduce another degree of freedom into the antenna design as the capacitance of the gap can be used to cancel out the inductance of the feed line allowing for improved impedance matching.

![Figure 32: Microstrip Feed Types](image)

Probes (Coaxial) Feeding

The patch can be fed from the underside using a probe feed. In this method, the center conductor of a coaxial cable is connected to the patch through the substrate while the outer conductor is connected to the ground plane. Impedance tuning is performed by adjusting the location of the feed on the patch in the same manner as with inset feeding. The input impedance $Z_{in}(R)$ is given as:

$$Z_{in}(R) = Z_0 = \cos^2 \left( \frac{\pi R}{L} \right) Z_{in}(0)$$  \hspace{1cm} (3.7)
Where $Z_0$ is the characteristic impedance of the coax feed which the antenna is to be matched to, $R$ is the distance of the feed from the edge of the patch, $L$ is the length of the patch along its resonant axis, and $Z_{in}(0)$ is the input impedance if the patch was fed at the edge. As with microstrip feeding, the probe can be trimmed so that it does not completely meet the patch for a coupled feed.

Probe feeding is useful for single antennas as they are easy to manufacture, impedance match, emit lower amounts spurious radiation, and possess a smaller profile. However, probe feeding requires a high amount of soldering and drilling precision, making it more difficult to tune after fabrication and to integrate into microstrip arrays. Moreover, probe feeding results in narrower bandwidths and the length of the feed is dependent on the substrate thickness. With a thicker substrate, a longer probe will be needed which will increase feed inductance and possible surface waves degrading impedance matching and antenna efficiency.

**Proximity Coupled Feeding**

Some patch antennas may use multiple layers of substrates. One example of this design is the proximity coupled fed antenna. This type of antenna consists of two different substrate layers where the patch is placed on the top and a microstrip feed is sandwiched between the two substrates. This type of antenna has the benefit of low spurious radiation and increased bandwidth as well as the ability to optimize antenna performance by using different substrates. However, precise alignment of the feed is required for proper operation and multilayer fabrication is more difficult.

**Aperture Feeding**

The aperture fed antenna is another coupled fed multi-substrate patch antenna. In this case, the microstrip feed line is placed at the bottom while the ground plane is sandwiched between the two substrates. An aperture cut into the ground plane allows energy to be transmitted from the microstrip feed line to the patch while shielding the patch from the rest of the feed circuitry. This antenna possesses the same benefits and shortcomings of the proximity coupled fed antenna. However, the aperture reduces spurious radiation and cross-polarization while increasing antenna bandwidth and improving polarization purity.

### 3.5.5 Transverse Modes

Patch antennas operate in the $TM_{10}$ fundamental mode. The electric field distribution on the patch along the resonant length possesses a maximum (+) electrical potential on one side of the patch, a minimum (−) on the other side, with no potential in the center (Figure 33). These maximas and minimas of the electric field constantly oscillate and change sides. Consequently, the current distribution forms a maximum in the center of the patch and is zero on both sides. The distribution of the electric field is not limited by the edges of the patch and the field will often extend past the patch edges in what are known as fringing fields. These fringing fields allow the patch antenna to radiate. A propagating electromagnetic beam in a waveguide can be described by its transverse mode. A transverse mode is the EM field pattern of radiation measured in a plane perpendicular to the propagation direction of the EM beam. These modes occur due to boundary conditions imposed on the beam by external physical structures such as a waveguide where each mode follows different propagation constraints. Four classifications of transverse modes exist [35]:

---

1. $TM_{10}$
2. $TM_{20}$
3. $TM_{21}$
4. $TM_{30}$

---
3.5. Patch Antenna Overview and Design Characteristics

- **Transverse Electromagnetic (TEM) Modes**: No electric or magnetic field in the direction of propagation.

- **Transverse Electric (TE) Modes**: No electric field in the direction of propagation. Also known as H-modes since there is only a magnetic field along the propagation direction.

- **Transverse Magnetic (TM) Modes**: No magnetic field in the direction of propagation. Also known as E-modes as there is only an electric field along the propagation direction.

- **Hybrid Modes**: Possess non-zero electric and magnetic fields in direction of propagation.

Modes are designated using their abbreviations with three subscript numbers (i.e. $TM_{xyz}$) which correspond the field variations in the x, y, and z direction respectively. The patch antenna in Figure 33 operates in the TM mode as magnetic fields in patch antennas are always transverse to the Z-axis (direction of propagation). Additionally, the “$z$” value can be omitted since only a phase variation exists in the electric field in the Z-axis (the electric field variation in the Z-axis can be considered negligible). The patch antenna in Figure 33 also indicates that an electric field variation only exists along the X direction. Therefore, “$x$” and “$y$” are “1” and “0” respectfully and the patch antenna is said to be operating in the $TM_{10}$ mode [47]. Due to their shape, the fundamental mode for round patches is $TM_{11}$. Higher modes also exist, although only a few of them are useful for patch antennas.

### 3.5.6 Inducing Circular Polarization

Without modification, patch antennas will generally operate in one fundamental mode and are inherently linearly polarized. Circular polarization can be induced in a patch antenna by making modifications to
the antenna so that the electric field varies in two orthogonal directions $90^\circ$ out of phase and equal in magnitude. This results in the simultaneous excitation of two transverse modes. For instance, a square patch similar to the one in Figure 33 will radiate in both the X and Y directions when operating in circular polarization (i.e. modes $TM_{10}$ and $TM_{01}$). Typical circularly polarized designs are dual feed and single feed antennas.

**Dual Feed**

Dual feeding is the simplest means for creating a circular polarized patch antenna (Figure 34). In this design, two feeds are placed on the patch along perpendicular axes. An external polarizer is then used to feed the patch with two signals possessing equal magnitudes but $90^\circ$ out of phase. While this method easily and reliably enables circular polarization, extra electronic components are needed, increasing the antenna’s profile while introducing additional losses and complexity into the design. The axial ratio bandwidth of this type of antenna is determined by the bandwidth of the external polarizer.

![Figure 34: Dual Feed Antennas](image)

**Single Feed**

Circular polarization can also be induced within patch antennas using a single feed. Single feed patches are considered to be some of the simplest circularly polarized antennas. These antennas also retain the benefits of their linearly polarized counterparts of being easy to manufacture, low in cost, and compact. Slots, shapes, and other perturbation segments are cut into the patch to split the E-field into two orthogonal modes required for circular polarization. A wide variety of simple to complex perturbation segments exist (Figure 35) and the dimensions of each segment must be fine-tuned to the design frequency. Perturbation segments also increase the antenna’s impedance bandwidth but are more difficult to optimize and possess lower axial ratio bandwidths than dual feed antennas.

3.6 Antenna Design

The ADS-B antenna was designed using ANSYS HFSS (High Frequency Structural Simulator). Parametric studies were used to experiment with different antenna dimensions and to optimize each design.
3.6. Antenna Design

3.6.1 Design Methodology

The first step to designing a circular polarized patch antenna was to analytically size a linearly polarized microstrip antenna. After determining the type and dimensions of the patch from the desired thickness and substrate, a perturbation segment would be introduced into the patch. After which, parametric studies would be used to optimize the patch, feed position, and the dimensions of the perturbation segment. Analytic sizing was only performed on the first antenna design iteration as any additional changes to the antenna’s substrate, thickness, and patch properties were re-optimized using HFSS parametric studies in latter iterations. Each design iteration is referenced by a designator for convenience (I1, I2, etc.).

3.6.2 Iteration 1: Initial TMM6 Antenna Design

A number of patch antenna designs were considered for the ADS-B antenna and due to the profile requirements and time constraints, a single substrate coaxial fed design was selected. The single probe feed provided the most compact design and would be compatible with the payload’s structure. Due to time constraints, multi-layered designs were not considered. The first antenna (Figure 36) was sized using TMM6 substrate ($\varepsilon_r = 6$) with a thickness ($h$) of 1.5 mm for a circular and rectangular patch. TMM6 was selected as a substrate due to its high dielectric constant and its thermal coefficients of
expansion ($\Lambda_L = 18$ ppm/k) and $\epsilon_r$ ($\Lambda_{\epsilon_r} = -11$ ppm/K) [48]. Additionally, TMM6 has been used in other L-band antenna designs [44], [48]. The radius ($r$) of a circular patch can be estimated as follows:

$$r = \frac{F}{\left\{1 + \frac{2h}{\pi r \epsilon_r} \ln \left( \frac{\pi r}{2h} \right) + 1.7726 \right\}^{\frac{1}{2}}}$$

(3.8)

Where:

$$F = \frac{8.791 \times 10^9}{f_r \sqrt{\epsilon_r}}$$

(3.9)

and $f_r$ is the design frequency. Due to the fringing effect of the electric field, the patch is electrically larger than its physically size since the electric field extends past the boundaries of the patch. To correct for this effect, the effective dimension for the patch is then calculated for use. The effective radius ($r_e$) of a circular patch antenna is [49]:

$$r_e = r \left\{1 + \frac{2h}{\pi r \epsilon_r} \ln \left( \frac{\pi r}{2h} \right) + 1.7726 \right\}^{\frac{1}{2}}$$

(3.10)

Additionally, the width ($W$) and effective length ($L_e$) for a rectangular patch can be estimated as follows.

$$W = \frac{c}{2f_r}\sqrt{\frac{2}{\epsilon_r + 1}}$$

(3.11)

where $c$ is the speed of light ($2.99792458 \times 10^8$ m/s). The effective length to be used factors in a calculated effective dielectric constant $\epsilon_{reff}$ and a length correction $\Delta L$.

$$\epsilon_{reff} = \frac{\epsilon_r + \frac{1}{2} + \frac{\epsilon_r - 1}{2} \left[1 + 12 \frac{h}{W}\right]^{-\frac{1}{2}}}{2}$$

(3.12)

Where $\frac{W}{h} > 1$

$$\frac{\Delta L}{h} = 0.412 \frac{(\epsilon_{reff} + 0.3) \left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258) \left(\frac{W}{h} + 0.8\right)}$$

(3.13)

The effective length can then be calculated:

$$L_e = \frac{c}{2f_r \sqrt{\epsilon_{reff}}} - 2\Delta L$$

(3.14)

Initial sizing indicated that for the operating frequency of 1090 MHz, either a 61.5 mm diameter circular patch or a 73.4 \times 50$ mm rectangular patch would be required. Since the rectangular patch was almost as wide as the permissible substrate size, the circular patch was chosen for the subsequent designs. Circular polarization (RHCP) was induced by cutting a cross slot cut into the patch. Contrary to common cross slot designs which maintained a constant cross thickness, it was found that greater control could be leveraged over the antenna’s properties by adjusting both thicknesses and length of each cross arm (Figure 37).

HFSS parametric studies were performed to optimize the dimensions of the cross slot to achieve a minimum boresight axial ratio, and the input probe position for impedance matching. Near the end of the design process, the thickness of the substrate was also increased to 0.68 mm due to difficulty with obtaining an optimal axial ratio. Figure 36 shows the first successful modeled antenna on a 10
mm thick aluminum thick base fed using a Teflon coax input (Design: I1). Additionally, Table 4a shows the dimensions of this antenna based on the template shown in Figure 37 and Table 4b indicates the performance of this design. While this antenna possessed a good axial ratio, it suffered from two main issues. First, it was discovered after design and simulation that the substrate thickness was not a standard thickness that could be manufactured. Secondly, this antenna also suffered from a very low peak gain. As such, the antenna had to be redesigned using standard Rogers thicknesses and with increased antenna gain. It should be noted that at the time of the design of this antenna, it was believed that constraining the bandwidths would reduce antenna based noise. However this design point was moot due to the fact that the ADS-B payload contained internal bandpass filters.

