Abstract

Government transport regulations and practical considerations limit the height of terminal antennas used in satellite-based mobile asset tracking applications to less than 2.5 cm. For Orbcomm systems that operate at 138 MHz (downlink) and 150 MHz (uplink), this implies an antenna that is just over one-hundredth of a wavelength tall. Achieving good efficiency and operating bandwidth with such an ultra low profile antenna is fundamentally difficult. Here we consider the possibility of using a multi-arm normal mode cylindrical helix antenna to achieve a significant fraction of the performance of a full size Orbcomm reference antenna in a more compact form. In order to simplify impedance matching, we introduced an internal magnetic coupling loop that can be increased or decreased in radius in order to achieve a good match. In order to identify the optimum design, we assessed the radiation efficiency and bandwidth of the antenna as a function of the key design parameters (helix height, radius, number of arms, number of turns, feed loop radius, presence or absence of crossbars that connect the arms at the top of the helix) using simulations and validated the results by measuring the performance of selected hardware prototypes. Further, we developed an equivalent circuit model that allows one to extract key design information much more quickly than would be possible by simulation or measurement. Our design curves show that bandwidth can only be increased with height and is independent of radius at ULP height. We found that efficiency increased significantly with helix radius but the presence of crossbars yielded only a marginal increase in efficiency. A pair of ULP antennas is required for uplink and downlink. To connect two such antennas to a standard Orbcomm modem with a single antenna port, we considered three duplexer designs: a carrier operated relay, a diplexer filter, and a novel complementary feed. The latter is both simpler and more effective than the first two. Over a 72-hour period, our best ULP antenna design (four arms, no cross bar, helix radius ~0.04 wavelengths) was able to exchange 45% of the messages that a standard Orbcomm reference antenna could.
Preface

This thesis is original, independent work by the author, Yin He, under the supervision of Prof. David G. Michelson. The ANSYS HFSS simulations were conducted in collaboration with Prof. Deniz Bolukbas. Several results from this thesis project has been published in conference proceedings:


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### Glossary

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<td>ADS</td>
<td>Advanced design system</td>
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<tr>
<td>COR</td>
<td>Carrier operated relay</td>
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<td>GPS</td>
<td>Global positioning system</td>
</tr>
<tr>
<td>HFSS</td>
<td>High Frequency Structural Simulator (ANSYS software)</td>
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<td>IC</td>
<td>Integrated circuit</td>
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<tr>
<td>KVL</td>
<td>Kirchhoff’s Voltage Law</td>
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<td>M2M</td>
<td>Machine to machine</td>
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<tr>
<td>OC</td>
<td>Open circuit</td>
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<tr>
<td>PCB</td>
<td>Printed circuit board</td>
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<tr>
<td>S/C</td>
<td>Subscriber communicator</td>
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<tr>
<td>SC</td>
<td>Short circuit</td>
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<td>SMT</td>
<td>Surface mount technology</td>
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<td>ULP</td>
<td>Ultra low profile</td>
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I would like to thank Rob Calis of Inevitable Technologies for bringing such an interesting research topic. It is very exciting to work on a research project with such a clear industrial application and I look forward to seeing my design on top of semitrailers.

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For my mother and father
Chapter 1: Introduction

For a variety of practical reasons, it may be necessary to realize a Low Profile antenna, i.e., a radiating element that is less than a tenth of a wavelength tall. There are several well-known engineering design challenges associated with such electrically small antennas: 1) The height limitation of low profile antennas makes them unable to fill a Chu sphere, which degrades their bandwidth. This means that practically, there is a limit to the bandwidth of the antenna. 2) Building a physically small matching network to improve the small antenna efficiency over a wide range of frequencies is not practical. 3) Close proximity to a large ground plane reduces the radiation resistance of the antenna and degrades the radiation efficiency. Additionally, the losses introduced by external matching networks often reduce total efficiency further.

In this thesis, we are motivated by the need of the Orbcomm satellite communications system for antennas with operating frequencies of 138 and 150 MHz that are less than one inch (2.54 cm) tall. We refer to such antennas as Ultra Low Profile (ULP) because they are just over one-hundredth of a wavelength tall – up to an order of magnitude shorter than a conventional Low Profile antenna. The challenges associated with the design and implementation of ULP antennas are even harsher than those associated with Low Profile antennas. Very few examples of such antennas have been reported in the literature.

1.1 Satellite Based M2M using Orbcomm System

During the past decade, M2M (machine-to-machine) communication via satellite has revolutionized the fields of mobile asset tracking and monitoring of fixed assets in remote locations. Industry leaders predict that several hundred thousand satellite modems based upon the Orbcomm, Globalstar and Inmarsat satellite communications systems will be deployed worldwide during the next five years in order to meet the growing demand. In 2013, Orbcomm’s M2M business was valued
at $74.21 million USD. The focus of the satellite-based M2M industry is transitioning from monitoring of large fixed assets (e.g., oil wells and hydroelectric dams) to monitoring of small fixed assets and tracking of mobile assets (e.g., shipping containers, semi trailers and heavy construction vehicles). Practical considerations and government transportation regulations limit the height of terminal antennas installed atop shipping containers and semi trailers to less than one inch (2.54 cm). For Orbcomm-based satellite communications systems that operate at VHF frequencies (137 – 138 MHz for receive and 148 – 150.05 MHz for transmit), 2.54 cm corresponds to a terminal antenna height of just over one-hundredth of a wavelength. However, designing an ultra low profile antenna that achieves good efficiency and the required operating bandwidth is fundamentally difficult.

The Orbcomm satellite communication system is a frequency division duplex (FDD) system. When the system was conceived in the early 1990’s, they chose to operate at VHF because: 1) the bands at 138 and 150 MHz are allocated to satellite communications on a worldwide basis and 2) it allowed them an incredible business advantage over other systems by being able to sell inexpensive transmitters (with expected cost of $50 to $200 US) that utilize ubiquitous and commercially available radio electronics [5]. Furthermore, since the electronics had already been tested in the marketplace, they would not require years of specialized testing. The uplink and downlink bands are a standard duplex pair that is based on the 12 - 15% operating bandwidth of a typical dipole antenna. A half-wave dipole or quarter-wave monopole antenna is generally suitable for fixed asset monitoring of sensors in remote areas such as oil pipelines and propane tanks. As the range of applications expand to tracking, monitoring, messaging for mobile assets like trucking and heavy construction equipment fleets, weather reports and fleet monitoring for the fishing industry, much smaller antennas are required. Smaller antennas generally incur performance tradeoffs.
Currently available low profile Orbcomm antennas are between 3 - 4 inches or 7.5 - 10 cm tall. Because they exceed the 2.54 cm limit, they cannot be mounted on the roof of a semi trailer and must be mounted on the front vertical face where their view of the sky and the Orbcomm satellite constellation is severely obstructed. For best performance, the antenna should be mounted on the roof of the semi trailer but this requires an antenna which is much shorter than current antennas. The origin of the 2.54 cm height limitation comes from US Department of Transportation (DOT) regulations as well as practical limitations. The US DOT sets the maximum allowable height of semitrailers from 4.11 m to 4.27 m [6] and width of 2.6 m. While clearance height is managed by each state to comply with clearance of tunnels, we consider 4.11 m as the minimum height required for all interstate traffic.

The height of a roof-mounted antenna is restricted because the semitrailers have an arched roof with a height of 3.81 cm so that rain will easily drain rather than pool. This transverse arch is tapered 121.92 cm from the front of the trailer to provide a smooth transition. This taper provides room at the front of the trailer for a 2.54 cm tall ULP antenna system. Figure 1 shows that the maximum dimensions of the entire receiver system (including battery, Orbcomm subscriber communicator modem, and ULP antenna) are 2.54 cm tall by 40.64 cm wide by 2.6 m long. Orbcomm recommends that the entire antenna system should fit within a space of 10.16 cm by 30.48 cm because when the antenna system gets too big, it gets harder to affix the antenna to the rooftop. Realistically, the antenna will be less than 2.54 cm tall to fit inside of a rugged radome. In this thesis, we consider a ULP antenna to be less than 2.54 cm tall.
An overview of messaging within the Orbcomm satellite network is presented in Figure 2. The ULP antenna is attached to an Orbcomm’s subscriber communicator (S/C) modem. When the S/C originates a message, it transmits to a satellite that receives, reformats and relays the message to a Gateway Earth Station, (GES). The GES transmits the message over a dedicated line to the Gateway Control Centre (GCC) that places the message on the public switched network for delivery to the recipient subscriber’s Internet provider. The recipient receives an email. The recipient can respond by sending an email, which goes through the ground segment to the GES to be picked up by a satellite, and transmitted to the S/C. More information about the Orbcomm messaging protocol is given in Appendix A.

Ideally, a ULP antenna should be: 1) simple and cost effective, 2) work with current subscriber communicators as there are already hundreds of thousands deployed, 3) power efficient as it is mobile, 4) have a small footprint as required by Orbcomm, and 5) easy to manufacture and install.
1.2 Previous Work and Limitations

We are not aware of any antennas on the market, nor any reported in the literature that meet the ULP height requirement. The Chu-Harrington limit sets an upper limit on the bandwidth of an electrically small antenna of given size [7][8]. While the Chu-Harrington and Gustafsson limits predict that a ULP antenna with a diameter of a sixteenth of a wavelength should be able to achieve a bandwidth of 2 MHz at 150 MHz, they do not explain how to realize this result. Current volumetric antennas reported in the literature at VHF frequencies are still too tall for our application: Best’s folded spherical helix antenna is 0.06λ tall [9], Lim et al.’s multiple folded antenna is 0.045λ tall [10], and Mehdipour et al.’s spherical wire antenna is 0.08λ tall [11].

To capture the maximum volume within a structure that is suitable for inexpensive manufacture, we consider the design of ULP cylindrical helix antennas. The axial mode helix is a popular design for antennas whose diameter is much larger than height [12][13][14][15][16]. However, the beam pattern and a circumference on the order of a wavelength make this design unsuitable for Orbcomm radiation and size requirements. We consider a normal mode cylindrical
helix antenna that meets Orbcomm’s omnidirectional radiation pattern requirement. However, its performance at ULP height has not been previously reported.

Feeding an electrically small antenna is challenging as improper technique will further reduce bandwidth, efficiency, or both. Previously proposed methods for feeding helices cannot be easily incorporated into a ULP helix. At our extremely flat aspect ratio, the lengths of the transmission length in Dinh et al.’s tap feed [17] and the crossed transmission line feed described by Sakaguchi et al. [18] would become substantial. Incorporating external components for impedance matching would introduce undesirable losses like Yu et al.’s [19] inductive feed.

Finally, simulating the antenna and setting up the environment according to typical online tutorials is not trivial in Ansys HFSS. While HFSS is an industry standard for antenna simulation, we also need a method of verifying the results.

1.3 Objectives and Approach

In this thesis, we take a different approach from the PIFA and planar antenna design that are currently popular in Orbcomm applications. We consider volumetric antenna design for ULP height following research indicating their improved performance over planar antennas because they are more effective at utilizing the Chu sphere [20]. Specifically, we seek to determine whether a cylindrical helix antenna, which combines favourable antenna properties with ease of manufacture, can still provide acceptable bandwidth and efficiency when its height is reduced to one hundredth of a wavelength and it becomes an Ultra Low Profile antenna.

We have further sought to: 1) demonstrate the merits of using a coupling loop feed (in place of the more conventional transmission line feed) to simplify impedance matching to the ULP cylindrical helix, 2) determine how bandwidth and radiation efficiency depend upon the dimensions and configuration of the ULP cylindrical helix, and 3) assess the merits of alternative schemes for
connecting a pair of ULP cylindrical helix transmitting and receiving antennas to a transceiver. We verified the accuracy of our results by demonstrating good agreement between the results that we obtained through: 1) numerical simulations (using ANSYS HFSS), 2) our equivalent circuit model for the loop coupled cylindrical helix, and 3) field testing our prototype ULP cylindrical helix antennas.

In particular, we recognize the limitations of feeding helices using techniques reported in the literature and propose a magnetic feed loop that simplifies coupling and impedance matching to a cylindrical helix antenna. We design an equivalent circuit model representation of the ULP and demonstrate that it can be used to identify the best ratio of feed loop to helix diameter orders of magnitude faster than modeling software or measurements can. The limitations of feed loop design and strategies for overcoming them are discussed in more detail in Chapter 2.

By using electromagnetic simulation software (Ansys HFSS) and taking real measurements, we investigate how simple changes in the geometry of normal mode multi-arm cylindrical helix antennas such as number of arms, height, radius, and presence or absence of crossbars can improve radiation efficiency and bandwidth. Our design curves extend current understanding of normal mode multi-arm cylindrical helical antenna and demonstrate that it is possible to achieve high radiation efficiency, contrary to the popular belief from proponents of axial mode helices.

Finally, we discuss various methods of connecting two ULP antennas to an Orbcomm modem as the 2 MHz theoretically achievable bandwidth barely covers either the uplink or downlink band. We envision a two-antenna system, one for transmit and the other for receive, with a duplexer that connects them to a single antenna port on the Orbcomm modem. The additional duplexer is required as it is highly unlikely that vendors will modify their modem design to include separate transmitting and receiving ports when there are already hundreds of thousands of such conventional Orbcomm modems already in use. Another option is to realize a dual resonance ULP antenna solution which
does not require a duplexer, but this was not studied due to lack of time. A variety of system level parameters are considered for the duplexer: bandwidth, isolation, power handling, switching delay, insertion loss, power consumption, price, and size. We compared three alternative duplexer concepts: 1) a conventional carrier operated relay (COR), 2) a conventional diplexer filter, and 3) a novel approach that we call a “complementary feed” that offers very good performance at low cost.

1.4 Thesis Outline

This remainder of this thesis is structured as follows.

In Chapter 2, we begin by reviewing the fundamental limitations of ULP antenna design pertaining to bandwidth, impedance matching, and radiation efficiency. We pursue our first objective by proposing the use of a simple magnetic feed loop to simplify impedance matching and proposing a corresponding equivalent circuit model to assist in the selection of the appropriate feed loop size for a given helix. We then pursue our second objective by presenting design curves showing how simple changes in the geometry of normal mode multi-arm cylindrical helix antennas such as number of arms, height, radius, and presence or absence of crossbars can affect the radiation efficiency and bandwidth. We also describe our simulation and measurement techniques.

In Chapter 3, we pursue our third objective and compare three methods for connecting uplink and downlink ULP antennas to the single antenna port on the Orbcomm modem: a conventional carrier operated relay (COR), a conventional diplexer filter, and a novel complementary feed.

In Chapter 4, we draw conclusions and offer directions for future work.
Chapter 2: Analysis and Design of Ultra Low Profile Cylindrical Helix Antennas

2.1 Introduction

A variety of antenna designs may meet Orbcomm’s requirements for a one-inch-tall ULP terminal antenna that operates at 138 and 150 MHz. Current low profile VHF antenna designs are mostly Inverted F and Planar Inverted F Antennas (PIFA) that are almost triple the height limitation, Table 1. Here, we take a different design approach from planar antennas as research suggests that volumetric antennas that effectively utilize the Chu sphere have improved performance. While this method allows reduced height, antennas with our extreme aspect ratios have not been previously reported in the literature. Employing Chu-Harrington, Gustafsson, and Kraus approximations, we predict that a ULP antenna should be able to achieve a bandwidth of 2 MHz with a radius of at least 8 cm and resonant frequency of 150 MHz. However, they do not explain how to realize this result and their methods do not account for such close proximity to the ground plane.

After selecting the size, we maximize power radiated by the antenna through: 1) proper impedance matching and 2) improving radiation efficiency. In particular, we introduce an internal magnetic coupling loop with adjustable radius for best impedance matching and develop an equivalent circuit model that allows key design parameters to be estimated much more quickly than would be possible by simulation or measurement. Further, we optimize bandwidth and radiation efficiency at ULP height by varying the geometry of the cylindrical helix such as height, radius, number of arms, number of turns, and presence or absence of crossbars in order to identify the optimum design.

