A Low-Multipath Wideband GPS Antenna With Cutoff or Non-Cutoff Corrugated Ground Plane

Francesca Sciré Scappuzzo (1, 2)

(1) Physical Sciences Inc., Andover, MA 01810, USA
(2) Eidgenössische Technische Hochschule (ETH), 8092 Zürich, Switzerland fss@psicorp.com

and

Sergey N. Makarov (3)

(3) Worcester Polytechnic Institute, Worcester, MA 01609, USA
makarov@wpi.edu

Accepted for publication: 25 Sep 2008
Paper No. AP0711-1113.R1

IEEE Transactions on Antennas and Propagation

Via Manuscript Central Website

October 2008
A Low-Multipath Wideband GPS Antenna With Cutoff or Non-Cutoff Corrugated Ground Plane

Francesca Sciré Scappuzzo(1,2)

(1) Physical Sciences Inc., Andover, MA 01810, USA
(2) Eidgenössische Technische Hochschule (ETH), 8092 Zürich, Switzerland fss@psicorp.com

and

Sergey N. Makarov(3)

(3) Worcester Polytechnic Institute, Worcester, MA 01609, USA
makarov@wpi.edu

Abstract

We compare the performance of the classic GPS choke ring ground plane with a new shallower design. Both the choke ring and the new ground plane are here designed to operate uniformly between 1.15 GHz and 1.60 GHz while maintaining the required low-multipath performance in the whole bandwidth. To achieve reception in this wide range of frequencies, the radiating element chosen is a droopy bowtie turnstile.

A classic choke ring is composed of deep concentric rings on a flat circular metal ground plane. We have modeled the choke ring as a metal corrugated surface of depth \(d\) such that \(\lambda/4 \leq d \leq \lambda/2\), operating at cutoff: the multipath suppression (i.e., the proper RHCP/LHCP pattern shaping) is obtained by eliminating the surface wave on the ground plane in a certain frequency band. When shallower concentric rings of depth \(d \leq \lambda/4\) are used, the corrugated surface operates at non-cutoff: the required multipath rejection can be achieved by cancellation (destructive interference) rather than suppression of surface waves. The comparisons performed in this study include theory, numerical simulations, and hardware tests. Our results show that both configurations, when properly optimized, are good candidates for reception of modernized GPS, GLONASS, and GALILEO satellite signals from GPS permanent stations.

Keywords: Global Positioning System, Multipath Interference, Receiving Antenna, Choke Ring Antennas, Jamming, Wideband Antennas, Multifrequency Antennas.

I. Introduction

Novel applications of GPS technology require innovations such as time-dependent, real-time, and near-real-time positioning at data acquisition times of only a few seconds. This fast observation rate enables the analysis of many dynamic phenomena that are often encountered in Earth and atmospheric sciences. For instance, in the new field of GPS seismology, the capability to detect and track the rapid propagation of seismic waves through entire continents with networks of GPS permanent stations with receivers acquiring data at 1 Hz rate (high-rate GPS), will provide new insights into earthquake processes and volcanic eruptions [1] [2] [3] [4] [5] [6] [7] [8] [9].

A major source of error in high-accuracy GPS is the interference of multiple reflections with the direct GPS signal, known as multipath [10] [11] [12] [13]. A variety of techniques (e.g. the narrow correlator spacing, multipath estimating delay lock loop, and strobe techniques) have been developed to mitigate multipath errors. For long delay multipath, the receiver itself can recognize the delayed (reflected) GPS signal and discard it. However, shorter delay multipath from signals reflected by the ground or other
nearby reflectors is harder to filter because it interferes with the direct GPS signal, causing effects almost indistinguishable from routine fluctuations in atmospheric delay. For observations made over 24 hours, multipath effects can sometimes be removed in post processing by averaging for networks of GPS reference stations [14]. Techniques using Signal-to-Noise Ratio information for carrier phase multipath mitigation have also been used in post-processing mode [15]. However, for time dependent measurements performed over few seconds (quasi-real-time) or in real-time (i.e. without post processing), an effective way to eliminate short delay multipath is to reject multiple reflections at reception using a low-multipath receiving antenna. One possible design solution is a GPS receiving antenna equipped with a ground plane designed to eliminate or reduce the effects of multiple reflections from the ground or nearby objects (e.g. trees, buildings, or metal structures). Alternative methodologies for the mitigation of multipath at reception include the design of GPS antenna arrays [16] [17].

GPS choke ring antennas are the state-of-the-art low-multipath antennas and they are used for GPS geodetic permanent networks employed in geodynamic studies [18]. These antennas are presently used for narrow band (24 MHz) dual-frequency operation and, in particular, work better at $L_2=1227.60$ MHz than at $L_1=1575.42$ MHz. However, choke ring antennas still have unsurpassed performance and they have been the multipath mitigation antennas of choice for dual frequency GPS geodetic applications for over 20 years.

Recent investigations were carried out to develop new dual-frequency designs for high-accuracy GPS [19] [20] [21], including antenna array solutions [22] [23] [24].

With the future introduction of Modernized GPS and GALILEO by year 2012, several more frequencies and signals will be available that will improve present high-accuracy GPS capabilities. According to information on frequency allocations available to date, Modernized GPS will have $L_1$, $L_2$, and an additional third frequency, $L_5=1176.45$ MHz, with $20$ MHz bandwidth; Galileo will feature 5 frequencies: $E_1=1589.742$ MHz ($4$ MHz bandwidth), $E_2=1561.098$ MHz ($4$ MHz bandwidth), $E_5a=1176.45$ MHz ($=L_5$), $E_5b=1207.14$ MHz (with $24$ MHz bandwidth for the whole $E_5$), and $E_6=1278.75$ MHz ($40$ MHz bandwidth). Future GNSS (Global Navigation Satellite Systems) high accuracy antennas will need to receive L-band signals within a large band, between $1.15$ GHz and $1.60$ GHz. The present dual frequency choke ring design is inadequate for future high accuracy applications.

There are some recent studies on novel triple-frequency GPS antenna designs that allow reception of $L_1$, $L_2$, and $L_5$ for Modernized GPS [25] [26] [27] [28] [29].