Table 4: I1 – Initial TMM6 Antenna

<table>
<thead>
<tr>
<th>(a) Design Properties</th>
<th>(b) Performance Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Substrate Type</strong></td>
<td><strong>Peak Gain</strong></td>
</tr>
<tr>
<td>TMM6 ($\epsilon_r = 6.0$)</td>
<td>-0.2187 dBiC</td>
</tr>
<tr>
<td><strong>Substrate Thickness ($h$)</strong></td>
<td>0.68 mm</td>
</tr>
<tr>
<td><strong>Substrate Size</strong></td>
<td><strong>Gain at Boresight</strong></td>
</tr>
<tr>
<td>$75 \times 75$ mm</td>
<td>-0.2474 dBiC</td>
</tr>
<tr>
<td><strong>Feed Position ($p$)</strong></td>
<td>11.82 mm</td>
</tr>
<tr>
<td><strong>Patch Diameter ($2r$)</strong></td>
<td>61.2 mm</td>
</tr>
<tr>
<td><strong>X₁ Length</strong></td>
<td><strong>Half Power Beamwidth</strong></td>
</tr>
<tr>
<td>17 mm</td>
<td>-45 to 50°</td>
</tr>
<tr>
<td><strong>X₁ Thickness</strong></td>
<td><strong>Reflection Coefficient at $f_r$</strong></td>
</tr>
<tr>
<td>2.98 mm</td>
<td>-24.3 dB</td>
</tr>
<tr>
<td><strong>X₂ Length</strong></td>
<td><strong>Impedance Bandwidth</strong></td>
</tr>
<tr>
<td>17.5 mm</td>
<td>($S_{11} \leq -10$ dB)</td>
</tr>
<tr>
<td><strong>X₂ Thickness</strong></td>
<td>10.6 MHz</td>
</tr>
<tr>
<td>1.65 mm</td>
<td><strong>Boresight Axial Ratio</strong></td>
</tr>
<tr>
<td><strong>Cross Slot Rotation ($\kappa$)</strong></td>
<td>45°</td>
</tr>
<tr>
<td></td>
<td><strong>Axial Ratio Beamwidth</strong></td>
</tr>
<tr>
<td></td>
<td>($P_{AR} \leq 3$ dB)</td>
</tr>
<tr>
<td></td>
<td>-80 to 67.8°</td>
</tr>
<tr>
<td></td>
<td><strong>Axial Ratio Bandwidth</strong></td>
</tr>
<tr>
<td></td>
<td>($P_{AR} \leq 3$ dB at Boresight)</td>
</tr>
<tr>
<td></td>
<td>2.7 MHz</td>
</tr>
</tbody>
</table>
3.6.3 Iteration 2: TMM4 Antenna Design

Following the Initial I1 TMM6 design, a 75 \times 75 \text{ mm} TMM4 antenna was sized and optimized (Design: I2). TMM4 was used as its lower dielectric constant ($\epsilon_r = 4.5$) contributed to higher gain while keeping the patch diameter under 75 mm. Furthermore, material selection was constrained to the TMM line due to time constraints and since materials in this line possessed similar, stable thermal properties (TMM4: $\Lambda_L = 16$ ppm/k, $\Lambda_{\epsilon_r} = +15$ ppm/K) [48]. Antenna gain was further maximized by thickening the substrate to a manufactured thickness of 3.175 mm. Tables 5a and 5b show the dimensions and performance of the TMM4 antenna design respectively.

Table 5: I2 – TMM4 Antenna

<table>
<thead>
<tr>
<th>(a) Design Properties</th>
<th>(b) Performance Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Type</td>
<td>TMM4 ($\epsilon_r = 4.5$)</td>
</tr>
<tr>
<td>Substrate Thickness ($p$)</td>
<td>3.175 mm</td>
</tr>
<tr>
<td>Substrate Size</td>
<td>75 \times 75 \text{ mm}</td>
</tr>
<tr>
<td>Feed Position ($p$)</td>
<td>8.88 mm</td>
</tr>
<tr>
<td>Patch Diameter ($2r$)</td>
<td>73.0 mm</td>
</tr>
<tr>
<td>$X_1$ Length</td>
<td>17.49 mm</td>
</tr>
<tr>
<td>$X_1$ Thickness</td>
<td>2.38 mm</td>
</tr>
<tr>
<td>$X_2$ Length</td>
<td>14.14 mm</td>
</tr>
<tr>
<td>$X_2$ Thickness</td>
<td>0.81 mm</td>
</tr>
<tr>
<td>Cross Slot Rotation ($\kappa$)</td>
<td>45°</td>
</tr>
<tr>
<td>Peak Gain</td>
<td>5.0 dBiC</td>
</tr>
<tr>
<td>Gain at Boresight</td>
<td>4.87 dBiC</td>
</tr>
<tr>
<td>Half Power Beamwidth</td>
<td>$-45$ to $50^\circ$</td>
</tr>
<tr>
<td>Reflection Coefficient at $f_r$</td>
<td>-20.2 dB</td>
</tr>
<tr>
<td>Impedance Bandwidth ($S_{11} \leq -10$ dB)</td>
<td>22 MHz</td>
</tr>
<tr>
<td>Boresight Axial Ratio</td>
<td>0.07 dB</td>
</tr>
<tr>
<td>Axial Ratio Beamwidth ($p_{AR} \leq 3$ dB)</td>
<td>$-77.7$ to $7^\circ$</td>
</tr>
<tr>
<td>Axial Ratio Bandwidth ($p_{AR} \leq 3$ dB at Boresight)</td>
<td>5.4 MHz</td>
</tr>
</tbody>
</table>

The I2 TMM4 design showed significant improvement over the I1 TMM6 design with a peak gain of 5 dB, a boresight axial ratio of 0.07 dB, and good impedance matching and beamwidth. With an acceptable and feasible antenna design, the I2 antenna was added to the CanX-7 HFSS model used for the simulation of the S-band and UHF arrays in Chapter 2. Fabrication of this antenna would use an MCX input plug similar to SFL S-band patch antennas. The pin diameter of this feed was 1 mm and was updated in the model. The I2 antenna can be seen in Figure 38 in both standalone and attached configurations.

Figure 38: I2 – TMM4 Antenna
3.6. Antenna Design

Upon attachment to the CanX-7 bus, it was observed that the center frequency shifted from 1090 MHz to 1101 MHz (Figure 39). This change was most noticeable in the axial ratio bandwidth where the boresight axial ratio at 1090 MHz increased to 21.1 dB causing the antenna to lose its circular polarization. This shift was also reflected in the radiation patterns shown in Figure 40 where cross- and co-polarization emissions in the boresight direction were almost equal in magnitude and RHCP gain at boresight dropped to 2.09 dBi.

![Figure 39: I2 – TMM4 Antenna Axial Ratio](image1)

![Figure 40: I2 – TMM4 Antenna Radiation (Gain) Pattern](image2)

The I3 and I4 antennas were developed to either mitigate or work around the frequency shifting issue. At this time, the design was still subject to time constraints and that the frequency shifting phenomena was not largely understood. Consequently, a more heuristic approach was taken in the design of these later iterations.
3.6.4 Iteration 3: 80 X 80 mm TMM6 Antenna Design

It was hypothesized that the shift in center frequency was due to the patch diameter being nearly as large as the substrate. This would mean that the antenna would not be electrically self-contained. As such, a new antenna was designed and optimized to place as much space between the edge of the patch and the substrate as physically possible (Figure 41).

![Figure 41: I3 – 80 × 80 mm TMM6 Antenna](image)

In an attempt to maximize the space between the edge of the patch and substrate, the substrate was switched back to TMM6 and expanded to 80 × 80 mm (Design: I3); the largest antenna size physically possible on CanX-7. However, this increased size would require the form factor constraint to be declared “non-compliant”. TMM6 was used again as its slightly higher dielectric content would allow the patch diameter to be reduced without sacrificing two much gain. For more accurate modeling, the aluminum backing and Teflon based coaxial feed was removed and the antenna probe pin was fed directly in the standalone setup. The dimensions and performance of the I3 antenna are shown in Tables 6a and 6b. Optimization was originally conducted for both the 3.175 mm and 3.81 mm thicknesses; the latter being chosen as it was easier to optimize based on the initial I2 cross slot dimensions.

**Table 6: I3 – 80 × 80 mm TMM6 Antenna**

<table>
<thead>
<tr>
<th>(a) Design Properties</th>
<th>(b) Performance Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Substrate Type</strong></td>
<td>TMM6 ($\epsilon_r = 6.0$)</td>
</tr>
<tr>
<td><strong>Substrate Thickness ($p$)</strong></td>
<td>3.81 mm</td>
</tr>
<tr>
<td><strong>Substrate Size</strong></td>
<td>80 × 80 mm</td>
</tr>
<tr>
<td><strong>Feed Position ($p$)</strong></td>
<td>8.88 mm</td>
</tr>
<tr>
<td><strong>Patch Diameter (2$r$)</strong></td>
<td>62.4 mm</td>
</tr>
<tr>
<td><strong>$X_1$ Length</strong></td>
<td>17.49 mm</td>
</tr>
<tr>
<td><strong>$X_1$ Thickness</strong></td>
<td>2.38 mm</td>
</tr>
<tr>
<td><strong>$X_2$ Length</strong></td>
<td>14.14 mm</td>
</tr>
<tr>
<td><strong>$X_2$ Thickness</strong></td>
<td>0.81 mm</td>
</tr>
<tr>
<td><strong>Cross Slot Rotation ($\kappa$)</strong></td>
<td>45°</td>
</tr>
</tbody>
</table>

After optimization, the I3 antenna was also attached to the CanX-7 model and fed via a Teflon...
3.6. Antenna Design

coaxial feed. It was found that the center frequency of this antenna shifted to 1120 MHz. As with the I2 TMM4 antenna, boresight axial ratio at 1090 MHz increased to 20.0 dB (Figure 42). Again, the boresight gain decreased to 1.5 dBi with similar cross- and co-polarized emissions (Figure 43).

![Figure 42: I3 – TMM6 Antenna Axial Ratio](image)

![Figure 43: I3 – TMM6 Antenna Radiation (Gain) Pattern](image)

3.6.5 Iteration 4: Bus Designed TMM6 Antennas

The following antennas were developed parallel to each other and optimized while attached to the spacecraft bus. This was done because it was found that the main 3U bus had the greatest effect on the frequency shift. Frequency shifting was observed even when the deployable boom, sail modules, and communication antennas were removed from the model. Therefore, it was predicted that optimization on a simplified CanX-7 HFSS model would be sufficient to develop a functioning antenna. The TMM6 substrate was retained to keep the patch diameter small while the substrate size was returned to 75 x 75
mm to meet the original profile requirement. Two spacecraft bus optimized antennas were developed using 3.175 mm (Design: I4a) and 5.08 mm (Design: I4b) thicknesses. It was found during the test phase of the I3 antenna that the 3.81 mm thick TMM6 substrate was not available by manufacturers. Consequently, the 3.175 mm thickness was used to retain the original dimensions in an attempt to develop an antenna that resembled the I3 design that would function on the spacecraft bus and comply with the form factor requirement. At the same time, there was a desire to determine if the gain on a TMM6 could be increased further. Hence, the a 5.08 mm thick I4b antenna was also developed using the next standard thickness above 3.81 mm (Figure 44). Time constraints prevented additional substrate and design experimentation. While the I4a antenna induced circular polarization using the previous cross slot perturbation segment, the I4b antenna introduced a rotational offset into a narrower, elongated cross slot. This rotated cross slot was found to aid optimization with a thicker substrate. The dimensions of both antennas are shown in Table 7.

![Antenna Diagrams](image)

(a) I4a – 3.175 mm Thick  
(b) I4b – 5.08 mm Thick + Rotated Cross Slot

Figure 44: TMM6 Bus Designed Antennas

<table>
<thead>
<tr>
<th>Antenna</th>
<th>I4a: 3.175 mm Thick Fixed Cross Slot Antenna</th>
<th>I4b: 5.08 mm Thick Rotated Cross Slot Antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Type</td>
<td>TMM6 ($\epsilon_r = 6.0$)</td>
<td>TMM6 ($\epsilon_r = 6.0$)</td>
</tr>
<tr>
<td>Substrate Thickness ($t$)</td>
<td>3.175 mm</td>
<td>5.08 mm</td>
</tr>
<tr>
<td>Substrate Size ($L \times W$)</td>
<td>75 $\times$ 75 mm</td>
<td>75 $\times$ 75 mm</td>
</tr>
<tr>
<td>Feed Position ($p$)</td>
<td>8.90 mm</td>
<td>9.75 mm</td>
</tr>
<tr>
<td>Patch Diameter ($2r$)</td>
<td>63.2 mm</td>
<td>62.62 mm</td>
</tr>
<tr>
<td>$X_1$ Length</td>
<td>17.1 mm</td>
<td>24.0 mm</td>
</tr>
<tr>
<td>$X_1$ Thickness</td>
<td>2.41 mm</td>
<td>0.27 mm</td>
</tr>
<tr>
<td>$X_2$ Length</td>
<td>15.7 mm</td>
<td>7.0 mm</td>
</tr>
<tr>
<td>$X_2$ Thickness</td>
<td>0.82 mm</td>
<td>0.21 mm</td>
</tr>
<tr>
<td>Cross Slot Rotation ($\kappa$)</td>
<td>$45^\circ$</td>
<td>$53.5^\circ$</td>
</tr>
</tbody>
</table>
Table 8: TMM6 Bus Designed Antennas - Performance Parameters

<table>
<thead>
<tr>
<th>Antenna</th>
<th>I4a: 3.175 mm Thick Fixed Cross Slot Antenna</th>
<th>I4b: 5.08 mm Thick Rotated Cross Slot Antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Gain</td>
<td>4.0 dBiC</td>
<td>4.63 dBiC</td>
</tr>
<tr>
<td>Gain at Boresight</td>
<td>3.57 dBiC</td>
<td>3.98 dBiC</td>
</tr>
<tr>
<td>Half Power Beamwidth</td>
<td>77 to 70°</td>
<td>72 to 70°</td>
</tr>
<tr>
<td>Reflection Coefficient at $f_r$</td>
<td>-18.2 dB</td>
<td>-11.2 dB</td>
</tr>
<tr>
<td>Impedance Bandwidth ($S_{11} \leq -10$ dB)</td>
<td>20 MHz</td>
<td>35.5 MHz</td>
</tr>
<tr>
<td>Boresight Axial Ratio</td>
<td>0.02 dB</td>
<td>0.34 dB</td>
</tr>
<tr>
<td>Axial Ratio Beamwidth ($p_{AR} \leq 3$ dB)</td>
<td>-49 to 78.1°</td>
<td>-50 to 71°</td>
</tr>
<tr>
<td>Axial Ratio Bandwidth ($p_{AR} \leq 3$ dB at Boresight)</td>
<td>4.8 MHz</td>
<td>7.5 MHz</td>
</tr>
</tbody>
</table>

Table 9 summarizes the compliance selected I4b antenna design to the stated design requirements. Additionally, the performance plots of this antenna design can be found in Figure 45. These plots verify the performance parameters recorded in Table 8.

