ANTENNA PERFORMANCE ANALYSIS FOR THE NATIONWIDE DIFFERENTIAL GLOBAL POSITIONING SYSTEM

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Ian Matthew Barton
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This thesis entitled
ANTENNA PERFORMANCE ANALYSIS FOR THE NATIONWIDE
DIFFERENTIAL GLOBAL POSITIONING SYSTEM

by
IAN MATTHEW BARTON

has been approved
for the School of Electrical Engineering and Computer Science
and the Russ College of Engineering and Technology by

Chris G. Bartone
Associate Professor of Electrical Engineering and Computer Science

Dennis Irwin
Dean, Russ College of Engineering and Technology
As part of ongoing efforts to improve the Nationwide DGPS (NDGPS), an Antenna Performance Analysis (APA) is conducted by Ohio University for the U.S. Coast Guard (USCG) with the assistance from the U.S. Department of Transportation Volpe Center. The APA conducted has four main parts: 1) generation of generic measurement accuracy vs antenna performance mapping function using a multipath model; 2) collection of antenna anechoic chamber radiation patterns; 3) prediction of NDGPS installed multipath performance using the multipath model, a reflection model, and measured antenna performance data; and, 4) field data validation, for the predictions in step 3. Five commercially available GPS antennas were selected by the USCG for assessment. The various trade-offs in system performance are discussed in this thesis, including the gain uniformity across the coverage volume and the ability to calibrate timing errors inherent in the antenna. The antenna anechoic chamber measurement accuracy is also considered as part of this research effort.

Approved:

Chris G. Bartone

Associate Professor of Electrical Engineering and Computer Science
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<th>Definition</th>
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<tr>
<td>APA</td>
<td>Antenna Performance Analysis</td>
</tr>
<tr>
<td>AUT</td>
<td>Antenna Under Test</td>
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<tr>
<td>C/A</td>
<td>Course Acquisition</td>
</tr>
<tr>
<td>Copol/Xpol</td>
<td>Co-polarization / Cross Polarization</td>
</tr>
<tr>
<td>CMC</td>
<td>Code Minus Carrier</td>
</tr>
<tr>
<td>CNR</td>
<td>Carrier to Noise Ratio</td>
</tr>
<tr>
<td>dBi</td>
<td>dB isotropic</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DGPS</td>
<td>Differential GPS</td>
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<tr>
<td>D/U</td>
<td>Desired to Undesired Ratio</td>
</tr>
<tr>
<td>ESL</td>
<td>Electro-Science Laboratory at Ohio State University</td>
</tr>
<tr>
<td>FAA</td>
<td>Federal Aviation Administration</td>
</tr>
<tr>
<td>HZA</td>
<td>High Zenith Antenna</td>
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<tr>
<td>I</td>
<td>In-phase</td>
</tr>
<tr>
<td>LNA</td>
<td>Low Noise Amplifier</td>
</tr>
<tr>
<td>LAAS</td>
<td>Local Area Augmentation System</td>
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<tr>
<td>LHCP</td>
<td>Left Hand Circular Polarization</td>
</tr>
<tr>
<td>GPS</td>
<td>Global Positioning System</td>
</tr>
<tr>
<td>IMLA</td>
<td>Integrated Multipath Limiting Antenna</td>
</tr>
<tr>
<td>MLA</td>
<td>Multipath Limiting Antenna</td>
</tr>
<tr>
<td>NEC</td>
<td>Numerical Electromagnetics Code</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>NGS</td>
<td>National Geodetic Survey</td>
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<tr>
<td>NDGPS</td>
<td>Nationwide DGPS</td>
</tr>
<tr>
<td>PAC</td>
<td>Pulse Aperture Correlator</td>
</tr>
<tr>
<td>PTC</td>
<td>Positive Train Control</td>
</tr>
<tr>
<td>Q</td>
<td>Quadrature phase</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
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<tr>
<td>RHCP</td>
<td>Right Hand Circular Polarization</td>
</tr>
<tr>
<td>RMS</td>
<td>Root Mean Square</td>
</tr>
<tr>
<td>SGH</td>
<td>Standard Gain Horn</td>
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<tr>
<td>SMR</td>
<td>Signal to Multipath Ratio</td>
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<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
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<tr>
<td>SWR</td>
<td>Standing Wave Ratio</td>
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<tr>
<td>USCG</td>
<td>United States Coast Guard</td>
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<tr>
<td>WAAS</td>
<td>Wide Area Augmentation System</td>
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1 Introduction

Over the past decade, a number of differential GPS (DGPS) systems have been developed for use in the United States (US) by Government and commercial entities. The Federal Aviation Administration has developed the Wide Area Augmentation System (WAAS), (RTCA DO-229C 2001) and is developing the Local Area Augmentation System (LAAS), (RTCA DO-245 2001). The US Coast Guard (USCG) in cooperation with other Government agencies has developed the Nationwide DGPS (DOT 2001). Commercial DGPS systems include OmniSTAR (Fugro 2005) Virtual Base Station (VBS) and OmniSTAR HP and Starfire from Navcom (Navcom 2005), to mention a few. These Government and commercial DGPS systems dramatically increase the positioning and timing accuracy of GPS to enable such applications as landing aircraft, precision farming, mining, and land survey.

The USCG in cooperation with the US Department of Transportation and other Government Agencies are pursuing improvements to the maritime DGPS and NDGPS services to enable high accuracy. This effort is proceeding along with re-capitalization to replace aging equipment within the DGPS/NDGPS infrastructure. The overall goal of 10 to 20 cm for long baselines and < 5cm for short baselines (<40km) is sought (FHWA-RD-02-110). This system may provide new navigation capabilities for transportation applications for positive train control (PTC) (FRA 2002), automated highway systems, guidance for snowplows, and numerous other applications (Arnold 2001).
In order to achieve the anticipated high accuracies from the modernized NDGPS, a high performance GPS antenna should be selected for the ground-based reference stations. As part of the USCG efforts to re-capitalized the NDGPS system, an effort was pursued to replace the current DGPS/NDGPS reference station antennas with a high performance commercially available survey type antenna that would mitigate multipath.

Multipath remains one of the largest sources of error in DGPS systems and can contribute ten meters or more (Braasch et al 2001) to an error budget if not mitigated. Multipath mitigation approaches can be broken down into three main categories: radio-frequency (RF) approaches, receiver processing signal (at the correlation/signal tracking level) approaches, and a post-correlation processing approaches. The first category involves mitigating multipath using antenna techniques before the signal enters the receiver. The second category involves the correlator and bandwidth technologies, such as narrow correlator (Van Dierendonck et al 1992) and strobe correlator technology (Veitsel 1998). The third category involves processing the observed measurements to remove multipath, such as estimating the multipath from the signal-to-noise ratio (SNR) or incorporating measurements from multiple antennas (Yang 2004), or smoothing (Hatch 1982).

This thesis concentrates on multipath mitigation at the RF level through antenna performance characterization. In order to reduce multipath, most GPS antenna designs seek to limit the antenna gain in the lower hemisphere. Ground-based DGPS reference stations are consistently exposed to signal multipath from the ground, which arrives at the
antenna in the lower hemisphere of the antenna radiation patterns. The antenna pattern
would ideally provide uniform gain over positive elevation angles and have no gain in the
lower hemisphere (Braasch 1994).

Several multipath limiting antenna designs exist including ground planes, arrays, novel
printed circuit designs, or the use of multiple antennas to form an antenna system.
Ground plane designs include a metal sheet, corrugated designs such as the choke ring
(Tranquilla et al 1994), and exotic ground planes with varied impedance from the edge
(Rao 2000). Printed circuit board designs have included the spiral pinwheel type design
from NovAtel (Kunsyz 2000) and the shorted annular ring (Boccia 2001). Other designs
seeking to approximate the ideal pattern have using vertically stacked arrays, such as the
dB Systems Integrated Multipath Limiting Antenna (IMLA) (Thornberg et al 2003) and
the BAE SYSTEMS Model ARL-1500 (Lopez 2001), each developed as part of the FAA
LAAS program.

The IMLA is actually an antenna system consisting of two separate antennas: a high
zenith antenna (HZA) which covers from zenith to an elevation angle of 30 degrees and
the multipath limiting antenna (MLA) which receives from 35 degrees to approximately 2
degrees above the horizon (Thornberg et al 2003). The HZA is a Right Hand Circularly
Polarized (RHCP) antenna consisting of a cross-V dipole element with a reflector bowl
and surrounded by RF absorber, and a choke ring on the edges. The MLA is a vertically
polarized antenna consisting of a linear array of cylindrical dipoles. The vertical
polarization allows it to better reject ground multipath, since the vertical reflection is weaker due to the Brewster angle (Aloi 2000). The IMLA antenna system provides multipath rejection on the order of 35 to 40 dB (Thornberg et al 2003) but requires a non-standard setup with two RF inputs and two receivers.

The BAE Systems Model ARL-1500 is a single input antenna phased array design that can receive the RHCP signal. This design comes very close to approximating the ideal gain pattern in computer simulation, but it would require calibration of code delay variations on the order of 0.5 meters (Lopez 2001).