The rest of this chapter is organized as follows. In Section 2.2, we discuss the limitations of planar antennas. In Section 2.3, we consider the fundamental challenges of electrically small antennas. In Section 2.4, we give the Orbcomm antenna requirements. In Section 2.5, we examine the
design of a multi-arm normal mode cylindrical helix antenna fed by a coupling loop. In Section 2.6, we discuss proper impedance matching techniques for the feed loop. We discuss the steps taken to ensure accurate measurement and simulation results in Section 2.7. In Section 2.8, we optimize the bandwidth and radiation efficiency at ULP height by manipulating antenna geometry.

Table 1- Example of commercial Orbcomm antennas

<table>
<thead>
<tr>
<th>MobileMark</th>
<th>Multiband</th>
<th>InevTech</th>
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<tbody>
<tr>
<td><strong>Bands</strong></td>
<td>Cellular, WIFI, GPS</td>
<td>Cellular, WIFI, GPS</td>
</tr>
<tr>
<td><strong>Height (cm)</strong></td>
<td>38</td>
<td>7.62</td>
</tr>
<tr>
<td><strong>Polarization</strong></td>
<td>Circular</td>
<td>Linear</td>
</tr>
<tr>
<td><strong>VSWR (:1)</strong></td>
<td>2</td>
<td>2.5</td>
</tr>
<tr>
<td><strong>Gain (dBi)</strong></td>
<td>2</td>
<td>4.5</td>
</tr>
<tr>
<td><strong>Performance compared to reference λ/2 whip</strong></td>
<td>85%</td>
<td>60%</td>
</tr>
</tbody>
</table>

2.2 Analysis of Planar Antennas

Planar Inverted F Antenna (PIFA) designs are currently popular for vehicular applications because they are low profile and the truck or semi trailer provides a natural large ground plane. A PIFA is shown in Figure 3 with dimensions length $L_1$, width $L_2$, shorting post width $W$, feed point distance $D$ from the shorting post, and height $H$ from the ground plane. It sits on top of a dielectric with permittivity $\varepsilon_r$. The feed point distance $D$ tunes the input impedance and the further away from the short, the higher the impedance. PIFA is a $\lambda/4$ antenna meaning the length between the edge and shorting area $W$ is $\lambda/4$ and required for radiation. PIFA size in relation to resonant frequency can be
roughly approximated by (1) [21]. Although height is not included in the equation as it is negligibly small and affects bandwidth [22].

\[ L1 + L2 = \frac{\lambda}{4} = \frac{c}{4f\sqrt{\varepsilon_r}} \]  \hspace{1cm} (1)

At 150 MHz using FR4 material (\(\varepsilon_r = 4\)), a variety of PIFA designs are possible by applying (1), some even meeting Orbcomm requirements. A special case occurs when the shorting strip becomes a shorting pin (\(W = 0\)) and the sum of \(L1\) and \(L2\) equals 25 cm. The largest bandwidth is achieved for the widest shorting strip (\(L2 = W\)) and \(L1\) is 25 cm.

![Figure 3- Typical PIFA with shorting pin](image)

Size reduction techniques are well documented in the literature but come at a cost of performance degradation [9] because PIFAs, like all electrically small antennas, suffer from narrow bandwidth due to their small size. For instance, \(L1\) can be decreased by introducing capacitive loading that will reduce the reactive part of the impedance by adding a shunt capacitor to the open end to balance the capacitance of the feed [23]. Capacitive loading reduces radiation efficiency and bandwidth. Other size reduction techniques include: folding configurations, surface etching, shorting walls or pins, and loading with high dielectric material. A variety of folding configurations have reduced the length to \(\lambda/8\) [10] or even \(\lambda/8.6\) and \(\lambda/11.7\) [11] with increased height and decreased bandwidth (2.98\% and 1.15\%, respectively).

To improve planar antenna performance, researchers have long understood the need for volumetric antennas that more efficiently utilize the Chu sphere [20]. Volumetric versions of planar
designs, such as 3D fractal antennas, have been considered [24][25][26][27]. Best showed that a four-arm hemispherical spiral antenna with an antenna height of 0.06\(\lambda\), has an input resistance of 43 \(\Omega\), and is 1.5 times the Chu limit [9]. Lim et al.’s multiple folded antenna is 0.045\(\lambda\) tall [10]. Mehdipour et al. fabricated a spherical wire antenna 0.08\(\lambda\) tall with input resistance nearly 50 \(\Omega\) and is 2.09 times the Chu limit [11]. However, antennas with 0.01\(\lambda\) heights have not been reported in literature. We propose the use of volumetric antennas for ULP design.

### 2.3 Fundamental Challenges of Electrically Small Antennas

Electrically small antennas are defined by their size and self-resonant frequency. The formal definition of electrically small antennas is a boundary limit with \(ka = 1\) [28][29], where \(k\) is \(2\pi/\lambda\) and \(a\) is the radius of a sphere encompassing the maximum dimension of the antenna. This limit was chosen to correspond with the dimensions of a sphere with radius of \(\lambda/2\pi\), or a radiansphere, which is the near field boundary for an electrically small antenna [30]. The definition of \(ka < 0.5\) [31] is also used.

All electrically small antenna designs are derivatives of the simple straight-wire dipole, monopole, or loop antennas, [31], including PIFA antennas. For instance, at 150 MHz, a straight wire half dipole (\(\lambda/2\)) antenna is 2 m tall, which is impractical for mobile applications. Researchers have reduced the length of wire by introducing a ground plane and winding the wire into a meander line or coil structure [32]. However, design trade-offs must be considered as antennas become smaller.

Electrically small antennas are limited by three fundamental design factors. The most important is that an electrically small antenna of given dimensions has an upper limit on bandwidth that cannot be exceeded. After setting the size, we maximize antenna radiation from the available source power by properly impedance matching between the antenna and source and improving the antenna’s radiation efficiency. If a small antenna has the same radiation efficiency and VSWR
relative to operating bandwidth, it will radiate the same total power as a larger sized antenna [31]. We employ this approach in our ULP design.

### 2.3.1 Fundamental Relationship Between Q and Bandwidth

Q is the quality factor of an antenna and is the ratio of energy stored to the energy dissipated as a function of frequency, (2) [7]. Q provides a quantitative measure of antenna performance relative to fundamental limits and is inversely proportional to the antenna’s -3 dB bandwidth, (3). The significance is that no electrically small antenna will exhibit a Q less than its lower bound and sets the maximum achievable bandwidth [8][30][28][33]. The actual Q may be orders of magnitude higher than that calculated making bandwidth estimation difficult. Researchers have developed increasingly sophisticated methods for estimating Q. Here,

\[
Q = \frac{2\omega_0 \max \left(W_E, W_M\right)}{P_A} \tag{2}
\]

\[
Q \approx \frac{1}{B} \text{ for } Q \gg 1 \tag{3}
\]

where \(W_E\) and \(W_M\) are the time averaged stored electric and magnetic energies, \(P_A\) is the received power, \(\omega_0\) is the resonant frequency, and \(B\) is the 3 dB transmission bandwidth.

Many applications use alternative definitions of bandwidth. We express bandwidth as the range over which the return loss is -10 dB or better. The cell phone industry sometimes uses a return loss of -6 dB or better instead. Researchers including Wheeler, Chu and Harrington, Gustafsson, Yaghjian and Best, among others, have developed methods of estimating minimum Q. Applying minimum Q, we can calculate bandwidth for a specific return loss by considering fractional matched VSWR bandwidth method, (4) and (5). This method considers a tuned antenna. Impedance is measured at the antenna’s feed point. The calculated bandwidth is independent of external matching components. This method only applies to antennas that exhibit a single resonance within its defined
operating bandwidth and approximations may not hold for multiple very closely spaced resonances [34]. Here,

\[ FBW = \frac{f_+ - f_-}{f_o} \quad (4) \]

\[ Q = \frac{2\sqrt{\beta}}{FBW} \quad (5) \]

where \( \beta = \frac{s-1}{2\sqrt{s}} \leq 1 \) and \( s \) is the VSWR.

**Wheeler Q**

Wheeler was the first to note the fundamental relationship between small antenna size and bandwidth [28]. He defined radiation power factor (RPF) as the ratio of the radiated power to reactive power using simple capacitive and inductive antenna models. He determined that RPF was directly proportional to the physical antenna volume. Defining \( Q \) as the inverse of RPF in his circuit model implies that RPF represents bandwidth. He was the first to observe for antennas that best utilize their minimum enclosing sphere of influence will have small \( Q \) and higher bandwidths compared to other geometries within the same volume. Wheeler’s work inspired the advancement of small antenna theory.

**Chu-Harrington Q**

Chu-Harrington considers \( Q \) based on spherical waves to express the stored and radiated energies outside the smallest circumscribing sphere of an antenna [7][8]. The Chu-Harrington limit imposes a fundamental lower bound on \( Q \) as a function of its size \( (a) \) and operating frequency \( (k) \), where

\[ Q_{Chu} = n_r \left( \frac{1}{(ka)^3} + \frac{1}{ka} \right) \quad (6) \]
and $n_r$ is the antenna’s radiation efficiency. However, Thiele [35], Thal [36], and Folz and McLean [37] noted that Chu’s approach restricts antenna radiation to a sphere and may not be suitable for practical small antennas [38].

**Gustafsson Q**

It is widely accepted that the Gustafsson bounds are comparable with classical bounds for spherical geometries and provide stricter bounds for non-spherical geometries. Gustafsson bounds hold for arbitrary antenna volumes making this method suitable for our unconventional aspect ratios.

Gustafsson et al. calculate the minimum $Q$, gain, and bandwidth of small antennas based on the scattering properties of small particles (their electro and magneto-static polarizability dyadic) [20][39] [40]. Equation 7 relates the gain and bandwidth while (8) relates directivity and $Q$.

$$\min \{ (1 - |\Gamma|^2 G) \cdot B \leq \frac{\pi^2}{\lambda_0^3} (\hat{\gamma}_e^\ast \cdot \hat{\gamma}_e \cdot \hat{\gamma}_e + \hat{\gamma}_m^\ast \cdot \hat{\gamma}_m \cdot \hat{\gamma}_m) \} \quad (7)$$

where $\gamma_e$ and $\gamma_m$ are the electric and magnetic polarizability dyadic, $\eta$ is the absorption efficiency, $\lambda_o$ is the average wavelength, $G$ is antenna gain, $\Gamma$ is reflection coefficient at the feed point, $B$ is fractional bandwidth. Here,

$$\frac{D}{Q} \leq \eta \frac{k_0^3}{\lambda_0^3} (\gamma_1 + \gamma_2) \quad (8)$$

where $\gamma_1$ and $\gamma_2$ are the largest two eigenvalues of $\gamma_\infty$, $D$ is directivity, and $k$ is the wavenumber.

**Yaghjian and Best Q**

Yaghjian and Best estimate $Q$ using the antenna’s feed point impedance [41] in (9). We employ this $Q$ estimation for HFSS simulations.

$$Qz(\omega_0) \approx \frac{w_0}{2R_0} \sqrt{ \left( \frac{dR(\omega_0)}{d\omega} \right)^2 + \left( \frac{dX(\omega)}{d\omega} + \frac{|X(\omega_0)|}{\omega_0} \right)^2 } \quad (9)$$
where $R(\omega) = R_0(\omega)$ and $X(\omega)$ are the resistance and reactance of the untuned antenna, respectively.

### 2.3.2 Q and Bandwidth Limitations of ULP Antennas

In Figure 4, we compare Q and bandwidth using the Chu-Harrington and Gustafsson limits for ULP cylinder and hemispherical geometries at various radii. Results indicate that to achieve a bandwidth of at least 2 MHz, which barely covers Orbcomm’s passband, requires an antenna radius of at least 8 cm. Typically, a bandwidth larger than the channel bandwidth is desirable as small antennas usually suffer from detuning in the operating environment [31]. Furthermore, the estimated Q does not take into account either the practical wire arrangement or the effect of wire interaction, which may be substantial given the small size. We investigate ULP antennas further in Section 2.8.2.

Q was estimated by considering a frequency of 150 MHz, a fractional matched VSWR of 1.925:1 (or a return loss of -10 dB or better), and 100% efficiency. The cylindrical Gustafsson Q was generated using Matlab code for linearly polarized antennas composed of non-magnetic materials and circumscribed by various geometries that was supplied by Gustafsson [42]. The radius is restricted to 0.16 m to maintain a normal mode helix design. For larger radii, or circumference $> 2/3\lambda$, the antenna will become an axial mode helix.

Since the Chu-Harrington limit is based on spherical waves, the Q of a sphere is the same as a cylindrical volume circumscribed within it. The Gustafsson limit provides a stricter bound for cylindrical shapes. As expected, the cylindrical shape has a higher Q and lower bandwidth than that of a sphere because it occupies less volume. For a hemispherical antenna, both Chu and Gustafsson predict similar Q’s, comparing (6) and (10).

\[
Q_{\text{Gust sphere}} = \frac{1}{n_r (ka)^3} \quad (10)
\]
2.3.3 Impedance Match

Optimal impedance matching is achieved when the total incident power from the source is transferred to the antenna. The total power delivered to the antenna is a function of the reflection coefficient and characterized by the antenna’s voltage standing wave ratio, 

\[ VSWR(\omega) = \frac{1 + |\Gamma(\omega)|}{1 - |\Gamma(\omega)|} \tag{11} \]

\[ \Gamma(\omega) = \frac{Z_A(\omega) - Z_{ch}}{Z_A(\omega) + Z_{ch}} \tag{12} \]

\[ P_A = (1 - |\Gamma|^2 P_{INC}) \tag{13} \]

where \( \Gamma \) is the reflection coefficient, \( Z_A \) and \( Z_{ch} \) are the antenna and characteristic impedances, \( P_A \) and \( P_{INC} \) are the accepted and incident powers.

Impedance matching an electrically small antenna with an external matching network comprised of lossy reactive components is non-ideal as ohmic loss within the matching network contributes to low overall efficiency. For small antennas, it is more desirable to impedance match by
incorporating a low loss feed into the antenna’s structure. Section 2.6 examines an internal loop feed for impedance matching.

### 2.3.4 Radiation Efficiency

Radiation efficiency is defined as the total power radiated divided by the power accepted by the antenna, (14). It is well known that decreasing antenna height reduces radiation efficiency due to the cancelation of opposing currents on the ground plane [43]. We can improve radiation efficiency by simply changing the number of arms and turns of the helix structure [9]. Here,

\[
\eta = \frac{P_{\text{rad}}}{P_A} = \frac{R_{\text{rad}}}{R_{\text{rad}} + R_{\text{ohmic}}} \tag{14}
\]

\[
R_{\text{rad}} = \frac{1}{2} l^2 P_{\text{rad}} \tag{15}
\]

\[
R_{\text{rad} N\text{Arm}} = N^2 R_{\text{rad 1 Arm}} \tag{16}
\]

where \(P_{\text{rad}}\) and \(P_A\) is the power radiated and power accepted, \(R_{\text{rad}}\) and \(R_{\text{ohmic}}\) are the radiated and ohmic resistances, and \(N\) is the number of arms.

Increasing the number of turns increases the amount of current flowing into the antenna and increases the radiation resistance. Because ohmic resistance increases linearly, efficiency improves. While increasing the number of arms increases the radiation resistance and ohmic resistance at the same time, high efficiency is achievable because radiation resistance is increased by a power relationship while ohmic resistance increases linearly as depicted in (16). Equation 16 is the radiation resistance of a loop in free space.

There are well known methods for achieving high radiation efficiency while achieving a better impedance match, such as folded arms and shunt or parallel matching stubs. Section 2.8.2 examines the effect of folded arms, as well as number of arms and turns on ULP antennas.
2.4 Orbcomm Antenna Requirements

Orbcomm antenna requirements are listed in Table 2. For ease of manufacturability and improved ruggedness, we propose to print the cylindrical helix antenna on a PCB board to stabilize the arms of the helix and allow it to be easily assembled with machine screws or rivets.