3.7 Summary

A 1090 MHz L-band patch antenna was successfully designed for use on the CanX-7 ADS-B payload. This patch antenna utilized a single substrate, single coaxial feed design which utilizes a rotated cross slot to induce RHCP. TMM6 was selected for use as the antenna substrate due to its high thermal tolerance and dielectric constant. A high dielectric constant allowed the patch to be kept small relative to the substrate to ensure proper operation of the antenna. Additionally, the use of increased substrate thicknesses was found to improve the antenna gain. It was also discovered that an antenna must be optimized when mounted on an approximation of the spacecraft bus to account for center frequency shifting so it would be able to retain circular polarization and optimal gain. This is especially important as the aircraft mounted ADS-B transmitters are linearly polarized and the ADS-B link possesses little redundancy. It is worth thoroughly investigating this frequency shift as it may be possible to further optimize antenna design on future SFL spacecraft. For future antenna design projects with less constrained time lines, it may be worth investigating additional substrate lines as well as more complicated antenna designs such
as those incorporating multiple substrates, coupled feeds, or apertures.

Future work involving this antenna includes fabrication and performance verification. During fabrication, further changes to the antenna design may be required to account for unforeseen manufacturing and installation related discrepancies. As with the S-band and UHF arrays on CanX-7, antenna performance must be verified through pattern measurement. Antenna measurement is expected to be performed using the procedure outlined in Chapter 2 along with greater emphasis on the axial ratio measurement. Following antenna measurement, the ADS-B payload must also undergo electromagnetic compatibility (EMC) testing to ensure that its operation is not adversely impacted by other onboard electronics.

Table 9: Requirement Verification Matrix for the Selected I4b Antenna Design

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Requirement</th>
<th>Compliance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating Frequency</td>
<td>1090 MHz</td>
<td>Compliant – 1090 MHz</td>
</tr>
<tr>
<td>Form Factor</td>
<td>Max. 75 × 75 mm</td>
<td>Compliant – 75 × 75 mm</td>
</tr>
<tr>
<td>Thickness</td>
<td>≤ 12.2 mm</td>
<td>Compliant – 5.08 mm</td>
</tr>
<tr>
<td>Feed</td>
<td>Must fit in existing payload structure</td>
<td>Compliant – single probe feed</td>
</tr>
<tr>
<td>Polarization</td>
<td>Circular</td>
<td>Compliant – RHCP with $\rho_{AR} \leq 3$ dB at boresight</td>
</tr>
<tr>
<td>Thermal Stability</td>
<td>Operating temperature between -20 and 60°C</td>
<td>Compliant – TMM6 capable of operating in temperature range without significantly affecting simulated antenna performance</td>
</tr>
<tr>
<td>Gain and Beamwidth</td>
<td>Should be maximized without interfering with other design requirements</td>
<td>Compliant</td>
</tr>
<tr>
<td>Magnetic Properties</td>
<td>Non-magnetic materials</td>
<td>Compliant</td>
</tr>
</tbody>
</table>
3.7. Summary

(a) Frequency VS Reflection Coefficient

(b) Radiation (Gain) Pattern [Red Solid: RHCP, Blue Dotted: LHCP]

(c) Axial Ratio Pattern

(d) Frequency VS Axial Ratio

Figure 45: Performance of the Selected I4b Antenna Design
Chapter 4

Global Navigation Satellite System Based Attitude Determination

In recent years, small satellites have proliferated due to their rapidly expanding capabilities. Such satellites are capable of fulfilling complex missions at a greatly reduced cost as compared to traditional spacecraft. However, due to their size, such satellites are heavily constrained in terms of available mass and volume which limits the quantity and types of attitude sensors available for use. Many small satellites use an attitude determination and control system (ADCS) that employ two external vector measurements along with a means of measuring or estimating angular rates. A common ADCS sensor setup on low Earth orbiting satellites consists of sun sensors and a three-axis magnetometer. Unfortunately, the operation of such sensors may be limited by unfavorable environmental conditions or operating scenarios. For example, sun sensors cannot operate when the satellite is in eclipse. At these points, the spacecraft’s attitude estimates may degrade heavily. Since some spacecraft may be required to perform attitude correction or use payloads during these times, a means of reducing attitude estimation degradation without adding additional mass, volume, or cost to the spacecraft is in demand.

If the spacecraft is equipped with a single global positioning system (GPS) antenna, it is possible to use GPS measurements and telemetry to partially supplement the denied sensor measurements. From knowing the positions of all spacecraft involved along with some metric which relates the angle of the satellite to GPS line-of-sight to the signal strength, a set of attitude measurements can be formed. These can then be entered into the observation model of an attitude estimator to provide additional attitude estimate correction.

4.1 Literature Review

An attitude determination method based on a single GPS antenna has been tested on a variety of different spacecraft. In their paper, Axelrad and Behre [50] demonstrates their technique of GPS attitude determination using datasets from three satellite experiments: Cryogenic Infrared Spectrometers for the Atmosphere-Shuttle Planet Satellite (CRISTA-SPAS), GPS Attitude/Navigation Experiment (GANE), and GPS-based Meteorology (GPS/MET). After obtaining GPS signal to noise ratio (SNR) measurements, a least squares method was employed to estimate the pointing vector of the GPS antenna. With this method, the GPS antenna boresight vector was estimated with a root-mean-square (RMS) error of
3.2° to 11.9°. This attitude determination scheme was also considered for use on the FedSat (Federation Satellite) microsatellite [51] and its performance was compared against an alternate signal-to-noise weighted line-of-sight vector method created by NASA JPL (Jet Propulsion Laboratory). It was found through simulation that the Axelrad and Behre method was superior to the JPL method; achieving an accuracy of ±5° as opposed to ±30°. Another successful implementation of this attitude determination method was on the 2010 University of Texas FASTRAC (Formation Autonomy Spacecraft with Thrust, Rehav, Attitude and, Crosslink) mission. Here, GPS SNR measurements were combined with three-axis magnetometer measurements in a norm constrained Kalman filter [52] to allow for autonomous control of thruster operations onboard the spacecraft. The attitude estimation error of this filter possessed a standard deviation of approximately 10° [53].

4.2 Chapter Overview

The work in this chapter builds upon the previous implementations of single antenna GPS augmented attitude estimators. In previous applications of this method, the GPS would be intended for use as a primary ADCS sensor. Consequently, the receiver and antenna would be selected and configured to provide precise SNR or carrier to noise density ratio (C/N0) telemetry prior to launch. Conversely, this chapter focuses on the on-orbit implementation and calibration of online estimation methods. This is performed to supplement an existing set of attitude sensors on a spacecraft not originally designed nor intended to use its GPS subsystem for attitude determination. Therefore, the estimation method will need to account for lower precision C/N0 telemetry as well as limitations in the GPS antenna’s radiation pattern. The University of Toronto’s Space Flight Laboratory’s (SFL) CanX-4 and CanX-5 mission provide an excellent test bed for extensions to the GPS based attitude determination concept. The algorithms presented herein are evaluated using flight data collected from the CanX-5 spacecraft as part of an extended mission phase.

An extended Kalman filter (EKF) is selected for attitude estimation due to its ease of use and low computational overhead. Four types of measurements are considered: sun vector, magnetic field vector, spacecraft body rate, and GPS C/N0. A rotation-matrix based parameterization is selected to avoid the singularities or norm-constraints associated with more traditional representations such as Euler angles or unit quaternions [54]. Attitude error and update terms are treated as small rotations, with the motion and measurement equations linearized about the current attitude solution. The attitude estimates are then compared against CanX-5’s onboard attitude solutions. Simulated and naturally occurring periods of eclipse without sun sensor data are used to test the effectiveness of the EKF during periods of denied ADCS sensor measurements.

4.3 Spacecraft Attitude Dynamics

4.3.1 Skew-Symmetric Cross-Operator

The skew-symmetric cross-operator \((\cdot)^x\) converts a \(3 \times 1\) column vector \(\mathbf{a} = \begin{bmatrix} a_1 & a_2 & a_3 \end{bmatrix}^T, \mathbf{a} \in \mathbb{R}\) into a \(3 \times 3\) skew-symmetric matrix.
Chapter 4. Global Navigation Satellite System Based Attitude Determination

\[
\mathbf{a}^x = \begin{bmatrix}
0 & -a_4 & a_2 \\
-a_3 & 0 & -a_1 \\
a_2 & -a_1 & 0
\end{bmatrix}
\]  
(4.1)

4.3.2 Reference Frames

Reference frames are essential in describing the orientation of an object in 3-dimensional space. A reference frame is defined as three unit length physical vectors orthogonal to each other. For instance, reference frame \( \mathcal{F}_1 \) is defined by unit vectors \( \mathbf{x}_1, \mathbf{y}_1, \text{ and } \mathbf{z}_1 \). Whereas:

\[
\mathcal{F}_1 = \begin{bmatrix}
\mathbf{x}_1 \\
\mathbf{y}_1 \\
\mathbf{z}_1
\end{bmatrix}
\]  
(4.2)

Where the vector matrix, \( \mathcal{F}_1 \) is referred to as a vectrix. The following describes common reference frames used for astronautical applications.

Body Fixed Frame

The origin of the body fixed frame is located at the center of the spacecraft. As seen in Figure 46 for CanX-5:

- The \( x_b \)-axis is pointing out the +X face of the spacecraft
- The \( y_b \)-axis is pointing out the +Y face of the spacecraft. This is also where the GPS antenna on CanX-5 is located
• The $z_b$-axis is pointing out the $+Z$ face of the spacecraft. The magnetometer is mounted at the end of a boom protruding out from this face.

While a second reference frame could have been established at the GPS antenna, GPS calibrations and measurements were taken relative to the orientation of the body frame to avoid the need for any rotational transformations. Translational transformations were accounted for; the GPS antenna’s coordinates in the body fixed frame were: $[-0.05, 0.1, 0.05]^T$ m. Additionally, magnetometer, rate sensor, and sun sensor measurements were obtained through the spacecraft’s ADCS in the body fixed frame.

**Earth Centered Earth Fixed**

The Earth-centered, Earth-fixed (ECEF) (Figure 47) originates at the center of the Earth. Its axes are defined as:

![Figure 47: ECEF Frame [55]](image)

- The $x_{ecef}$-axis is pointing out from where the Prime-Meridian intersects the Equator
- The $z_{ecef}$-axis is pointing upwards through the Earth’s geographical north pole
- The $y_{ecef}$-axis is orthogonal to the $x_{ecef}$ and $z_{ecef}$-axes as per the right hand rule ($y_{ecef} = z_{ecef} \times x_{ecef}$)

The location each axis is fixed to the Earth’s surface and the frame rotates with an angular rate of $\omega_{ecef} = 9.2921 \times 10^{-5}$ rad/s about the $z_i$ axis of any Earth centered inertial frame.

**Earth Centered Inertial**

The Earth-centered inertial (ECI) Frame is similar the ECEF frame except that its axes are not fixed to the surface of the Earth and the frame does not rotate. There are many types of ECI frames in use in which the direction of the $x$ and $z$-axes are defined. CanX-5’s ADCS outputs attitude estimates in the J2000 frame defined below [56]:

• The $x_{ecef}$-axis is pointing up from where the Prime-Meridian intersects the Equator
• The \( x_i \)-axis is pointing out from where the Prime-Meridian intersects the Equator. This point is defined with the Prime-Meridian and Equator at 12:00 Terrestrial Time on 1 January 2000.

• The \( z_i \)-axis is pointing upwards through the Earth’s geographical north pole.

• The \( y_i \)-axis is orthogonal to the \( x_i \) and \( z_i \)-axes as per the right hand rule (\( y_i = z_i \times x_i \)).

### 4.3.3 Representations of Rotations

The orientation of an Earth orbiting spacecraft with respect to the Earth can be described in terms of the orientation of the body fixed reference frame with respect to an ECEF or ECI frame as a rotation. This orientation is also known as the spacecraft’s attitude. The following sections describe methods of describing rotations between reference frame \( F_2 \) and \( F_1 \) in rigid body systems.

#### Rotation Matrices

A rotation matrix is an element in the special orthogonal group of order 3 (\( SO(3) \)). These are used to transform vector coordinates between reference frames or to represent rotations within a frame [56].

\[
SO(3) = \{ R \mid R \in \mathbb{R}^{3\times 3}, R \text{ is } \perp \text{ and } \det(R) = 1 \} \tag{4.3}
\]

Where \( \mathbb{1} \) is an identity matrix. Observe the following vector \( \mathbf{r} \) in two reference frames.

\[
\mathbf{r} = F_2^T \mathbf{r}_2 = F_1^T \mathbf{r}_1 \tag{4.4}
\]

The relationships of these coordinates via a rotation matrix (\( C_{21} \)) is obtained by:

\[
\mathbf{r}_2 = C_{21} \mathbf{r}_1 \tag{4.5}
\]

Successive rotations can be performed by stacking the corresponding rotation matrices. For example, a rotation for vector \( \mathbf{r} \) from frames \( F_1 \) to \( F_3 \) is performed as:

\[
\mathbf{r}_3 = C_{32} C_{21} \mathbf{r}_1 \tag{4.6}
\]

#### Euler Angles

Any rotation matrix can be represented as chain of three individual rotations known as principal rotations. One common set of these rotations are known as Euler angles. These can be represented as the principal rotations shown below [57].