In the present study, the authors carry out extensive numerical analysis in order to

a) expand the band and performance of classic choke ring antennas from dual-frequency to wideband
b) develop a novel wideband ground plane, shallower than the choke ring design

both for Modernized GPS and GALILEO applications.

a) Using a basic electromagnetic theoretical analysis, the principles of choke ring ground planes operation are here explained using the theory of metal corrugated surfaces. Choke ring ground planes block the propagation of plane waves and surface waves along the ground plane, thus reducing side lobes and back lobes, preventing multipath at certain specific frequencies (operation at cutoff). However, these deep corrugated designs (with corrugation depth $d$~61 mm and ring diameter $D$~380 mm) are inevitably massive, heavy, and relatively expensive to manufacture. Other choke ring modifications are compared in the literature [30] [31] [32] [33] [34] [35] [36] [37] [38] [39] [40] [41] [42] [43] [44] [45] [46] [47] [48].
In this study we have optimized the choke ring performance to operate uniformly over the band from 1.15 to 1.60 GHz, with a slightly deeper design, with corrugation depth $d\approx65$ mm.

b) We also introduce a new wideband low-multipath high-accuracy GPS antenna that features a non-cutoff frequency selective surface (FSS) ground plane design. This non-cutoff ground plane is a shallow corrugated surface, with corrugation depth $d\approx25$ mm and ring diameter $D\approx340$ mm, optimized to control the propagation of surface waves, rather than suppressing them (as the choke ring antennas do at cutoff). The non-cutoff ground plane is intended to operate in such a way that the Line-Of-Sight (LOS) signal from the antenna located above the non-cutoff FSS ground plane and the surface waves propagating along the ground plane cancel each other out at the rim. This is achieved by tuning the surface wave propagating on the ground plane to be a certain angle out of phase with the LOS signal at the edge of the surface, causing destructive interference and eliminating the unwanted signals for low elevation angles (below $5^\circ$ elevation angle). Signals from low elevation angles are mainly electromagnetic noise, such as multipath, or interference sources, which tend to cause jamming. Because of its non-cutoff operation, this ground plane is smaller than a conventional choke ring, especially in its depth, allowing a lighter and cheaper design. From our study it emerged that this design is more robust and might require cheaper materials and less precise manufacturing processes.

II. Theory of Metal Corrugations for Antenna Ground Planes

In this section we summarize the basic theory of metal corrugated surfaces as it relates to the choke ring design. Based on this theory, in Section V we will introduce the new wideband, shallower, low-multipath ground plane.

A. Plane waves and metal corrugations of variable depth

A planar corrugated surface of exactly resonant depth, $d=\lambda/4$, shown in Fig. 1 blocks the propagation of an oblique plane wave whose wave vector is perpendicular to the corrugation independently from the direction of the electric field. In order to demonstrate this property, we compare the performance of a solid metal surface (Fig. 1a) with a corrugated metal surface obtained combining multiple quarter-wave, parallel-plate resonators (Fig. 1b). There are two types of TEM plane waves, labeled as type I and type II in Fig. 1, that could potentially propagate along both surfaces. Oblique plane waves of type I are often designated as a TE to $z$ or, in short, TE$_z$ plane waves. Similarly, the oblique plane waves of type II are the TM to $z$ or TM$_z$ plane waves.

The metal surface shorts out type I waves since $E_t=0$, but still allows the propagation of type II. Similarly, for a corrugated surface, the metal teeth require $E_t=0$ at top and still short out the type I very efficiently. At the same time, the waveguide openings require $H_z=0$ at their top (according to the elementary theory of a quarter-wave resonator) and will short out the signal II, for which it must be $H_z\neq0$. Thus, neither type I waves nor type II waves can propagate along the resonant corrugation. Such reasoning was used in a classic reference [49] devoted to artificial soft and hard surfaces. One condition for it to be valid is the geometry inequality [49]: $g+t<\lambda/2$, where $g$ is the width of the trough and $t$ is the width of the tooth. This inequality implies that the parallel-plate waveguides cannot be very wide. The definition of a soft surface given in [49] implies that, for two plane waves propagating along the corrugated surface and perpendicular to the corrugation, one has boundary conditions in the form $E_z(z)=0$ and $H_x(z)=0$. These conditions are again equivalent to shorting out type I waves and type II waves. Thus, a quarter wave corrugated surface is a soft surface.

1 These surfaces are introduced by analogy with acoustics: the soft (absorbing) surface means zero excess pressure (velocity potential is zero - Dirichlet boundary condition); the hard surface means zero acoustic velocity (gradient of velocity potential is zero - Neumann boundary condition).
Since any oblique plane wave with wave vector perpendicular to the corrugations is a combination of type I and II waves (TE\(_z\) and TM\(_z\) modes), we conclude that a planar corrugated surface with corrugation depth \(d\) exactly quarter wavelength blocks out any oblique plane wave whose wave vector is perpendicular to corrugation, independent of the direction of the electric field. In that sense, it becomes a polarization-independent soft surface, according to terminology adopted in Ref. [49].

To extend this result to any corrugation depth, we formalize the problem as shown in Fig. 2: the approach is to replace the entire corrugated surface with one boundary condition that involves surface impedance. The surface impedance relates the external fields \(E_x\) and \(H_y\) in the upper half space to each other. The average surface impedance of metal corrugation \(Z_S\) with infinitesimally small teeth can be written in the form

\[
Z_S = Z_0 \frac{Z_{\text{bottom}} + jZ_0 \tan kd}{Z_0 + jZ_{\text{bottom}} \tan kd}
\]

(1)

where \(Z_0 = 377\Omega\) is the impedance of free space, \(k = 2\pi/\lambda\) is the wave number, and \(d\) is the corrugation depth, with \(Z_{\text{bottom}}\) being zero for metal bottom corrugation, in which case one has \(Z_S = jZ_0 \tan kd\). To account for finite, but a relatively small tooth width, \(t\), one may write (see Ref. [50])

\[
Z_S = jZ_0 \frac{g \tan kd}{g + t}
\]

(2)

An oblique plane wave of type I or, which is the same, a TE\(_z\) plane wave of Fig. 2, cannot exist for the same reason as given above, for any corrugation depth. Nor can the oblique plane wave of type II (or TM\(_z\)), since it does not have a \(E\)-field component \(E_x\) in the propagation direction, which is required by the corresponding boundary condition when the surface impedance is different from zero.

**B. Canonic surface wave solution for planar metal corrugation**

We investigate the propagation of type III waves on the corrugated surface shown in Fig. 2. The potential solution has an \(E\)-field component in the direction of propagation and it is, therefore, a TM to \(x\) wave, and simultaneously TM to \(z\) wave. Such a wave is expected to decay far from the surface: this means that the solution is becoming a surface non-leaky wave. The surface wave is created by any non-trivial value of the surface impedance, \(Z_s\). We can express the boundary condition for this wave in the form

\[
Z_S = \frac{E_x(z = 0)}{H_y(z = 0)}
\]

(3)

It is important to remember that \(E_x\) and \(H_y\) denote the field components of the external surface wave field in the opening of the parallel plate resonator, not the inner field in the resonators.