Table 2- Orbcomm antenna requirements

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Uplink: 148 – 150.05 MHz (Bandwidth = 2.05 MHz) Downlink: 137.0 – 138.0 MHZ (Bandwidth = 1 MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna Pattern</td>
<td>5° - 45° elevation, 85% of time Omnidirectional</td>
</tr>
<tr>
<td>Polarization</td>
<td>Right hand circularly polarized (RHCP)</td>
</tr>
<tr>
<td>Dimensions of Antenna</td>
<td>Height: 2.54 cm Width and length: flexible</td>
</tr>
<tr>
<td>Gain</td>
<td>-14 dBi (min)</td>
</tr>
<tr>
<td>VSWR (TX)</td>
<td>2.0:1</td>
</tr>
<tr>
<td>VSWR (RX)</td>
<td>1.5:1</td>
</tr>
<tr>
<td>Impedance</td>
<td>50 Ω</td>
</tr>
</tbody>
</table>

2.5 Normal Mode Cylindrical Helix Antenna (NMCHA)

It is clear from Section 2.3.2 and [44] that spherical helical antennas are most effective at filling the volume of a Chu sphere and exhibit the widest possible bandwidth. However, these designs and their derivatives are not suitable for our applications due to: 1) the height limitation, 2) the complex shape makes them difficult and expensive to be manufactured, and 3) the shape may change as the truck drives down the highway affecting resonant frequency. Reducing the height by half produces a hemispherical antenna that sits on a ground plane and similar efficiency is achievable compared to their spherical counterparts [9][45][46][47][48] but they are still too tall for our applications. For hemispherical antennas, Jobs et al. showed in [49] that a 0.001λ decrease in height shifts the resonant frequency by 78% of the bandwidth.
To capture the maximum volume at ULP height, we consider the design of cylindrical helix antennas. The axial mode helix is a well-studied design for antennas whose diameter is much larger than height [12][13][14][15][16]. However, they do not meet our ULP’s omnidirectional radiation pattern and small size requirements. To meet the requirements, we consider the design of a normal mode cylindrical helix antenna. At our extreme aspect ratios, the performance of these antennas is not generally known and the shortest reported antennas are four to five times our height limitation.

Another benefit of the cylindrical helix antenna design is simplicity in manufacturing. The helix design can be produced on a PCB board and the cylindrical shape is achieved by securing it end to end. The resonant frequency will remain constant by simply affixing the helix to the ground plane with cost effective screws or rivets.

Section 2.5.1 introduces the geometry of cylindrical helix antennas and Section 2.5.2 examines how simple changes in the antenna geometry affects bandwidth, gain, and radiation efficiency. Proper feeding technique is considered in Section 2.6.

2.5.1 Characteristics of a Helical Antenna

A normal mode cylindrical helix antenna is defined by its geometry and radiation mode. This antenna has a circumference much smaller than a wavelength (circumference $< 2/3\lambda$). As a result, its current distribution is nearly sinusoidal similar to a long straight wire antenna. We employ Kraus’ method to examine the optimal size of normal mode cylindrical helix antennas for circular polarization in the general case. Since Kraus’ method does not account for proximity to the ground plane, further analysis is required for ULP antennas, see Section 2.8.
$L = \sqrt{H^2 + (2\pi rN)^2}$ \hspace{1cm} (17)

The geometric parameters of a normal mode cylindrical helix antenna are shown in Figure 5. The helix arm length is related to height and radius according to (17). The helix has two special cases. When the pitch angle is zero, the antenna becomes a loop; when the pitch angle is 90°, the antenna becomes a linear antenna.

A small helix can be modeled as a number of small loops and short dipoles connected in series, Figure 6. We assume that current is uniform in magnitude and in phase over the entire length of the helix. For the normal mode case, the far-field pattern is independent of the number of turns and can be represented by a single small loop and short dipole.

The far field of a small loop only has an $E_{\phi}$ component while the far field of a short dipole only has a $E_{\theta}$ component, (18) and (19). From the axial ratio (AR), three special cases of polarization...
exist, (20): 1) When $E_\phi = 0$ (resembling a dipole) the axial ratio is infinite and we have linear vertical polarization; 2) When $E_\theta = 0$ (resembling a loop) the axial ratio is 0 and we have linear horizontal polarization; and 3) When $|E_\theta| = |E_\phi|$, axial ratio is 1 and circular polarization occurs, (21) [50]. The Kraus and Wheeler methods of approximating circular polarization for helical antenna yield similar results [12]. Here,

$$E_\phi = \frac{120\pi^2 L \sin \theta}{r} \frac{A}{\lambda^2}$$  \hspace{1cm} (18)$$
$$E_\theta = j \frac{60\pi L \sin(\theta)}{r} \frac{P}{\lambda}$$ \hspace{1cm} (19)$$
$$AR = \frac{|E_\theta|}{|E_\phi|} = \frac{P \lambda}{2 \pi A} = \frac{2P \lambda}{\pi^2 D^2}$$ \hspace{1cm} (20)$$
$$\pi D = C = \sqrt{2P \lambda}$$ \hspace{1cm} (21)$$

where the area of the loop is $A = \frac{\pi D^2}{2}$. 

Maintaining a helix arm length of $\lambda/4$ and combining (17) and (21), circular polarization is achievable for a variety of helix geometries, Figure 7. A helix structure with one turn and 8 cm radius is consistent with Chu-Harrington and Gustafsson limits. As the antenna gets shorter, the pitch $P$ decreases, $E_\theta$ approaches 0, and polarization becomes linear and horizontal. Kraus’ method does not account for proximity to the ground plane and we further examine the characteristics of ULP antennas in Section 2.8. By approximating the helix as a small loop and short dipole following Kraus’ method [51], the radius of the ULP is calculated as 8 cm which is consistent with Chu-Harrington and Gustafsson limits.
2.5.2 Design Freedoms

The performance of a cylindrical helical antenna can be improved simply by modifying its geometry. Design parameters such as height, radius, number of arms, number of turns, the presence or absence of crossbars, can be considered to optimize bandwidth, radiation resistance, and efficiency.

For a self-resonant frequency of 150 MHz, the length of the arm was maintained at $\lambda/4$. For a single arm helix, the arm must wrap fully around the cylinder and crossbars are not required, Figure 8. The multi-arm case supports a fractional number of turns, with and without crossbars.

From Section 2.3.4, it is expected that radiation resistance and efficiency will improve as the number of turns and arms increase. From Section 2.5.1, the presence of crossbars contribute a negligible amount of radiation power as the antenna gets shorter and as $E_\theta$ approaches 0.

The helical arms contribute to inductive loading. Inductance can be tuned by changing the number of turns and size. As height decreases, capacitance dominates over inductance. Also, the arms get closer together creating mutual inductance between the arms and interaction with the ground plane. Previous equations for cylindrical helix antennas may no longer hold for height limited antennas.
Crossbars connect the arms at the top and contribute to capacitive loading at the ends of the helix. From the perspective of an equivalent circuit, crossbars connect the arms in parallel and increasing the number of arms decreases the overall inductance. However, for height-constrained antennas, the length of the crossbar may become a substantial part of the arm length and contribute to transmission line effects, and (21) may not hold. Maximum capacitance is obtained with infinite crossbars that resemble a disk.

Wheeler states that there is an optimum number of turns and a corresponding optimum shape for the self-resonance case to obtain maximum radiation [12]. Through simulations, measurements, and mathematical models, we endeavour to determine the optimum shape and feed mechanism.

![Geometries for normal mode cylindrical helix: 1) single arm, N turns, 2) multi-arm, fractional turns, 3) multi-arm, N turns, crossbar, and 4) multi-arm, fractional turns, crossbar](image)

### 2.6 The Feed Loop

Feeding an electrically small antenna is challenging as improper technique will further reduce the antenna’s bandwidth, efficiency, or both. Feeding the antenna through an external matching network is not ideal because its loss resistance may exceed the radiation resistance of the antenna and reduce overall efficiency. It may also increase the size of the antenna. Improper network matching will further reduce the bandwidth, which, in the case of our Orbcomm antenna, barely covers a passband. There is also a physical limitation that prevents direct transmission line feeding to a multi-arm helix. Our simulations suggest that feeding just one of the multi arm helix creates an uneven current distribution between the arms and produces a skewed radiation pattern.
We present an internal magnetic feed loop that couples to the helix arms like a transformer for an even current distribution, creating an omnidirectional radiation pattern. This method of magnetic coupling affords ease of impedance matching and limitations can be mitigated through proper feed loop design. We show that the loop-fed helix can be represented using a simple equivalent circuit model that is in excellent agreement with both measured and simulated Ansys HFSS results. Applying this model, the optimal feed loop size for a given helix is easily predictable, and much faster than complicated modeling software or measurements. The feed loop has some potential limitations such as energy storage and coupling variations but these can be mitigated through proper feed loop design.

A variety of methods have been proposed to feed helical antennas. Our loop feed does not require the transmission length like Dinh et al.’s tap feed [17]. At our aspect ratio with larger diameter than radius, the length of a crossed transmission line feed described by Sakaguchi et al. [18] would become substantial. Our solution does not require an external lossy capacitor like Yu et al.’s [19] inductive feed.

2.6.1 Equivalent Circuit Model

We derive the equivalent circuit model for a loop-fed one-arm helix antenna and discuss the relationship between voltage, current, and impedance. In the next section, we show this simple model is in excellent agreement between measured and simulated results. Its main benefit is that calculations are orders of magnitude faster than complicated modeling software or measurements.
Figure 9- Thevenin equivalent model of inductively coupled feed loop to antenna

Applying (22) to (28), we derive the inductance, capacitance, and resistances.

\[ L_p' = N^2 r \mu_0 \mu_r \left[ \ln \left( \frac{8r}{a} \right) - 2 \right] \]

- Inductance of loop feed in free space \([52]\). Inductance is directly proportional to number of turns \((N)\) squared and radius \((r)\) and thickness of feed loop \((a)\) has little effect.

\[ L_s' = \frac{\mu_0 \mu_r N^2 A}{h} \]

- Inductance of the helix is proportional to the area \((A = \pi r^2)\), the number of turns squared, and indirectly proportional to height \((h)\).

\[ L_i = \frac{\mu_0 r}{4} \]

- The DC self-inductance of the wire. The total inductance of the loop feed is the sum of inductance in free space and the inductance of wire. This also applies to total inductance of helix.

\[ C_{1,2} = \frac{1}{L_{1,2}'(2\pi f)^2} \]

- Capacitance of feed and antenna are calculated knowing its inductance at resonant frequency \((f)\). Feed capacitance comes from close proximity to ground plane.

\[ R_\Omega = \frac{2\pi r}{\sigma s} \]

- Both feed loop and antenna have ohmic resistance, which is dependent on the radius, conductivity \((\sigma)\) and skin depth of wire \((\delta)\).

\[ s = \pi (a^2 - (a - \delta)^2) \]

\[ \delta = \frac{2\rho}{\omega u} \]

\[ R_5 = R_\Omega + R_{rad} \]

- Total resistance of the antenna is the sum of ohmic resistance and radiation resistance. The feed loop only has ohmic resistance.
Measurement and simulations confirm that the radiation resistance of the helix should be modeled as a loop in free space. Radiation resistance is directly proportional to circumference to the power of four, \( C \), and number of turns squared.

The voltage and current relationships of the loop-fed helix can be examined by applying Kirchhoff’s Voltage Law (KVL) to both the primary and secondary sides of the circuit, (29) to (31). The current in the antenna, \( I_5 \), peaks at resonance. A negative sign indicates that the current is flowing counterclockwise. The current is directly proportional to frequency and the mutual impedance \( M \).

Here,

\[-V_{in} + I_p Z_p + I_p j \omega L'_p + I_5 j \omega M = 0 \tag{29}\]

\[I_5 Z_S + I_5 j \omega L'_S + I_p j \omega M = 0 \tag{30}\]

\[I_5 = \frac{-j \omega MV_{in}}{\omega^2 (M^2 - L'_p L'_S) + j \omega (Z_S L'_S + Z_p L'_p) + Z_S Z_p} \tag{31}\]

where \( Z_p = R_\Omega Z_p = R_p / j \omega c_p \), and \( Z_S = R_S + \frac{1}{j \omega c_S} \).

We define the impedance of the loop-fed helix as the impedance seen by the modem, also known as the feedpoint impedance. Equation 32 defines the impedance of the antenna, \( Z_o \), as the sum of the feed loop and helix. Magnetic coupling increases the overall antenna impedance by a factor of \((\omega M)^2\). Adjusting the size of feed loop inside a fixed helix changes both antenna resistance and reactance where: 1) the reactance of feed loop \( X_p \) changes linearly, and 2) mutual impedance \( M \) changes by a power of two. Here,

\[Z_o = \left( R_p + \frac{\omega^2 M^2 R_S}{|Z_S|^2}\right) + j \left( X_p - \frac{\omega^2 M^2 X_S}{|Z_S|^2}\right) \tag{32}\]

where \(|Z|^2 = R_S^2 + X_S^2\), \( X_p = \omega L'_p / \frac{1}{j \omega c_p} \), and \( X_S = \omega L'_S - \frac{1}{\omega c_S} \).
We determine the relationship between mutual impedance $M$ and the effect of feed loop size to helix in Section 2.6.2. In Section 2.6.3, we employ the equivalent circuit model to calculate the appropriate feed loop size to helix with comparable accuracy but orders of magnitude more quickly than conventional electromagnetic simulation software. Finally, in Section 2.6.4 we employ the equivalent circuit model to assess the limitations of the feed loop in terms of bandwidth, frequency, and efficiency.

### 2.6.2 Estimating Coupling Coefficient and Mutual Inductance

Estimating the mutual impedance $M$ is difficult because 1) our coupling does not follow the 1:n transformer relationship, and 2) coupling is difficult to predict due to close proximity to ground plane. Employing Ansys Maxwell software, we derive the relationship of $M$ for various feed loop sizes for a range of helix sizes.

The relationship between mutual impedance $M$ and the coupling coefficient $k$ is defined in (33). In our case, mutual coupling is inductive and the relationship between indirect inductive coupling and direct inductive coupling follows (34). The coupling coefficient $k$ varies between 0 (no coupling) and 1 (tight coupling) [53]. Our simulations show that the coupling is loose and we cannot use the ideal transformer relationships between the input and output voltages and currents that are a function of the turns ratio [54]. Here,

$$M = k \sqrt{L_p' L_s'} \quad (33)$$

$$L_p = L_p' - M \ ; \ L_s = L_s' - M \ ; \ L_M = M \quad (34)$$

where $M$ is the mutual inductance, $k$ is the coefficient of coupling, and $L_p$ and $L_s$ are the self-inductances of the coils.

Due to the close proximity to the ground plane, we use Ansys Maxwell software to predict mutual coupling $k$ for various feed loop to helix sizes. For each case, the feed loop is concentric to the
helix. Trendlines in Figure 10 show a clear linear relationship for $\ln(k)$ for various feed loop radii per helix size, which allows us to estimate $M$.