1. A rotation \( \phi \) about the transformed reference frame’s x-axis (roll):

\[
C_x(\phi) = \begin{bmatrix}
1 & 0 & 0 \\
0 & \cos \phi & \sin \phi \\
0 & -\sin \phi & \cos \phi
\end{bmatrix} \tag{4.7}
\]
2. Next, a rotation $\theta$ about the resulting reference frame’s $y$-axis (pitch):

$$
C_y(\theta) = \begin{bmatrix}
\cos \theta & 0 & -\sin \theta \\
0 & 1 & 0 \\
\sin \theta & 0 & \cos \theta
\end{bmatrix}
$$

(4.8)

3. Lastly, a rotation $\psi$ about a reference frame’s $z$-axis (yaw):

$$
C_z(\psi) = \begin{bmatrix}
\cos \psi & \sin \psi & 0 \\
-\sin \psi & \cos \psi & 0 \\
0 & 0 & 1
\end{bmatrix}
$$

(4.9)

A common sequence of rotations for aerospace applications is the 3-2-1 sequence also known as the “yaw, pitch, roll” convention. In this case, a rotation matrix from reference frames $F_1$ to $F_2$ is defined as:

$$
C_{21}(\theta_1, \theta_2, \theta_3) = C_x(\theta_1)C_y(\theta_2)C_z(\theta_3) = \begin{bmatrix}
\frac{c_2c_3}{s_2} & \frac{c_2s_3}{s_2} & \frac{-s_2}{s_2} \\
\frac{s_1s_2c_3-c_1s_3}{s_2} & \frac{s_1s_2s_3+c_1c_3}{s_2} & \frac{s_1c_2}{s_2} \\
\frac{c_1s_2c_3+s_1s_3}{s_2} & \frac{c_1s_2s_3-c_1c_3}{s_2} & \frac{c_1c_2}{s_2}
\end{bmatrix}
$$

(4.10)

Where $s_i = \sin \theta_i$ and $c_i = \cos \theta_i$. Euler angles present a very intuitive and easy to use representation for the attitude of objects in 3D space. However, Euler angles as well as other methods which use three parameters to describe rotation matrices are prone to singularities. For instance, a singularity occurs in the 1-2-3 rotation matrix shown above when $\theta_2 = \pi/2$. At this point, the roll and yaw angles are associated with the same rotation and cannot be determined uniquely.

When the angles associated with the 3-2-1 Euler sequence are infinitesimally small ($\theta_1, \theta_2, \theta_3 \ll 1$), the small angle approximation can be used. In this case $c_i \approx 1$, $s_i \approx \theta_i$, and products of small angles ($\theta_i \theta_j \approx 0$) are neglected. This results in:

$$
C_{21} \approx \begin{bmatrix}
1 & \theta_3 & -\theta_2 \\
-\theta_3 & 1 & \theta_1 \\
\theta_2 & -\theta_1 & 1
\end{bmatrix}
$$

(4.11)

$$
\approx 1 - \theta^* 
$$

(4.12)

Where

$$
\theta = \begin{bmatrix}
\theta_1 \\
\theta_2 \\
\theta_3
\end{bmatrix}
$$

(4.13)

is known as a rotation vector. The same results are obtained regardless of the order of the rotations are performed. This rotation vector only occurs with small angles which approach vector like behavior.

**Unit Quaternions**

Expressing rotations as a set of unit quaternions (also known as Euler Parameters) is a popular means of avoiding the singularities present when using Euler angles. A unit quaternion $q$ consists of a real component $\eta$ and three imaginary components $\epsilon = \begin{bmatrix} \epsilon_1 & \epsilon_2 & \epsilon_3 \end{bmatrix}$ satisfying the constraint $q^Tq = 1$. 
As such, unit quaternions belong to the following set \( Q \) [56]:

\[
Q = \{ q | q^T q = 1, q = [\epsilon, \eta]^T, \epsilon \in \mathbb{R}^3 \text{ & } \eta = \mathbb{R} \}
\] (4.14)

The constraint imposed on unit length components also implies that:

\[
\epsilon_1^2 + \epsilon_2^2 + \epsilon_3^2 + \eta^2 = 1
\] (4.15)

Instead of breaking a rotation matrix down into three principal rotations, a unit quaternion represents the rotation as one rotation \( \varphi \) around a fixed axis of rotation \( a = [a_1, a_2, a_3]^T \). This rotation and axis are related to the quaternion by:

\[
\eta := \cos \left( \frac{\varphi}{2} \right), \quad \epsilon := a \sin \left( \frac{\varphi}{2} \right) = \begin{bmatrix}
\epsilon_1 \\
\epsilon_2 \\
\epsilon_3
\end{bmatrix}
\] (4.16)

A rotation matrix can then be expressed in terms of a unit length quaternion as [57]:

\[
C_{21} = (\eta^2 - \epsilon^T \epsilon) I + 2\epsilon \epsilon^T - 2\eta \epsilon \epsilon^T = \begin{bmatrix}
1 - 2(\epsilon_2^2 + \epsilon_3^2) & 2(\epsilon_1 \epsilon_2 + \epsilon_3 \eta) & 2(\epsilon_1 \epsilon_3 - \epsilon_2 \eta) \\
2(\epsilon_2 \epsilon_1 - \epsilon_3 \eta) & 1 - 2(\epsilon_3^2 + \epsilon_1^2) & 2(\epsilon_2 \epsilon_3 + \epsilon_1 \eta) \\
2(\epsilon_3 \epsilon_1 + \epsilon_2 \eta) & 2(\epsilon_3 \epsilon_2 - \epsilon_1 \eta) & 1 - 2(\epsilon_1^2 + \epsilon_2^2)
\end{bmatrix}
\] (4.17)

It should be noted that CanX-5’s ADCS outputs attitude estimates using quaternion parameterization.

### Linearizing Rotation Matrices

As indicated in previous sections, both Euler angles and quaternions possess characteristics which can complicate the linearization of rotations. Euler angle parametrization contains singularities which must be avoided by selecting an appropriate rotation sequence, while quaternions require a norm constraint. An alternative method is to linearize rotation matrices directly, avoiding the issue of singularities and constraints. The following linearization method by Barfoot [54] treats errors and update terms as small rotation vectors.

From first principles, a variable \( x \) in function \( f(x) \) can be perturbed slightly from its nominal value, \( \bar{x} \), by \( \delta x \) to represent a small change in the function. Assuming that \( \delta x \) is small and not constrained, this can be represented as a first-order Taylor-series expansion.

\[
f(\bar{x} + \delta x) \approx f(\bar{x}) + \frac{\delta f(\bar{x})}{\delta x} \bar{x} \delta x
\] (4.18)

Euler angles can be used to apply the same process to rotation matrices. These are favorable as they contain exactly three parameters which can be varied independently without any constraints.

Rotation matrix, \( C(\theta)v \) can be perturbed with respect to Euler angles \( \theta = [\theta_1, \theta_2, \theta_3]^T \), where \( v \) is an arbitrary constant vector. This results in the following first-order Taylor-series approximation.
The identity
\[
\frac{\delta(C(\theta)v)}{\delta \theta} = (C(\theta)v)^T S(\theta_2, \theta_3)
\]
is obtained by applying Euler’s theorem to a three angle Euler sequence. Euler’s theorem:
\[
C = \cos \phi \mathbf{1} + (1 - \cos \phi) \mathbf{a} \mathbf{a}^T - \sin \phi \mathbf{a}^x
\]
enables a rotation matrix \( C \) to be written in terms of an angular rotation \( \phi \) about an axis \( \mathbf{a} \). Furthermore,
\[
S(\theta_2, \theta_3) = \begin{bmatrix}
\cos \theta_2 \cos \theta_3 & \sin \theta_3 & 0 \\
-\cos \theta_2 \sin \theta_3 & \cos \theta_3 & 0 \\
\sin \theta_2 & 0 & 1
\end{bmatrix}
\]
relates the angular velocity to Euler-angle rates [54]. Returning to the first-order Taylor series approximation, \( v \) is arbitrary and can be dropped from both sides.
\[
C(\bar{\theta} + \delta \theta) \approx (1 - (S(\bar{\theta}_2, \bar{\theta}_3)\delta \theta)^x) C(\bar{\theta})
\]
Resulting in the product of a small rotation matrix, \( (1 - (S(\bar{\theta}_2, \bar{\theta}_3)\delta \theta)^x) \) and the unperturbed rotation matrix, \( C(\bar{\theta}) \). Or simplified as:
\[
C(\bar{\theta} + \delta \theta) \approx (1 - \delta \phi^x) C(\bar{\theta})
\]
with \( \delta \phi = S(\bar{\theta}_2, \bar{\theta}_3)\delta \theta \).

4.4 GPS Based Attitude Determination

The global positioning system (GPS) is a satellite navigation system that provides location and time information anywhere on or near the Earth. This system was conceived by the US military in the 1970’s with the first satellite launched in 1978 and the system declared fully operational on April 27, 1995 [58]. Due to its accuracy and ease of use, GPS has become an attractive option for orbit determination in low Earth orbiting spacecraft especially in microsatellites and nanosatellites.

The use of GPS based attitude determination is not a novel concept. Fine three-axis attitude estimates have been produced through the use of multiple GPS antennas mounted on spacecraft in non-parallel orientations. In this setup, measurement of phase differences between these antennas would provide the attitude estimate. This method was used on the REX-II (Radiation Experiment Satellite II) spacecraft to provide estimates with a standard deviation within 1.66° [59].

Of course, many small satellites simply do not have the space, mass, or need for multiple GPS antennas and often only possess a single GPS antenna for orbit determination. Fortunately, a single
GPS antenna can still be used to augment attitude estimates.

4.4.1 Attitude Estimation With a Single GPS Antenna

The $C/N_0$ reported for each GPS satellite tracked can be correlated to the directionally dependent radiation pattern of the onboard GPS antenna. This enables the off-boresight angle ($\alpha$) or its cosine ($\cos(\alpha)$), between the line-of-sight (LOS) vector to each GPS satellite and the antenna’s boresight to be employed as a measurement. $\cos(\alpha)$ is recommended as opposed to $\alpha$ as it does not require any additional resources to map and allows the cosine to be excluded in a sensor observation model. Notably, this single antenna method cannot provide a full three-axis attitude solution, although some spacecraft can obtain a three-axis solutions if they possess multiple GPS antennas.

The relation between the measured $C/N_0$ and $\cos(\alpha)$ is quantified by means of a mapping function. This can be obtained through one of two methods [50]. The first method can be performed prior to launch when the GPS antenna has been integrated into the spacecraft and involves directly measuring the radiation pattern of the GPS antenna to scale $C/N_0$ measurements. This can be done using the same techniques described in Chapter 2. The second method consists of obtaining the mapping function through on-orbit calibration and is recommended since it takes into account signal variation from multipath and the surface geometry of the spacecraft (if present). Additionally, the latter method provides a more accurate means of mapping $C/N_0$ to $\cos(\alpha)$ which can be updated if necessary. The algorithm for on-orbit calibration consists of the following steps [50]:

1. Obtain a flight calibration dataset containing $C/N_0$ measurements, LOS vectors in a fixed or inertial measurement frame, and spacecraft attitude in body frame.

2. Compute normalized LOS vectors in the spacecraft’s body frame and adjust $C/N_0$ measurements for free space path loss (FSPL).

3. Compute $\cos(\alpha)$ for each GPS satellite contact using the normalized LOS vector and known antenna boresight vector in the body frame.

4. Assemble all calculated $\cos(\alpha)$ values based on measured $C/N_0$ bins. The $\cos(\alpha)$ mean and standard deviation for each $C/N_0$ bin is calculated.

5. Compute mapping function by fitting a curve to the mean $\cos(\alpha)$ values for each $C/N_0$ bin. The standard deviation is used for weighting if solving for the boresight vector directly.

A correction factor is added to the $C/N_0$ measurements to adjust for variations in FSPL and GPS satellite transmission levels. This correction factor ($S_{FSPL}$) for GPS satellite contact $j$ is computed as:

$$S_{FSPL}^j(dB) = 20\log_{10}R^j$$  \hspace{1cm} (4.27)

Where $R$ is the distance between CanX-5 and each GPS satellite in kilometers. This equation is a truncated version of the FSPL equation in Chapter 2; the frequency dependent component and constant are dropped as only the range between the spacecraft change.
4.5 Modeling

The state to be estimated is the rotation matrix between the J2000 ECI frame \( (i) \) and the spacecraft’s body frame \( (b) \) shown in Figure 46. The rotation matrix is defined as:

\[
C_{bi} = F_b \cdot F_i^T
\]

(4.28)

Where \( F_b \) and \( F_i \) are the vectrixes defining the body and ECI frames respectively [54]. Again, the J2000 frame has been selected for these calculations as it is the coordinate system used by CanX-5’s ADCS.