Most commercially available GPS antennas for ground reference stations consist of a single element with some type of ground plane or a printed circuit design, with the choke ring as the most popular. These designs typically have a peak gain at the zenith angle and taper off gradually towards the horizon. These antennas are mass-produced and therefore quicker to deploy than large-aperture phased array designs but have lower performance.

It should be noted that in general, multipath performance is only one consideration for an antenna. The performance of an antenna can be characterized in terms of many parameters including: gain performance for availability, group/phase delay performance, gain ripple, polarization, susceptibility to electromagnetic interference, sensitivity to environmental conditions (temperature, etc.), noise figure, bandwidth, standing wave ratio (SWR), cost, etc. Multipath mitigation was the primary focus of this thesis.
1.1 Scope

The goal of this APA was to determine which antennas out of five commercially available models that the USCG chose was most suitable for a ground reference station antenna. In order to do this, a multipath model was implemented to generate generic error curves showing the maximum ground multipath error that can occur for a given signal-to-multipath ratio (SMR) and satellite elevation angle and antenna height. Inherent assumptions in these curves included good antenna siting over a perfectly flat ground with no nearby obstacles that produce additional multipath. These curves show to what level of multipath mitigation the antenna system must have in order to stay below a certain performance (code or carrier) error level. Next, the multipath model was used to predict actual code and carrier errors versus satellite elevation angle for a given antenna height using antenna patterns measured at the Antenna Anechoic Chamber and a reflection model. Prior to the final measurements, the antenna test range was validated at the L band using antennas measured at other test ranges. An additional set of tests was performed to determine the feasibility of code and phase delay calibration. All of this information was considered in the final antenna selection.

This thesis is laid out in the following manner. First, the antenna performance criteria are discussed in detail in Chapter 2. In Chapter 3, a validation study of the Antenna Anechoic Chamber for the L Band is presented. Following that is a description of the data collection in the chamber including gain and phase patterns and group delay estimation is
presented in Chapter 4. Chapter 5 details the multipath error model including the reflection model, the receiver model, the generic performance mapping function, and the multipath error predictions for the antennas included in the analysis. In Chapter 6, the results from field testing that was performed in order to validate the model predictions is presented followed by some concluding remarks in Chapter 7 and some recommendations for future work in Chapter 8.
2 GPS Antenna Performance Criteria

The antenna performance criteria for this APA included: multipath rejection over all elevation angles, adequate gain with low variation over the coverage volume, and phase and group delays that can easily be calibrated. A suitable antenna would have a good balance in all of these criteria.

2.1 Ground Multipath Rejection

Multipath is a dominant error source in most DGPS architecture, especially when the reference station is ground based, as is the case for the NDGPS. Low frequency ground multipath can contribute 3 meters of the system error budget if not properly mitigated (Van Dierendonck et al 1992).

Ground reflection multipath can be mitigated, to a certain level, using a radio frequency (RF) approach whereby the desired (D) signal, at positive elevation angles, is passed to the GPS receiver with reasonable gain, and the undesired (U) signal, at negative elevation angles, is attenuated by the antenna pattern gain. The merit of this RF multipath mitigation technique can be generally characterized in terms of the D/U ratio of the antenna; for the purpose of this thesis, the D/U is calculated as the RHCP gain at positive elevation angles, in units of dB, subtracted from the RHCP gain at the corresponding negative elevation angle, in units of dB. (For this effort, a single-ground multipath bounce was assumed so the negative elevation angle in the D/U is exactly opposite (with the exception of the sign) as that for the positive elevation angle.)
From a GPS receiver perspective, often the multipath performance is characterized in terms of a fixed SMR ratio (Van Dierendonck et al 1992), (Braasch 1992), (Van Nee 1995). This SMR is at the point within the GPS receiver is just prior to the detector (i.e., code or carrier loop decision device). The D/U of an antenna effectively characterizes the antenna performance; however, additional factors may influence the final SMR presented to the GPS measurement decision devices (e.g., reflection, obstructions, additional signal processing techniques, etc.). Differences in the desired signal or undesired multipath wave characteristics will generally produce a D/U that is different from the SMR (Bartone et al 2005).

2.2 Phase Delay

Since an antenna’s received phase does not remain constant as a function of look angle, an antenna will induce relative phase delays on the accumulated Doppler measurements as a function of look angle for the reception of a GPS signal. For most high quality off-the-shelf GPS antennas produced for surveying and DPGS reference stations, phase variations are at the millimeter level and rarely exceed 1 cm. Phase delays may be calibrated out of the measurements (Mader 1999). Phase delay corrections for most commercial L1 and L2 survey type antennas may be obtained from the National Geodetic Survey (NGS) website (NGS 2005). For a DPGS network, repeatability at the manufacturing level is an important performance criterion in effective application of these calibrations. Such a study was beyond the scope of the work performed here.
2.3 Group Delay

Group delay is the measure of time delay that a narrowband signal packet or envelope undergoes (Haykin 1989). For GPS, the signals experience group delay from propagation through dispersive media such as the ionosphere, at the antenna, and in propagation through waveguides and the receiver front-end (Misra 2001), (Spilker 1977), (van Graas et al 2004). Antenna group delay is generally a function of look angle and polarization. If the antenna group delay is not calibrated at the receiver end, it may induce timing errors of several centimeters on pseudorange measurements (van Graas et al 2004). The scope of this work was not to provide calibration-level measurements, but to verify that the timing delays are well behaved and can be easily calibrated in future work.

2.4 Gain

All antennas in the APA were active antennas, i.e. they each contain a low noise amplifier that was powered by a DC voltage applied through the coaxial cable. The amplifier power source, usually 5-7 Volts, is most commonly contained in the receiver. The antenna gain should provide an adequate CNR to the receiver over the entire sky. Large gain variations over the coverage volume can cause the receiver to loose lock on the satellite or cause cross-correlation errors.

Fixed reception pattern GPS antenna patterns typically have a roll-off beyond 90 degrees from boresight in order to reduce reception in the lower hemisphere and reduce ground multipath interference. However, if the roll-off is too steep at positive elevation angles, it can effectively mask satellites that are low in the sky.
3 Anechoic Chamber Validation at L Band

3.1 Anechoic Chamber Details

Since the Ohio University Antenna Anechoic Chamber was a new facility, a validation study was performed in which pattern data collected at Ohio University were compared to patterns measured at other facilities. This ensures that the data collected at the chamber compares favorably with other antenna test ranges and can be trusted with a reasonable degree of certainty for antennas comparable to those in the APA.

The Ohio University Antenna Anechoic Chamber is a fully anechoic test chamber built by ETS-Lindgren with high performance RF shielding and extra-high performance pyramidal microwave absorber. The shielding isolates the chamber from outside interference and likewise stops any radiators in the chamber from becoming an interference source to the outside environment. The chamber’s shielding performance is specified to provide 100 dB of isolation from 200 kHz to 10 GHz, and it has been tested from 900 MHz to 3.0 GHz. However, two 12” x 12” access ports are currently open for the penetration of control and RF cables. In the future, all cabling entering the room will be filtered and the access ports will be sealed.

The walls, ceiling, and floor of the chamber are covered with microwave absorber, which reduces multipath reflections within the chamber, enhancing the measurement accuracy and validity. The performance of each absorber piece is listed in Table 3-1. The
rectangular chamber is 26 ft. long x 13 ft. wide x 13 ft. tall (interior shield-to-shield dimensions). Also included are some absorber pieces on the floor with a walking surface. The "walkable" sections have 12” thickness pyramidal absorber pieces within the walkable structure which are physically 18” high. A performance specification for the walkable surface is not listed by the manufacturer. The chamber is laid out with just enough walking surface to allow the operator to walk to the scanner without destroying the pyramidal absorber. Additionally, a small 2’x2’ section on the back wall, which is part of the door structure, has 8” absorber where the door handle is located.

<table>
<thead>
<tr>
<th>Location</th>
<th>Absorber Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>End Walls</td>
<td>18” Thickness (&lt;-40 dB Reflectivity from 1-2 GHz)</td>
</tr>
<tr>
<td>Side Walls &amp; Door</td>
<td>12” Thickness (&lt;-35 dB Reflectivity from 1-2 GHz)</td>
</tr>
<tr>
<td>Ceiling</td>
<td>12” Thickness (&lt;-40 dB Reflectivity from 1-2 GHz)</td>
</tr>
<tr>
<td>Floor</td>
<td>18” Thickness (Unknown Reflectivity from 1-2 GHz)</td>
</tr>
<tr>
<td>Walkway (Floor)</td>
<td>8 Pieces of 12” Walkable (Unknown Reflectivity)</td>
</tr>
<tr>
<td>Corners</td>
<td>Lossy Block (Unknown Performance)</td>
</tr>
</tbody>
</table>

### 3.2 Scanner Details

Inside the chamber is a hybrid near-field antenna test system built by Antcom, as shown in Figure 3-1. This system is capable of planar scans (76 x 76 inch) for high gain narrow beam antennas, spherical scans (360° x 360°) for low gain broad beam antennas, and cylindrical scans (360° x 76 inch) for fan beam antennas and linear arrays. The system can take measurements in the near-field of electrically-large antennas and compute the far-field pattern or it can operate in a far-field mode.
The system consists of four main components: the scanner (6x6x6 feet), the motion control subsystem, a network analyzer, and a high-performance personal computer. The seven-axis scanner positions the antenna being tested and the probe in space. The motion control subsystem precisely moves the scanner by controlling the servo motors on the scanner. The network analyzer performs the measurements by sending a signal to the probe and measuring the test antenna’s response. The computer interfaces with the control subsystem with a control card, controls the network analyzer, and records the measurements.