![Graph showing trendlines for ln(k) vs. radius for different feed loop sizes](image)

**Figure 10- Mutual coupling between various feed loop to helix radii has clear trend**

### 2.6.3 Impedance Matching with Feed Loop

From (32), as the size of the feed loop changes with respect to helix, the impedance of the antenna changes: linear change in feed loop reactance $X_P$, and helix impedance $X_S$ changes by a factor of $M^2$. The Smith charts in Figure 11 show impedance matching for various feed loop sizes to 8 cm radius helix for one-arm (top) and 4 arms (bottom). There is excellent agreement between the simulated results from Ansys HFSS software, the measured results, and the calculated results from the equivalent circuit model. We apply this model to approximately predict the optimum feed loop radii for all helix radii that is orders of magnitude faster than Ansys HFSS and measurements. Applying the equivalent circuit model, we further show the effects of bandwidth on impedance matching.
Figure 11- Impedance matching of various feed loop sizes to 8 cm radius ULP for one-arm (top) and 4 arms (bottom)
As the radius of the feed loop changes with respect to helix, the impedance of the antenna changes according to Figure 11 for one-arm (top) and 4-arm (bottom) helices. The general impedance properties are consistent with previous work [55] that describes a loop coupled resonant antenna as represented by a series RLC circuit. At DC, the loop coupled resonant antenna can be represented as an open circuit, which is slightly rotated about the Smith chart due to finite wire lengths. As feed loop radius increases inside of a fixed helix size, the constant resistance circle increases by an exponential relationship defined by $M$. The constant resistance circle differs between one-arm and 4 arms due to the proximity of the arms. Simulated and calculated results were rotated slightly about the Smith chart to correspond with measured results, with real cable lengths. The optimum feed loop size is 6.5 cm for one-arm 8 cm helix and 5.5 cm for 4-arm 8 cm helix.

For any sized helix, there is an optimum feed loop size. The close agreement between measured, simulated, and model impedances gives us confidence in approximating the optimum feed loop size using the equivalent circuit model. Figure 12 shows the return loss for a range of loop sizes for a one-arm 8 cm radius helix (top) and confirms the optimum feed loop is 6.5 cm.

The equivalent circuit model can approximately predict optimum feed loop radius for all helix radii, Figure 12 (bottom). For instance, the optimum feed loop size of a one-arm 7 cm ULP helix is 5.5 cm, and for 9 cm is 7.5 cm. It even predicted an optimum loop size of 4.75 cm for 6 cm helix radius, which was initially missed in the simulation sweep with large step size. Return loss values were scaled to correspond with simulated results. Predicting the best loop radius requires 14 to 18 Ansys HFSS simulations and each simulation takes about 20 minutes on a Xeon-based 8-core HP z800 workstation. Predicting the optimum feed loop radius takes a few seconds with our model. We demonstrate that optimum feed loop radius prediction is possible and is faster than Ansys HFSS simulation and measurements.
Figure 12- Equivalent circuit model accurately predicts optimal feed loop for a one-arm, 8 cm radius helix is 6.5 cm radius (top), and predicts optimum feed loop size to various helix sizes for one-arm ULP (bottom).
Proper impedance matching with the feed loop to helix ensures maximum power transferred to the antenna. Mismatches in feed loop radius degrade bandwidth for constant one-arm helix radius of 8 cm, using the equivalent circuit model in Figure 12. To achieve the widest bandwidth, it is not necessary to optimize the feed loop at 6.5 cm but rather around 7 cm. Reducing antenna efficiency can give the appearance of increasing bandwidth beyond the Chu-Harrington and Gustafsson limits but this is generally undesirable, as will be discussed further in Section 2.6.4.

![Graph](image)

**Figure 13- Bandwidth for various feed loop radii to one-arm 8 cm radius helix**

### 2.6.4 Limitations of the Feed Loop

We discuss two limitations of the feed loop and suggest ways to reduce their effects. There is a concern that the coupling loop has the potential to degrade bandwidth and efficiency of the antenna. As the ULP antenna is already bandwidth limited, we want to prevent further bandwidth degradation. Another limitation is that the coupling coefficient varies with placement of the feed loop to the helix and changes the resonant frequency. Shifting the resonant frequency of a narrowband antenna can cause it to detune, which is not ideal. Through proper feed loop design, we demonstrate that it is possible to mitigate these limitations through simulations and measurements.
Mitigating Energy Storage in Feed Loop

We can minimize energy storage in the feed loop by designing it as a thick strip rather than a wire. The current in a wire is considered with respect to its volume and current in a strip is expressed with respect to its surface area. Excessive energy storage of a feed loop is reduced considering a wide strip compared to a wire design of the same thickness. Magnetic energy stored in the feed loop is a function of its inductance, $L$, and current, $I$, where $W_m = \frac{1}{2} LI^2$ [56]. Inductance is represented by $L = \mu \frac{N^2}{l}$, where $S$ is the area between the conductors. Current is a function of the wire structure. The total current in a wide strip with width, $w$, is $I_w = \int_w J_s dw = J_s w$ and the total current in a wire is $I_s = \int_s Jds = J\pi r^2$, where $J_s$ and $J$ are surface and volume current density respectively, $w$ is width, and $r$ is radius. For same current density and same thickness of wire where $J_s = J$ and $r = \frac{w}{2}$, total current is $I_s = J\pi \frac{w^2}{4}$. For wide $w$, total current increases linearly with width of strip and width-squared for wires. Clearly, there is more total current in wires than in strips and more energy is stored in wire verses strip feed loops. In our design, the feed loop is made of a 1 mm wide copper strip with almost negligible thickness.

![Diagram of current in wire and strip]

**Figure 14- Comparison of total current in wire to strip**

Another method of minimizing energy stored in the feed loop is to ensure its length does not support a standing wave. We design the helix arm a quarter of a wavelength long to support a standing wave at 150 MHz. The length of the feed loop is much shorter to ensure that it does not resonate at the frequency of interest.
Applying the above principles, we examine the effects of additional energy storage in the feed loop compared to the no feed loop case in terms of bandwidth, resonant frequency shift, and efficiency. The condition without feed loop considers the equivalent circuit model of the helix in its natural resonance plus an ideal matching network with largest possible bandwidth, natural self-resonant frequency, and largest efficiency. The condition with feed loop considers the equivalent circuit model of a loop fed helix in Figure 9 and represents practical parameters. The difference between the two conditions represents the effects of energy storage with feed loop. From Table 4 and Figure 15, the addition of a feed loop contributes very little to energy storage and there are minimal differences between bandwidth and frequency shift. Efficiency is consistently about 40% due to the cancellation of opposing fields as a result of close proximity to the ground plane. In Section 2.8.2, we increase efficiency by increasing the number of arms.

Table 3- Bandwidth, frequency, and efficiency with and without feed loop conditions

<table>
<thead>
<tr>
<th>Helix Radius (cm)</th>
<th>Loop Radius (cm)</th>
<th>Unloaded -10 dB BW (MHz)</th>
<th>Loaded -10 dB BW (MHz)</th>
<th>Δ -10 dB BW (%)</th>
<th>Unloaded Freq. (MHz)</th>
<th>Loaded Freq. (MHz)</th>
<th>Δ Freq. (%)</th>
<th>η Unloaded (%)</th>
<th>η Loaded (%)</th>
<th>Δ η (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>4.75</td>
<td>0.40</td>
<td>0.40</td>
<td>0.00</td>
<td>202.10</td>
<td>203.2</td>
<td>0.54</td>
<td>68.90</td>
<td>38.70</td>
<td>43.83</td>
</tr>
<tr>
<td>7</td>
<td>5.50</td>
<td>0.40</td>
<td>0.30</td>
<td>0.57</td>
<td>174.10</td>
<td>174.6</td>
<td>0.29</td>
<td>70.47</td>
<td>40.50</td>
<td>42.53</td>
</tr>
<tr>
<td>8</td>
<td>6.50</td>
<td>0.45</td>
<td>0.30</td>
<td>0.33</td>
<td>151.90</td>
<td>153.3</td>
<td>0.92</td>
<td>71.87</td>
<td>42.02</td>
<td>41.53</td>
</tr>
<tr>
<td>9</td>
<td>7.50</td>
<td>0.60</td>
<td>0.40</td>
<td>0.33</td>
<td>136.10</td>
<td>136.6</td>
<td>0.37</td>
<td>72.96</td>
<td>43.47</td>
<td>40.42</td>
</tr>
</tbody>
</table>
Preventing Resonant Frequency Shift in Feed Loop

We can prevent the resonant frequency from shifting by securing the feed loop concentric to the helix. Figure 16 shows that moving the feed loop by ±5 mm and ±10 mm off centre changes both the coupling coefficient $k$ and the resonant frequency. The shift in resonant frequency at 10 mm is 2 MHz; this is substantial given the narrow bandwidth. Furthermore, the amount and direction of frequency shift is unpredictable, either higher or lower. We verified resonant frequency stability through measurement.
Figure 16- Moving feed loop by 5 and 10 mm off centre (top) shifts resonant frequency (bottom)

2.7 Methodology

Simulating the antenna is not trivial and many measurements were taken to ensure proper setup of the software environment. Because the antenna is so close to a ground plane, the usual design equations no longer hold and simulations can be used to quickly assess how antenna geometry affects resonant frequency, bandwidth, and efficiency. Here, we introduce six simulation design decisions that are not obvious from typical online tutorials. Three types of antenna measurement techniques helped verify simulations: a bench test ensures self-resonance at the desired frequency, a range test verifies radiation, and a message test confirms communication with the Orbcomm satellite system. Section 2.6.3 confirms our methods by showing excellent agreement between simulation and
measurement results. The lessons learned can be applied to simulation and measurement of any antenna.

2.7.1 The Simulation Environment

Figure 17 shows the ULP antenna, a loop fed cylindrical helix structure, within its simulation environment in Ansys HFSS. We present six simulation design decisions that are unobvious from online tutorials and include relevant references. Each design decision was verified by measurement.

1. Geometric properties of the helix and loop structure representing the ULP antenna are listed in Table 4. The helix and loop structures are created by rectangular helix library rather than equation-based curve because as height increases, the helix defined by an equation-based curve resembles a cone shape.

In practice, the helix and loop structures are realized on two PCB boards to facilitate manufacturing. We chose a board thickness of 28 mil for its sturdiness and bendability. The dielectric was not modeled.

Figure 17- Simulation environment setup in Ansys HFSS with best practices (numbered)
### Table 4- ULP antenna parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>copper</td>
</tr>
<tr>
<td>Substrate Dielectric</td>
<td>4</td>
</tr>
<tr>
<td>Substrate Height</td>
<td>28 mil</td>
</tr>
<tr>
<td>Trace Thickness</td>
<td>1.4 mm</td>
</tr>
<tr>
<td>Feed Loop Trace Width</td>
<td>1 mm</td>
</tr>
<tr>
<td>Helix Trace Width</td>
<td>3 mm</td>
</tr>
</tbody>
</table>

2. The antenna is fed with a coaxial cable and wave port rather than a lumped port following measurement results. The copper wire is simulated with a horizontal length of 5 cm, or 0.025λ.

3. One end of the feed loop connects to coax cable while the other end terminates to ground, Figure 18. Otherwise, the feed loop incorrectly resembles a helix. The feed loop is modeled as a rectangular helix in HFSS with pitch 0 and 0.98 turns.

![Figure 18- Lumped port feed (left) and wave port with coaxial cable feed (right)](image)

4. Following measurements, the impedance of length and diameter of the coax cable are removed by deembedding during post processing. Deembedding the coax cable accurately measures impedance at the input of the antenna. The outer core of the coaxial cable is modeled as Teflon while the inner core material is perfect electric conductor (PEC), as suggested by Ansys [57].

5. The outer box in Figure 17 is the radiation boundary box and filled with air. It is placed λ/4 away from any radiating surfaces of the antenna, as suggested by Ansys [58]. This box represents the boundary where HFSS absorbs the wave, essentially modeling the boundary as
an infinitely open space. The smaller box is the virtual object that is filled with vacuum and placed at least $\lambda/10$ away from any radiating element. The virtual object increases accuracy and reduces simulation time [59]. In the basic case, the radiation boundary box is required.

6. Proper collection of S11 parameters is a two-step process. First a coarse frequency sweep finds the resonant frequency of the antenna followed by a fine sweep targeting the bandwidth. Data is collected at the resonant frequency.

2.7.2 Practical Antenna Measurement Techniques

Bench Test

Bench test measures the self-resonating frequency and bandwidth of the prototype antenna. The S11 parameters were measured using an Agilent N9912A FieldFox analyzer with the antenna outdoors and on a 1 m by 1 m copper ground plane. This allowed us to avoid reflections from walls, floor and ceiling of the room [60]. Smith chart and return loss graphs were produced and examined.

When calibrating the FieldFox, losses were removed by calibrating to the end of the cable. Phase stable cables were used to provide electrical stability to flexion and temperature. An RF choke at the input of the ULP eliminates surface currents on the cable. Measurement error may be attributed to the addition of an N connector used while calibrating but not measuring, which is negligible.

Outdoor Range Test

An outdoor range test is used to measure the gain of the antenna under test. The range test is synonymous with tilt test or slant range, Figure 19. Tilting ensures reflection free propagation from buildings and other objects. While the setup is similar to [61], the ULP antenna sits on a large ground plane (1 m by 1 m) that further minimizes ground reflections. A reference discone antenna (model DS150S) attached to a mast is located 10 m away from the ULP to satisfy the far field requirement of $10 \, m \gg \frac{2D^2}{\lambda}$, where $D$ is the maximum linear dimension of the antenna (16 cm). As the mast is raised,
gain is measured from 14.5° to 39° above the horizontal. An Agilent N9912A FieldFox analyzer is connected to the reference discone as transmitter and an Agilent Spectrum Analyzer N9342C connects to ULP as receiver. Knowing the gain of the discone from the datasheet as 5.12 dB, the power received by the ULP antenna from the spectrum analyzer, the transmit power of the FieldFox as 0 dBm, the gain of the antenna can be calculated using the Friis transmission equation, where

$$G_{ULP} = P_{ULP} - P_{DISCONE} - G_{DISCONE} - 20\log\left(\frac{\lambda}{4\pi r}\right)$$  \hspace{1cm} (35)$$

![Figure 19- A tilt test setup reduces reflections](image)

**Message Test with Orbecomm Satellite System**

The ultimate test is a 72-hour message test with the Orbecomm satellite system, Figure 20. This length of time is required for the satellite constellation to repeat itself. The setup follows Orbecomm’s antenna validation setup where the performance of a ULP antenna is compared to a reference λ/2 antenna. The antennas connect via Stelcomm modems and Stelcomm message testing software is run on the computer. The antennas sit on a 2 m by 2 m ground plane on a four-storey roof, away from other radiating elements.
2.8 Design Curves

We conduct two simulation experiments in Ansys HFSS to examine the manner in which performance varies with antenna radius and height. Firstly, the simulation results are validated showing good agreement between input impedance with measurements of 4-arm ULP antenna, Figure 21. From these results, we choose feed loop radius as 0.8 times helix radius showing optimum feed condition, Figure 11. Antennas with this extreme aspect ratio have not been previously reported in literature.

The results presented here are for best-matched cases. In all cases, we maintain a self-resonance of 150 MHz by keeping helix arm length at $\lambda/4$. We follow the simulation method in Section 2.7 using a lumped port. $Q_{\text{lim}}$ is the Chu-Harrington limit in (6) and $Q$ is found using Yaghjian and Best’s method in (9).

Figure 21- Loop-fed 4-arm ULP (left) and its model in Ansys HFSS (right)
### 2.8.1 Investigation of Antenna Height

We examine the effects of number of arms and the presence or absence of crossbars on performance parameters such as $Q$, bandwidth, efficiency, and radiation resistance, as height reduces and the aspect ratio becomes unconventional. From Section 2.3.4, increasing the number of arms with the presence of crossbars should increase the antenna’s radiation resistance and radiation efficiency. We examine how simple changes in geometry impact ULP antenna performance, $0.01\lambda$.

![Figure 22- Height investigation parameters as it affects pitch (left) and radius (right)](image)

This experiment varies the height while keeping the number of turns constant ($N = 1$). To maintain constant arm length of $\lambda/4$, the radius and pitch of arms varied according to Figure 22. The length of crossbars is not included in the arm length. Bandwidth, efficiency, and radiation resistance are collected at resonance. Height is examined below $0.12\lambda$ because impedance mismatch was observed between feed loop and helix beyond this height.