4.5.1 Discrete Time Motion Model

The discrete time motion model for CanX-5 using rotation matrix parameterization [54] is shown below:

\[
C_{b_{k+1}i} = \Psi_{b_{k-1}b_k} C_{b_{k-1}i}
\]

(4.29)

Where \( C_{b_{k+1}i} \) is the inertial to body rotation matrix at time step \( k \), and \( \Psi_{b_{k-1}b_k} \) is the sampled data rotational kinematics matrix given by:

\[
\Psi_{b_{k-1}b_k} = e^{-\psi_k} \cos \psi_k \mathbf{1} + (1 - \cos \psi_k) \begin{pmatrix} \psi_k \\ \psi_k \end{pmatrix} \begin{pmatrix} \psi_k \\ \psi_k \end{pmatrix}^T - \sin \psi_k \begin{pmatrix} \psi_k \\ \psi_k \end{pmatrix} \times
\]

(4.30)

And:

\[
\psi_k = \omega_{b_{k-1}i} T_k + \delta \psi_k
\]

(4.31)

Where \( \omega_{b_{k-1}i} \) is the angular velocity of the body frame with respect to the ECI frame at time step \( k - 1 \), \( T_k \) is the sampling period at time step \( k \), and \( \delta \psi_k \) is the rotational process and rate sensor noise.

Angular Rate Measurement Considerations

This subsection shows ways of augmenting the state to account for rate sensor biases or to indirectly estimate angular rates from control torques for integration into other spacecraft and systems. However, it should be noted that in this experiment, the simulated body rate measurements do not possess any bias and flight measurements are bias corrected by CanX-5’s ADCS computer. Consequently, these augmentations were not applied to CanX-5’s motion model.

Generally, there are two ways to obtain the angular rates of a satellite. The first is direct measurement using rate sensors (Figure 48), while the second involves expanding the attitude estimator to estimate the rates using the attitude control inputs applied to the spacecraft. In the case of rate sensors, the continuous time measurement model is given below [60].

\[
\omega(k) = g(k) + e(k) + \eta(k)
\]

(4.32)

Where \( g \) is the output vector of the rate sensors, \( e \) is the sensor bias, and \( \eta \) represents Gaussian white measurement noise. In the case of the motion model above, the Gaussian white measurement noise is merged with process noise in \( \delta \psi \). When using direct measurement, it is necessary to account for the effects of sensor bias which are often time and environmentally dependent. This can be done by
expanding the estimated state to include biases.

\[
\begin{bmatrix}
C_{b_k i} \\
\epsilon_k
\end{bmatrix} = \begin{bmatrix}
\Psi_{b_k b_{k-1} i} C_{b_{k-1} i} \\
\epsilon_{k-1} + v_k
\end{bmatrix}
\] (4.33)

Where:

\[
\psi_k = (\omega_{b_{k-1} i} + \epsilon_{k-1}) T_k + \delta \psi_k
\] (4.34)

And the random walk of the bias is approximated by the Gaussian function: \( v_k \sim \mathcal{N}(0, W) \). Conversely, for spacecraft without rate sensors, the state can be expanded as follows to estimate the angular rates from control torques applied to the spacecraft [57].

\[
\begin{bmatrix}
\psi_k \\
C_{b_k i}
\end{bmatrix} = \begin{bmatrix}
T_k \left( I^{-1} \left[ \omega_{b_{k-1} i} \omega_{b_{k-1} i} + u_{k-1} \right] \right) + \delta \psi_k
\end{bmatrix} (4.35)
\]

Where \( I \) is the inertial matrix of the spacecraft, \( u_k \) is the control input, and the state is augmented to include \( \psi_k \) separately. In these two cases where the state is augmented, it is necessary to check that the system is still observable.

### 4.5.2 Discrete Time Observation Models

Discrete time observation models were created for the three-axis magnetometer, sun sensors, and the GPS antenna on CanX-5.
Three-Axis Magnetometer

A three-axis magnetometer (Figure 49a) measures the intensity and direction of the geomagnetic field surrounding the spacecraft. This instrument contains three orthogonal sensors which allow the calculation of the magnetic field vector expressed in the spacecraft body frame. The observation model for the magnetometer is:

\[ y_{B_k} = C_{bk} \rho_{B_k} \quad (4.36) \]

Where \( y_{B_k} \) is the normalized magnetic field measurement in the body frame at time step \( k \) and \( \rho_{B_k} \) is the normalized magnetic field in the ECI frame calculated from the International Geomagnetic Reference Field (IGRF) model [56]:

\[ \rho_{B_k} = a \sum_{n=1}^{N} \sum_{m=1}^{n} \left( \frac{a}{r} \right)^{n+1} \left( g_n^m \cos m\phi + h_n^m \cos m\phi \right) P_n^m (\cos \theta) \quad (4.37) \]

In the IGRF model, \( a \) is the mean radius of the Earth, \( r \) is the distance from the center of the Earth, \( \phi \) is the longitude east of the Prime Meridian, and \( \theta \) represents the colatitude (90° – latitude). A single three-axis magnetometer measurement by itself can only provide orientation with respect to Earth about two orthogonal axes [61].

Sun Sensors

Sun sensors (Figure 49b) directly measure the line-of-sight vector from the spacecraft to the Sun in the body frame. This is compared to the known Sun vector in the ECI frame to obtain an attitude estimate using [56]:

\[ y_{S_k} = C_{bk} \rho_{S_k} \quad (4.38) \]

where \( y_{S_k} \) is the normalized Sun vector measurement in the body frame at time step \( k \) and \( \rho_{S_k} \) is the normalized known Sun vector in ECI.

GPS Antenna (Iterative GPS Boresight Estimation)

The GPS observation model compares the known orientation of the GPS antenna boresight vector in the spacecraft body frame and the observation of this vector in the ECI frame [54]. This model is represented as:

\[ \rho_{GPS_k} = C_{bk}^T y_{GPS_k} \quad (4.39) \]

where \( \rho_{GPS_k} \) is the normalized measured GPS antenna boresight vector in the ECI frame at time step \( k \) and \( y_{GPS_k} \) is the normalized known antenna boresight in the body frame. On CanX-5, the GPS antenna is mounted on the +Y face (Figure 46) and \( y_{GPS_k} \) is expressed as:

\[ y_{GPS_k} = \begin{bmatrix} 0 \\ 1 \\ 0 \end{bmatrix} \quad (4.40) \]

Previous implementations of single GPS antenna attitude estimation [50][51], have sought to directly estimate the GPS boresight from the measured GPS line-of-sight and \( \cos(\alpha) \) by minimizing the cost
function:

\[ J(\rho_{GPS_k}) = \frac{1}{2} \sum_{j=1}^{M} \frac{1}{(\sigma_k^j)^2} \left| \cos(\alpha_k^j) - \mathbf{l}_k^T \rho_{GPS_k} \right|^2 \]  

(4.41)

Where, \( \mathbf{l}_k^j \) is the normalized LOS vector for each GPS contact \( j \) in the ECI frame at time step \( k \), \( M \) is the total number of acquired GPS satellites, and \( \sigma_k^j \) is the measurement standard deviation. This cost function is solved iteratively with an initial condition obtained by normalizing \( \mathbf{b}_k^0 \frac{\mathbf{\hat{b}}_{GPS_k} b_{k}^0}{|b_{k}^0|} \) [50]:

\[ \mathbf{b}_k^0 = (\mathbf{A}_k^T \mathbf{W}_k \mathbf{A}_k)^{-1} \mathbf{A}_k^T \mathbf{W}_k \mathbf{c}_k \]  

(4.42)

Where:

\[ \mathbf{A}_k = \begin{bmatrix} \mathbf{l}_k^1 \ T \\ \vdots \\ \mathbf{l}_k^M \ T \end{bmatrix}, \quad \mathbf{c}_k = \begin{bmatrix} \cos(\alpha_k^1) \\ \vdots \\ \cos(\alpha_k^M) \end{bmatrix}, \quad \mathbf{W}_k = \text{diag}\left\{ \frac{1}{(\sigma_k^1)^2}, \ldots, \frac{1}{(\sigma_k^M)^2} \right\} \]  

(4.43)

The measurement variances to be used are obtained from the standard deviation of \( \cos(\alpha) \) values in each \( C/N_0 \) bin from the mapping function. A gradient search is then performed on two spherical angles to enforce a normal constraint and return a GPS boresight unit vector.

\[ \rho_{GPS_k} = \begin{bmatrix} \sin(\theta_k)\cos(\theta_k) \\ \sin(\theta_k)\sin(\theta_k) \\ \cos(\theta_k) \end{bmatrix} \]  

(4.44)

The resultant ECI GPS boresight vector estimate can then be used as a measurement. However, solving the minimization problem at each time step increases computational overhead.

**GPS Antenna (Integrated \( C/N_0 \) Model)**

To reduce computational overhead, the GPS observation model can be augmented to use \( \cos(\alpha_k^j) \) directly. This variable is related to the normalized GPS LOS and antenna boresight vectors (Figure 50) by:

\[ \cos(\alpha_k^j) = \mathbf{l}_k^T \cdot \mathbf{y}_{GPS_k} \]  

(4.45)
The GPS observation model can then be modified as follows:

\[ \cos(\alpha^j_k) = Y^T_k C^T_{b_k} y_{GPS_k} \] (4.46)

In this case, \( \cos(\alpha^j_k) \) is chosen as the measurement value rather than \( \alpha^j_k \) or \( C/N_0 \) for two reasons. First, the use of this variable greatly simplifies the model and its integration into an estimator. Secondly, the \( C/N_0 \) to \( \cos(\alpha^j_k) \) mapping function is dependent on the characteristics of each antenna, its integration, and the operating environment; which may not be accurately known until after spacecraft launch. The current setup keeps the mapping function out of the observation model, removing the need to re-linearize every time the function is updated.

Full Observation Model

The individual sensor observation models can then be concatenated into a full observation model.

\[
\begin{bmatrix}
Y_{B_k} \\
Y_{S_k} \\
\cos(\alpha^j_k)
\end{bmatrix} = \begin{bmatrix}
C_{b_k} y_{B_k} \\
C_{b_k} y_{S_k} \\
1^T_k C^T_{b_k} y_{GPS_k}
\end{bmatrix} + n^j_k
\] (4.47)

Where \( n^j_k \) is the total measurement noise from the attitude sensors.

4.6 Estimation

The extended Kalman filter (EKF) (Figure 51) is a standard state estimation technique, suitable for systems with nonlinear motion and observation models. The EKF is comprised of a predictor and corrector stage that estimate a time-varying state while modeling the uncertainty arising from process and measurement noise [57], [62]. The predictor stage uses the motion model to propagate the prior known state from the knowledge of body rates. The corrector stage then corrects the predictor’s propagator state using sensor data provided by the observation model. The EKF is a popular method for spacecraft attitude estimation as it can be operated online with low computational overhead while providing adequate accuracy for most spacecraft applications. Alternative estimation methods were also considered for use, however the EKF was selected for a number of reasons. First, spacecraft attitude determination requires a fast online solution since the estimate may need to be used for real time attitude control, ruling out any batch estimators. Computational overhead and response times were also important factors as small satellites often possess very little processing power and memory. The unscented Kalman filter (UKF) was considered as it was reputed to possess greater performance on non-linear systems. However, it was found through literature review that on simple, single body spacecraft kinematics with quaternion or Euler angle parameterization the higher computational overhead on a UKF outweighed any increase in accuracy [60], [62]. Since the motion and observation models use rotation matrix parameterization, which can be seen as an intermediary between quaternions and Euler angles, it was hypothesized that similar performance characteristics would be encountered on CanX-5. Sliding window type filters were also ruled out based on computational overhead and from the experience that low amounts of measurements lead to poorly conditioned Hessians, hindering the performance and speed of the filter.
4.6.1 Motion Model Linearization

The linearization of the motion model consists of perturbing the state slightly from its nominal value. This is done using the small rotation matrix approximation discussed previously in Equation 4.26 [54]:

\[
C(\bar{\theta} + \delta\theta) \approx (1 - \delta\theta^x) \ C(\bar{\theta})
\] (4.48)

This can be applied to the motion model (Equation 4.29) to yield:

\[
(1 - \delta\phi_k^x) \ C_{b,i} \approx (1 - \delta\psi_k) \ C_{b,i} \ C_{b,k-1} \ (1 - \delta\phi_{k-1}^x) \ C_{b,k-1,i}
\] (4.49)

The nominal solution (operating point of the linearization) can be subtracted off:

\[
C_{b,i} = C_{b,i} \ C_{b,k-1,i}
\] (4.50)

Leaving the linearized motion model:

\[
\delta\phi_{b,i} = \delta\psi_{k} + \delta\phi_{k-1} + \delta\phi_{b,k-1,i}
\] (4.51)