The parallel plate resonators have all been replaced by the corresponding boundary condition on the top of corrugation. This is a reasonable approximation commonly used in practice. (If necessary, the exact mathematical analysis of the external field coupled to all modes in the inner corrugation waveguide can be performed using the so-called space harmonic approach — see Ref. [51] for a planar corrugation, or Ref. [52] for a cylindrical corrugation). Following Harrington [53], we seek a solution for the surface wave of type III in the form
\[ E_x = E_{x0} \exp(-j k_x x - \alpha z) \]
\[ E_z = E_{z0} \exp(-j k_x x - \alpha z) \]
\[ H_y = H_{y0} \exp(-j k_x x - \alpha z) \quad (4) \]

with an \textit{a priori} unknown positive decay factor \( \alpha \). From Maxwell’s equations the curl of the \( H \) field is related to the \( E \) field as:

\[ \nabla \times \vec{H} = j \omega \varepsilon \vec{E}, \quad \nabla \times \vec{H} = \left[ -\frac{\partial H_y}{\partial z}, 0, \frac{\partial H_y}{\partial x} \right] = [\alpha, 0, -j k_x] H_{y0} \exp(-j k_x x - \alpha z) \quad (5) \]

That yields

\[ E_{x0} = \frac{\alpha}{j \omega \varepsilon} H_{y0} \quad E_{z0} = -\frac{k_x}{\omega \varepsilon} H_{y0} \quad (6) \]

The next step is to substitute Eq. (6) into the vector Helmholtz equations: either into \( \Delta \vec{E} + k^2 \vec{E} = 0 \) or into \( \Delta \vec{H} + k^2 \vec{H} = 0 \). This gives us the dispersion relation in the form

\[ k^2 - k_x^2 + \alpha^2 = 0 \quad (7) \]

The decay factor \( \alpha \) is yet to be determined. To find \( \alpha \), we substitute the anticipated solution (4) into the boundary condition (3) and use (6) to obtain

\[ \frac{\alpha}{j \omega \varepsilon} = -Z_0 \frac{Z_{\text{bottom}} + jZ_0 \tan kd}{Z_0 + jZ_{\text{bottom}} \tan kd}, \quad \alpha = -j \omega \varepsilon Z_0 \frac{Z_{\text{bottom}} + jZ_0 \tan kd}{Z_0 + jZ_{\text{bottom}} \tan kd} \quad (8) \]

The result simplifies for \( Z_{\text{bottom}} = 0 \) (metal corrugation with no bottom load) and takes the form

\[ \alpha = \omega \varepsilon Z_0 \tan kd = k \tan kd \quad (9) \]

where \( k \) is such as \( k_x = k \sqrt{1 + \tan^2 kd} \). When the corrugation teeth are considered to be of finite width, \( g \), an approximate solution can be obtained using the average surface impedance Eq. (2). It has the form [53]

\[ k_x = k \sqrt{1 + \left( \frac{g}{g + t} \right)^2 \tan^2 kd} \quad (10) \]

The propagation or phase speed of the surface wave thus becomes

\[ c_p = \frac{\omega}{k_x} = c \frac{k}{k_x}. \]

Now we summarize the results for surface wave propagation on a planar corrugated surface in the form of a chart that uses the corrugation depth \( d \) as a varying parameter.
Case a). Zero corrugation depth, $d = 0$. The decay factor $\alpha$ from Eq. (9) becomes zero and the $E$-field component in the propagation direction given by Eq. (6) is also zero. The surface wave transforms into a non-decaying plane wave of type II. The corrugated surface becomes a solid metal surface.

Case b). Shallow corrugation depth, less than quarter wavelength, $0 < kd < \pi/2$ or $d < \lambda/4$. The decay factor $\alpha$ from Eq. (9) is positive. The slow surface wave can propagate, with the propagation speed being less than speed of light in the medium, i.e.: $c_p = c \cdot \left(\sqrt{1 + \tan^2 kd}\right)^{-1} < c$. In this case (the surface wave non-cutoff region), the propagation speed depends on the corrugation depth. We use this property as an effective mechanism to control the surface wave propagation in the new non-cutoff ground plane design described in the present paper (section V). No other waves of type I or type II can propagate along the shallow corrugated surface.

Case c). Medium corrugation depth, greater than or equal to quarter wavelength, $\pi/2 \leq kd < \pi$ or $\lambda/4 \leq d < \lambda/2$. The decay factor $\alpha$ from Eq. (9) is negative and the decaying surface wave cannot exist. No type I or II or III waves can, therefore, propagate along the surface. This case – the surface wave cutoff region – is often implemented in practical applications, such as GPS choke ring antennas.

Case d). Deep corrugation depths, greater than or equal to half wavelength, $kd > \pi$ or $d > \lambda/2$. The process repeats periodically according to the sign of the tangent function, as in case b) and case c).

C. Kildal’s correction for concentric metal corrugation
The corrugated surface in the form of a disk whose cross-section is shown in Fig. 3 may require a certain correction to the cutoff frequency compared to the case of a planar corrugation if the antenna height is significant [49]. First, the rings with smaller radii correspond to a coaxial waveguide rather than to a parallel-plate waveguide. It becomes apparent, from symmetry considerations that a TE$_{11}$ mode is excited there, but not the TEM coaxial mode. For the TE$_{11}$ mode, the guide wavenumber is approximately equal to (see Refs. [49], [54]):

$$k_{TE_{11}} \approx k \sqrt{1 - \frac{1}{k^2 \rho_0^2}} \quad \text{when} \quad g + t \ll \rho_0$$

where $\rho_0$ is the parametric distance of the corrugations from the centre of the ground plane (see Fig. 3). Equation (2) for the surface impedance should then be replaced by

$$Z_S = jZ_0 \tan k_{TE_{11}} d = jZ_0 \tan \left( kd \sqrt{1 - \frac{1}{k^2 \rho_0^2}} \right)$$

The rest of the solution remains the same, so we can express the cutoff surface condition as

$$kd \sqrt{1 - \frac{1}{k^2 \rho_0^2}} = \frac{\pi}{2} \Rightarrow d = \frac{\lambda}{4} \sqrt{1 - \left(\frac{\lambda}{2\pi \rho_0}\right)^2} \geq \frac{\lambda}{4}$$

Thus, the cylindrical corrugations closer to the antenna should theoretically have a slightly larger depth $d$ in order to assure the cutoff for all the corrugation rings at the same frequency.
Another question of interest in Fig. 3 is the phase of the surface wave, $\Theta_s$, that travels all the way through the corrugation toward the end of the ground plane. This phase should be compared to the phase $\Theta_0$ of the line-of-sight signal from the same antenna, propagating in the same direction. With reference to Fig. 3 and for a constant corrugation depth, one has

$$\Theta_s = \int_{c/2}^{G/2} k_r(r)dr, \quad k_r = k \sqrt{1 + \tan^2 \left( kd \sqrt{1 - \frac{1}{k^2 r^2}} \right), \quad \Theta_s \approx k \left( \frac{G}{2} - \frac{c}{2} \right)$$

(14)

where $r$ is the radial distance across the corrugation. A typical value for the lower integration limit is $c \leq \lambda/2$. The concentric geometry generally diminishes the corrugation effect.