As expected, as antenna height increases and fills more of the Chu sphere volume, $Q$ approaches its limit (Figure 23), efficiency nears 100% (Figure 24), and both radiation resistance (Figure 25) and bandwidth (Figure 26) improve significantly.
Figure 23 - Antenna Q (Best) versus absolute Q limit (Chu-Harrington) at various heights

Figure 24 - Antenna efficiency increases with height and number of arms
At ULP height, we observe poor performance in terms of large Q and efficiency as well as low radiation resistance and bandwidth. Performance improves by adding more arms and filling the Chu sphere. For instance, Q improves from 34 times the limit for one-arm helix to 12 times the limit for 4-arm helix with crossbar. Efficiency for one-arm helix compared with 4 arms with crossbar improves from 44% to 61%. The simulated radiation resistance increases more sharply than that expected of resonant helical antennas, \( R_{rad} = (25.3 \frac{h}{\lambda})^2 \) [62].

However at ULP height, bandwidth remains just shy of 2 MHz. While increasing the number of arms improves bandwidth, there is physically not enough space for more than four arms. Thinner arms is possible with the tradeoff that the amount of power supported by the antenna will also be reduced. The thinnest arm that can still support 2 A of transmit current is 1 mm.

If the restriction on height can be relaxed, bandwidth can increase significantly and efficiency nears 100%. For instance, at a height of just 0.056\( \lambda \) for 4-arm helix without crossbar configuration, bandwidth is 6.5 MHz with 100% efficiency.
Figure 25- Radiation resistance increases with number of arms and height

Figure 26- Max achievable bandwidth increases with number of arms and height
2.8.2 Investigation of Helix Size with Constant Height

This experiment investigates changing helix size at ULP height and compares the effects of the number of arms and the presence of crossbars to performance parameters such as resonant frequency, Q, bandwidth, efficiency, and radiation resistance. For an arm length of \( \lambda/4 \), as helix radius increases, the number of turns decreases, Figure 27. We examine radii between 4 cm (0.02\( \lambda \)) and 16 cm (0.08\( \lambda \)) because below this range, the antenna does not resonate and above this range the antenna is no longer a normal mode helix. The geometry for 8 cm radius at ULP height is similar to the geometry where height is 0.01\( \lambda \) in Section 2.8.1, and yield similar results.

![Figure 27- Changing radius affects number of turns while length and height are constant](image)

The resonant frequency of the no-crossbar case can be easily predicted as \( f = c/4l \), or about 150 MHz, Figure 28. The resonant frequency of the crossbar case does not follow the same trend. Figure 29 shows the change in resonant frequency for various feed loop radii 0.6 to 0.9 times the helix radius. Because the frequencies are all very similar to without helix cases (Theory No Crossbar), we conclude that little energy is stored in the feed loop compared to the helix. It is easier to design ULP antennas without crossbars because resonant frequency can be more easily predicted.
Figure 28- Resonant frequency is calculable for no-crossbar compared with crossbar case

Figure 29- Resonant frequency for various feed loop radii for 4 arm ULP is calculable
Figure 30 - Antenna Q (Best) versus absolute Q limit (Chu-Harrington) at various radii

Figure 31 - Efficiency is dependent on number of arms at various radii
As in Section 2.8.1, antenna performance improves with the number of arms, the presence of crossbars, and the ability to fill a Chu sphere. As the radius increases and at ULP height, the antenna becomes unable to fill a Chu sphere and Q increases dramatically above its limit, Figure 30. While increasing the number of arms and the presence of crossbars lowers Q, space issues prevent us from accommodating more than four arms. We observe Q between 2-arm crossbar and 4 arms no-crossbar
is similar. As radius increases, simulated Q is relatively stable compared with the Q limit (Chu-Harrington) that changes with frequency.

As the number of arms increases to fill a Chu sphere, efficiency increases, but there is a maximum limit, Figure 31. For instance, a 4-arm ULP antenna can achieve a maximum efficiency of 80%. When the antenna size is a factor, the maximum limit for 2 arms occurs at 7 cm radius and 4 arms occurs at 11 cm radius provides design guidelines. The efficiency for 8 cm radius is slightly higher than efficiency at 0.01λ in Section 2.8.1 because this antenna is slightly taller.

Radiation resistance increases more slowly when increasing radius compared with increasing height, Figure 32. Below 0.055λ radius, the radiation resistance is very similar between 2 and 4 arms, with and without crossbars. At 0.01λ, or 8 cm, the radiation resistance for 4 arms is 8 Ω.

The number of arms does not affect the maximum achievable bandwidth, which remains just shy of 2 MHz regardless of radius size, Figure 33. This is because antenna Q is relatively stable. This bandwidth limitation illustrates the fundamental flaw in antennas with aspect ratios where the diameter is much larger than the height. At ULP height, it is very difficult to achieve theoretical bandwidth. Simulations show that Q and bandwidth are both stable as radius increases. From Figure 28 and Figure 29, we conclude that significant energy is not stored in the feed loop. In order to increase the bandwidth, the height must be increased.

### 2.9 Antennas for Orbcomm Frequencies

We present prototype single resonance 4-arm ULP antennas tuned for Orbcomm frequencies, Figure 34. The receive ULP (137.5 MHz) has helix radius 7.8 cm and feed loop radius 6 cm. The transmit ULP (149 MHz) has helix radius 7.25 cm and loop radius 4.5 cm. The difference in size is within simulation results suggesting feed loop radii 60 - 80% of helix radius yield similar operating frequencies. Using the Bench Test method in Section 2.7.2, we measure the receive ULP bandwidth
of 1.44 MHz (at return loss better than -10 dB), which is within Orbcomm requirement. The height limitation restricts the transmit ULP to a bandwidth of 0.72 MHz, Figure 36, which does not meet Orbcomm uplink bandwidth requirement of 2.05 MHz. The maximum antenna gain for both antennas was measured to be -10 dBi using the Range Test method, which is consistent with simulation results of -11 dBi. The measured gain exceeds the Orbcomm requirement of -14 dBi.

Figure 34- Prototype ULP antennas on ground plane tuned for Orbcomm frequencies
Figure 35 - Receive ULP antenna input impedance (top) and return loss (bottom)
Figure 36- Transmit ULP antenna input impedance (top) and return loss (bottom)
2.10 Discussion

It is well known that volumetric antennas that effectively utilize the Chu sphere have best performance. While this method has greatly helped researchers to reduce heights, antennas with our extreme aspect ratios have not been previously reported in literature. For small antennas, there is a well-known size to bandwidth limitation that cannot be exceeded. Applying Chu-Harrington, Gustafsson, and Kraus methods revealed that a cylindrical antenna with a height of 2.54 cm and a radius of at least 8 cm will achieve a 2 MHz bandwidth at 150 MHz, which just covers Orbcomm’s bandwidth requirement. However, they do not specify how to realize these antennas. To capture the maximum volume, we considered the design of cylindrical helix antennas. We have demonstrated that these ultra low profile (ULP) antennas at one hundredth of a wavelength tall are suitable for applications such as Orbcomm. Our design goal was to achieve a substantial fraction of the performance of a full size Orbcomm reference antenna.

The type of feed can further reduce a ULP antenna’s bandwidth, efficiency or both. We introduced an internal magnetic coupling loop with adjustable radius chosen to achieve the best...
impedance match. In order to simplify design calculations, we developed an equivalent circuit model that allows extraction of key design parameters to be estimated much more quickly than would be possible by simulation or measurement.

To identify the optimum helix design, we assessed the radiation efficiency and bandwidth of the antenna as a function of helix height, radius, number of arms, number of turns, feed loop radius, presence or absence of crossbars that connect the arms at the top of the helix using simulations and validated the results by measuring the performance of selected hardware prototypes. As expected from Section 2.3.4, increasing the number of turns and arms increases the volume and improves performance. At ULP height, the presence of crossbars contributes minimally to radiation by showing similar radiation resistance and efficiency compared to no the crossbar case, as expected from Section 2.5.1. Due to close proximity to the ground plane, the radiation resistance cannot be determined from the helical antenna formula and long crossbars make resonant frequency unpredictable. Clearly, it is easier to design for the no-crossbar case.

Our design curves show the bandwidth of a ULP antenna is insensitive to increasing radius, and remains just shy of 2 MHz. Increasing the number of arms improves the bandwidth but this strategy is limited by the available space. This bandwidth limitation illustrates the fundamental flaw in height-limited antennas with aspect ratios where the diameter is much larger than the height. Efficiency increases significantly with radius but there is a limit defined by the number of arms. If the restriction on height can be relaxed, bandwidth can increase significantly and efficiency nears 100%.

Our best 4-arm ULP antenna design without crossbars for downlink is 7.8 cm radius helix with 6 cm feed loop and for uplink is 7.25 cm helix radius with 4.5 cm feed loop. The bandwidths are 1.44 MHz (meets requirements) and 0.72 MHz (smaller than requirement), for receive and transmit frequencies. The gain is -10 dBi, which exceeds Orbcomm’s -14 dBi minimum.
Chapter 3: Duplexers for Orbcomm ULP Antennas

3.1 Introduction

Electrically small antennas are compatible with low-power time division duplex (TDD) systems that have a single relatively narrow passband. Conversely, the majority of long-range systems are high power frequency division duplex (FDD) systems with a frequency separation between the uplink and downlink passbands. The uplink and downlink bands are generally a standard duplex pair that is based on the 12 - 15% operating bandwidth of a typical dipole antenna.

A half-wave dipole or quarter-wave monopole antenna is generally suitable for fixed asset monitoring of sensors in remote areas such as oil pipelines and propane tanks. As the range of applications expands to tracking, monitoring, messaging for mobile assets like trucking and heavy construction equipment fleets, weather reports and fleet monitoring for the fishing industry, much smaller, low-profile antennas are required.

To meet the requirements of mobile asset tracking, designs that use small antennas with relatively narrow bandwidths must use a two-antenna system or a single dual resonant antenna. If these antennas can be directly connected to the transmitter and receiver, the problem is solved. In the case of legacy systems, the designer must provide a method of connecting both antennas to a single antenna port on the modem. The device used to combine this system is known as a duplexer, Figure 38.

![Figure 38- Block diagram of modem (a), ideal two-antenna solution (b), two-antennas with duplexer solution (c)](image-url)
There are several types of FDD systems that may encounter the same small antenna problem. For instance, as the size of antennas used in cellular phones (GSM) and land mobile radios decrease, the operating bandwidth becomes substantially less than that of the reference dipole antenna and a suitable duplexer is required. There are a variety of ways to isolate the transmitted and received signal: frequency (FDD), polarization (antenna), or timing (TDD). There are two well-known schemes for realizing a duplexer, each has its strengths and limitations. For instance, a carrier-operated relay (COR) employs the timing-based technique, has the widest bandwidth, but requires external power for operation. The diplexer filter is a frequency-based solution, is bandwidth limited, and has high insertion loss. There is no universal solution.

Before moving forward, we recognize existing solutions. For instance, Reactel can tune their duplexers for Orbcomm frequencies but their specialized filters are expensive (several hundred dollars each), heavy, and double the height requirement (5.5 cm tall). Plus, they detune easily. We propose that a specialized duplexer design will meet Orbcomm cost and complexity.

![Figure 39- Diplexer from Reactel Inc. with one antenna port and two output ports](image)

Initially, our objective was to determine the feasibility of realizing a duplexer for connecting a pair of ULP antennas for Orbcomm’s FDD system. We found neither the standard COR nor diplexer filter entirely satisfactory in meeting the system requirements. We propose an alternative scheme, a “complementary feed” that meets the requirements at a fraction of the cost and complexity.
The remainder of this chapter is organized as follows. In Section 3.2, we discuss the unique system level requirements for the Orbcomm system. We design a carrier-operated relay in Section 3.3 and discuss methods to reduce power consumption. In Section 3.4, we design a diplexer and examine its statistical yield for manufacturability. Finally in Section 3.5, we design a far simpler duplexer solution called a complementary feed and examine its performance.

3.2 Orbcomm Duplexer Requirements

The key parameters of a duplexer are listed below and the affected duplexer design(s) is identified. Radar charts in Figure 41 present the limitations in a graphical manner. We consider any weakness as 50% and no weakness as 100%. The large size of the corresponding radar chart shows that the complementary feed is a potentially better solution than the COR and diplexer filter.

**Bandwidth**

We consider the bandwidth limitations of the duplexer. Because the COR is a switch based solution, bandwidth is not a limitation. Bandwidth issues greatly affect the design of the diplexer filter, Section 3.4.2, and complementary feed to a lesser extent, Section 3.5.2.

**Isolation**

Isolation is a measure of the attenuation between ports that should not be internally connected. For full duplex systems, isolation reduces interference caused by: 1) receiver desensitization from unwanted frequencies entering a receiver’s upper stage passbands, and 2) transmitter noise as a result of transmission power leaking into other nearby frequencies. Isolation is not an issue for our application because: 1) Orbcomm is a half duplex system, and 2) the modems already include internal duplexers with high isolation. This greatly relaxes the requirements of the duplexer design, particularly the diplexer filter, Section 3.4.2 and complementary feed, Section 3.5.2.
Power Handling

The maximum transmit power of the Orbcomm modem is typically 5 to 10 W (depending on manufacturer), Appendix D. Orbcomm requires the diplexer to handle a maximum RF power of 5 W or 37 dBm. For a 50 Ω line, this means 22.4 V. The receive signal generally falls between -100 and -115 dBm. This affects COR switch selection, Section 3.3.1.

Switching Delay

Unlike voice systems, data systems transmit as soon as a connection is made which leaves little guard time for TDD systems. There are a variety of modem vendors for Orbcomm and the backoff time between switch open (duplexer inside the modem) and data transmit is not standard. Examination of Orbcomm’s Acquire signal, Figure 40, shows that the guard time is restricted to 1 symbol rate. This reveals a guard time of 416 us for OG1 and 208 us for OG2 systems. Our duplexer must switch faster than 208 us. This only affects carrier sensing based solutions like the COR, Section 3.3.1.

Insertion Loss

Insertion loss is the reduction of signal power from adding a device in a transmission line and it is measured in dB. Adding a duplexer will reduce power delivered to the transmit antenna and from the receive antenna. From the minimum link margin in Table 11 in Appendix B, the maximum insertion loss should be well below 5 dB. This affects diplexer filter design the most, Section 3.4.2, and COR switch selection to a lesser extent, Section 3.3.1.
Power Consumption

Power efficiency is pivotal as the receiver system is battery operated. The modem uses a 12 V battery. This only impacts COR component selection, Section 3.3.2.

Price and Size

We expect to deploy hundreds of thousands of duplexers with the ULP antennas so cost is a concern. The COR and diplexer require specialized components while the complementary feed uses regular RF equipment.

Orbcomm expects the duplexer and antenna solution to be no taller than 2.54 cm and fit within a space of 10.16 cm by 30.48 cm. Duplexer size is generally negligible compared to the size of the antenna.
3.3 Carrier Operated Relay

In the last section, we noted that while the COR has largest bandwidth, it uses carrier sensing to distinguish between the transmit and receive signals and is limited in almost every system parameter. Power handling, switching delay, insertion loss, and price can be addressed through proper switch selection to meet Orbcomm requirements. Proper selection of circuitry can minimize power consumption but not totally eliminate it. We consider power scavenging from other sources as an alternative to batteries.
Two types of COR designs are possible: a carrier-operated relay and a carrier-operated switch. For the case of a carrier-operated relay, no inverter (3) is required because the relay only requires power to close the normally open connection. A carrier-operated switch requires constant power to close all connections and an inverter is required. In Section 3.3.3, we analyze the power consumption of both systems.

The COR operates as follows. In the absence of a transmitted signal, the COR remains in the normally closed (NC) position and connects the radio to the receive antenna. When the radio transmits, the signal that appears at the output of the envelope detector (1) exceeds a voltage threshold and the comparator (2) causes the RF relay or switch (4) to shift to the normally open (NO) position and connects the radio to the transmitting antenna. An inverter (3) is required at the NC position for RF switches.