Where \(\delta\phi_k = \bar{S}_k \delta\theta_k\) and \(\bar{S}_k = S_k(\bar{\theta}_2, \bar{\theta}_3)\). This leaves the following Jacobians (\(H\)) with respect to the small rotation vector (\(\phi\)) and process/rate sensor noise (\(\psi\)):

\[
H_{\phi,k} = \bar{S}_{b,i} \ C_{b,k-1,i} \ C_{b,k-1,i}, \ H_{\psi,k} = 1
\] (4.52)

4.6.2 Observation Model Linearization

The linearization process for the three-axis magnetometer and sun sensor observation models are the same. As with the motion model, the rotation matrices in these observation models (Equations 4.36 and 4.38) were perturbed as such:
\[
\begin{bmatrix}
\delta y_{B_k} \\
\delta y_{S_k}
\end{bmatrix}
= \begin{bmatrix}
(1 - \delta \phi_k^x) \mathbf{C}_{b_k i \rho_{B_k}} \\
(1 - \delta \phi_k^x) \mathbf{C}_{b_k i \rho_{S_k}}
\end{bmatrix} (4.53)
\]

The following nominal solutions were then subtracted off:

\[
\begin{bmatrix}
\delta y_{B_k} \\
\delta y_{S_k}
\end{bmatrix}
= \begin{bmatrix}
\mathbf{C}_{b_k i \rho_{B_k}} \\
\mathbf{C}_{b_k i \rho_{S_k}}
\end{bmatrix} (4.54)
\]

Leaving the linearized magnetometer and sun sensor observation models:

\[
\begin{bmatrix}
\delta y_{B_k} \\
\delta y_{S_k}
\end{bmatrix}
= \begin{bmatrix}
(\mathbf{C}_{b_k i \rho_{B_k}})^x \\
(\mathbf{C}_{b_k i \rho_{S_k}})^x
\end{bmatrix} [\delta \phi] (4.55)
\]

the GPS observation model was linearized in the same manner using \(\cos(\alpha_k^i)\) as the measurement.

\[
\overline{\cos(\alpha_k^i)} + \delta \cos(\alpha_k^i) = l_k^T \left( (1 - \delta \phi_k^x) \mathbf{C}_{b_k i \rho_{Y_{GPS_k}}} \right) y_{GPS_k} (4.56)
\]

Again, the nominal solution was subtracted out:

\[
\overline{\cos(\alpha_k^i)} = l_k^T \mathbf{C}_{b_k i Y_{GPS_k}} (4.57)
\]

Leaving the linearized observation model:

\[
\delta \cos(\alpha_k^i) = -l_k^T \mathbf{C}_{b_k i Y_{GPS_k}} \delta \phi_k (4.58)
\]

To account for noise, the \(n^i_k\) term of the observation model was replaced by \(n^i_k + \delta n^i_k\) where \(\mathbf{n}^i_k = 0\) due to the zero-mean noise assumption. This produces a full nominal observation model.

\[
\begin{bmatrix}
\delta y_{B_k} \\
\delta y_{S_k} \\
\cos(\alpha_k^i)
\end{bmatrix}
= \begin{bmatrix}
\mathbf{C}_{b_k i \rho_{B_k}} \\
\mathbf{C}_{b_k i \rho_{S_k}} \\
l_k^T \mathbf{C}_{b_k i Y_{GPS_k}}
\end{bmatrix} (4.59)
\]

And a full linearized observation model:

\[
\begin{bmatrix}
\delta y_{B_k} \\
\delta y_{S_k} \\
\delta \cos(\alpha_k^i)
\end{bmatrix}
= \begin{bmatrix}
(\mathbf{C}_{b_k i \rho_{B_k}})^x \\
(\mathbf{C}_{b_k i \rho_{S_k}})^x \\
-1 l_k^T \mathbf{C}_{b_k i Y_{GPS_k}}^x
\end{bmatrix} [\delta \phi_k] + \delta n^i_k (4.60)
\]

The Jacobians (\(G\)) with respect to the small rotation vector (\(\phi\)) and sensor noise (\(n\)) are obtained as:

\[
G_{\phi,k} = \begin{bmatrix}
G_{\phi,k}^B \\
G_{\phi,k}^S \\
G_{\phi,k}^{\cos(\alpha^i)}
\end{bmatrix} = \begin{bmatrix}
(\mathbf{C}_{b_k i \rho_{B_k}})^x \\
(\mathbf{C}_{b_k i \rho_{S_k}})^x \\
-1 l_k^T \mathbf{C}_{b_k i Y_{GPS_k}}^x
\end{bmatrix} (4.61)
\]

\[
G_{n,k} = G_{n,k}^B = G_{n,k}^S = G_{n,k}^{\cos(\alpha^i)} = 1 (4.62)
\]
4.6.3 Extended Kalman Filter Setup

Predictor

The EKF’s predictor propagates the state from the previous time step to the current time step using body rate measurements and the motion model:

$$\hat{C}_{b_k} = \Psi_{b_k} b_{k-1} \hat{C}_{b_{k-1}}$$

(4.63)

The covariance prediction is then performed using:

$$\hat{P}_k = H_{\phi,k} \hat{P}_{k-1} H_{\phi,k}^T + H_{\psi,k} Q_k H_{\psi,k}^T$$

(4.64)

Where the Jacobians $H_{\phi,k}$ and $H_{\psi,k}$ are determined in the previous section. The initial estimate $\hat{C}_{b_0}$ either consisted of a truth value (for simulation datasets) or a sun sensor and magnetometer measurement based TRIAD (TRIaxial Attitude Determination) solution. The initial covariance is assumed to be:

$$\hat{P}_0 = \text{diag} \{0.1, 0.1, 0.1\} \text{rad}^2$$

(4.65)

The diagonal elements of the process noise covariance matrix $Q_k$, are found as the variances of CanX-5’s rate sensor errors. These variances are assumed to remain constant over time. Based on the motion model (Equation 4.29), process noise is defined by:

$$\delta \psi_k = \delta \omega_{b_{k-1}} T_k$$

(4.66)

As such, $Q_k$ is scaled by the sample time.

$$Q_k = \text{diag} \left\{ (\sigma_{\omega_x})^2, (\sigma_{\omega_y})^2, (\sigma_{\omega_z})^2 \right\} T_k^2$$

(4.67)

Corrector

To begin the corrective step of the EKF, the sensor measurements ($y_k$), observation coefficient matrices, and sensor covariances were stacked as follows:

$$y_k = \begin{bmatrix} y_{B_k} \\ y_{S_k} \\ \cos \alpha_k^1 \\ \vdots \\ \cos \alpha_k^M \end{bmatrix}$$

(4.68)

$$G_{\phi,k} = \begin{bmatrix} G_{B_{\phi,k}} \\ G_{S_{\phi,k}} \\ \cos \alpha_k^1 \\ \vdots \\ \cos \alpha_k^M \\ G_{\phi,k} \end{bmatrix}$$

(4.69)
4.6. Estimation

\[
\mathbf{G}_{n,k} = \text{diag}\left\{ G_{n,k}^B, G_{n,k}^S, G_{n,k}^{\cos\alpha^1_k}, \ldots, G_{n,k}^{\cos\alpha^M_k} \right\}
\]
(4.70)

\[
\mathbf{R}_k = \text{diag}\left\{ R_k^B, R_k^S, R_k^{\cos\alpha^1_k}, \ldots, R_k^{\cos\alpha^M_k} \right\}
\]
(4.71)

Where components were included or excluded based on the availability of sensor measurements per time step. As with \( \mathbf{Q}_k \), the components of \( \mathbf{R}_k \) consisted of the sensor variances, and were assumed to be constant over time. The Kalman gain and covariance correction were [54]:

\[
\mathbf{K}_k = \mathbf{P}_k^{-1} \mathbf{G}_{\phi,k}^T \left( \mathbf{G}_{\phi,k} \mathbf{P}_k^{-1} \mathbf{G}_{\phi,k}^T + \mathbf{G}_{n,k} \mathbf{R}_k \mathbf{G}_{n,k}^T \right)^{-1}
\]
(4.72)

\[
\mathbf{P}_k = (1 - \mathbf{K}_k \mathbf{G}_{\phi,k}) \mathbf{P}_k^{-1}
\]
(4.73)

To ensure that a valid rotation matrix is obtained at every time step, the corrector equation was arranged as follows:

\[
\delta\mathbf{\xi}_k = \mathbf{K}_k (\mathbf{y}_k - \hat{\mathbf{y}}_k)
\]
(4.74)

Where:

\[
\hat{\mathbf{C}}_{b_k} = \Xi_k \hat{\mathbf{C}}_{b_k}^{-1}
\]
(4.75)

\[
\Xi_k = e^{-\delta\mathbf{\xi}_k} = \cos \delta\mathbf{\xi}_k \mathbf{I} + (1 - \cos \delta\mathbf{\xi}_k) \left( \frac{\delta\mathbf{\xi}_k}{\delta\xi_k} \right)^T - \sin \delta\mathbf{\xi}_k \left( \frac{\delta\mathbf{\xi}_k}{\delta\xi_k} \right)
\]
(4.76)

and \( \delta\mathbf{\xi}_k = |\delta\mathbf{\xi}_k| \). The “predicted measurement” was given by:

\[
\hat{\mathbf{y}}_k = \begin{bmatrix}
\hat{\mathbf{y}}_{B_k} \\
\hat{\mathbf{y}}_{S_k} \\
\vdots \\
\hat{\mathbf{y}}_{M_k}
\end{bmatrix} = \begin{bmatrix}
\hat{\mathbf{C}}_{b_k}^{-1} \mathbf{B}_{B_k} \\
\hat{\mathbf{C}}_{b_k}^{-1} \mathbf{B}_{S_k} \\
\vdots \\
\hat{\mathbf{C}}_{b_k}^{-1} \mathbf{B}_{M_k} \\
\end{bmatrix}
\]
(4.77)

4.6.4 TRIaxial Attitude Determination Method

The TRIAD (TRIaxial Attitude Determination) method is a simple attitude estimation that only uses two non-parallel external vector measurements. It has been included here as it has been used to calculate initial conditions to initiate the EKF. The initial condition is calculated using sun sensor (\( \mathbf{\hat{s}} \)) and magnetometer (\( \mathbf{\hat{b}} \)) measurements. The steps which comprise the TRIAD method are [57]:

1. Define an intermediate reference frame, \( \mathcal{F}_i^T = [\mathbf{x}_i, \mathbf{y}_i, \mathbf{z}_i] \) Where:

\[
\mathbf{x}_i = \mathbf{\hat{s}}_i, \mathbf{y}_i = \frac{\mathbf{\hat{s}}_i \times \mathbf{\hat{b}}_i}{||\mathbf{\hat{s}}_i \times \mathbf{\hat{b}}_i||}, \mathbf{z}_i = \mathbf{x}_i \times \mathbf{y}_i
\]
(4.78)

2. Express the components of the newly formed reference frame, \( \mathcal{F}_i^T \) in terms of \( \mathcal{F}_i^T \) and \( \mathcal{F}_b^T \):

\[
\mathbf{x}_i = \mathcal{F}_i^T \mathbf{x}_{i,i} = \mathcal{F}_i^T \mathbf{\hat{s}}_{i,i}, \mathbf{y}_i = \mathcal{F}_i^T \mathbf{y}_{i,i} = \mathcal{F}_i^T \frac{\mathbf{\hat{s}}_{i,i} \mathbf{\hat{b}}_{i,i}}{||\mathbf{\hat{s}}_{i,i} \times \mathbf{\hat{b}}_{i,i}||}, \mathbf{z}_i = \mathcal{F}_i^T \mathbf{z}_{i,i} = \mathcal{F}_i^T \mathbf{x}_{i,i} \mathbf{y}_{i,i}
\]
(4.79)
And:

\[ X_t = F_{x,b}^T x_{t,b} = F_{x,b}^T \hat{s}_b, \quad Y_t = F_{y,b}^T y_{t,b} = F_{y,b}^T \hat{s}_b \hat{b}_b \]
\[ Z_t = F_{z,b}^T z_{t,b} = F_{z,b}^T x_{t,b} y_{t,b} \]  

(4.80)

3. Solve for \( C_{bt} \) where:

\[ C_{it} = [x_{t,i}, y_{t,i}, z_{t,i}], \quad C_{bt} = [x_{t,b}, y_{t,b}, z_{t,b}] \]  

(4.81)

And:

\[ C_{bt} = C_{bt} C_{it}^T \]  

(4.82)

Where: \( \hat{b}_b = \mathbf{y}_{B_k}, \hat{b}_i = \mathbf{\rho}_{B_k}, \hat{s}_b = \mathbf{y}_{S_k}, \) and \( \hat{s}_i = \mathbf{\rho}_{S_k} \)

4.7 Experimental Scenarios

The EKF was validated using three datasets. The first dataset comprised of simulated ADCS and GPS \( \cos(\alpha) \) measurements with artificial noise imparted. The purpose of this was to validate the functionality of the EKF and establish baseline expectations. This was then followed by two flight datasets containing flight sensor measurements and attitude estimates from CanX-5’s ADCS to validate filter performance in practice. It should be reiterated that flight body rate measurements were bias corrected via the on-board computer and no additional bias correction was implemented into the EKF.