The principle components of the scanner are shown in Figure 3-1. The waveguide probe in the rear of the chamber functions as a linearly polarized radiation source for pattern
measurements. It can be moved from side to side, up and down, as well as rotated. In front of the probe is the mounting for the antenna-under-test (AUT) on the dielectric tower. The AUT can also be rotated, moved front to back, and turned about a vertical axis with the arm below. In addition, the rotation center of the AUT can be adjusted for maximally flat phase. With these seven axes of motion, the scanner can trace out three different scanning surfaces: planar, cylindrical, and spherical as shown in Figure 3-2.

![Antenna Pattern Scanning Surfaces](image)

Figure 3-2: Antenna Pattern Scanning Surfaces (Antcom 2003)

Broad-beam antennas such as a GPS antenna are usually measured in the spherical mode in order to capture all of the radiated energy.

The measurement system has a specific coordinate system as described from the perspective shown in Figure 3-1. The motion control platform has seven axes of motion
with four linear drives and three rotational drives. Three of the linear drives correspond with the three rectangular axes: X, Y, Z. The X-axis originates from the center and extends to the right side of the room, with positive orientation in the right direction. The Y axis is vertical with up in the positive sense. The Z axis extends from the back to the front of the room, and the whole arm below the AUT can move up and down this axis. The first rotational drive is behind the probe, allowing the operator to change the probe polarization. The second rotational drive is behind the AUT, also allowing the rotation of the AUT and controlling the phi angle on a spherical measurement. The third drive rotates the whole arm below the AUT, controlling the theta angle on a spherical measurement. The last linear drive controls the linear distance from the AUT to the center of rotation for the arm, thus allowing the phase center to be aligned to center of rotation.

The measurement surfaces shown in Figure 3-2 can be realized as follows. The planar scan surface is accomplished by moving the probe in the X-Y plane, i.e. in the plane of wall behind the probe as shown in Figure 3-1. A cylindrical scan is realized by scanning along the X axis, which forms the axis of the cylinder, (i.e. across the room shown in Figure 3-1) and turning the AUT with the AUT rotation motor to turn around the cylinder. The spherical mode is realized by turning the arm about the Y axis for a polar cut and rotating the AUT around for a different track around the sphere.
3.3 Network Analyzer Details

The network analyzer is a measurement device that measures the parameters of an RF two-port network. For antenna measurements, the AUT and the probe are connected to separate ports. The test range utilizes an Agilent 8753ES S-parameter vector network analyzer, located in the control room, which is capable of measurement from 30 kHz to 6 GHz (Agilent 2001). For antenna pattern measurement, the forward transmission parameter is of primary interest because it contains a relative power level and a phase term. The reflection term is of interest because it can help the user determine the resonant bandwidth of the antenna in the form of a SWR.

3.4 Measurement Methods

The measurements were performed in a spherical near-field mode in increments of 10 degrees in azimuth and 1 degree in theta. The spherical scan mode is ideal for GPS antennas, since they have a wide beamwidth. The probe used in the measurements was an open-ended rectangular waveguide probe model number WR650, which is usable in the 1.12 to 1.7 GHz range. The antennas were measured in a near-field mode even though the AUT and the probe were separated by a distance of 70 inches.

All data was collected in a RHCP method. Since the probe is linearly polarized, this is really a software option. The software collects data for one linear polarization and then the orthogonal polarization. After the program finishes scanning in the first linear polarization, it rotates the probe by 90 degrees and repeats the scan for the orthogonal polarization. The conversion to circular polarization is a complex addition with a 90
degree phase shift on one of the polarizations to remove the phase shift from the probe rotation.

The data were collected in a co-polarization/cross-polarization (Copol/Xpol) mode rather than an $E_\theta/E_\varphi$ mode. For the Copol/Xpol mode, the probe was rotated the same azimuth angle as the AUT for each cut in the scan. This keeps the same polarization at boresight for each cut. In contrast, in an $E_\theta/E_\varphi$ mode, the probe remains stationary during each linear scan.

For gain calculations, a gain-transfer method was utilized in the scanner software. A standard gain horn (SGH) was scanned in the same manner as the AUT. The gain horn measurement was used to convert the relative power measurements from the AUT scan to an absolute gain measurement by a substitution (Balanis 1989):

$$G_{AUT} = G_{SGH} + P_{AUT} - P_{SGH}$$  \hspace{1cm} (3.1)

where: $G_{AUT}$ is the gain of the AUT

$G_{SGH}$ is the gain of the SGH

$P_{AUT}$ is the power measurement of the AUT

$P_{SGH}$ is the power measurement of the SGH

3.5 Chamber Validation Results

Five different broad-beam L-band antennas were used as part of the validation. The antennas were selected because they resonate at GPS frequencies, are broad-beam,
pattern data from another source was readily available, and the antennas were on-hand at Ohio University. Two of the antennas were NovAtel “pinwheel” type antennas, the GPS-600 and the GPS-702. The next two were dB Systems High Zenith Antennas (HZA) developed as part of the IMLA for the FAA LAAS program. The last antenna is a monopole tuned at GPS L1 with a two foot ground plane.

3.5.1 NovAtel Pinwheel

3.5.1.1 NovAtel GPS-600

In Figure 3-3, pattern data on the NovAtel GPS-600 collected at Ohio University are plotted with all azimuths against pattern data from the manufacturer. NovAtel data was provided courtesy of NovAtel (Kunysz 2003) in digit format. Note that data from NovAtel is from a single cut and does not account for variation in azimuth angle. This 360 degree cut is folded in half and plotted over 180 degree span to show two traces: each trace is one half of the elevation cut. The patterns show reasonable agreement, but the Ohio University data shows considerable more variation at below 20 dB and in the LHCP patterns. Much of the variation can be attributed to reflections in the chamber, where the main beam is pointed at the back wall or away from the probe. The variation in the pattern is a combination of chamber effects and the actual pattern variation against azimuth. Note that the measured gain was 31 dB isotropic (dBi) which is within the published gain range of 33.5 ± 3 dBi.
Pattern data for L2 is plotted in Figure 3-4 with similar results. Chamber effects are again noticeable at gain levels below -20 dB. Also, the LHCP pattern data appears to be less accurate than the RHCP data. Absolute gain was not measured in the L2 test.
3.5.1.2 **NovAtel GPS-702**

The GPS-702 model was also measured and compared to manufacturer data as shown in Figures 3-5 to 3-8. Data was provided for the GPS-702 in polar plot format (Kunysz 2003) in color. In Figure 3-5, cuts from all azimuths are plotted in a polar format with RHCP in red and LHCP in blue for comparison. These data compare favorably with the manufacturer’s published patterns shown in Figure 3-6.
In Figure 3-7, the measured L2 relative gain patterns are plotted. These also compare favorably with the manufacturer’s pattern shown in Figure 3-8.
Figure 3-7: GPS 702 Pattern Measured at Ohio University at L2

Figure 3-8: GPS 702 Manufacturer’s Pattern at L2 (Kunysz 2003)
3.5.2  dB Systems High Zenith Antennas (HZA)

3.5.2.1  200-HZA

Next, pattern data for dB Systems HZA antennas are shown. The 200-HZA is a heli-antenna design. The data from dB Systems was provided as part of an acceptance package (dB Systems 1999). In Figure 3-9, measured RHCP patterns are plotted against the manufacturer’s data. In the figure, two orthogonal cuts are shown plotted instead of a single cut. Again, the pattern data are a close match, but some chamber effects are evident around the back lobes where the low front-to-back ratio makes it difficult to separate the actual gain from the back wall reflection.

Figure 3-9: dB Systems 200-HZA Pattern
3.5.2.2 200A-HZA

The next plots show a different HZA version which is a cross-V dipole design, the dB Systems 200A-HZA. It is also an RHCP antenna, but the plots show linear polarization data. In Figure 3-10, a relative gain pattern in the vertical polarization is shown. In Figure 3-11, the horizontal relative gain pattern is plotted. The results are similar for both measurements, but the backlobes are much higher than the manufacturer’s pattern. This is most likely due to the back-wall reflections, and could be minimized with by averaging two patterns taken at two separation distances that differ by \( \frac{1}{4} \) wavelength.