The rest of this section is laid out as follows. In Section 3.3.1 we discuss relay and switch selection for COR and use the selected components to design a COR for Orbcomm in Section 3.3.2. Section 3.3.3 discusses the power budget while Section 3.3.4 considers power scavenging from the GPS as an alternative to batteries.
3.3.1 Relay and Switch Selection for COR

We compare RF relay and switch technologies against Orbcomm’s requirements, Section 2.4. There are a variety of technologies to choose from: electromechanical relays, semiconductor-based devices, and MEMs switches. The difference is that the switching mechanism in electromechanical relays is moveable while semiconductor-based devices are not. Examples of semiconductor-based devices include: PIN diodes, gallium-arsenide (GaAs) FETs, and silicon-on-insulator (SOI) or silicon-on-sapphire (SOS) transistors as the active switching devices [63]. Recently, MEMs switches are becoming an attractive alternative due to low insertion loss, low power consumption, and ease of integration.

FET based switches from fourteen manufacturers and distributors were compared, Table 5. Insertion loss and VSWR were read from performance curves on the datasheet. Max RF Input Power is the absolute maximum rating of the device. Switching Time and Current Draw are typical values from the datasheet unless otherwise stated. Isolation is considered for RF-In to RF-Out. The same is recorded for electromechanical relays and PIN diodes in Table 6.

We find that FET based semiconductor switches offer the best value in today’s market. Table 5 shows that with proper switch selection, we can easily meet and exceed Orbcomm requirements for power handling, switching delay, insertion loss, and price. With volume pricing, switches can cost
less than a dollar. The top 5 switch choices are highlighted and presented in order from lowest price: 1) HMC544E, 2) MASWSS02041, 3) MASWSS018, and 4) RFSW8000. Additionally, this technology consumes very little power, only 0.0002% (10 uW/5 W) of the total transmit power of the modem, Appendix D.

From Table 6, it is clear that electromechanical relay and PIN diode technologies are not suitable for Orbcomm because they are expensive, have high switching delay, and high power consumption (2.5% of the total transmit power of modem). MEMs devices are unattractive due to their large control voltage. As RF MEMs technology matures and with volume pricing, latching MEMs technology may be a candidate in the future. In writing this thesis, we hope to motivate the research and development of normally closed RF MEMs technology for VHF applications with lower control voltage.
## Table 5- Comparison of switch technology with top choices highlighted

<table>
<thead>
<tr>
<th>Component</th>
<th>Max RF Input Power (dBm)</th>
<th>Insertion Loss (dB)</th>
<th>Switching Time 90/10% RF to 10/90% RF (ns)</th>
<th>Current Draw (mA)</th>
<th>VSWR (:1)</th>
<th>Volume Price (USD)</th>
<th>Unit Price (USD)</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF Micro Devices</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>RFSW80000</td>
<td>40</td>
<td>0.3</td>
<td>100</td>
<td>5 uA</td>
<td>1.12</td>
<td>$0.97 ea for 100+</td>
<td>$0.68 ea for 750+</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>$2.12 for 100+</td>
<td>$1.4 for 750+</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>$3.03 ea for 25</td>
<td></td>
</tr>
<tr>
<td>FMS2031-001</td>
<td>41</td>
<td>0.5</td>
<td>90</td>
<td>&lt;5 uA</td>
<td>1.12</td>
<td>$0.8 ea for 100+</td>
<td>$0.56 ea for 750+</td>
</tr>
<tr>
<td>RF1200</td>
<td>38</td>
<td>137.6 MHz: 0.29</td>
<td>5 us</td>
<td>20 uA</td>
<td>137.6 MHz: 1.06</td>
<td>$0.8 ea for 100+</td>
<td>$1.14 ea for 25</td>
</tr>
<tr>
<td></td>
<td></td>
<td>148.8 MHz: 0.29</td>
<td></td>
<td></td>
<td>148.8 MHz: 1.04</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Hittite</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HMC595E 3 W</td>
<td>39</td>
<td>0.25</td>
<td>80</td>
<td>0 V: 10 uA</td>
<td>1.07</td>
<td>$1150.00 for 500</td>
<td>$1.53 ea for 10</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>3 V: 2 uA</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>8 V: 40 uA</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HMC574MS8 5 W</td>
<td>39</td>
<td>0.25</td>
<td>80</td>
<td>10 uA</td>
<td>1.04</td>
<td>$137.00 for 500</td>
<td>$1.82 ea for 10</td>
</tr>
<tr>
<td>HMC544E</td>
<td>39</td>
<td>0.25</td>
<td>70</td>
<td>3 V: 0.5 uA</td>
<td>1.03</td>
<td>$590 for 500</td>
<td>$0.85 ea for 10</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>5 V: 2 uA</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MASW-007588</td>
<td>37 at 3 V</td>
<td>0.5</td>
<td>55</td>
<td>15 uA</td>
<td>&lt;1.22</td>
<td>$1.58 ea for 100</td>
<td>$3.27</td>
</tr>
<tr>
<td></td>
<td>39 at 5 V</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MASW-007921</td>
<td>38 at RFC to RX path</td>
<td>0.43</td>
<td>21</td>
<td>2 uA</td>
<td>1.08</td>
<td>$1.15 for 500 - 999</td>
<td>$1.09 for 1000 - 2499</td>
</tr>
<tr>
<td></td>
<td>40 at RFC to TX path</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>$0.89 for 2500 - 2999</td>
<td>$1.52</td>
</tr>
<tr>
<td>MASW-008853</td>
<td>38</td>
<td>0.25</td>
<td>70</td>
<td>5 uA</td>
<td>1.07</td>
<td>$2.15 ea for 100</td>
<td>$2.69</td>
</tr>
<tr>
<td>MASWSS0181</td>
<td>38</td>
<td>0.4</td>
<td>40</td>
<td>50 uA (max)</td>
<td>1.22</td>
<td>$0.98 ea for 100</td>
<td>$0.98</td>
</tr>
<tr>
<td>MASWSS0204</td>
<td>38</td>
<td>0.25</td>
<td>70</td>
<td>5 uA</td>
<td>1.07</td>
<td>$0.72 ea for 100</td>
<td>$0.84</td>
</tr>
<tr>
<td>Skyworks</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>SKY-13329-321LF</td>
<td>39.5</td>
<td>0.36</td>
<td>200</td>
<td>200 uA</td>
<td>1.094</td>
<td>$1.39 ea for 100</td>
<td>$1.262 ea for 500</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>$1.147 ea for 1000</td>
<td>$1.85</td>
</tr>
</tbody>
</table>

2 Minimum order is $100 and $30 shipping. 3 Manufacturer has confirmed that switches work at frequencies of interest though datasheets do not specify. The external DC blocking capacitors used on the RF ports should be high enough to present a low Xc, like 1 nF.
Table 6- Comparison of alternative switching technologies: electromechanical relays, reed relays, PIN diodes, and MEMs

<table>
<thead>
<tr>
<th>Component</th>
<th>Max RF Input Power (dBm)</th>
<th>Insertion Loss (dB)</th>
<th>Switching Time 90/10% RF to 10/90% RF (ms)</th>
<th>Operating Power (mW)</th>
<th>VSWR (:1)</th>
<th>Volume Price (USD)</th>
<th>Unit Price (USD)</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Electromechanical Relay</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Panasonic ARE13A24Z</td>
<td>40</td>
<td>0.04</td>
<td>on: 10 off: 5</td>
<td>200</td>
<td>1.05</td>
<td>$3.31 for 800</td>
<td>$4.37 for 400</td>
</tr>
<tr>
<td>Panasonic ARE104H</td>
<td>40</td>
<td>~0</td>
<td>on: 10 off: 6</td>
<td>200</td>
<td>~1</td>
<td>$4.23 for 100</td>
<td>$3.74 for 500</td>
</tr>
<tr>
<td>Panasonic ARS144H</td>
<td>40</td>
<td>~0</td>
<td>on: 10 off: 6</td>
<td>200</td>
<td>~1</td>
<td>$7.86 for 100</td>
<td>$6.94 for 500</td>
</tr>
<tr>
<td><strong>Reed relay</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>SIL05-1B90-71Q (Reed relay) normally closed</td>
<td>40</td>
<td></td>
<td>on: 0.7 off: 1.5</td>
<td>125</td>
<td></td>
<td></td>
<td>$6.00</td>
</tr>
<tr>
<td>DIL series reed relay</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>PIN Diodes</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>SKY12208-306LF (Switch)</td>
<td>48.75</td>
<td>0.1</td>
<td>85 ns</td>
<td>1.4 W</td>
<td>1.02</td>
<td>$9.35 for 100</td>
<td>$7.90 for 500</td>
</tr>
<tr>
<td><strong>MEMs</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>RMSW220HP</td>
<td>40</td>
<td>0.3</td>
<td>on: &lt;10 us off: &lt;2 us</td>
<td>&lt;1 nW +/-90 V</td>
<td>1.04</td>
<td></td>
<td>$20</td>
</tr>
</tbody>
</table>
3.3.2 A Basic COR Circuit for Orbcomm

Using off the shelf components, we designed a prototype COR for evaluation. The output control signals (measured on Agilent InfiniiVision DSO7034B digital storage oscilloscope) are shown in Figure 43. The COR is powered by a bench power supply at 5 V. We first consider the component selection: envelope detector, comparator and inverter, and switch. The presence of a discharging resistor is not obvious but required to facilitate connection between any S/C and COR. Finally, we ensure the COR is compliant with Orbcomm requirements.

Envelope Detector

A large resistor is placed in series with the envelope detector consisting of a diode, resistor, and capacitor. A 1N5711 Schottky diode reduces switching delay and power consumption and is designed for VHF applications. The large resistor in series with the envelope detector (1 kΩ) acts as high impedance looking into the envelope detector and minimizes the transmitted power lost to the RF sensing circuit. The large resistor also reduces impedance mismatch so that the S/C sees a VSWR of 1.05:1 (1 kΩ//50 Ω). We suggest using even larger resistors to reduce impedance mismatch and power dissipation.

Comparator and Inverter

The comparator, LM311, protects inverter and switch from voltage spikes. An internal BJT in the comparator connects a pull up resistor to 5 V to ensure voltage high. A 5 V voltage regulator powers the comparator and inverter (74HCT04). A threshold voltage for the comparator is chosen above ground.

RF Switch

The RF switch is RFSW8000, third choice in switch selection as per Table 5.
Discharging Resistor

A discharge resistor to ground (1 kΩ) is required as some modems (Stellar DS300) do not discharge their internal capacitance. Without the discharging resistor, charge will build up on the source side of the Schottky diode causing the envelope detector to turn ON during both positive and negative cycles. Adding the resistor before the diode creates a path for capacitor discharge when the diode is OFF. We suggest larger resistors for the voltage divider to reduce power dissipation. We don’t need to know the size of the internal capacitor as it likely differs between vendors.

The location of the discharging resistor affects input resistance seen by the S/C. For best impedance match, resistance should be close to 50 Ω to be consistent with characteristic line impedance. However, adding 50 Ω right after the radio is impractical because the S/C will see 25 Ω (50 Ω antenna in parallel with 50 Ω resistor). We suggest adding the discharging resistor after the high impedance resistor to eliminate the changes in input resistance seen by the S/C, Figure 43. For modems with internal discharge resistors, this solution looks like a regular voltage divider.

COR Analysis

In the last section, we demonstrated that with proper switch selection, we can easily meet and exceed Orbcomm requirements for power handling, switching delay, insertion loss, and price. We ensure that the entire COR system still meets these requirements.

The total VSWR of the circuit and RFSW8000 switch is 1.12:1, which is within Orbcomm requirements in Table 3. The typical switching delay is 302 ns (100 ps Schottky diode, 200 ns for LM311, 1.67 ns for 74HCT04, 100 ns for RFSW8000) while the absolute maximum propagation delay is about 639 ns (100 ps Schottky diode, 200 ns for LM311, 139 ns for 74HCT04, 300 ns for RFSW8000) which is orders of magnitude lower than OG2’s delay of 208 us. In the next section, we examine ways to mitigate power consumption of COR.
3.3.3 Power Budget and Cost

While we can easily meet or exceed most system level requirements, power consumption can be reduced but not totally eliminated. Interestingly, the switch consumes very little power, only 0.0002% of the 5 W transmit power of the modem, and the driving circuitry consumes more power. We compare the power efficiency of three driving circuitry designs: voltage regulator, Zener diode, and combination comparator and inverter integrated circuit (IC). We demonstrate that with proper design of driving circuitry, we can reduce power consumption and cost.
The power consumed by the basic COR circuit is shown in Table 7.1). The typical power dissipated is 2% (of the 5 W transmit power) of which 90% is consumed by the comparator and voltage regulator. As the components age, the absolute maximum power dissipated will increase to 36%. The total component cost of a basic COR is $2.74 ($1.39 for switch, $0.13430 for 74HCT04, $0.53 for LM311, $0.02 for capacitor, $0.63 for 1N5711, $0.035 for resistors).

We replace the voltage regulator with a Zener diode and comparator with an AND gate, Table 7 2) and Figure 44. The comparator can be easily replaced with an AND gate because our testing did not reveal any voltage spikes from the transmitter. This indicates electrical components do not need to be protected from large differential input voltage. The new circuit will save 99% typical power compared with the basic COR circuit. Its total cost is $2.41 ($1.39 for switch, $0.13430 for 74HCT04, $0.11874 for Zener diode, $0.13827 for resistors, $0.63 for 1N5711). The Zener diode dissipates less power during typical operation but its absolute power is higher as components age.

The lowest power option consists of a combination comparator and inverter IC that can be powered from the modem’s 12 V battery, Figure 45. From Table 7 3), the design consumes 0.086% of the transmit power and 0.14% as components age. The power consumption of this design is consistent with measurement results (at 2.8 mW). The total cost of low power COR is only $2.19 ($1.39 for switch, $0.14 for LM358, $0.007 per resistor, $0.63 for 1N5711).

With proper design of driving circuitry, we can reduce power consumption and cost. Option 3 is the most power efficient, cheapest, and the clear choice. Additionally, Option 3 is versatile. Using professional batteries, like professional 3*1.5 V AA Duracell Procell Alkaline battery at 2557 mAh and 0.1 A load, the batteries will last about 1 year, Table 8. This is a very conservative estimate as there is no data for small currents like 0.3 mA for LM358.
### Table 7 - Power budget of three COR designs

<table>
<thead>
<tr>
<th>Component</th>
<th>Absolute Max Power Dissipated</th>
<th>Typical Power Dissipated</th>
</tr>
</thead>
<tbody>
<tr>
<td>1) Basic COR circuit (Figure 43)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$2.74 ea</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Inverter 74HCT04</td>
<td>750 mW</td>
<td>610 uW</td>
</tr>
<tr>
<td>Comparator LM311 + V&lt;sub&gt;Ref&lt;/sub&gt; resistors</td>
<td>500 mW + 8.3 mW (calculated)</td>
<td>52.8 mW + 8.3 mW (calculated)</td>
</tr>
<tr>
<td>5 V voltage regulator LM7805</td>
<td>581 mW (calculated)</td>
<td>31.7 mW</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>1.8 W</strong></td>
<td><strong>93.4 mW</strong></td>
</tr>
<tr>
<td>2) Zener Diode powered Inverter + Switch (Figure 44)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$2.41 ea</td>
<td></td>
<td></td>
</tr>
<tr>
<td>2 Inverters 74HCT04</td>
<td>750 mW*2</td>
<td>610 uW*2</td>
</tr>
<tr>
<td>5 V Zener diode PLVA650A</td>
<td>250 mW</td>
<td>105 uW</td>
</tr>
<tr>
<td>Resistor (70Ω)</td>
<td>350 mW</td>
<td>4.9 uW</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>2.1 W</strong></td>
<td><strong>1.3 mW</strong></td>
</tr>
<tr>
<td>3) Low Power Comparator and Inverter + Switch (Figure 45)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$2.19 ea</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LM358</td>
<td>6.6 mW</td>
<td>3.6 mW</td>
</tr>
<tr>
<td>2 Resistors (~100kΩ each)</td>
<td>0.4 mW</td>
<td>0.4 mW</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>7 mW</strong></td>
<td><strong>4 mW</strong></td>
</tr>
</tbody>
</table>

Prices from Digikey referenced Nov. 2015.