4.7.1 Simulation Datasets

A simulated dataset is initially used to validate the EKF and to set expectations for the effect of GPS augmentation on attitude estimates. This is useful as simulated measurements allow full control over measurement noise properties and will ensure that sensor measurements are free from external biases. Lastly, since the sensor measurements are calculated on the CanX-5’s attitude estimates, the reliability of the attitude truth is ensured.

Simulated unbiased rate sensor data as well as magnetometer, sun sensor, and GPS \( \cos(\alpha) \) measurement data were generated at a sample period of 5 seconds. The reference truth state used consists of attitude and angular rate data gathered from CanX-5 while it was in three-axis pointing mode; the GPS antenna was pointed towards zenith to gather data (later known as Set 1). This data was then used to generate the simulated magnetic fields, and Sun vectors in both the ECI and body frames. The simulated sun sensor measurements were generated over the entire span of the reference dataset. As such, there were no loss of sun sensor measurements at the eclipse times present in the reference set.

GPS contacts were simulated based on the orbit position of CanX-5 along with GPS ephemeris data. For an ideal simulation, all contacts outside of the GPS antenna’s field of view (±80° off boresight) were dropped. The number of satellites in contact at each time step is shown in Figure 52 and ranged from 4 to 14. This number was favorable since the GPS antenna was pointed towards zenith when the truth dataset was gathered. The off-boresight \( (\alpha) \) angle for each line-of-sight (LOS) vector as well as its cosine, \( \cos(\alpha) \) could then be calculated with knowledge of the GPS antenna boresight and LOS vectors.
4.7. Experimental Scenarios

Zero-mean Gaussian noise was then imparted directly into the rate sensor measurements and the cos(α) measurements. The standard deviation of the additive rate sensor noise was 0.1°/s, a value chosen to be representative of CanX-5 rate sensor errors. The standard deviation of the cos(α) noise was assumed to be 0.99 based on experimentation and validation with mapping functions. Based on telemetry from CanX-5, it was determined that both the sun sensor and magnetometer data possessed errors with a standard deviation of approximately 1°. Noise was imparted by creating a small 3-by-1 rotation vector for each time step out of zero-mean Gaussian random values (\( \mathbf{e}_k \)). This vector was converted to a rotation matrix using small angle parameterization (\( \mathbf{C}_{\text{err}_k} \approx \mathbf{1} - \mathbf{e}_k \hat{x} \)) and multiplied by either the normalized true Sun and magnetic field vectors in the body frame (i.e. \( \mathbf{y}_{B_k} = \mathbf{C}_{\text{err}_k} \mathbf{y}_{B_{\text{true}}} \)).

This zero-mean Gaussian noise model was used so that the simulated sensor measurements would possess realistic variances while remaining free from any unknown biases that may occur in flight datasets. Such biases, should they exist, may skew estimate accuracy and affect initial validation of the EKF if not corrected for in the motion and observation models.

4.7.2 Flight Data Sets

Two sets of flight data from previous on-orbit operations were used to further assess the performance of the EKF. These sets consist of magnetometer and Sun vectors in both the body and ECI frames as well as bias corrected body rate measurements at a sample rate of 2 seconds. For both sets, the GPS receiver provided the orbital position of CanX-5 and \( C/N_0 \) values at a rate of 5 seconds. LOS vectors were then calculated from the reported orbital position and GPS ephemeris data. One notable factor in these datasets was that \( C/N_0 \) reporting were constrained to integer values between 25 and 51 dB-Hz (i.e. a reported \( C/N_0 \) value of 51 dB-Hz could be higher). These limitations in reporting were due to the use of a compressed range log from the receiver. Although, it may be possible to obtain more accurate \( C/N_0 \) values from a full range log, the restricted values were used as they were more reflective of CanX-5’s
operating state during nominal operations and were the only measurements available at the time of this analysis.

**Set 1 - Attitude Hold**

The first set (Set 1) consists of CanX-5 in attitude hold mode with its GPS antenna pointing towards zenith. Set 1 occurs on 09/19/14 between the times of 07:45:04 and 10:58:44 UTC. This set also contains two areas of eclipse occurring at 08:44:36 - 09:18:24 UTC and 10:22:22 - 10:56:10 UTC where measurements from the sun sensors are unavailable.

**Set 2 - Orbit Maintenance Maneuver**

The second set covers an orbit keeping maneuver on 10/31/14 between 17:26:04 and 17:31:54 UTC. The maneuver begins with an attitude slew to an operations attitude after 62 seconds, followed by an 8.5 second long CNAPS thrust and 2 minute attitude hold after 140 seconds, ending with a slew back to the nominal attitude with a 3 minute attitude hold at 184 seconds. CanX-5’s attitude change in Set 2 is shown in Figure 53 where roll, pitch, and yaw are the spacecraft’s attitude about the x, y, and z axes respectively in the J2000 ECI frame. The entire span of Set 2 is sunlit and eclipse performance is assessed by manually denying sun sensor measurements to the EKF.

![Figure 53: CanX-5 Attitude in Set 2](image)

**4.8 Mapping Function**

The sunlit portion of Set 1 was used as the calibration dataset to create a 4\(^{th}\) order polynomial mapping function from $C/N_0$ to $\cos(\alpha)$. This was performed by binning measured $C/N_0$ values along with reference $\cos(\alpha)$ values. Reference $\cos(\alpha)$ values were calculated from the true spacecraft attitude,
known GPS boresight in the body frame, and GPS LOS vectors calculated from external GPS ephemeris data and orbital position reported by the GPS receiver. Figure 54 shows the calibration bin distribution of the $C/N_0$ values. The distribution of each bin were fairly wide as $C/N_0$ reporting was limited to integer values hence the minimum bin size was 1 dB-Hz. Furthermore, the receiver possessed a $C/N_0$ reporting range. Any $C/N_0$ measurements falling outside would be rounded to either the minimum or maximum of the reporting range, leading to a wider variance in the lowest and highest bins. The observed reverse slant was attributed to the free space path loss adjustment whereas, unadjusted bins were typically vertical. Since the mapping function was created by fitting a curve to the mean $\cos(\alpha)$ of each $C/N_0$ bin, accuracy of the function is dependent on the $\cos(\alpha)$ spread per bin.

![Figure 54: Calibration Bin Distribution](image)

Figure 55 shows the mapping function obtained from FSPL corrected $C/N_0$ measurements along with indicated performance. Performance was calculated from the error between $\cos(\alpha)$ estimates obtained through the mapping function and reference $\cos(\alpha)$ values calculated from the calibration dataset. It was found through experimentation that when using the mapping function, corrected $C/N_0$ values under 130 dB-Hz had to be dropped to obtain acceptable accuracy. This was attributed to two reasons. First, as evidenced in Figure 54, bins lower than this value possessed a very large spread with high inaccuracy. Secondly, values below 130 dB-Hz are related to contacts which fell outside the GPS antenna’s designed field of view ($80^\circ$). While it would be possible for the GPS antenna to occasionally establish contacts outside its field of view, $\cos(\alpha)$ measurements from these contacts were extremely inaccurate and unreliable. Furthermore, due to the receiver imposed maximum limit on $C/N_0$ reporting, the higher $C/N_0$ bins also possessed greater spread and would decrease accuracy. These bins could not be dropped without affecting the EKF’s ability to obtain GPS measurements at regular intervals. In general, a non-range limited $C/N_0$ set with higher precision will enable the selection of smaller bin sizes and reduce the variance in each bin.
4.9 Simulated Data Results

For initial validation, the EKF was fed four simulated sensor datasets representative of different operation scenarios, described in the following subsections. For GPS cases, the \( \cos(\alpha) \) observation model was used. Root-mean-square (RMS) errors were calculated and expressed in terms of a small rotation vector between the estimated and true rotation matrices. The components of that error vector with respect to the body frame axes: \( \theta_x \), \( \theta_y \), and \( \theta_z \) were tabulated in Table 10. The \( \theta_x \), \( \theta_y \), and \( \theta_z \) angular errors for each case were also plotted in the figures below as solid lines. Moreover, the estimated standard deviation (\( \pm 3\sigma \)) envelopes for each axis, representing the estimate confidence of the EKF, were also plotted in the figures as dotted lines.

Table 10: Simulated Attitude Estimation Errors

<table>
<thead>
<tr>
<th>Case</th>
<th>( \theta_x )</th>
<th>( \theta_y )</th>
<th>( \theta_z )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetometer + Sun Sensors</td>
<td>0.72</td>
<td>0.92</td>
<td>0.58</td>
</tr>
<tr>
<td>Magnetometer Only</td>
<td>3.03</td>
<td>5.85</td>
<td>1.65</td>
</tr>
<tr>
<td>Magnetometer + GPS</td>
<td>1.14</td>
<td>2.41</td>
<td>0.76</td>
</tr>
<tr>
<td>Magnetometer + Sun Sensors + GPS</td>
<td>0.68</td>
<td>0.86</td>
<td>0.56</td>
</tr>
</tbody>
</table>

Nominal Operations (Magnetometer + Sun Sensors)

This case represents nominal operation of the EKF using only information provided from CanX-5’s purpose built ADCS sensors that are operating unimpeded without any GPS augmentation. As such, all sunlit cases running with only magnetometer and sun sensor data are referred to as “nominal operations”
and are intended to provide an experimental baseline. Overall, estimates from this set proved to be very accurate (Figure 56) with few overconfident areas (errors which fell outside of the ±3σ envelope). Some jitter in the error curve was also observed, manifesting as sudden irregular spikes at 7000 seconds in \( \theta_y \) and 11500 seconds in both \( \theta_x \) and \( \theta_y \). These jitters were likely a product of the truth dataset which was derived from flight data and subject to inherent ADCS inaccuracies. It should also be noted that a nominal case (where both magnetometer and sun sensor measurements are used) spanning this period of time for CanX-5 is not encountered during flight. In its actual orbit, the spacecraft would encounter periods of eclipse where sun sensor measurements would be unavailable and the sun sensors disabled by the ADCS.

![Figure 56: Simulated – Magnetometer + Sun Sensors](image)

**Eclipse Operation (Magnetometer Only)**

The following scenario was run with the EKF attitude estimates only receiving correction from magnetometer measurements. This test was performed to demonstrate the drop in estimator performance when one of CanX-5’s attitude sensors are denied functionality, such as the sun sensors in eclipse. As shown in Table 10, \( \theta_x \) and \( \theta_z \) RMS errors triple uniformly. Sharper increases in \( \theta_y \) RMS error and ±3σ was likely due to the fact that a single magnetometer measurement can only provide correction primarily in two axes [61]. In this case, magnetometer correction was potentially limited in the y-axis. This was further supported by the fact that the ±3σ envelop for \( \theta_y \) converges much later than those corresponding to the other two axes (Figure 57). Lastly, periodic dilation and contraction of the ±3σ envelop was observed in all axes at approximately 3000 second intervals; about half of CanX-5’s orbital period. It was found that these were a product of the magnetometer correction resulting from fluctuations present in the IGRF magnetic field model. This behavior could also be seen in other cases as well.
Chapter 4. Global Navigation Satellite System Based Attitude Determination

GPS Correction During Eclipse (Magnetometer + GPS)

Figure 58 illustrates the results of an eclipse scenario where the estimator is receiving measurements from the GPS antenna along with existing magnetometer correction. With GPS correction, the RMS errors are two to three times lower than the magnetometer-only estimates during simulated eclipse. It was observed that the $\theta_y$ errors were still elevated compared to the other two axes. This reflects the scenario seen in the eclipse case where y-axis is aligned with the magnetic field, limiting magnetometer correction in this direction. However, as indicated in previous sections, corrections via GPS measurements were also limited to two axes as well. Since the antenna boresight was aligned with the spacecraft’s y-axis, the corrective ability of GPS along that axis was also diminished. Furthermore, the changes in the GPS uncertainty and accuracy could also be affected by the number of acquired GPS satellites available per each measurement period. This was supported by increased error at 3500 and 6500 seconds, which were times corresponding to decreasing numbers of acquired GPS satellites (Figure 52).

GPS Correction During Nominal Operations

GPS cos($\alpha$) measurements are then combined with both magnetometer and sun sensor measurements to demonstrate the effect of GPS correction during nominal (sunlit) operations. As shown in Table 10 and Figure 59, RMS error is slightly lower than the nominal case. Additionally, the total error curve appears lower in overall magnitude with some reductions in error curve anomalies.