![Figure 3-10: dB Systems 200A-HZA, Vertical Polarization Cut](image_url)
3.5.3  **L1 Monopole with Two Foot Ground Plane**

A monopole with a two foot diameter ground plane was simulated in Win-NEC Pro by Ohio University Research Engineer Joseph Kelly and built by Ohio University Research Engineer Curtis Cohenour under an FAA research project. In Figure 3-12, the simulated pattern was plotted over the measured far-field pattern. Pattern differences do exist, but they mainly involve the depth of nulls and the strength of lobes. The simulated pattern always has some simulation error that is not representative of real antenna. In addition, the chamber data will reflect the imperfections in the antenna construction as well as effects from the chamber.
3.6 Conclusions of Chamber Validation

As a result of this validation study, it was apparent that the chamber effects were noticeable in the measurements on the antenna backlobes. In order to mitigate these effects, ¼ wavelength averaging was employed in the final APA measurements to average out the chamber effects in the pattern (Hansen 1988). Additionally, antenna specific consideration to accuracy in the backlobes (and sidelobes) of the AUT should be investigated in a final data test. Near-field effects were found to be negligible for antennas for the type in the assessment.
4 Chamber Measurements for the Test Antennas

This thesis describes an APA to choose the best performing antenna among five commercially produced L1/L2 survey-type antennas. For this performance analysis, the USCG provided five antennas shown in Figure 4-1: Ashtech Geodetic III L1/L2 GPS antenna, Ashtech Choke Ring L1/L2 GPS antenna, Leica Geosystems AT503 L1/L2 Choke Ring GPS antenna, Trimble Zephyr Geodetic L1/L2 GPS antenna, and the NovAtel GPS 702 Pinwheel antenna. The actual antenna models are not linked to the results in this document, and will instead be noted as A, B, C, D, and E in no particular order. The approach to selecting the performance criteria are linked to the major error sources in ground-based DGPS systems that are affected by antenna performance: weak signal, multipath, group delays, phase delays, as discussed in Chapter 2. However, antenna bandwidth performance was not included in the assessment.
Figure 4-1: Included Antennas (from top then left to right): Leica Geosystems AT-503 with Radome Placed to the Left of the Antenna, Ashtech Choke Ring, Trimble Zephyr Geodetic, NovAtel GPS-702, Ashtech Geodetic III

4.1 Gain & Phase Measurements

4.1.1 Gain & Phase Measurement Procedures

In order to reduce chamber effects, the scanner and all access ports were covered with absorber as shown in Figure 4-2 and the entire walking surface was removed for improvements of 1 or 2 dB on the backlobes where the measurement is 25 to 30 dB down from the peak.
4.1.1.1 AUT Mounting

For this test, the AUT rotation motor was removed to obtain more accurate measurements on the back lobes of the antenna, as shown in Figure 4-2. Instead, a 1.25 inch thick piece of acrylic was bolted to the dielectric tower in place of the AUT rotation motor. This mount has a 5/8” x 11tpi tapped hole for inserting a threaded rod at a height of 18.5” above original motor mount.

4.1.1.2 Hardware Setup

Hardware connections are shown in Figure 4-3. The probe is connected to port 1 of the network analyzer with a ferrite RF isolator. On port 2, the AUT was connected with a
bias-t to apply filtered DC power to the built in low-noise amplifier (LNA). An isolator was connected after the bias-T to reduce standing waves.

![Figure 4-3: Hardware Test Setup](image)

4.1.1.3 Scan Procedure

The scan mode was the *Eth/Ephi spherical direct far-field* with linear polarization selected. The following resolution was used.

- 2 cuts or phi points (The same cut was repeated).
- 361 points per cut or theta points (1 deg increments)

In this mode, the linear probe was held stationary while the AUT was rotated. The arm sweeps around 360 degrees for each elevation cut. For the second scan, the probe was rotated by 90 degrees to give the orthogonal linear polarization. The scan was repeated to give to measure the repeatability of the measurement.
In order to give reasonable L band far-field measurements, the separation distance was increased to a maximum of 85 inches for the first scan and shortened by a quarter wavelength for subsequent scans. This translates to an average separation of 11.2 and 8.7 wavelengths at L1 and L2, respectively.

The separation distance was changed by a quarter wavelength with a manual command to the software, and the scan was repeated. The four scans were averaged to reduce measurement noise and chamber effects. Finally, the scans were repeated with the standard gain horn in order to do a gain substitution.

**Power considerations:** The power output was increased to a maximum of 10 dBm. A bias-T was added to power the LNA on active antennas. Isolators are added at both the probe and the AUT ports.

**Bias Considerations:** All antennas were biased 12 volts, which was within the manufacturer’s recommended voltage range for every antenna.
**Phase Center calibration:** The phase centers for the antennas were approximated by the NGS phase calibration data and then calibrated for maximum phase flatness using the iterative software routine for phase center calibration.

![Phase Pattern](image)

*Figure 4-4: Phase Pattern after Phase Center Calibration*

An example of a calibrated phase pattern is shown in Figure 4-4. Note that the 90 degree phase difference between the Eph and Eth patterns is caused by the rotation of the probe.

4.1.2 **Measured Patterns**

4.1.2.1 **Polarization**

Antenna pattern data were collected with a linear polarized waveguide probe for two orthogonal directions (i.e., theta and phi coordinates). A simple 90 degree rotation of the probe changed the polarization to be aligned with the theta or phi direction. In order to
realize RHCP patterns, the linear polarization responses were added with a 90 degree phase shift on one of the polarizations to remove the phase shift caused by the probe rotation. (A 180 degree phase shift on the opposite polarization results in circular polarization response in the opposite sense, i.e. left hand circular polarization (LHCP)).

The relations are shown in Equations 4.1 and 4.2:

\[
E_{RHCP} = \frac{(E_{\theta} + jE_{\phi})}{\sqrt{2}} \text{ (Volts)} \quad (4.1)
\]

\[
E_{LHCP} = \frac{(E_{\theta} - jE_{\phi})}{\sqrt{2}} \text{ (Volts)} \quad (4.2)
\]

where \( E_{RHCP} \) is the complex RHCP antenna response in Volts

\( E_{\theta} \) is the complex vertical polarization response in Volts

\( E_{\phi} \) is the complex horizontal polarization response in Volts.

Figure 4-4 shows the relative gain patterns for circular and linear polarizations for Antenna E. The antenna pattern data plotted were collected in two linear polarizations: \( \theta \) and \( \phi \), which are labeled Eth and Eph, respectively. The RHCP curve was calculated using Equation 4.1 and the LHCP curve was calculated using Equation 4.2. Note that the RHCP pattern is 3 dB higher than the linear polarization curve at 0 degrees and the LHCP curve is 24 dB down from the linear polarization curves.
4.1.2.2 RHCP Patterns

The normalized measured gain patterns at the L1 and L2 bands for antennas A, B, C, D, and E are shown in Figures 4-6 and 4-6, respectively. Note that a single azimuth cut is shown, but all the antennas are highly symmetrical in azimuth. The absolute measured gains are shown in Table 4-1.
For the L1 patterns in Figure 4-5, there is one grouping with slow roll-off consisting of antennas A, B, and E. Antenna E has lower backlobes than A and B, though. The other grouping has a relatively steep roll-off consists of antennas C and D. Antenna C’s rolloff of about 17 dB at the horizon is relatively steep for a reference station antenna and could significantly affect reception for low elevation satellites.
RHCP Amplitude Patterns L2

Figure 4-7: RHCP Normalized Gain Patterns at L2

Table 4.1: Absolute Gain of GPS Antenna Tested Including Antenna Pre-amplifier Gain

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Antenna L1 Gain (dBi)</th>
<th>L2 Gain (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>46.5 dB</td>
<td>46.5 dB</td>
</tr>
<tr>
<td>B</td>
<td>42 dB</td>
<td>44.3 dB</td>
</tr>
<tr>
<td>C</td>
<td>36 dB</td>
<td>34.1 dB</td>
</tr>
<tr>
<td>D</td>
<td>33.5 dB</td>
<td>32.6 dB</td>
</tr>
<tr>
<td>E</td>
<td>54.1 dB</td>
<td>53.8 dB</td>
</tr>
</tbody>
</table>

4.1.2.3 Desired-to-Undesired Ratio

The D/U ratios were computed from the chamber gain patterns (using the data via Figure 4-6) at L1 is shown in Figure 4-8. The plot field of view was reduced from 180 degrees to 90 degrees to show only one azimuth for ease of comparison. Antenna C has the best D/U ratio at L1, followed by antenna D, and antenna E. However, antenna C’s high D/U
comes with the price of relatively low reception gain at low elevation angles, see Figure 4-6.

![Figure 4-8: D/U Ratio for Antennas Tested at L1](image)

The D/U ratio at L2 is shown in Figure 4-8. Again, with respect to D/U, antenna C performs the best; with antenna D second best, followed by antenna E.
The D/U ratio of an antenna provides an indication of the antenna’s multipath rejection, but it does not include factors such as the polarization change or attenuation of multipath reflections. As part of the APA, the measured D/U data, presented in Figures 4-7 and 4-8 are further investigated with respect to the reflections effects from the ground. The next section will present these details which add fidelity to the multipath model predictions.