### Table 8 - Power budget of the COR designs using external batteries

<table>
<thead>
<tr>
<th>Component</th>
<th>Absolute Max Power Dissipated</th>
<th>Typical Power Dissipated</th>
</tr>
</thead>
<tbody>
<tr>
<td>1) Basic COR circuit (Figure 43)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Inverter 74HCT04</td>
<td>750 mW</td>
<td>732 uW</td>
</tr>
<tr>
<td>Comparator LM311</td>
<td>500 mW</td>
<td>66 mW</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>1.3 W</strong></td>
<td><strong>67 mW</strong></td>
</tr>
<tr>
<td>2) Zener Diode powered Inverter + Switch (Figure 44)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Inverter 74HCT04</td>
<td>750 mW*2</td>
<td>732 uW*2</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>1.5 W</strong></td>
<td><strong>1.5 mW</strong></td>
</tr>
<tr>
<td>3) Low Power Comparator and Inverter + Switch (Figure 45)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>LM358</td>
<td>6.6 mW</td>
<td>1.8 mW</td>
</tr>
<tr>
<td><strong>Total</strong></td>
<td><strong>6.6 mW</strong></td>
<td><strong>1.8 mW</strong></td>
</tr>
</tbody>
</table>
3.3.4 Power Scavenging

Proper switch selection allows us to easily meet or exceed most system level requirements. Interestingly, most power is consumed by the driving circuitry. In the last section, we demonstrated that with proper driving circuitry design, we can reduce power consumption and cost. In fact, batteries will last about one year. However, changing batteries is never ideal. Power scavenging from the GPS is a
viable no-battery solution as it only consumes 1.3% to 2.3% of the total GPS power and only increases the component count by one, Figure 46. The ferrite bead jointly powers the comparator and inverter combo circuits in Option 3 while suppressing high frequency electromagnetic interference from the GPS. Design considerations such as location of COR and ferrite bead selection are discussed below.

**Figure 46- Schematic of power scavenging from GPS antenna**

Since it is up to the customer to install the COR and antennas, the antenna designer can only make recommendations. We consider the location of the COR if the S/C and antennas are spatially distant in relation to cable cost. Placing the COR close to the S/C requires three long cables for the antennas whereas placing the COR close to the antennas only requires two long cables for the S/C.

A ferrite bead is a passive device that removes noise from a circuit in the form of heat [64]. The crossover frequency should be selected near the GPS frequencies (1227.60 MHz and 1575.42 MHz). Below the crossover frequency, the ferrite bead acts as a normal inductor and inductive reactance increases above resistance. At the crossover frequency where resistance becomes greater than inductance and noise is suppressed because energy is dissipated as heat rather than being reflected back into the GPS antenna port. Above the crossover frequency, the self-capacitance of the ferrite bead increases over resistance and impedance decreases. Since the GPS antenna is not operational in this capacitive region, impedance mismatch does not need to be considered. Impedance mismatch can be
reduced in the inductive region by selecting ferrite beads with sharp impedances at the crossover frequency such as: BLM18HB331SN1D, MMZ1005F121E, and MMZ1005D121E.

The drawback to power scavenging is that the GPS must now be consistently ON whereas previously, its state was controlled by user application. COR Option 3 with power scavenging only requires 13 components: (2) SMT devices (RF switch and dual op amp), (1) SMT diode, (1) SMT capacitor, (8) SMT resistors, and (1) Ferrite bead.

3.4 Diplexer Filter

The main advantage of the diplexer filter in comparison with COR is that it does not require power. However, insertion loss is an issue and will reduce power radiated by the antenna and lowers overall efficiency, (14). To a lesser extent, bandwidth and price are also limiting factors of the diplexer filter. Because isolation is not a factor in a half duplex system, diplexer design is greatly simplified. We employ statistical yield analysis to demonstrate that a diplexer design with lossy lumped element components is realizable while maintaining relatively low insertion loss and high manufacturing yield.

The rest of this section is organized as follows. We select a diplexer configuration in Section 3.4.1 following simulation results. We conduct yield analysis to account for and minimize the uncertainty attributed to real component tolerances in Section 3.4.2 and present our final design.

3.4.1 Overview of Diplexer Technology

A diplexer multiplexes two frequency bands onto a third. There are two possible configurations of diplexers bandpass.bandstop and lowpass/highpass, Figure 47. From simulation, we select a configuration that mitigates bandwidth limitation while keeping low component count and price.

We simulate ideal diplexer characteristics using Diplexer Network Designer from Tonne Software using elliptical filters for their steep roll off because transmit and receive bands are only 11.5
MHz apart. Selecting elliptical filter design mitigates bandwidth limitation. We compare the return loss of ideal bandpass/bandstop and lowpass/highpass diplexer configurations of orders 2 and 3.

We observe that responses from bandpass/bandstop configurations have steeper transitions than lowpass/highpass configurations for the same order, as seen in Figure 48 and Figure 49. The ideal bandpass/bandstop diplexer configuration of order 2 was selected and statistical yield analysis was conducted mimicking real component tolerances.

![Diplexer configurations: lowpass/highpass (top), and bandpass/bandstop (bottom)](image)

**Figure 47-** Diplexer configurations: lowpass/highpass (top), and bandpass/bandstop (bottom)
Figure 48- Order 2 (top) and 3 (bottom) lowpass/highpass diplexers have flat transitions
Figure 49- Bandpass/bandstop diplexer of order 2 has ideal steep transitions
3.4.2 Statistical Design for Diplexer

Designing for statistical yield is important as we will potentially produce hundreds of thousands of diplexers and need to account for and minimize the uncertainty of component tolerances. There are various ways to approach the problem: brute force, where the designer iteratively simulates component variation based on trial and error; synthesis, where the designer inputs optimization goals and lets the computer iteratively simulate component variation; or analysis, where the designer inputs component variation and lets the computer generate a statistical yield based on optimization goals. We conduct yield analysis using Agilent’s Advanced Design System software (ADS) as it contains Murata capacitor and Coilcraft inductor libraries with real tolerances. Our study follows the analysis approach and we set an optimization goal of -1 dB for the insertion loss. Finally, we present a diplexer design with high statistical yield and flat sensitivity histogram using just 8 components.

Yield Analysis and Sensitivity Histogram for Diplexer

Yield analysis is the process of varying a set of parameter values, based on specified probability distributions, to determine the number of possible combinations resulting in satisfying the specifications. The design is simulated over a number of trials (as specified by ADS confidence tables) in which the yield variables have specific probability distribution functions. The number of passing and failing trials are recorded and yield is estimated. While yield analysis indicates whether there is a problem in the overall circuit due to the statistical nature of tolerances, it does not identify the responsible components. Sensitivity histograms examine how the tolerances of individual components affect the specifications. Through tuning, the components can ideally become insensitive to component tolerance. After setting up the ADS environment, we conduct yield analysis for a range of component tolerances and select one with good manufacturability prospects.
Figure 50- Yield analysis setup in ADS
Figure 51- Sensitivity histogram setup examines component tolerance to insertion loss

Figure 50 shows the diplexer and yield analysis setup in ADS and Figure 51 shows sensitivity histogram setup for bandpass and bandstop filter. The yield controller was set up for 250 trials to achieve a confidence level of 95% for a yield around 86% to 94% with an error of +/- 4%. The optimization goal is set at insertion loss better than -1 dB for transmit and receive bands.

Yield analysis was conducted for a range of component tolerances, Table 9. Studies show a yield of 94.8% is achievable for component tolerances of 2%, indicating good manufacturability. By incorporating real component tolerances (mimicking capacitors from Murata and inductors from Coilcraft), Table 10, the highest yield achievable is 97% with a confidence level of 95%, Figure 52.
The sensitivity histograms of the bandstop and bandpass filters have flat responses in Figure 53, indicating immunity to component tolerances.

Table 9- Comparison of component tolerances to manufacturing yield for diplexer

<table>
<thead>
<tr>
<th>Component Tolerances (%)</th>
<th>Yield (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>2</td>
</tr>
<tr>
<td>5</td>
<td>28.8</td>
</tr>
<tr>
<td>2</td>
<td>94.8</td>
</tr>
</tbody>
</table>

Table 10- Final diplexer design with component tolerances resembling real components

<table>
<thead>
<tr>
<th>Component</th>
<th>Value (pF)</th>
<th>Tolerance (+/- pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1</td>
<td>500</td>
<td>2%</td>
</tr>
<tr>
<td>L2</td>
<td>12.1</td>
<td>2%</td>
</tr>
<tr>
<td>L3</td>
<td>17</td>
<td>2%</td>
</tr>
<tr>
<td>L4</td>
<td>500</td>
<td>2%</td>
</tr>
<tr>
<td>C1</td>
<td>3</td>
<td>0.05</td>
</tr>
<tr>
<td>C2</td>
<td>129.91</td>
<td>0.1</td>
</tr>
<tr>
<td>C3</td>
<td>91</td>
<td>2</td>
</tr>
<tr>
<td>C4</td>
<td>2.7</td>
<td>0.05</td>
</tr>
</tbody>
</table>
Our final diplexer design is a bandpass/bandstop configuration of order two. The components are listed in Table 10. Yield analysis and sensitivity histograms show that the design has a high yield of 97% at a confidence level of 95% and flat sensitivity histograms indicating immunity to component tolerances. All eight components can fit on a single PCB board no bigger than 2.54 cm by 2 cm and trace impedance is negligible as they will be much shorter than $\lambda/20$ [65]. The trace should be designed wide enough to handle the specified power.

Figure 52- Yield analysis increases to 97% with real component tolerances
Figure 53- Flat sensitivities of bandstop (top) and bandpass (bottom) filters indicate immunity to component tolerances

3.5 Complementary Feed

Dissatisfied with the high power consumption of the COR and the insertion loss of the diplexer filter, we propose a novel complementary feed design that exploits a property of the ULP and other resonant antennas. This solution is simple and cost effective because it does not require specialized hardware like the other two. In fact, the complementary feed solution only requires a T-connection and a $\lambda/2$ length of coaxial cable. Its only weakness seems to be bandwidth but we show with proper cable selection, bandwidth will not become an issue. The complementary feed meets all of our system
requirements. Finally, we demonstrate successful satellite communication with working ULP prototypes and complementary feed.

The rest of this section is as follows. In Section 3.5.1, we discuss the working principle of the complementary feed. In Section 3.5.2, we analyze the bandwidth and in Section 3.5.3, we demonstrate successful satellite communication with working ULP antennas and complementary feed solution.

3.5.1 Overview of Complementary Feed

The typical reason not to connect two wideband antennas to a common port is that the input port will see a mismatch over the combined frequency band. This decreases the amount of power delivered to the antenna and creates reflections that can damage the modem. Schroeder’s theoretical complementary pair antennas [66] cancels reactance over the entire bandwidth but they are difficult to realize. We find this is not the case for narrowband antennas. By considering the complementary feed like a stub matching network, two cables are required for impedance matching in the general case. We define the complementary frequency of the transmitting antenna as the receiving frequency, and the complementary frequency of the receiving antenna as the transmitting frequency. By exploiting the property that our antennas have a complementary frequency almost at open circuit, only one cable is required. The complementary feed is a simple duplexer solution that requires only a T-connector and \( \lambda/2 \) cable.

Complementary feed matching resembles a stub-matching network where the end of the stub is an antenna instead of the typical open or short circuit. Figure 54 shows complementary feed setup where two ULP antennas are connected together via a T connector to cables of lengths \( d \) and \( L \). In the transmit case, the transmit ULP acts as the load, length \( d \) is chosen to cancel the reactance of the transmit ULP, and length \( L \) is chosen to match the real part of the load to characteristic line impedance.
This ensures the transmit ULP is matched and receives maximum power from the S/C. In the receive case, the purposes of lengths \(d\) and \(L\) are reversed.

Our problem is more complex than single stub matching because we are dealing with two frequencies. The typical solution is to incorporate more stubs. In the next section, we examine the bandwidth of a stub. In the general case, at least two lengths of cable are required to match the ULP antennas.

We observe that the complementary frequency of our ULP antenna is almost at open circuit, Figure 35 and 36, which helps to simplify our duplexer. At transmit, the impedance of the receive ULP (complementary frequency) acts as an open circuit and all of the power flows out the transmit ULP. In terms of a stub matching network at transmit, length \(d = 0\) because the reactance of the transmit ULP is zero and only length \(L\) is required, (36) and (37). Because the impedance of the receive ULP is almost at open circuit at transmit frequencies, length \(L\) becomes a multiple of \(0.5\lambda\). The opposite is true for receive ULP for length \(d\). For practical purposes, we account for the distance between the antennas by setting one cable length to \(0.5\lambda\) and the other to 0. Which cable is the longer one does not matter. By exploiting the property that our antennas have a complementary frequency almost at open circuit, only one cable is required.

Practical cables have a frequency dependent loss which may affect the calculated impedance. The total cable assembly loss for LMR-400 is low at 0.1 dB [67]. Here,

\[
Z_{149\text{ULP}+L} = 50 \times \left( \frac{1 + \Gamma_{149\text{ULP}} e^{-2j\beta L}}{1 - \Gamma_{149\text{ULP}} e^{-2j\beta L}} \right)
\]

(36)

\[
\frac{1}{Z_{\text{Total}}} = \frac{1}{Z_{137.5\text{ULP}}} + \frac{1}{Z_{149\text{ULP}+L}}
\]

(37)

where \(\beta L = 2\pi d \frac{\lambda_0}{\lambda}\) and \(d\) is the length of the complementary feed line, in metres.
3.5.2 Bandwidth Analysis of Complementary Feed Design

It is well known that a single stub will achieve a perfect match at only one frequency. For wideband matching, several stubs are required. The robustness of our design is challenged when we consider the possibility that the complementary frequency of the antenna does not occur exactly at open circuit. Using theoretical calculations, we examine how the complementary feed supports a wide range of complementary frequencies, or a large bandwidth.

![Diagram](image)

**Figure 54**- Complementary feed with length \( d \) and \( L \) connected to ULP antennas

**Figure 55**- Case 1 (top) shows transmission line length affects the impedance seen by the S/C, Case 2 (bottom) shows the impedance seen by the S/C is unaffected by transmission line

From transmission line theory, we examine the input impedance seen by the S/C as the open circuit impedance varies in parallel with 50 \( \Omega \). We consider two modes of operation. Case 1 in Figure 55 (top) shows the antenna in receive mode with the complementary frequency of the transmit antenna as open circuit. Variations in the length of the transmission line affects the impedance of the open circuit and
Figure 56- Combined active and complementary frequencies seen by the S/C
the impedance seen by the S/C. Case 2 in Figure 55 (bottom) shows the antenna in transmit mode where the complementary frequency of the receive antenna as open circuit. Here, variations in the length of the transmission line do not affect input impedance. Only Case 1 can affect bandwidth.

Figure 56 shows input impedance as seen by the S/C in Case 1. The bandwidth is about 112 MHz (at a return loss of -10 dB or better), which indicates a variety of complementary frequencies are supported by the complementary feed solution. We conclude that the complementary feed is very robust. The complementary feed is the clear choice because it does not require external power like the COR and there is no insertion loss like the diplexer filter.

### 3.5.3 ULP Antennas with Complementary Feed for Orbcomm

In this section, we ultimately validate the design of the ULP antennas and complementary feed solution through Orbcomm message testing. We also confirm that the complementary feed solution does not degrade the bandwidth of either antenna with proper cable length.