III. Corrugated Surfaces in GPS Ground Planes

A. Choice of the corrugation depth

The idea of the corrugated surface as a ground plane for an antenna is perhaps best explained on the base of Fig. 4 that is partially adopted from Ref. [55]. Let’s consider a monopole-like antenna located in the center of a metal or a corrugated finite ground plane, as shown in Figs. 4a) and 4b), respectively. A metal ground plane supports plane waves of type II that travel all the way toward the surface edge. These plane waves are then radiated as if the edge were essentially a point source. The edge radiation is also directed downwards, which produces a significant antenna backlobe. Due to reciprocity, the same pattern holds for the receiving antenna that does not only receive the useful signal from a positive elevation angle but also all the electromagnetic noise emitted by the ground and possible interference sources located close to the horizon.

The cutoff corrugated surface shown in Fig. 4b suppresses both plane and surface wave and thus prevents edge radiation. Hence the backlobe radiation and the radiation at very low-elevation angles (from horizon) are significantly reduced. Indeed, the present property of the corrugated surface holds only for a certain band of frequencies.

A crucial point that we would like to clarify in this paper is that this metal corrugation is inherently broadband, even though the metal corrugation is frequently associated with exactly $\lambda/4$ corrugation resonance – when every corrugation groove acts as a quarter wave open-closed resonator. This is the case for classic choke ring ground planes. One important property revealed from our discussion is that the metal corrugation actually is functioning not only close to the resonant point (as usually assumed) but also in the entire frequency domain that approximately satisfies the inequality $\lambda/4 < d < \lambda/2$, where $d$ is the corrugation depth and $\lambda = c/f$ is the wavelength. In other words, if lower and upper band frequencies are given by $f_L, f_U$, respectively, then the (planar) corrugation depth should satisfy inequality $\lambda_L/4 \leq d \leq \lambda_U/2$ in order to suppress surface waves and any other plane waves propagating perpendicular to corrugation. This result makes the choke ring a good candidate for multi-frequency and multi-band applications.

According to the theory of planar corrugation, the conventional GPS choke ring antenna becomes a cutoff corrugated surface when the depth of the corrugation becomes greater than the largest quarter wavelength – in this case – the quarter wavelength in L2 GPS band centered at 1.227 GHz.

$$d \geq \lambda_{1.227\text{GHz}}/4 = 61\text{ mm}$$

(15a)

At the same time, the depth still should be less than $\lambda/2$ at the highest frequency – in this case, the half wavelength in the L1 band centered at 1.575 GHz, which yields
\[ d \leq \frac{\lambda_{1.575\text{GHz}}}{2} = 95\text{mm} \]  

(15b)

In fact, these values need to be slightly modified using the corrections to planar corrugation discussed above in section C. A certain value in between these two should be chosen. It makes sense to choose a lower depth in order to reduce the ground plane height. In particular, the depth of 65 mm gives good results for a wideband base station GPS turnstile according to full-wave numerical simulations, as it will be shown in the following.

**B. Choice of the ground plane diameter**

The optimal size of a geodetic GPS antenna (flat metal or choke ring ground plane) can be chosen on the base of a well-known observation for planar metal ground planes. We consider a numerical example of a half-wave strip dipole spaced a quarter-wavelength apart from the square ground plane of a variable size. Such a dipole may model one wing of a turnstile antenna for circular polarization. We varied the ground plane size, \( G \), from \( 0.5\lambda \), to \( 2.0\lambda \), in intervals of \( 0.5\lambda \), and looked at two radiation patterns: the total gain in the \( E \)-plane of the antenna (the \( xy \)-plane) and total gain in the \( H \)-plane (the \( yz \)-plane). The corresponding results are shown in Fig. 5 and are summarized in Table 1. Note that, in the plots, the value of \( \theta = 0 \text{ deg} \) corresponds to zenith.

It is evident from Fig. 5 that the radiation in the backward direction is quite significant. As expected, the back-lobe decreases with an increasingly large ground plane. However, this decline is not monotonic. The front-to-back ratio for this antenna model is given in Table 1 for a square ground plane.

<table>
<thead>
<tr>
<th>Ground plane size, ( G )</th>
<th>Front-to-back ratio, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5( \lambda )</td>
<td>~8 dB</td>
</tr>
<tr>
<td>1.0( \lambda )</td>
<td>~14 dB</td>
</tr>
<tr>
<td>1.5( \lambda )</td>
<td>~21 dB</td>
</tr>
<tr>
<td>2.0( \lambda )</td>
<td>~21 dB</td>
</tr>
</tbody>
</table>

The performance obviously increases as the size of the ground plane increases. From the view point of performance/size ratio, the most beneficial is the ground plane of size \( G \approx 1.5\lambda \), since the incremental performance deteriorates for larger values of \( G \). Similar results can be derived for a circular ground plane of diameter \( G \).

For the classical dual-band GPS frequencies of operation, where the wavelengths are 19 cm and 24 cm, the optimal ground plane size is thus 28.5 cm at L1 and 36 cm at L2. This is perhaps one reason why the size of choice for classic geodetic antennas traditionally has been about 37 cm. For the GPS choke rings, a similar observation can be made (see Ref. [30]).