4.1.3 Data Accuracy

4.1.3.1 Gain Accuracy

Antenna pattern measurements contain error particularly in measuring the back beams of the antenna. In a high gain part of the pattern, errors are dominated by the accuracy of
the network analyzer’s capability. However, this is normally limited to a fraction of a dB for the power ranges used in the test (i.e. 0 to -25 dB at the beam peak), as shown in Figure 4-10. This measurement uncertainty is statistically derived (Agilent 2001) from a calculated root mean squared value, but the manufacturer describes this measurement uncertainty as a “worst case,” not a specific statistical relation. If the measurement error is assumed to have a Gaussian distribution, then the final measurements which are derived from an average of four independent measurements would have uncertainties exactly half of what is shown in Figure 4-10. In order to maintain reasonable measurement accuracy, care was taken to keep the transmission coefficient above -60 dB during the measurements. The transmission coefficient is affected by several factors including the cable losses, the gain of the probe, the signal spreading loss, and the gain of the AUT. For L1, spreading losses were 43 dB and the combination of cable losses and gain of the probe were 12 to 13 dB, which resulted in a total link loss of 55 dB. Losses for L2 were a few dB less than L1. The differences in the peak transmission coefficient for each AUT were primarily due to differences in the LNA gains. If for example an AUT has an integrated amplifier gain of 35 dB, then the net transmission coefficient peaked at -20 dB. An antenna with a higher gain LNA resulted in measurements with a higher transmission coefficient, which resulted in antenna patterns with lower measurement noise.
In low gain areas of the pattern where the main beam of the antenna is pointed at the back wall, chamber effects can dominate the error effects since the measurement system does not have time-gating capability. These effects can be accurately modeled as a single reflection from the back wall of the chamber. The error from chamber effects can be bounded using equations for constructive and destructive interference for monochromatic coherent wave interference, which result in the plot shown in Figure 4-10.
The upper curve in Figure 4-11 is the maximum gain error resulting from a constructively interfering reflection for a given relative power ratio for the reflected wave. The lower curve shows the maximum gain error resulting from a destructively interfering wave. For a reflection at 0 dB down from the direct ray, there can be full cancellation of the signal or a 3 dB gain constructive gain.

The absorber manufacturer guarantees a reflected signal level of 35 dB or lower in the L Band at normal incidence for the 18 inch absorber on the back wall. The reflected signal will also incur an additional 6 dB spreading loss, since it must travel twice as far as the
direct ray. This puts the reflected signal level at a 41 dB loss from the direct signal at the AUT location.

In order to complete the link budget, the actual collected pattern must be used to determine the signal ratios for a particular AUT. The antenna gain at the measurement angle must be compared to the measured gain at the angle facing the back wall of the chamber. This front-to-back gain ratio added to the reflection loss of 41 dB determines the relative reflection power level. Equation 4.3 summarizes this relative reflection level prediction:

\[
R/D = G(\theta + 180) - G(\theta) + RPL \quad (4.3)
\]

where

- \( R/D \) is the relative reflection power (dB)
- \( G \) is the measured pattern gain (dB)
- \( RPL \) is reflection propagation loss (i.e. 41 dB)

The results of this error assessment are shown in Figure 4-12 for Antenna E and Figure 4-13 for Antenna D.
Figure 4-12: Accuracy of Antenna E Pattern vs. Angle

Figure 4-13: Accuracy of Antenna D Pattern vs. Angle
From Figures 4-12 and 4-13 it can be observed that when the main beam is pointed towards the probe (i.e. away from the back wall) the measurement accuracy is on the order of 0.1 dB or less. As expected, the error begins to grow rapidly around the back side of the antenna (e.g. from 135 to 180 degrees). Recall that the patterns were taken at two points in the chamber separated by a quarter wavelength. The quarter wavelength difference also provides an indication of the maximum chamber effects. The difference of the patterns shown in Figures 4-14 and 4-15 are comparable, for antennas D and E, respectively.

Figure 4-14: Measured Gain Pattern Variation for Antenna E for $\lambda/4$ Separation Change
Since all of the antenna types tested had a fairly broad beamwidth, the backwall was the dominate error source. As the sharpness of the antenna pattern begins to increase (e.g., narrow beam width, or very sharp roll-off), the multipath reflection off the side walls, ceiling, and floor could be more significant.

### 4.1.3.2 Phase Accuracy

Similarly, the maximum phase error due to chamber effects can also be estimated. For maximum phase error, the reflected wave is assumed to be 90 degrees out of the phase with the direct wave. A curve was generated for maximum phase error due to a single
reflected wave as plotted in Figure 4-16. Note that the constructive and destructive interference curves are symmetrical.

Figure 4-16: Maximum Phase Error vs. Relative Reflection Power Level
Predicted uncertainties are shown in Figures 4-17 and 4-19 and the pattern difference measurements shown in Figures 4-17 and 4-19 are on the same order of magnitude, for antennas E and D, respectively.

![Phase Accuracy Assessment](image)

Figure 4-17: Accuracy of Antenna E Phase Pattern vs. Angle
Figure 4-18: Antenna E Phase Pattern Difference
Figure 4-19: Accuracy of Antenna D Phase Pattern vs. Angle
4.2 Group Delay

4.2.1 Test Setup

An additional set of tests was completed in order to assess the antenna group delay performance in the chamber. The measurements should not be interpreted as calibration grade. Instead, they were completed in order to look for large or abnormal behavior in the antenna group delay performance. In order to measure group delay, phase measurements must be taken at multiple frequency points with a stable measurement device. The network analyzer was stabilized with a 10 MHz rubidium frequency source, which have typically short-term stabilities on the order of $10^{-12}$ over a 100 second
period (Symmetricom 2003); the rubidium frequency source was used to add long-term repeatability to the measurements.

The test setup was similar to the gain and phase tests, with the following changes. Walking surface was put back in the chamber from the door to the scanner for ease of entry. In addition, the scan setup was changed to 5 degree increments in theta over a range of 180 degrees with 21 frequency points with a 1 MHz spacing.

Each azimuth was scanned twice by rotating the AUT by 180 degrees in azimuth and repeating the scan. This procedure was repeated for the orthogonal plane with the antenna by turning the AUT by 90 degrees, which makes a total of four 180 degree polar cuts.

### 4.2.2 Processing

Since the raw phase patterns were linear polarization, the data was transformed to circular polarization. Next, the phase of patterns was unwrapped across frequency at every look angle to span beyond 360 degrees. The slope of the phase with respect to frequency was approximated using a discrete difference at every look angle with an assumed constant group delay over the band, according to Equation (4.3):

\[
\tau_g \approx \frac{c}{N-1} \sum_{n=1}^{N-1} \left( \frac{\phi_{n+1} - \phi_n}{\omega_{n+1} - \omega_n} \right) \quad \text{(m)} \quad (4.3)
\]

where \(N\) is the number of discrete frequency points

c is the speed of light in free space (m/s)
$\phi_n$ is the unwrapped phase measurement at the nth frequency point (rad)

$\omega_n$ is the angular frequency at the nth frequency point (rad/sec).

With the stated typical RMS phase noise of the Agilent 8753ES network analyzer of 0.038 degrees (Agilent 2001), the 1-$\sigma$ value for a group delay measurement can be calculated under the assumption of Gaussian measurement error and independent frequency measurements, which are differenced to obtain a group delay using Equation (4.3); this calculated group delay measurement noise was estimated to be 4.5 cm 1-$\sigma$. By averaging over 21 frequency points, the group delay measurement error can be reduced to about 1 cm.

The absolute group delay was not measured because the bias of the measurement system has not been determined. Instead the variation of group delay with respect to look angle was examined. For each measurement cut, the group delay variation is determined by taking the difference of the group delay with respect to the group delay at boresight. It is this variation that causes relative timing differences in between pseudorange measurements, since the antenna will receive signals from satellites at different regions of the sky.

Each of the four group delay variation measurements are plotted for Antenna D in Figure 4-21. The plot legend indicates the position of the antenna connector when viewed from the behind the antenna. For instance, the “left” position means that the connector was in
the nine o’clock position and the “right” position indicates that the connector was in the
three o’clock position. Ideally, the left and right curves as well as the up and down
curves would be identical, since they are measurements at the same look angle. A
significant level of error exists in the measurements, which is a combined effect of clock
noise and chamber effects. Chamber effects appear to have a large effect near 80 and 90
degrees, since chamber effects cause uncertainty as high as ±0.5 degrees at the theta
angle of 90 degrees. For more on chamber effects, see the section 4.3.

Figure 4-21: Raw Group Delay Difference Measurements for Antenna D
After the group delay differences were processed, the up/down and left/right azimuths were averaged to reduce the measurement error. Next, these two azimuth cuts were averaged, as shown in Figure 4-22.