4.10 Flight Set 1 Results

To ease analysis, a subsection of Set 1 was used to perform flight validation of the EKF. Ranging from 08:18:24 to 09:25:04 UTC, this subsection contains one sunlit and one eclipse section. The EKF for
both cases were initialized using a magnetometer and sun sensor based TRIAD estimate as the initial condition. Two test cases were run, a nominal operations case with only the sun sensor and magnetometer and a case with additional GPS correction via the \( \cos(\alpha) \) observation model (Table 11). As with the simulated data, the estimate error (solid line) and the \( \pm 3\sigma \) envelope (dotted line) representing estimate confidence were plotted.
Table 11: Set 1 Attitude Estimation Errors

<table>
<thead>
<tr>
<th>Case</th>
<th>RMS Errors [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\theta_x$</td>
</tr>
<tr>
<td>Nominal (Sunlit)</td>
<td>0.5</td>
</tr>
<tr>
<td>Nominal (Eclipse)</td>
<td>3.00</td>
</tr>
<tr>
<td>Nominal (Combined)</td>
<td>2.16</td>
</tr>
<tr>
<td>Nominal + GPS (Sunlit)</td>
<td>0.49</td>
</tr>
<tr>
<td>Nominal + GPS (Eclipse)</td>
<td>2.73</td>
</tr>
<tr>
<td>Nominal + GPS (Combined)</td>
<td>1.97</td>
</tr>
</tbody>
</table>

4.10.1 Nominal Operations (Sun Sensor + Magnetometer)

The nominal operations case possessed a low amount of error under sunlit conditions. However, as shown in Figure 60, the error increased heavily during eclipse due to the absence of sun sensor measurements and contributed disproportionately to the overall RMS. Regardless, the error curve for much of the set remains within the $\pm 3\sigma$ envelop with the exception of some overconfident values around 3000 seconds. Lastly, $\theta_x$ errors appeared to be the greatest implying limited magnetometer or GPS correction in that direction. Conversely, the $\theta_z$ error remains the lowest and the least perturbed by eclipse conditions.

![Figure 60: Set 1 - Nominal Operations](image)

4.10.2 Nominal Operations + GPS

As with the simulated set, GPS measurements (Figure 61) served to slightly increase estimate accuracy during sunlit conditions. However, improvements to the estimate during eclipse were also limited to only
4.11. Flight Set 2 Results

a few hundredths of a degree. The error curve indicates that the areas containing overconfident values still form but are reduced in magnitude. However the GPS measurements appear to shift the eclipse section of the curve in the positive direction. At first, it was believed that the limited correction was due to the sparse frequency of GPS measurements and the low precision of $C/N_0$ measurements. Low $C/N_0$ precision would have impact accuracy in two ways: First, the inaccuracies would directly impact the corrective ability of the GPS measurements themselves. Secondly, the limited precision forces the use of larger bin sizes when creating the mapping function, reducing the reliability of $C/N_0$ to $\cos(\alpha)$ correlation.

While the above factors certainly contribute to the corrective ability of the GPS, another possible cause of inaccuracy was found after comparing these results with simulation and Set 2 data. Recall, that estimation error for flight sets was determined using attitude estimates output by CanX-5 as an attitude truth set. As Set 1 contains a naturally occurring period of eclipse where CanX-5’s attitude solution was calculated without its sun sensors. Therefore, it was likely that the attitude solution of the EKF was in fact more accurate than the truth set from CanX-5, skewing the calculated errors. This phenomena was negligible for the attitude truths from the sunlit section of Set 1 and Set 2 which were obtained when CanX-5 was using both its magnetometer and sun sensors. This case effectively illustrates the challenges of accurately accessing eclipse performance of the EKF using eclipse flight data.

![Figure 61: Set 1 – Nominal Operations + GPS](image)

**Figure 61: Set 1 – Nominal Operations + GPS**

### 4.11 Flight Set 2 Results

Since Set 2 was comprised of completely sunlit data, EKF performance was analyzed in the same manner as the simulated dataset. To assess EKF performance during eclipse maneuvers, sun sensor data was excluded from the EKF corrector’s input. The results for each experiment case were tabulated in
Table 12: Set 2 Attitude Estimation Errors

<table>
<thead>
<tr>
<th>Case</th>
<th>RMS Errors [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\theta_x$</td>
</tr>
<tr>
<td>Magnetometer + Sun Sensors</td>
<td>1.53</td>
</tr>
<tr>
<td>Magnetometer + Sun Sensors + GPS</td>
<td>1.53</td>
</tr>
<tr>
<td>Magnetometer Only</td>
<td>20.34</td>
</tr>
<tr>
<td>Magnetometer + GPS</td>
<td>12.83</td>
</tr>
</tbody>
</table>

4.11.1 Magnetometer + Sun Sensors (Nominal Operations)

As with previous datasets, a nominal operations case using only sun sensor and magnetometer measurements was run to provide an experimental baseline. Since the spacecraft was actively slewing and thrusting, the error in all axes were higher than those in Set 1 and the simulated set. The error and $\pm 3\sigma$ envelope were plotted in Figure 62 and generally show converging values with reasonable confidence. Some increases in error could be seen at approximately 65 and 180 seconds which correspond to the two attitude slews in the set.

![Figure 62: Set 2 – Magnetometer + Sun Sensors](image)

4.11.2 Magnetometer + Sun Sensors + GPS (Nominal Operations + GPS)

As with previous datasets, GPS measurements had little effect on overall errors when sun sensors were active. Figure 63, confirms that the overall error curves remained virtually unchanged from the nominal
4.11. Flight Set 2 Results

operations case. It was also found that the $\theta_y$ error experienced an increase on the order of thousandths of a degree over the nominal case. This difference was negligible and arose as a product of the $\cos(\alpha)$ variance being tuned for maximum error reduction in the eclipse case.

Figure 63: Set 2 – Magnetometer + Sun Sensors + GPS

4.11.3 Magnetometer Only (Eclipse Conditions)

As Set 2 did not possess any periods of eclipse, eclipse performance was assessed by manually excluding sun sensor measurements. As expected, this case possess the highest amount of error out of all the Set 2 cases; $\theta_x$ and $\theta_y$ errors were almost equally high. This was due to the limited two-axis correction that can be provided with magnetometer measurements impeding correction after the maneuver was started. Figure 57 also indicates that most of the estimates are overconfident; with errors deviating from the $\pm 3\sigma$ envelope after the first slew. Given the shorter time span of Set 2, dilation and contraction of the $\pm 3\sigma$ envelope also occurred mainly at the two attitude slews.

4.11.4 Magnetometer + GPS (Eclipse + GPS)

The GPS measurements when applied to the eclipse case had a greater effect on accuracy than in Set 1; almost halving the RMS error from the eclipse case in all axes. Figure 65 shows that although there exist periods with overconfident attitude estimates between the two slews, the GPS measurements were nearly able to bring the estimate errors back within the $\pm 3\sigma$ envelopes after the second slew. There were two reasons for this increase in performance over Set 1. First, it was believed that a shorter dataset reduces the amount of time available for errors to propagate. The second reason related to the inaccuracy issue discussed in Set 1. Set 2 was a completely sunlit maneuver which meant that the attitude truth was collected using both CanX-5’s magnetometer and sun sensors. This meant that the accuracy of the error calculations were much higher than those in Set 1. Additionally, the trend of eclipse
Figure 64: Set 2 – Magnetometer Only

improvement from the results agree with those obtained from the simulated dataset despite the fact that Set 2 covered numerous attitude slews and used flight measurements. These observations support the hypothesis regarding the eclipse inaccuracies in the Set 1 truth set and legitimizes GPS augmentation as a viable means of improving attitude estimation accuracy during eclipse.

Figure 65: Set 2 – Magnetometer + GPS
4.12 Summary

An extended Kalman filter was created to assess the additional corrective ability of $C/N_0$ based measurements from a single GPS antenna when combined with existing sun sensor and magnetometer measurements. This augmentation was performed to determine if it would be possible to retain some attitude estimation accuracy during periods of eclipse when the spacecraft’s sun sensors were denied functionality. Results obtained using simulated sensor measurements indicated that under experimental conditions, the estimator’s accuracy was increased by a factor of two to three during simulated periods of eclipse; contrasted to magnetometer only estimates. However, it was found that during sunlit operations, the contributions of GPS measurements were limited due to the availability of more accurate measurements from the purpose built ADCS sensors. Despite this, GPS measurements would still provide minor contributions to the general accuracy and confidence of the resulting sunlit estimates.

When introduced to a lengthy attitude hold flight dataset, it was initially found that the GPS measurements only provided minimal correction during eclipse. However, further evaluation with shorter flight datasets employing more accurate reference truth attitudes, revealed that GPS measurements could significantly reduce large amounts of error even during attitude slewing in eclipse conditions. Nonetheless, accuracy degradation could be attributed to the sparse GPS sampling period and low precision of the $C/N_0$ measurements. The low precision of the $C/N_0$ measurements not only detracted directly from the accuracy of the EKF correction but also from the accuracy of the $C/N_0$ to $\cos(\alpha)$ mapping function. More specifically, the minimum size of the $C/N_0$ bins that could be reliably used to fit the function. In the case of CanX-5, while the attitude references were accurate, the $C/N_0$ bin size was fixed at 1 dB-Hz. Based on estimation results on CanX-5 as well as experiments from other missions, it is predicted that using higher resolution $C/N_0$ measurements with an unconstrained range would increase estimation accuracy. In addition, it was found that using less accurate estimates from the satellite gathered under eclipse conditions as a truth dataset also had the potential to misrepresent EKF performance.

Moving forward, it is recommended that the EKF with the $C/N_0$ observation model be retained with more precise $C/N_0$ measurements taken at a higher sample rate to increase estimation accuracy. $C/N_0$ precision can be increased by enabling the full range log on CanX-5’s GPS receiver. This logging mode records $C/N_0$ at a resolution of 0.1 dB-Hz with no measurement range constraint.
Chapter 5

Conclusion

This thesis has documented the modeling and measurement of the uplink and downlink antenna arrays on the CanX-7 nanosatellite, the development of an L-band antenna for the CanX-7 ADS-B payload, and the development and assessment of a GNSS (Global Navigation Satellite System) based attitude determination system. The author’s contributions to CanX-7, enabled the nanosatellite to meet its communications and ADS-B payload requirements. Furthermore, the work towards the GNSS attitude determination system will enable current and future GPS equipped nanosatellite missions to perform operations in eclipse or in the event of sun-sensor or star tracker failure. Moreover, these contributions have addressed and provide solutions to communications and attitude determination based challenges encountered by small satellites.

CanX-7 is equipped with UHF and S-band antenna arrays to receive uplinked commands and to transmit downlinked telemetry and payload data. Omnidirectionality in the design of these arrays to allow the spacecraft to successfully maintain communications independent of attitude. This performance parameter is quantified by coverage, the percentage of the antenna’s radiation pattern possessing enough signal strength to close the communications link with sufficient margin. Before performing any physical measurement, the performance of the antenna arrays on CanX-7 is simulated using a finite element solver. This practice reduces costs and provides baseline expectations for real world antenna measurement. Simulations were performed on CanX-7 in its drag sails stowed and deployed configurations. The resultant radiation confirmed that the deployed sails had a reflective or scattering effect on the radiation patterns. Antenna measurement then comprised of measuring a number of two dimensional cross sections of the real world radiation pattern and approximating the coverage. Measurement was not performed for the sail-deployed case due to the increased logistics of transporting and testing the sails along with the limited utility of guaranteeing communications during deorbiting. It is found from both simulation and measurement that the uplink and downlink antennas possess sufficient coverage for CanX-7’s communications subsystem to function properly and meet requirements.

The knowledge gained from antenna simulation and measurement was applied toward designing an L-band antenna for the secondary ADS-B payload on CanX-7. Due to space constraints and lack of fine pointing, the antenna design was constrained profile wise. Additionally, the antenna was required to be circular polarized to enable it to receive information from numerous linearly polarized transmitting antennas with minimal polarization mismatch loss. The antenna redesign was performed through a finite element solver and resulted in a single feed, single substrate patch antenna. The substrates used for
the antenna designs were selected based on the sensitivity of the dielectric constant and expansion on the temperature. Circular polarization was induced by cutting a cross slot into the patch to cause the antenna to radiate in two different modes. While intermediary designs did obtain a suitable amount of antenna gain to ensure proper operation of the receiver, it was found that the center frequency of the antenna would shift when mounted on the spacecraft bus. This shift resulted in a significant decrease in antenna gain and loss of circular polarization. The final design iterations were sized and optimized while attached to the spacecraft bus to compensate for frequency shifting. Future work for this antenna includes fabrication and antenna measurement to validate its performance.

Lastly, a single onboard GPS antenna was successfully augmented to function as a coarse attitude determination method. A mapping function representing the antennas radiation pattern, enabled carrier to noise density ratio measurements to be correlated to the off-boresight elevation of each contacted GPS satellite. The corrective ability of this method was accessed by integrating GPS measurements along with existing sun sensor and magnetometer measurements into an extended Kalman filter onboard the CanX-5 spacecraft. This augmentation was performed in the context of accessing the ability of the spacecraft to retain some attitude estimation accuracy during periods of eclipse when the spacecraft’s sun sensors have been denied functionality. It was found when evaluated using simulated sensor measurements, estimator accuracy increased by a factor of two to three during simulated periods of eclipse where sun sensor data was unavailable; contrasted to magnetometer only estimates. Conversely, GPS measurements would only provide minimal improvement during sunlit conditions. The EKF was then tested using flight datasets containing naturally occurring or artificial periods of eclipse. These tests revealed that GPS augmentation would reduce estimation error even in sets where the spacecraft was undergoing attitude slews. Tests with flight data also predicted that the sampling period, GPS measurement precision, and range affect the accuracy of the GPS based estimate correction. Additionally, the lower accuracy of the reference attitudes in the natural eclipse sets may cause estimation errors to be over reported. This work is significant for SFL spacecraft as current and future GPS enabled spacecraft now possess an additional means of attitude determination, enabling greater safety and flexibility in mission planning and operations.
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