**IV. Optimization of a Cutoff Choke Ring Over a Wide Frequency Band**

**A. Low-multipath GPS antenna specifications**

GPS satellites transmit Right-Hand-Circular-Polarized (RHCP) electromagnetic signals to allow discrimination between direct GPS signals (RHCP signal) and reflected/scattered signals, which become mainly Left-Hand-Circular-Polarized (LHCP) signals after bouncing off reflecting objects near the receiving antenna. This allows multipath attenuation to be achieved by polarization discrimination [11]. To achieve the required positioning accuracy, ideally a GPS antenna should have a cross-polarization rejection ratio (or polarization isolation) \( \geq 15 \text{ dB} \) for signals incoming at any positive elevation angle (attenuation on the order of 10 dB is typical).
Because the antenna is required to receive signals from all the satellites of the constellation in view at a given time for good positioning performance (including GPS signals incoming from low elevation angles near the horizon), the antenna pattern needs to be virtually hemispherical. However, to reject electromagnetic noise (multipath and other interference such as jamming signals) mainly coming from low elevation angles, a very sharp slope is desirable near the horizon. For this reason, a high accuracy antenna should have a uniform antenna pattern from zenith (90° elevation angle) down to 5° elevation angle, while it should reject all signals at negative or elevation angle lower than 5° for any azimuth angle. These are two very difficult requirements to achieve simultaneously, and impose some engineering compromise within the design process. Moreover, the back radiation needs to be less than -5 to -10 dB, in order to decrease the effects of the reflections from the ground.

All the previous requirements are established mainly to achieve effective multipath mitigation. However, for optimal GPS signal reception the antenna is required to have a near hemispherical antenna radiation pattern for the upper hemisphere. The radiation pattern variation over the main beam needs to be ≤ 6-8 dB to obtain good coverage of the signals from GPS satellites. The group delay response should be nearly uniform for all angles of incidence in most of the upper hemisphere to avoid timing errors. The antenna also needs to have a very stable phase center for good positioning performance. Because the signal received by the antenna is very weak, the antenna needs good impedance matching. Finally, future GNSS will require operation in a wide bandwidth for reception of Modernized GPS, GLONASS, and GALILEO.

All these strict specifications required for future high-accuracy GNSS antennas are summarized in Table 2. (See also Refs. [56] and [57].)

Table 2. Specifications for High-Accuracy GNSS Antennas

<table>
<thead>
<tr>
<th>List of Antenna Specifications</th>
<th>GPS Antenna Performance Required</th>
</tr>
</thead>
<tbody>
<tr>
<td>FOR MULTIPATH MITIGATION</td>
<td></td>
</tr>
<tr>
<td>Antenna pattern (low elevation performance)</td>
<td>Slope at 0-5 deg ≥ 1 dB / 1 deg</td>
</tr>
<tr>
<td>Front-to-Back Ratio</td>
<td>≤ -5 to -10 dB</td>
</tr>
<tr>
<td>Polarization</td>
<td>RHCP</td>
</tr>
<tr>
<td>Polarization Isolation</td>
<td>≥ 15 dB</td>
</tr>
<tr>
<td>Polarization (Cross Polarization Rejection Ratio)</td>
<td>(ideally for any elevation angle)</td>
</tr>
<tr>
<td>FOR OPTIMAL GPS SIGNAL RECEPTION</td>
<td></td>
</tr>
<tr>
<td>Near Hemispherical Pattern</td>
<td>Upper hemisphere</td>
</tr>
<tr>
<td>Reception of Weak GPS Signals</td>
<td>Good impedance matching</td>
</tr>
<tr>
<td>Radiation Pattern Variation Over the Main Beam</td>
<td>≤ 6-8 dB</td>
</tr>
<tr>
<td>Group Delay Response (for Timing)</td>
<td>Uniform for all angles</td>
</tr>
<tr>
<td>Phase Center</td>
<td>3mm RMS phase deviation at L1 [30]</td>
</tr>
<tr>
<td>Bandwidth of Operation</td>
<td>1.15 GHz ≤ f ≤ 1.60 GHz</td>
</tr>
<tr>
<td></td>
<td>L1, L2, L5, E1, E2, E5a, E5b, E6</td>
</tr>
</tbody>
</table>

B. Choke ring configurations
After having established the desired antenna performance, we now consider the choke ring basic ground plane configuration as in Fig. 6. We intend to compare the performance of choke rings with different numbers of rings to see whether they meet the requirements. We will also evaluate which configuration among the modeled ones is the best for GPS applications. Table 3 provides the parameters for some selected geometry designs shown in Fig. 7. The number of corrugation rings is the rounded ratio of the
choke ring radius, \(G/2\), minus the cavity radius, \(R\), to the ring period, \(g+t\). It is equal to 2-5 for all the geometries listed in Table 3. One variable is the corrugation depth \(d\). Since we are interested here in a larger bandwidth from 1.15 GHz to 1.60 GHz that covers L1, L2, and L5 bands, Eq. (15a) should be modified to

\[
d \geq \frac{\lambda_{1,15\,\text{GHz}}}{4} = 65 \text{ mm}
\]  

(15c)

The presence of a sufficiently large and deep central cavity is a necessary condition for good choke ring operation with the turnstile element used as a radiator. Further shaping of the cavity bottom (e.g., a conical bottom) has been found to have little or even negative influence on the patterns. For \(v\), \(s\), and \(t\) from Fig. 6, Table 3 gives some most appropriate parameter values. The turnstile droop angle \(\beta\) should be close to 30-45 deg. The presence of the turnstile balun affects the polarization isolation at higher elevation angles and cannot be ignored.

### Table 3. Geometric Parameters for Selected Antenna Ground Plane Configurations

<table>
<thead>
<tr>
<th>Geometry</th>
<th>Ring diameter (G), mm</th>
<th>Cavity radius (R), mm</th>
<th>Bowtie half-wing (v), mm</th>
<th>Separation from choke (s), mm</th>
<th>Ring period (g+t), mm</th>
<th>Tooth thickness (t), mm</th>
<th>Ring depth (d), mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>Flat ground plane (Fig. 7a)</td>
<td>360</td>
<td>--</td>
<td>30</td>
<td>--</td>
<td>14</td>
<td>100</td>
<td>6</td>
</tr>
<tr>
<td>Two rings choke ring (Fig. 7b)</td>
<td>360</td>
<td>80</td>
<td>30</td>
<td>14</td>
<td>14</td>
<td>14</td>
<td>6</td>
</tr>
<tr>
<td>Three rings choke ring (Fig. 7c)</td>
<td>360</td>
<td>80</td>
<td>30</td>
<td>14</td>
<td>14</td>
<td>14</td>
<td>6</td>
</tr>
<tr>
<td>Four rings choke ring (Fig. 7d)</td>
<td>360</td>
<td>80</td>
<td>30</td>
<td>14</td>
<td>14</td>
<td>14</td>
<td>6</td>
</tr>
<tr>
<td>Five rings choke ring (Fig. 7e)</td>
<td>360</td>
<td>60</td>
<td>30</td>
<td>14</td>
<td>14</td>
<td>14</td>
<td>6</td>
</tr>
</tbody>
</table>