![Antenna Group Delay Variation Azimuth Averaging](image)

**Figure 4-22: Antenna Group Delay Variation Azimuth Averaging**

Without adding the fidelity of an azimuth specific group delay variation curve, the azimuth performance was averaged. For the final average, each side of the graph was averaged together. The results of the final averaging are shown for antennas D and E in Figure 4-23. From this figure, it appears that both antennas have well-behaved group delay variation that can be calibrated out of the pseudorange measurements.
4.2.3 Measurement Validation

In order to validate the measurements, two key factors were examined: repeatability and a match with other chamber group delay measurements performed at another facility. First, to illustrate repeatability, the difference of two group delay measurements performed with the same setup is shown in Figure 4-24. The measurements were made back-to-back. Certainly, a more rigorous approach would difference many more than just two measurements. However, in the interest of time and because the measurements were not performed for calibration purposes, this measurement does suffice. The 1-σ value for this measurement is on the 1 cm level, which matches the prediction in the previous section.
Another test was completed to access the severity of chamber reflection effects on the group delay measurement. A group delay measurement was completed at two separation distances a quarter wavelength apart. The results of these measurements are plotted in Figure 4-25. The measurements have similar trends, but noticeable differences of 5 to 10 cm are apparent. Chamber effects may be partially compensated for by mapping the chamber multipath for calibration grade measurements.
A comparison of data from a set of group delay tests performed at The Ohio State University Electro-Science Lab (ESL) on antenna D was done. The ESL uses time-gating to drastically reduce chamber effects. A plot of the average group delay across azimuth is shown in Figure 4-26.
The two data sets appear to be a close match when averaged across azimuth. Note that many more azimuth cuts were done at the OSU test than the measurements for the APA and the measurements were done at 1 degree increments instead of 5 degrees.

4.3 Conclusions of Chamber Measurements for Test Antennas

Based on D/U, Antenna C appeared to have the best multipath rejection followed by antennas D and E. However, antenna C had a high gain roll-off above the horizon, which can result in poor tracking at low elevation angles.

Gain and phase patterns errors were dominated primarily by chamber reflections in the backlobes. D/U curves calculated based on the antenna pattern measurements were also
dominated by the same errors. These errors can be modeled and nearly bounded using a model of a single reflection from the back wall.

Antennas D and E had acceptable group delay variation on the same order of magnitude, but the measurements were not calibration grade. Calibration-grade group and phase delay measurements are possible with current network analyzer, but chamber effects must first be addressed.
5 Multipath Error Modeling

5.1 Reflection Model

The propagation model is a first-order ray-based flat earth model. The model geometry is illustrated in Figure 5-1. The wave is modeled as a uniform plane wave where the multipath signal was modeled as a single specular reflection from a flat and level ground. This reflection will lead to an attenuation, change in polarization, and path delay for the multipath signal (Aloi 1999). The specular reflections are modeled using Fresnel reflection coefficients based on the electromagnetic properties of the ground: electric permittivity, magnetic permeability, and conductivity (Balanis 1989).
In the classical reflection coefficient formulation, polarization is defined with respect to the plane of incidence. The plane of incidence is parallel to the incident ray and normal to the reflecting surface. Referring to the geometry shown in Figure 5-1, the plane of incidence is the surface of the paper. The perpendicular component is defined as into the page, and the parallel component is directed towards the upper left-hand corner for the direct ray. For the reflected ray, the parallel component is defined as pointing towards the upper right-hand corner. The parallel and perpendicular components relate to the $\theta$ and $\phi$ antenna polarizations, respectively.

It should also be noted that typically the reflections at boundaries are characterized with respect to the angle of incidence. The angle of incidence is the angle from “normal” (i.e., perpendicular) to the boundary surface. Thus, for a typical ground based GPS receiver, an angle of incidence of 0 deg corresponds to an elevation of 90 deg, and an angle of incidence of 90 deg corresponds to an elevation angle of 0 deg.

Reflection coefficients are calculated based on equations for oblique incidence (Balanis 1989). The parallel and perpendicular reflection coefficient equations are given in Equations (5.1) and (5.2), respectively:

\[
\Gamma_{\text{parallel}} = \frac{-\eta_{\text{air}} \cos(\theta_i) + \eta_{\text{ground}} \cos(\theta_i)}{\eta_{\text{air}} \sin(\theta_i) + \eta_{\text{ground}} \cos(\theta_i)} \quad (5.1)
\]

\[
\Gamma_{\text{perpendicular}} = \frac{\eta_{\text{ground}} \cos(\theta_i) - \eta_{\text{air}} \cos(\theta_i)}{\eta_{\text{ground}} \sin(\theta_i) + \eta_{\text{air}} \cos(\theta_i)} \quad (5.2)
\]
where \( \eta_{\text{air}} \approx \sqrt{\frac{\mu_0}{\varepsilon_0}} \approx 377[\Omega] \) is the intrinsic impedance of air [\( \Omega \)]

\[
\eta_{\text{ground}} = \sqrt{\frac{j\omega \mu_0 \mu_r}{\sigma_{\text{ground}} + j\omega \varepsilon_0 \varepsilon_r}}
\]
is the intrinsic impedance of the ground [\( \Omega \)].

\[
\cos(\theta_i) = \sqrt{1 - \sin^2(\theta_i)}
\]
is the cosine of the transmission angle [unitless]

\[
\sin(\theta_i) = \frac{\gamma_{\text{air}}}{\gamma_{\text{ground}}} \sin(\theta_i)
\]
is the sine of the transmission angle [unitless]

\[
\gamma_{\text{air}} = j\frac{2\pi}{\lambda}
\]
is complex propagation constant of air [Np/m + j rad/m]

\[
\lambda_{\text{ground}} = \sqrt{j\omega \mu_0 \mu_r (\sigma_{\text{ground}} + j\omega \varepsilon_0 \varepsilon_r)}
\]
is the complex propagation constant of the ground [Np/m + j rad/m]

Note that reflection coefficients must be calculated as complex numbers, which can be converted to an amplitude and phase. The reflection coefficients for both parallel and perpendicular polarizations under “wet earth” (\( \varepsilon_r = 10, \mu_r = 1, \sigma = 1.2\text{e-2} \text{[Mho/m]} \)) and “dry earth” (\( \varepsilon_r = 6.2, \mu_r = 1, \sigma = 5.5\text{e-4} \text{[Mho/m]} \)) are plotted in Figures 5-2 and 5-3, respectively; these are plotted with respect to the angle of incidence. Both conditions have a clear Brewster angle for the parallel polarization case at: 10 degrees in elevation for wet ground, and 22 degrees in elevation for dry ground. The parallel polarization component goes to zero and undergoes a 180 degree phase shift. For an incoming RHCP wave, reflections at elevation angles below the Brewster angle will be right-hand elliptically polarized, and reflections above the Brewster angle will be left-hand elliptically polarized. From a signal level perspective (independent of time delay considerations) this does imply that the performance of the antenna at low elevation
angles is the most critical, more so than its performance at higher elevation angles.

(Time delay considerations will be discussed in the next section.)

Figure 5-2: Reflection Coefficients for Wet Ground
The analytical reflection coefficient factors can be combined with the measured antenna pattern data from the anechoic chamber data to provide a Combined Signal-to-Multipath Ratio (SMR). First the complex desired and undesired signals are calculated in volts in Equations (5.3) and (5.4), respectively:

\[
D(\alpha) = \frac{1}{\sqrt{2}} \left( 10^{G_{\text{par}}(\pi/2-\theta)/20} e^{j\Phi_{\text{par}}(\pi/2-\theta)} + j10^{G_{\text{perp}}(\pi/2-\theta)/20} e^{j\Phi_{\text{perp}}(\pi/2-\theta)} \right)
\]

\[
U(\alpha) = \frac{1}{\sqrt{2}} \left( 10^{G_{\text{par}}(\theta-\pi/2)/20} \Gamma_{\text{par}} e^{j\Phi_{\text{par}}(\theta-\pi/2)} - j10^{G_{\text{perp}}(\theta-\pi/2)/20} \Gamma_{\text{perp}} e^{j\Phi_{\text{perp}}(\theta-\pi/2)} \right)
\]

where \( D(\alpha) \) is the desired signal [volts]
\( U(\alpha) \) is the undesired signal [volts]

\( \alpha = \pi / 2 - \theta \) [rad] is the elevation angle

\( G_{\text{par}}(\theta) \) is the parallel polarization antenna gain [dB]

\( G_{\text{perp}}(\theta) \) is the perpendicular polarization antenna gain [dB]

\( \Phi_{\text{par}}(\theta) \) is the parallel polarization antenna phase [rad]

\( \Phi_{\text{perp}}(\theta) \) is the perpendicular polarization antenna phase [rad]

Equations (5.3) and (5.4) can be converted to a decibel scale as in Equations (5.4) and (5.5):

\[
S(\alpha) = 20 \log_{10} |D(\alpha)| \quad \text{[dB]} \quad (5.4)
\]

\[
M(\alpha) = 20 \log_{10} |U(\alpha)| \quad \text{[dB]} \quad (5.5)
\]

The SMR is calculated in decibels in Equation (5.6):

\[
\text{SMR}(\alpha) = S(\alpha) - M(\alpha) \quad \text{[dB]} \quad (5.6)
\]

Results for the wet ground condition at the L1 band are shown in Figure 5-4 for the five antennas tested. For simplicity, antenna pattern variations in azimuth were not considered since the antenna patterns were nearly uniform in azimuth.
Figure 5-4: Combined SMR Wet Ground for Antennas Tested at L1 Band

Figure 5-5: Combined SMR Wet Ground for Antennas Tested at L2 Band
The time delay of a ground-bounce wave at an antenna can be simply models as a function of the antenna height above the boundary and the elevation angle to the far-field source (i.e., SV) as shown in Equation (5.7).

\[ T_d = 2h \sin(\theta_{el}) \quad [m] \quad (5.7) \]

where: 

- \( T_d \) is the time delay, [m]
- \( h \) is the height of the antenna reference point, [m]
- \( \theta_{el} \) is the satellite elevation angle

Figure 5-6 shows the ground bounce path delay in units of ns normalized to one meter in antenna height.
5.2 Receiver Model

A receiver multipath model is an effective tool to predict code and carrier phase error (Van Nee 1995). A multipath error model was used to deterministically simulate code error and carrier error for a narrow-correlator receiver based on the SMR, time delay, and correlator spacing (Dickman et al 2003). The direct (i.e., D) and multipath signals (i.e., U) are modeled in Equation (5.8) as:

\[ D = R(\tau) \quad \text{and} \quad M = aR(\tau - \delta)e^{j\varphi} \quad (5.8) \]
Where D is the direct signal, \( R(\tau) \) is the autocorrelation function, \( \alpha \) is the inverse of the SMR, \( \delta \) is the delay, and \( \phi \) is the phase difference between the direct and multipath signals as calculated from the time delay.