A setup of ULP antennas with complementary feed for message testing is shown in Figure 57, following the procedure in Section 2.7.2. The length of the complementary feed is 0.5λ. Figure 58 shows that there is no bandwidth degradation when the ULP antennas combine with a complementary feed of length \( d = 0.5\lambda \). S11 parameters of each antenna and combined with complementary feed are shown in Appendix C. The receive ULP antenna has a bandwidth of 0.7 MHz which covers the entire receive passband according to Orbcomm’s channelization plan [68]. The transmit ULPs have a bandwidth of 0.8 MHz, which covers 42% of Orbcomm’s channelization plan.
Figure 57- ULP antennas with complementary feed on ground plane for message test

Figure 58- Return loss of calculated and measured complementary feed show excellent agreement at $d = 0.5\lambda$. 
Our 72-hour message tests reveal the performance of prototype ULP antennas with complementary feed is 45% of a reference $\lambda/2$ antenna, an average of 6 messages per hour, Figure 59. Results are consistent and were repeated 3 times. For testing, we used a known 125% Orbcomm antenna and scaled the results, as the reference $\lambda/2$ antenna was unavailable.

![Figure 59- Screenshot of 72 h message test with reference Orbcomm antenna transmitting 1331 messages (left) and ULP antennas with complementary feed transmitting 480 messages (right)](image)

### 3.6 Discussion

Our objective in this chapter was to determine the best way to realize a duplexer for connecting a pair of ULP antennas to a standard Orbcomm modem with a single antenna port. Initially, we sought to determine the feasibility of realizing a duplexer using standard COR or diplexer filter methods. We compared the designs to a variety of system level parameters such as bandwidth, isolation, power...
handling, switching delay, insertion loss, power consumption, price, and size. However, we found neither method was entirely satisfactory in meeting the requirements. Our complementary feed is a novel duplexer solution that meets the requirements at a fraction of the cost and complexity.

The COR has the largest bandwidth because it uses carrier sensing to distinguish between transmit and receive signals. Proper selection of FET based switching technology allows us to easily meet or exceed most system level parameters like power handling, switching delay, insertion loss, and price. While MEMs technology switches much faster, it was not yet a suitable candidate requiring excessive switching voltage. As technology matures, latching MEMs may be a consideration in the future. Interestingly, the switch consumes very little power compared to the driving circuitry. We have demonstrated that proper selection of low power combination ICs reduced power consumption and cost. In fact, batteries would have lasted about one year. However, changing batteries is never ideal. We have suggested power scavenging from the GPS as an attractive alternative to batteries as the low power combination IC solution only consumes 1.3% of the total GPS power. The final COR design consists of 13 components: (2) SMT devices (RF switch and dual op amp), (1) SMT diode, (1) SMT capacitor, (8) SMT resistors, and (1) Ferrite bead.

We have also suggested a diplexer filter as an attractive solution over COR because it does not require power. However, insertion loss is an issue and will reduce power radiated by the antenna and lowers overall efficiency. To a lesser extent, bandwidth and price are also limiting factors of the diplexer filter. Because isolation is not a factor in the half duplex Orbcomm system, design of the diplexer filter was greatly simplified. Our final design is a bandpass/bandstop diplexer configuration with order two using elliptical filters. Yield analysis shows that the diplexer has a high yield of 97% at a confidence level of 95% (for an insertion loss of -1 dB or better). Flat sensitivity histograms indicate
immunity to component tolerances. All eight components can fit on a single PCB board no bigger than 2.54 cm by 2 cm and trace impedance is negligible as they will be much shorter than $\lambda/20$.

Dissatisfied with the high power consumption of the COR and insertion loss of the diplexer filter, we examine a novel complementary feed design that is simple and cost effective. By exploiting the property that the complementary frequency of the ULP antennas is almost at open circuit, this duplexer solution merely consists of a T-connector and $\lambda/2$ cable. We have calculated that with proper cable selection, complementary feed can tolerate a wide range of complementary frequencies with a large bandwidth of 112 MHz (considering a return loss of -10 dB or better). Furthermore, our theoretical calculations and measurements indicate that the complementary feed does not degrade the bandwidth of either antenna. Compared to the COR and diplexer, the complementary feed meets all of our system requirements.

Our final solution is a pair of ULP antennas connected to the modem through a complementary feed duplexer. Several tests, each conducted over a 72-hour period, showed that the solution consistently supported exchange of 45% of the test messages compared with a full size Orbcomm reference antenna, equivalent to 6 messages per hour.
Chapter 4: Conclusion

Government transportation regulations and practical considerations limit the height of terminal antennas for satellite-based mobile asset tracking applications to less than 2.54 cm. For Orbcomm systems that operate at 137 - 138 MHz (downlink) and 148 - 150.05 MHz (uplink), this implies an antenna that is just over one-hundredth of a wavelength tall. Designing an ultra low profile antenna that achieves good efficiency and operating bandwidth is fundamentally difficult. We have sought to overcome the limitations of conventional inverted-F and related planar antennas by using a multi-arm normal mode cylindrical helix antenna.

Researchers have long understood that volumetric antennas that effectively utilize the Chu sphere achieve the widest possible bandwidth. Several methods for estimating $Q$ have been developed by Wheeler, Chu-Harrington, Gustafsson, Yaghjian and Best, and Kraus, among others. From these, we conclude that a cylindrical antenna with a height of 2.54 cm and a radius of at least 8 cm will achieve a 2 MHz bandwidth at 150 MHz, which just covers Orbcomm’s bandwidth requirement. However, these limits do not specify how to realize these antennas.

Our design goal was to develop a ULP antenna that achieves a substantial fraction of the performance of a full size Orbcomm reference antenna. Dissatisfied with the limitations of conventional transmission line feeds, we introduced an internal magnetic coupling loop with adjustable radius allows us to easily achieve the best impedance match. In order to simplify design calculations, we developed an equivalent circuit model that allows design parameters to be estimated much more quickly than would be possible by simulation or measurement. In order to identify the optimum design, we assessed the radiation efficiency and bandwidth of the antenna as a function of helix height, radius, number of arms, number of turns, feed loop radius, presence or absence of crossbars that connect the
arms at the top of the helix using simulations and validated the results by measuring the performance of selected hardware prototypes.

Our design curves show that bandwidth increases significantly with height and is invariant of radius at ULP height. Efficiency increased significantly with helix radius and removing the crossbars yielded only a marginal decrease in efficiency. A pair of ULP antennas combined through a duplexer is required to support uplink and downlink. Our best 4-arm ULP antenna design without crossbars for downlink is 7.8 cm radius helix with 6 cm feed loop and for uplink is 7.25 cm helix radius with 4.5 cm feed loop. Several tests, each conducted over a 72-hour period, showed that this pair of antennas consistently supported exchange of 45% of the test messages compared with a full size Orbcomm reference antenna, or 6 messages per hour.

We began by considering two alternative duplexer designs: a carrier operated relay and a diplexer filter. We compared system level parameters such as bandwidth, isolation, power handling, switching delay, insertion loss, power consumption, price, and size. While a Carrier Operated Relay (COR) design has largest bandwidth, power consumption had been a concern. We have demonstrated that proper selection of switch and low power ICs reduced power consumption and component count. We also discussed the possibility of power scavenging from the GPS as an alternative to batteries. Our final COR design consists of 13 components. The design of the diplexer filter was an attractive solution over COR due to its passive nature but insertion loss had been a concern. Its design was greatly simplified because isolation is not a factor in the half duplex system. The final design is a bandpass/bandstop diplexer configuration with order two using elliptical filters. Our design has a high yield of 97% at a confidence level of 95% (for an insertion loss of -1 dB or better).

Dissatisfied with the cost and complexity of the COR and diplexer filter, we proposed a far simpler duplexer solution that we call a complementary feed that exploits the property that the
complementary frequency of the ULP antennas is almost at open circuit. This means ULP antennas can connect with Orbcomm modem via a T-connector and $\lambda/2$ cable. We have determined that complementary feed can tolerate a wide range of complementary frequencies with a bandwidth of 112 MHz (considering a return loss of -10 dB or better). Through calculations and measurements, we have demonstrated that with proper selection of cable length, the complementary feed does not degrade the bandwidth of either antenna. Because it offers low insertion loss, adequate bandwidth, large power handling capacity, is simple and inexpensive to manufacture and requires no external power, the complementary feed is the clear choice.

4.1 Limitations and Future Work

The equivalent circuit model in Section 2.6.1 applies to loop-fed single-turn helices. It also applies to loop-fed multi-arm helices with crossbar if the values for mutual inductance $M$ are recalculated. For multi-arm helices without crossbar, mutual inductance $M$ must be represented by a square matrix and the loss resistance of the helix changes due to the proximity effect. Understanding these relationships will allow us to quickly predict optimum loop size orders of magnitude more quickly than simulation.

Our investigation of ULP antennas in Section 2.8.2 showed that the performance (bandwidth, radiation resistance, and efficiency) of 2- and 4-arm geometries without crossbars are similar at 8 cm radius. Preliminary measurements and simulations suggest the possibility of tuning two arms to resonate at one frequency and the other two arms to another frequency. More research is required to evaluate the feasibility of this dual-band approach.

We did not observe any changes in the performance of the uplink and downlink ULP prototype antennas when the antennas were placed 18 cm apart. However, we have observed changes when they are brought closer together. Further research is required to fully understand this coupling effect.
References


[64] C. Burket, “All Ferrite Beads are Not Created Equal - Understanding the Importance of Ferrite Bead Material Behavior,” In Compliance Mag., pp. 18–26, 01-Aug-2010.


Appendices

Appendix A: Orbcomm Messaging Protocol

The S/C originated data transfer session begins when it receives control information from the satellite, known as Sync. The S/C sends a “request to send” signal to the satellite, called the Acquire Burst that is 3.3 ms long. If there is no interference and there are no time-overlapping bursts on the same receive channel, the satellite receives the Burst and sends control information for Slot Assignment to the S/C. The S/C waits a period of time to receive the Slot Assignment and if not received in time, the process will repeat at a later time. If the S/C receives a response, it transmits short bursts in one of three forms: a Data Report (up to 6 bytes), a request to send a larger amount of data, or control information. For larger message transfers, the satellite reserves a time slot and uplink frequency channel for the message. The S/C sends each message burst and waits for an acknowledgement from the GCC. The process repeats until the message is completely and successfully transferred from the S/C to GCC.

For the S/C terminated message transfer process, when the Gateway as a message addressed to an S/C, it sends an assignment packet to the appropriate satellite (designated according to geographical zones), which sends an assignment packet to the S/C. The S/C responds with an Acquire Burst and the Gateway will start sending message packets. The transfer is complete after the S/C acknowledges the final packet. In cases when the satellite is in view of an S/C but no Gateways, such as in the ocean, the satellite will store the packet (called a GlobalGram) in memory until it sees a Gateway.
Appendix B: Orbcomm Link Budget

Orbcomm’s link budget is shown in Table 11 for their Generation 1 (OG1) and Generation 2 (OG2) satellites. Each OG2 satellite acts as the functional equivalent of six OG1 satellites by providing quicker service, enhancing coverage at higher latitudes and allowing for larger message sizes and increased data rates.

Table 11- Link budget for Orbcomm Generation 1 and 2 (OG1 and OG2) satellites

<table>
<thead>
<tr>
<th></th>
<th>Uplink (149 MHz)</th>
<th>Downlink (137.5 MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>OG1</td>
<td>OG2</td>
</tr>
<tr>
<td>Tx Power Out (W)</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>VSWR</td>
<td>1</td>
<td>3</td>
</tr>
<tr>
<td>Tx out w/ Mismatch Loss (W)</td>
<td>5</td>
<td>3.8</td>
</tr>
<tr>
<td>Tx Power Out (dBW)</td>
<td>7</td>
<td>5.7</td>
</tr>
<tr>
<td>Tx Antenna Gain (dBic, including polarization loss)</td>
<td>-1</td>
<td>-8</td>
</tr>
<tr>
<td>Tx Other Loss (dB, e.g. pointing)</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Tx EIRP (dBW)</td>
<td>6</td>
<td>-2.3</td>
</tr>
<tr>
<td>Total Path Loss (dB)</td>
<td>-146.6</td>
<td>-145.9</td>
</tr>
<tr>
<td>Rx Antenna Gain (dBic, including polarization loss)</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>Manmade Noise (dB over thermal)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Rx Noise Temperature (dBK)</td>
<td>28.6</td>
<td>28.6</td>
</tr>
<tr>
<td>Rx G/T (dB)</td>
<td>-26.6</td>
<td>-26.6</td>
</tr>
<tr>
<td>Data Rate (kbps)</td>
<td>2.4</td>
<td>4.8</td>
</tr>
<tr>
<td>Rx Signal Power (dBW)</td>
<td>-138.6</td>
<td>-146.2</td>
</tr>
<tr>
<td>Rx Noise Power (dBW)</td>
<td>-166.2</td>
<td>-163.2</td>
</tr>
<tr>
<td>FEC Gain (dB)</td>
<td>0</td>
<td>2</td>
</tr>
<tr>
<td>Received Eb/N0 (dB)</td>
<td>26.6</td>
<td>18</td>
</tr>
<tr>
<td>Link Margin (dB)</td>
<td>16.3</td>
<td>7.7</td>
</tr>
<tr>
<td>Sensitivity (dBm)</td>
<td>-126.9</td>
<td>-125.9</td>
</tr>
</tbody>
</table>
Appendix C: Measured S11 Parameters of ULP Antennas

The measured S11 parameters for receive and transmit ULP antennas used in message testing are shown in Figure 60 and Figure 61, respectively. Figure 62 shows that there is no bandwidth degradation for a cable length of 0.5\( \lambda \) when the ULP antennas are connected by a complementary feed.
Figure 61- Transmit ULP antenna with 0.72 MHz bandwidth
Figure 62- No bandwidth degradation for receive and transmit ULP antennas with complementary feed where \( d = 0.5\lambda \).
Appendix D: Orbcomm Modems

Today, there are many Orbcomm modems on the market with varying levels of functionality. Functions like standby and sleep modes reduce power consumption when not transmitting or receiving; some modems have GPS while others do not. Table 12 samples some modems in each category and displays their power usage. The transmit power is typically 5 to 10 W and receive is negligible. This provides a power usage guideline for the duplexer.
### Table 12: Power usage of Orbcomm modems from various vendors

<table>
<thead>
<tr>
<th>Modem</th>
<th>Transmit (RF)</th>
<th>Receive (DC)</th>
<th>Standby (DC)</th>
<th>Sleep (DC)</th>
<th>GPS (RF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Digi m10¹</td>
<td>5 W max</td>
<td>960 mW max</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Stellar DS100</td>
<td>5 W max</td>
<td>924 mW nominal</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Stellar DS101</td>
<td>5 W max</td>
<td>540 mW nominal</td>
<td>-</td>
<td>60 uW</td>
<td>-</td>
</tr>
<tr>
<td>Quake DS300 (Stellar)</td>
<td>10 W max</td>
<td>1.02 W max</td>
<td>420 mW nominal</td>
<td>600 uW nominal</td>
<td>264 mW nominal, 300 mW max</td>
</tr>
<tr>
<td>Quake 1000</td>
<td>10 W max</td>
<td>840 mW</td>
<td>840 mW</td>
<td>60 mW max</td>
<td>N/A</td>
</tr>
<tr>
<td>Quake 4000</td>
<td>10 W max</td>
<td>840 mW nominal</td>
<td>420 mW nominal</td>
<td>300 uW nominal</td>
<td>156 mW nominal</td>
</tr>
<tr>
<td>OG2 Satellite Modem Orbcomm</td>
<td>24 W max</td>
<td>840 mW</td>
<td>120 uW</td>
<td>360 uW</td>
<td>420 mW</td>
</tr>
</tbody>
</table>

¹product discontinued on Sept. 30 2014  
Battery is considered 12 VDC  
Dash (-) means function is not available