### C. Results

We present here the complete modeling results for the cases given in Table 3 for a flat metal ground plane (Fig. 7a) and corrugated metal ground planes with different number of rings (Figs. 7b-7e), over the band 1.15 GHz to 1.60 GHz. Again, only the pattern information is given, for the absolute antenna gain (RHCP or LHCP). Phase center dynamics and the impedance matching problems of the corresponding feeding element are not reported here. In particular, in this subsection, the performance of a wide central-cavity choke ring with two rings (Fig. 7b), three rings (Fig. 7c), four rings (Fig. 7d), and five rings (Fig. 7e) is discussed. One reason for choosing the set of parameters from Table 3 is a desire to maximize the polarization isolation over the upper hemisphere. Diminishing a (relatively narrow) antenna backlobe at 180 deg was a secondary priority. A numerical FEM solution (Ansoft HFSS v. 11.1) is obtained after 18-20 mesh refinement steps, with the final mesh sizes on the order of 250,000-350,000 tetrahedra. The PML box size is \(550\times550\times550\) mm. Only the discrete frequency sweep has been used. The antenna with a balun is fed via two balun input terminations at the bottom using two orthogonal in-quadrature lumped ports.

The RHCP/LHCP patterns are given in Fig.8a to Fig. 8e. Every figure corresponds to a separate case from Table 3, in consecutive order. Variation of parameter \(d\) in Table 3 is the choke ring height variation; antenna height above the top of the ring always remains 14 mm. This data – choke ring depth – is labeled on top of every separate pattern plot for the choke ring. For the flat ground plane (Fig. 8a), the separation
distance from the ground plane is given. All gain curves are given for a droopy turnstile bowtie over the band 1.15-1.60 GHz, in steps of 0.05 GHz. The following observations can be made from the figures:

i. The flat ground plane of Fig. 7a has a poor polarization isolation (less than 10 dB) at elevation angles above 60 deg and a large LHCP backlobe (-7 dB or so) at 180° elevation angle, as shown in Fig. 8a. Furthermore, RHCP gain considerably changes with frequency, which points toward a potential problem with phase center fluctuations.

ii. The two-ring choke with a wide cavity of Fig. 7b improves the polarization isolation to a minimum of about 10 dB for elevation angles above 100 deg (see Fig. 8b). The LHCP backlobe still remains, about -12 dB. The RHCP gain variation greatly stabilizes with frequency.

iii. The three-ring choke with a wide cavity of Fig. 7c considerably improves the polarization isolation to a minimum of about 15 dB for elevation angles above 100 deg, as it can be observed in Fig. 8c. The LHCP backlobe is on the order of -15 to -17 dB. The RHCP gain variation greatly stabilizes with frequency.

iv. As evident in Fig. 8d, the four-ring choke with a ring wide cavity of Fig. 7d has performance quite similar to the three-ring choke (Fig. 8e). Therefore, one may conclude that further increase in the number of rings can only have a minor effect on the pattern performance.

v. Finally, the five-ring choke of Fig. 7e is quite similar in performance to the four-ring choke in Fig. 7d, as can be observed in Fig. 8e. Moreover, the pattern performance starts to degrade at certain antenna heights.

Thus the choke ring with three to four thick rings, tooth height of 65-69 mm, and the droopy turnstile element in a reasonably wide cavity is a good candidate to uniformly cover the band from 1.15 GHz to 1.60 GHz. It is important to emphasize that decreasing the radius of the central cavity may lead to a significant degradation in the antenna performance. It is also interesting to slightly reduce the tooth height, below the cutoff at 1.15 GHz, and then check its effect on the choke ring performance. We consider two cases: three-ring choke with a wide cavity (Fig. 7c), and four-ring choke with a wide cavity (Fig. 7d). It appears that the three-ring choke is very much immune to depth reduction below the cutoff; however, the four-ring choke provides a generally poor performance. This is in line with some observations made in Ref. [58]. It follows from these observations that the three-ring choke with a reduced height is probably also a good candidate for the wideband GPS antenna. The idea of choke ring depth reduction will be further explored in the next section, based on a different concept. Note that other efficient modifications of the choke ring ground plane are known from the literature (see [31] [32] [33] [34] [35] [36] [37] [38] [39] [40] [41] [42] [43] [44] [45] [46] [47]) – among them are a tapered choke ring, dual-depth choke ring, dielectric-filled choke ring, resistive-tapered ground plane, and horizontal corrugation. These designs may also provide a considerable reduction in size (see, for example, [38] [47], and [48], pp. 231-236).

V. One Concept of a Non-Cutoff Shallow Corrugated Surface

A. Non-cutoff surface wave cancellation

In this section, we discuss a new concept of corrugated surface operation: a non-cutoff FSS ground plane that may support slow surface waves, as shown in Fig. 9. According to this design, we let the surface wave be excited. The signal at low elevation angles is a combination of the direct LOS signal from the turnstile antenna (TM to z plane wave type II) and the slow wave supported by the corrugation surface (surface wave type III). The plane wave type I (TE to z) is not excited as it is shorted out by the corrugation.

We wish for the LOS signal and the surface wave to be cancelled at the end of the rim, i.e., have almost equal amplitudes and a phase difference equal to $\pi$. Such a cancellation would imply that the major source of edge diffraction, the quasi TM to z wave, may be reduced or eliminated. The idea of pattern control via
phase cancellation finds its roots in many other antenna setups; one of them is the Salisbury screen – the widely-known non-reflecting material introduced by MIT during WWII. The motivation for the non-cutoff FSS ground plane is a desire to reduce its height as the non-cutoff corrugated surface is more low profile than the deeper cutoff corrugation (choke ring design).

For a planar infinite corrugated surface one obtains the cancellation condition by comparison of the phases for both waves at the ring edge. It is assumed that the initial phase at the ring center is the same. The cylindrical amplitude divergence does not contribute to the phase difference. From Eq. (10), the phase cancellation condition is thus obtained in the form:

$$ k_s R - k_0 R = n \pi, \quad \sqrt{1 + \left( \frac{g}{g + t} \tan k_0 d \right)^2} - 1 = \frac{n \pi}{k_0 R}, \quad n = 1, 3, ... $$  \hspace{1cm} (16)

where:

\( R \) is the distance from the antenna to the rim of the ground plane (radius).