Carrier phase error was calculated from Equation (5.9):

\[
\theta_c = \arctan\left( \frac{\alpha R(\tau - \delta)\sin(\phi)}{R(\tau) + \alpha R(\tau - \delta)\cos(\phi)} \right)
\]  
(5.9)

After determining the carrier phase, the in-phase (I) and quadrature (Q) signal components are modeled by multiplying the received signal by sine and cosine replicas of the carrier. Early and late versions are formed using a bit shift register, and a code phase dot product discriminator was used to determine the multipath induced code error, as shown in Equation (5.10):

\[
D = (I_E - I_L)I_p + (Q_E - Q_L)Q_p
\]
(5.10)

With an infinite SMR, the discriminator output should be zero. A code loop low pass filter was applied to remove noise in the modeled code output.
5.3 \textit{Generic Performance Mapping Functions}

Using the multipath receiver error modeling, a series of curves were generated to illustrate what level of SMR would be necessary to meet a particular multipath error performance level under different conditions. (A “narrow correlator” (d=0.1) was used and a perfect electric conductor is assumed for these generic performance mapping functions.) A NDGPS antenna height of 40 feet (12.192 m), which is one of highest antenna heights, was selected to illustrate a worst case scenario. The predicted C/A code error is plotted in Figure 5.7. The predicted P-code error performance is plotted in Figure 5.8 (Bartone et al 2005).

![Figure 5-7: Predicted C/A Code Multipath Performance for Antenna Height of 40 feet (Bartone et al 2005).](image)
Figure 5-8: Predicted P Code Multipath Error Performance for Antenna Height of 40 feet (Bartone et al 2005)

Figure 5-9: Predicted Carrier Phase Multipath Error for Antenna Height of 40 feet (Bartone et al 2005)
The data presented in Figures 5-7 through 5-9 represent a minimum SMR (i.e., multipath mitigation) necessary to meet a measurement performance error specification for a given antenna height. Data for the code performance in Figures 5-7 and 5-8 indicate that for low elevations SVs, while the D/U may not be very good, the end result is little code tracking error because of the short time delay between the D and U signals; however, as the elevation angle increases slightly the tracking errors can rise rapidly for low D/U ratios. Also from Figures 5-7 and 5-8 one can conclude, depending upon siting, that if the back lobes are not controlled, this condition can produce large tracking errors since this is the case where the maximum time delay is present.

From Figure 5-9, carrier phase errors are dominated by SMR, not by the path delays. This does not reflect errors caused by drops in signal power in the nulls created by the multipath interference pattern. A worst case fading loss can be modeled as coherent monochromatic wave interference in the fully destructive case. Equation (5.11) represents the worst case fading with a direct signal D and an undesired interference signal U:

\[ T_v \cos(\omega t + \varphi) = D_v \cos(\omega t) + U_v \cos(\omega t + \pi(2n + 1)) = D_v \cos(\omega t) - U_v \cos(\omega t) \]  

(5.11)

where \( T_v \) is the total signal amplitude (Volts)

\( D_v \) is the direct signal amplitude (Volts)

\( U_v \) is the direct signal amplitude (Volts)

\( \varphi \) is phase difference between the two signals (radians)

\( \omega \) is the angular frequency (radians/sec).
If the Equation 5.12 is normalized to the direct signal by dividing by the direct signal and converted to dB, the equation simplifies to:

$$\text{Fading Loss} = 20 \log\left(\frac{V}{V_D}\right) = 20 \log(1 - \alpha) \quad (5.12)$$

Where $\alpha = 10^{-\frac{\text{SMR}}{20}}$

The worst-case signal loss due to multipath fading is plotted against SMR in Figure 5-10. This fading loss, at the carrier tracking loop, may produce cycle slip and thereby introduce additional tracking error.

![Figure 5-10: Carrier-to-Noise Ratio Loss from Signal Fading](image-url)
5.4 Prediction of Installed Performance

The SMR calculated from the antenna anechoic chamber data and the multipath reflection model were used to predict multipath errors versus satellite elevation for a typical NDGPS installed antenna height. Results for the antenna D and antenna E are plotted in Figures 5-11 and 5-12 for an antenna height of 20 feet (typical NDGPS GPS antenna height). From Figures 5-13, the C/A code error peaks at around 0.9 meters for antenna D at an elevation angle of about 8 degrees, and around 1 meter at an elevation angle of 50 degrees for antenna E for the 20 feet high installation over wet ground. (The large error for antenna E at an elevation angle of 90 deg is discounted due to the assessment of measurement accuracy in the chamber as predicted in Figure 4-11.) As shown in Figure 5-12, above 30 degrees in elevation, the carrier phase error is predicted at below 7mm and 19mm for antenna D and E, respectively.
Figure 5-11: Predicted C/A Code Errors at L1 for Antenna Height of 20 feet

Figure 5-12: Predicted Carrier Phase Errors at L1 for Antenna Height of 20 feet
In addition, multipath errors were predicted for the L2 band at the same installation height of 20 feet over wet ground. Results for the P code multipath error and carrier phase error are shown in Figures 5-13 and 5-14, respectively. From Figure 5-13, the P code multipath error peaks from about 10 to 15 degrees at about 0.5 meters and drops with higher elevation angles. (The large error around 80 to 90 degrees is again discounted due to the large measurement error.) Antenna E also has a peak error of about 0.5 meters, but the error stays at this level from 20 to 75 degrees. From Figure 5-14, the carrier phase error is limited above 30 degrees to about 3 mm for antenna D and 1 cm for antenna E.

![Pseudorange multipath error vs. elevation for L2](image)

Figure 5-13: Predicted P Code Errors at L2 for Antenna Height of 20 feet
Figure 5-14: Predicted Carrier Phase Errors at L2 for Antenna Height of 20 feet
6 Field Measurements

6.1 Site Selection

Careful site selection was important when using a first-order reflection model to avoid obstacles such as buildings, trees, and fences. A large flat and level reflecting surface at was necessary to represent the ground multipath model. Therefore, a beach on a relatively large lake was the best choice. From satellite photos (Terraserver 2005), Alum Creek Lake near Westerville, Ohio had an excellent site for the test. Figures 6-1 and 6-2 show photos of the test performed on the beach at Alum Creek State Park on March 11th, 2005 from 10 o’clock to 2 o’clock in the afternoon.

Figure 6.1: Field Test Site at Alum Creek State Park
6.2 Equipment and Procedures

6.2.1 Receiver

The receiver used for the test was a NovAtel OEM4 Propak G2 with a special firmware build, 2200S31, to allow the selection of a narrow correlator rather than the pulse aperture correlator (PAC). A special command was issued to the receiver in the software in order to select the narrow correlator.

The receiver was configured to log for range and time at a 1 Hz rate and the ephemeris and WAAS data when new data became available.
6.2.2 Mobile Test Setup

In order to set up a field test at a remote location, a mobile test setup was developed. This consisted of a deep-cycle battery connected to a power inverter, a laptop with the QNX operating system and a special prototype version of the LAAS code for data logging, and the receiver. All of the equipment was housed in the gray plastic flight case shown in Figure 6-1.

The antenna was mounted on a tripod with a tribrach for leveling purposes. In addition an aluminum extension rod added some additional height and separation from the tribrach. The antenna was set to a height of about 58.5 inches above the water surface.

6.3 Data Processing

6.3.1 CodeMinusCarrier Analysis

CodeMinusCarrier (CMC) analysis exposes multipath and group delay error terms (Misra 2001). The code and carrier measurements can be expressed in the following equations.

\[ p(t) = r(t) + c \cdot t_b(t) + I(t) + T(t) + \xi_p(t), \quad (6.1) \]

\[ \Phi(t) = r(t) + c \cdot t_b(t) - I(t) + T(t) + \lambda N + \xi_\Phi(t), \quad (6.2) \]

where \( p(t) \) is the code measurement [m],

\( r(t) \) is the geometric range [m],

\( c \) is the speed of light [m/s],
\( t_b(t) \) is the combined clock bias term of the receiver and the satellite (sec),

\( I(t) \) is the ionospheric error term [m],

\( T(t) \) is the tropospheric error term [m],

\( \xi_p(t) \) is an additional code measurement error term [m],

\( N \) is integer ambiguity [unitless],

\( \lambda \) is the signal wavelength [m]

\( \xi_\phi(t) \) is an the additional carrier measurement error term [m].

When the code measurement is subtracted from the carrier measurement, the clock bias and troposphere error terms are eliminated. This is illustrated in Equation (6.3).

\[
CMC = p(t) - \Phi(t) = 2*I(t) + \xi_p(t) - \xi_\phi(t) \quad (m) \tag{6.3}
\]

A polynomial fit can effectively eliminate the ionosphere error term in post-processing. By subtracting the fitted polynomial, the residual error terms are exposed. These are dominated by multipath error, group delay variation, and phase delay errors.

\[
CMC_{\text{residual}} = CMC - P(t) \approx \xi_p - \xi_\phi \quad (m) \tag{6.4}
\]

where \( P(t) \) is the fitted polynomial

Filtering is also applied to the residual to reduce receiver noise in the measurement.
6.3.2 Carrier-to-Noise Ratio Analysis

Signal fading from multipath is predicted based on path delays and predicted SMR (Hannah 2000) as shown Figure 5.10. These are computed based on a simple wave interference model in Equation (6.5):

\[ \Delta CNR = 20 \log_{10} \left( 1 + \frac{SMR^{-1} \cos(\phi)}{1} \right) \]  

(6.5)

where \( \phi \) is the path phase difference of the multipath signal and the SMR is unitless voltage ratio. Fading can be detected in the CNR data by applying a polynomial fit and examining the residual.

6.4 Results & Discussion

Since the receiver was on the west shore of the lake, the analysis was restricted to azimuths in the eastern half of the sky. For each antenna four data items will be shown: 1) sky plot for the data collection interval, 2) For two GPS SVs chosen for investigation, an estimated SMR vs measurement epoch, in units of seconds, for the antenna using the measured anechoic chamber data, 3) CMC time and frequency domain plot, and 4) CNR time and frequency domain plot.

For the CMC and CNR time and frequency domain plots, four subplots are shown within each figure. The top two subplots are from the prediction model (using the measured antenna pattern data, the multipath and reflection model components); the bottom two subplots are from the field measurement data; the left two subplots are in the time domain; and the right two subplots are in the frequency domain.
The specular reflection at the Earth surface was modeled using Fresnel reflection coefficients, as presented in Equations 5.1 and 5.2, based on the electromagnetic properties of fresh water ($\varepsilon_r = 79.8, \mu_r = 1, \sigma = 2.5e^{-3} [\text{Mho/m}]), (Aloi 1999). The data analysis attempts to match measured CmC and CNR data with CMC and CNR predictions from the model.

### 6.4.1 Antenna D Field Results

Data were collected on March 11, 2005 from approximately 10:00 am to 12:00 pm local time. The sky plot for the field data collected on antenna D is shown in Figure 6-3. PRN numbers are placed at the beginning of the SV track in the sky. Again, only SVs due east (azimuth angles 0 to 180 deg) were considered for data analysis due to the site location. Satellite tracks with minimal azimuth variation were chosen for the for the fastest changing multipath. Since SVs 24 and 28 had minimal azimuth variation, as compared to the other SVs, they were chosen for further investigation.
The estimated SMR variation over time, using data from the anechoic chamber tests and SV elevation angles are shown in Figure 6-4 for PRN 24.
The CmC from the field data measurements were compared to model predictions in Figure 6-5. Fading frequencies for the data are in the mHz range. The field data matches
reasonable well with the predictions from the model, except the field data has an additional low frequency component. In general, non-agreement between the model and measured data can results from the limitations of the simple ground multipath model. The model implemented only considers first-order specular reflections from a flat and level reflecting surface. Secondary reflections off of trees or land masses or man-made structures as well as non-flat earth or obstructions can produce additional higher order terms that appear in the data. These higher order terms can add (constructively and destructively) to produce some disagreements in the comparison. Additionally, antenna group and phase delays are not compensated for in this model. Overall, if these disagreements are assessed to be minimal or insignificant, good confidence can be obtained in model predictions.

Figure 6-6: CNR Fading Analysis for PRN 24 for Antenna D
In Figure 6-6, CNR fading is shown in the model (top subplots) and the field data (bottom subplots) in both the time (left subplots) and frequency (right subplots) domains. The CNR data matches the field data very closely, illustrating the dominance of the multipath from the lake.

Similar analysis is shown for PRN 24 in Figures 6-7 to 6-9. The CMC plots in Figure 6-8 match at the predicted fading frequency, but additional low frequency components dominate, possibly from secondary reflections that were not modeled or from antenna group delay variation that was not captured in the chamber tests. The CNR plots show a strong match between the model data and the field data.

Figure 6-7: Predicted SMR and SV Elevation Angle vs. Time for PRN 28 for Antenna D
Figure 6-8: CMC Analysis for PRN 28 for Antenna D

Figure 6-9: CNR Analysis for PRN 28 for Antenna D
6.4.2 Antenna E Field Results

Another data collection was performed on March 11, 2005 from approximately 12:00 pm to 2:00 pm local time with antenna E at the same site immediately after the first set. The tri-pod shown in Figure 6-1 was used in the same location. The GPS sky plot for the field data set is shown in Figure 6-10.

![Sky Plot March 11 2005](image)

Figure 6-10: Sky Plot for Antenna E Field Test

SVs due east (azimuth angles 0 to 180 deg) were considered for data analysis due to the site location. Since SVs 4 and 7 had minimal azimuth variation, as compared to the other SVs, they were chosen for further investigation. The analysis for PRN 4 is shown, beginning with the SMR, from the measured anechoic chamber data and elevation angles in Figure 6-11. The resulting error magnitude from a CMC analysis in Figure 6-12 is
lower than predicted, while the frequency error component is reasonably close; this could indicate a slight mis-model in the reflection model. At high elevation angles, the reflection area is closer to the ground surrounding the antenna, and may include effects from the shoreline and the tripod. The resulting CNR analysis in Figure 6-13 provides similar results.

Figure 6-11: Predicted SMR and Elevation Angle vs. Time for PRN 4 for Antenna E
Figure 6-12: CMC Analysis for PRN 4 for Antenna E

Figure 6-13: CNR Analysis for PRN 4 for Antenna E
The same analysis was performed for PRN 7, which had a much lower elevation profile. The predicted SMR and elevation profile are shown in Figure 6-14, with elevation angles ranging from 16 to 8 deg in elevation. The results of the CMC analysis shown in Figure 6-15 indicate a closer match for PRN7, although there are some additional low frequency terms. The CNR analysis in Figure 6-16 is a better match, particularly in the frequency domain.

Figure 6-14: Predicted SMR and SV Elevation Angle vs. Time for PRN 7 for Antenna E
Figure 6-15: CMC Analysis for PRN 7 for Antenna E

Figure 6-16: CNR Analysis for PRN 7 for Antenna E
7 Conclusions

An antenna performance analysis was performed on five different commercially available GPS antennas for consideration in the NDGPS architecture where high performance is being pursued. A multipath error model was used to generate performance error curves to estimate the minimum SMR needed to mitigate ground multipath errors to a specific system error performance level. Antenna pattern measurements were performed at the Ohio University Antenna Anechoic Chamber to characterize the antennas radiation characteristics. A SMR, using the measured antenna patterns were combined with a propagation model for wet and dry soil to predict performance for typical NDGPS installation heights. Results from these efforts indicated that while antenna C had the best multipath rejection, its gain at low elevation angles was sufficiently lower than other antennas. Antenna D and Antenna E were found to have good multipath rejection with adequate gain at low elevations.

In order to validate the prediction model implemented, a field data collection effort was performed at Alum Creek State Park in Ohio on a single day with antennas D and E. A code-minus-carrier analysis and signal fading analysis were performed and compared to model data. For the single ground bounce model, reasonably good agreement was obtained with the field test data, which confirmed that antenna D had lower multipath error than antenna E.
It should be noted that no final recommendation on which antenna should be selected for NDGPS fielding due to the many additional factors specific (i.e., susceptibility to electromagnetic interference, sensitivity to environmental conditions, noise figure, cost, etc.) in the introduction to this thesis. A conclusion was drawn based on the multipath mitigation performance of these antennas.
8 Recommendations for Future Work

Additional data collection at an actual NDGPS site would be useful to analyze the actual system performance for Antenna D and E. This would help check for unanticipated effects that were not included in the APA, such as antenna behavior in a high-power RF environment.

Additional group and phase delay calibration in a high fidelity method for group and phase calibration implementation should be performed prior to actual use of the final antenna selected. Methods for completing group and phase delay calibrations are documented in a 2004 Institute of Navigation paper (van Graas et al 2004).

Additional data collection should be performed over bandwidth. A simple SWR ratio measurement over frequency would suffice. No formal tests of the SWR were given to determine the antenna bandwidth and out-of-band rejection.
References


National Geodetic Survey website. “GPS Antenna Calibration.”