Equation (16) is indeed an approximation that does not take into account the corrections introduced by Eqs. (13) and (14) for the rings of constant curvature, variable non-zero incidence angles, divergence, and amplitude taper. At the same time, the corresponding full-wave simulation test performed for a finite exactly planar corrugated surface excited by a simple dipole indicated that Eq. (16) at \( n = 1 \) does predict the pattern cancellation at oblique radiation angles of about 18 dB for a single frequency band.

However, for a wideband operation, it has been found that not the ring gap that achieves the 180 deg phase shift at the rim but rather a 90 deg average phase shift over the entire band is best suited for pattern control. The exact reason for this observation is not yet completely clear. For example, at

$$ d = 33 \text{mm}, \quad g = 6 \text{mm}, \quad t = 4 \text{mm}, \quad R = 170 \text{mm} $$  \hspace{1cm} (17)

the average phase shift over the band is about 90° and the maximum phase shift at 1.60 GHz is about 150°. In practice, a slightly lower corrugation depth, \( d \), may be chosen. The dipole height above the corrugation is critical for the subsequent analysis.

**B. Antenna setup and related simulations/measurements**

A non-cutoff FSS ground plane GPS antenna built according to the corresponding numerical model is shown in Fig. 10. Fig. 11 depicts the antenna feed element – a droopy bowtie turnstile with a balun – to achieve wide bandwidth (Fig. 11a) and the rear of the ground plane with a quadrature hybrid (Fig. 11b). We have used a split-coaxial balun with four slots and two inner transmission lines. Details of the balun operation are beyond the purpose of this study and may be found elsewhere. The dipole height (from top) above the ground plane was optimized by numerical simulations and hardware experiments around 60 mm.

Fig. 12a shows RHCP and LHCP antenna patterns in the \( E \)-plane of one dipole (top) and in the \( D \)-plane, 45 deg cut between two dipoles (bottom). The left column gives the measured data for a droopy bowtie turnstile (realized gain); the right column shows the corresponding numerical simulations (Ansoft HFSS; absolute gain). There is good agreement between theory (simulations) and measurements. Note that the antenna still has a certain impedance mismatch (the return loss approaches -5 to -7dB at low frequencies and then decreases toward high frequencies).
Also note that the feeding dipoles perform well mostly within the band 1.20 GHz – 1.55 GHz – see Fig. 12a. If we add the patterns at 1.15 GHz and 1.6 GHz as shown in Fig. 12b, the antenna performance at zenith (polarization isolation) starts to become worse compared to the simulations. It is believed that a not perfect impedance matching and asymmetry of the feeding elements is the major reason for that.

Both in theory (simulations) and in experiment, the non-cutoff FSS ground plane has a somewhat larger RHCP gain at low elevation angles compared to that for the cutoff choke ring. It also has a faster decay of the RHCP gain at elevation angles just below the horizon. At the same time, the backlobe at 180° is higher than for the cutoff choke ring. It is believed that the performance of the present antenna can be further enhanced with proper improvement of the radiating element (e.g. better impedance matching).

Finally, Fig. 13 shows the relative size of the non-cutoff FSS ground plane versus a commercial dual-frequency choke ring GPS antenna. One could see a considerable reduction in the size, both in vertical and in horizontal dimensions. At the same time, the pattern performance approaches that of the deep wideband choke ring - see Fig. 8. Moreover, it is important to remember that for a choke ring to operate in the wideband 1.15-1.60 GHz the minimum corrugation depth allowed is d=65 mm, as demonstrated in Eq. (15c).

VI. Phase Center of Antennas with Ground Plane

The phase center of an antenna is the local center of curvature of the far-field phase front – in other words, it is the center of a sphere that is tangent to the far-field phase front for a given observation angle. In general an antenna phase center location can vary with elevation angle and frequency of operation, and its dynamics are critical for high-precision GPS applications [59].

For a simple dipole antenna without ground plane, the phase front is angle-independent and frequency-independent. The phase center coincides with the physical center of the antenna. For a dipole (and any other antenna) with an infinite reflector the phase centre is the center of the reflector ground plane, for any frequency and for any dipole height. In other words, it is located exactly in the middle between the dipole and its image. When the ground plane is of finite size or just small, the phase center is expected to be located somewhere between the ground plane and the dipole, since the effect of the ground plane is less profound. It is also expected to have an angular dependence, especially at low elevation angles.

Typically dual frequency (L1/L2) choke ring base station antennas maintain a stable phase center that has less than 1 mm drift with elevation angle. It is suggested in [60] to determine the best-fit phase sphere first by a least-mean square fit to the measured (or calculated) hemispherical phase data. Then a local rms deviation from that value can be estimated, giving the average error estimate. For a choke ring ground plane a rms phase deviation of about 5 deg (approximately 3 mm at L1) is typical [60].

In order to estimate the phase centre variation as function of frequency for the non-cutoff antenna design proposed in this study, we partially apply the results from [61] for wide-flare angle scalar corrugated horns (see Fig. 14-43, Ref. [61]). The distance of the phase center from the apex as a function of frequency can be estimated from the phase factor \( \Delta = a \cdot [\tan(\theta_f/2)] \), where \( a \) is the distance between the rim of the corrugations of the horn and the axis (in our case \( a \sim r =17 \text{ cm} \), where \( r \) is the radius of the ground plane), and \( \theta_f \) is the flare angle (in our case \( \theta_f \sim 85^\circ \)). According to Bird et al. in chapter 14 of Ref. [61], when \( \Delta \) is very small, the phase center is located on the axis of the horn, very close to the plane of the aperture. For larger values of the phase factor \( \Delta \), the phase center moves along the axis toward the throat and eventually becomes fixed at the horn apex for scalar horns with \( \Delta =0.7\lambda \). Within the wide range of frequencies covered by the non-cutoff FSS antenna design, from 1.164 GHz and 1.610 GHz, the ratio between the phase factor and the wavelength (\( \Delta/\lambda \)) varies between 0.60 and 0.84, respectively. At higher
frequencies, between 1.349 GHz (for which the ration is 0.70) and 1.610 GHz (for which this ratio is 0.84), the phase center position tends to be fixed at the horn apex and, therefore, its variation is negligible. At lower frequencies, between 1.164 GHz and 1.349 GHz, we can estimate the phase center position at each frequency and the total variation along the axis is about 20 mm. However, within each bandwidth (e.g. at L5, lowest bandwidth of interest, between 1.164 GHz and 1.188 GHz), the phase center variation is less than 2 mm. The phase center variation with elevation angle should follow that of wide-flare angle horns, at least at high elevation angles, but needs to be investigated separately for low elevation angles.

VII. Conclusions

In the present study we have compared the performance of the classic low-multipath GPS choke ring ground plane with a new shallower design. Both ground planes are here optimized to obtain wideband operation uniformly over 1.15 – 1.60 GHz. To achieve reception of GPS signals over such a large bandwidth we have chosen a simple wideband radiator, the droopy bowtie turnstile. From our simulations and measurements it emerges that:

- Both ground plane configurations (cutoff and non-cutoff), when properly optimized, can operate over the wide bandwidth required for acquisition of GNSS signals (such as Modernized GPS and GALILEO). They meet most other stringent GPS antenna performance requirements.
- The non-cutoff FSS ground plane GPS antenna discussed in the present study is less than 40 % smaller in volume than the equivalent choke ring antenna.
- The non-cutoff FSS antenna has a higher RHCP gain at low elevation angles compared to the choke ring design.
- The non-cutoff antenna has a faster decay of the RHCP gain at elevation angles just below the horizon.
- The deep choke ring antenna has a higher back lobe at 180 deg than the non-cutoff FSS antenna.

Further studies focused on the GPS/GNSS radiating element (rather than the ground plane as in this work) are underway to improve the existing droopy bowtie turnstile and to achieve a better impedance matching over the band for the present setup. A better radiating element will significantly improve the overall GPS antenna performance for the choke ring and for our novel non-cutoff FSS wideband ground plane design.

Because of its simple and elegant design, its smaller size, robustness and ease of manufacturing, the wide bandwidth, small back-lobe and side-lobe radiation, and excellent polarization isolation (especially at low elevation angles), the proposed non-cutoff FSS ground plane antenna is a very promising candidate for future GNSS high-precision and high-rate GPS applications.
Figure captions

Fig. 1. A corrugated surface on the base of quarter-wave resonators.

Fig. 2. Corrugated surface of arbitrary depth, corresponding boundary conditions, and TM surface wave.

Fig. 3. Cross-section of a corrugated disk. The star schematically denotes the antenna feed.

Fig. 4. A monopole-like like antenna in the middle of a finite ground plane; a) – finite metal ground plane; b) – finite cutoff corrugation surface. The signal diffracted by the ground plane edge is schematically shown.

Fig. 5. Radiation patterns of a horizontal dipole above a finite ground plane of variable size.

Fig. 6. Geometry and associated dimensions of the choke ring with a bowtie turnstile.

Fig. 7. Wideband choke ring geometries with different number of rings. All ground planes have diameter D=360 mm and corrugation depth $d=65-69$mm; Fig. 7a) – antenna with flat metal ground plane. Fig. 7b) – choke ring with two rings, Fig. 7c) – choke ring with three rings, Fig. 7d) – choke ring with four rings, and Fig. 7e) – choke ring with five rings. The other ground planes parameters are listed in Table 3. These ground planes were optimized for wideband operation between 1.15 and 1.60 GHz.

Fig. 8a. RHCP and RHCP gain curves (absolute gain) in the E-plane of one turnstile element – Design of Fig. 7a, Table 3 (flat ground plane). The gain curves are given for a droopy turnstile bowtie over 1.15-1.60 GHz, in steps of 0.05 GHz. Different plots correspond to different choke ring depths (see Table 3).

Fig. 8b. RHCP and RHCP gain curves (absolute gain) in the E-plane of one turnstile element – Design of Fig. 7b, Table 3 (two rings; wide central cavity). The gain curves are given for a droopy turnstile bowtie over 1.15-1.60 GHz, in steps of 0.05 GHz. Different plots correspond to different choke ring depths (see Table 3).

Fig. 8c. RHCP and RHCP gain curves (absolute gain) in the E-plane of one turnstile element – Design of Fig. 7c, Table 3 (three rings; wide central cavity). The gain curves are given for a droopy turnstile bowtie over 1.15-1.60 GHz, in steps of 0.05 GHz. Different plots correspond to different choke ring depths (see Table 3).
Fig. 8d. RHCP and RHCP gain curves (absolute gain) in the E-plane of one turnstile element – Design of Fig. 7d, Table 3 (four rings; wide central cavity). The gain curves are given for a droopy turnstile bowtie over 1.15-1.60 GHz, in steps of 0.05 GHz. Different plots correspond to different choke ring depths (see Table 3).

Fig. 8e. RHCP and RHCP gain curves (absolute gain) in the E-plane of one turnstile element – Design of Fig. 7e, Table 3 (five rings; wide central cavity). The gain curves are given for a droopy turnstile bowtie over 1.15-1.60 GHz, in steps of 0.05 GHz. Different plots correspond to different choke ring depths (see Table 3).

Fig. 9. Concept of phase cancellation for the slow surface wave and the LOS signal for the turnstile antenna above the non-cutoff ground plane. The phase cancellation is expected to take place close to the rim.

Fig. 10. Non-cutoff corrugated ground plane prototype. The ground plane diameter is 34 cm; it has a slightly frusto-conical shape, with a semi-flare angle close to 85 deg. The corrugation depth is kept the same along the radius (d = 25 mm).

Fig. 11. a) Radiating element used for the wideband non-cutoff FSS ground plane GPS antenna: droopy turnstile bowtie with a balun. b) Rear of the ground plane with a quadrature hybrid.

Fig. 12a. RHCP and LHCP antenna patterns in the E-plane of one dipole (top) and in the D-plane (45 deg cut between two dipoles). Left column – measured; right column – simulated. The measurement data is given for 1.20:0.05:1.55 GHz.

Fig. 12b. RHCP and LHCP antenna patterns in the E-plane of one dipole (top) and in the D-plane (45 deg cut between two dipoles). Left column – measured; right column – simulated. The measurement data is given for 1.15:0.05:1.60 GHz.

Fig. 13. Commercial dual-frequency choke ring antenna (left) and wideband GNSS non-cutoff FSS ground plane antenna (right). We have shown that for wideband applications the minimum corrugation depth of the choke ring, $d$, needs to be increased to $d=6.5$ cm; for dual-band applications the minimum corrugation depth is 6.1 cm.
Acknowledgement

The authors wish to thank Dr. Jonathan Towle for his help with the test measurements in anechoic chamber.

This work was supported by the National Science Foundation, under Grant Number: DMI-0450524. Any opinions, findings, and conclusions or recommendations expressed in this publication are those of the author(s) and do not necessarily reflect the views of the National Science Foundation.

References


[32] The University NAVSTAR Consortium (UNAVCO); on line at http://facility.unavco.org/project_support/permanent/equipment/antennas/ant_cals.html


