Design of a Compact Hemispiral GPS Antenna with Direction Finding Capabilities

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Abstract—A nearly hemispherical spiral, or “hemispiral,” antenna for global positioning system (GPS) reception is presented with design methodology and measurements. Its set of distinct radiation patterns allows receivers to estimate the angular direction of incoming signals in the presence of noise without the need for an array. It requires no dielectric or absorbing materials and may operate from a single RF path, making it ideal for cost-constrained platforms. The antenna’s compact size also compares favorably with known reconfigurable antenna designs.

Index Terms—Direction of arrival (DOA) estimation, global positioning system (GPS), signal processing antennas, radio direction finding, hemispherical spiral antennas.

I. INTRODUCTION

The estimation of a signal’s direction of arrival (DOA) can be valuable for many applications, especially if the information does not use additional data bandwidth. Specifically, DOA knowledge can provide validation for a true global positioning system (GPS) signal in the presence of spoofers, since the expected angle of arrival can be calculated for each satellite using the GPS almanac broadcast [1]. Mitigating the risks from spoofing and jamming is a growing concern in the GPS community [2], [3]. Most proposed solutions necessitate the use of digital beamforming networks with arrays, sometimes called controlled reception pattern antennas [4]. Unfortunately, these are difficult to implement on cost-sensitive platforms due to an expensive set of front end electronics—typically, low-noise amplifiers, mixers, oscillators, and filters proceed an ADC for each antenna in the array [5].

The motivation for this study stems from the desire to build a low-cost, space-efficient solution that can still offer the DOA functionalities provided by arrays. Work on this topic by past authors often involves some variant of the multiple signal classification algorithm, or MUSIC, originally derived for arrays and later modified for single antennas [6], [7]. The time-domain processing detailed in [7] requires that the incoming signal be sampled at a multiple of its own frequency, so sources of unknown frequency or with Doppler shifts cannot be identified—a critical problem for GPS. If time-averaging of received power is used without coherence, as in [8], multiple sources from different directions cannot be resolved reliably, especially if they vary slightly in frequency and amplitude. Additionally, the antennas used for these techniques, such as the ESPAR and the leaky-wave antenna, are larger because they generally rely on an array of undriven dummy elements to provide the required set of different radiation patterns [7]–[9]. The overall antenna volume would become larger still if two or more of these linearly polarized (LP) antennas are collocated to efficiently receive the right-hand circularly polarized (RHCP) signals from GPS satellites. The planar four-arm spiral antennas used in [10], [11] are electrically smaller, but each requires four RF inputs for operation, so the cost of electronics would be comparable to a four element array. Finally, prior single antenna solutions have not presented results for successfully identifying multiple sources, which is, as noted, difficult when modifying MUSIC for one RF input [7].

To address this need, we propose a hemispherical spiral (“hemispiral”) antenna for GPS capable of sensing DOA. Direction estimation is tested at L1 or 1.57542 GHz, where the most widely used civilian GPS signals reside, but the design and measurements encompass the L2 (1.2276 GHz) and L5 (1.17645 GHz) bands. The original contributions of this paper include:

- A GPS antenna to sense DOA that is smaller than its one- or two-port predecessors and requires significantly fewer electronic components than arrays
- Methodology and measurements to estimate the direction of more than one incoming source of varying frequencies

The details of the antenna design and simulations are presented in Section II. A brief overview of the modified algorithm is given in Section III. The performance of a prototype, measured in an anechoic chamber, is analyzed in Section IV.

II. ANTENNA DESIGN APPROACH

For the purposes of the DOA method used in this study, the most critical design requirement for the antenna is the ability to provide enough sufficiently different radiation patterns. Our modified two-arm Archimedean spiral will support distinct patterns based on the phasing of its arms. When operated in the usual way, the arms are fed 180° out of phase and a broad circularly polarized (CP) pattern is produced, which we will refer to as the differential pattern. Feeding an in-phase signal on the arms produces a deep broadside null, which will be called the common-mode pattern. Adjusting the phase from 0° in either direction steers the null and effectively changes

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the pattern. The rest of the section details the justifications for the most distinct part of the antenna, its shape, and the design tradeoffs made to receive signals efficiently in all operation modes. The reference frequency is L1 unless otherwise noted, but the spiral is sized for L5 as its lowest design frequency.

A. Dome profile

The outer shape of the design is a dome that is nearly a hemisphere, hence the name of this antenna. While this design could be implemented with a planar spiral, and arrays of flat spirals have certainly been proposed for DOA [12], we assert that a larger field of view (FOV) may be achieved with the dome. Under normal circumstances, i.e. no jammers, a GPS antenna ideally offers good RHCP reception from all directions above the horizon. To that end, we aimed to use the available solution volume more efficiently, with the expectation that the increased gain of the forward radiation would improve the axial ratio (AR) across a larger range of look angles and increase our FOV. The link between antenna equivalent volume and gain (and bandwidth) has long been observed and characterized [13], [14]. The idea of conforming a spiral to a hemisphere is not necessarily unique [15], [16], but publications with prototypes and documented measurements are rare. Conforming the spiral to a cone, however, is a known method for reducing the planar spiral’s back lobe [17], so this was used as a point of comparison.

To substantiate our hypothesis, three Archimedean spirals were simulated in HFSS® with all geometries identical except the profile [18]. One was flat, serving as a reference. Side views of the other two, a cone and a dome, are shown in Figure 1. All three have a maximum z-dimension of 0.25λ, with an infinite ground plane at that distance. Figure 2 shows the resulting AR for two different modes versus angle off the z-axis (θ), averaged across all azimuthal cut-planes in φ. For the differential pattern, Figure 2a, used in normal operation, the dome shape provides significantly lower mean AR over elevation. This delta decreases and overall AR increases as we approach the common-mode pattern, where AR is roughly the same for all three profiles. Figure 2b shows an in-between mode with an obvious null. To a receiver, this results in decreasing RHCP power as input phasing approaches 0°.

B. Matching network

Given the observed AR increase over certain pattern modes, it is critical to minimize other losses. The discussion in this section addresses the two-wire matching section and its role in balancing performance across these modes.

The hemispiral is implemented with controlled distance to a ground plane slightly larger than its own outer diameter, which could represent the side of any platform that it would be mounted to for operation. Ordinarily, for the differential-mode pattern, this plane should be at a distance of 0.25λ to align the phase of the reflected back lobe [19]. In this state, the hemispiral’s low input reactance allows the use of a simple two-wire quarter-wave transformer from the outputs of a matched rat-race coupler behind the ground plane, leading to a single RF port on the receiver [20]. However, as we deviate from 180° to generate the other patterns, the two-wire segment is no longer a transmission line due to the increasing common-mode components. At the extreme of zero phase, the same signal is seen on each arm of the two-wire plus the hemispiral. The separate impedances seen by the differential and common-mode signal components vary with the antenna’s distance to ground and, equivalently, the two-wire length [21]. Figure 3 shows the real and imaginary parts of the common- ($Z_{cm}$) and differential-mode ($Z_{diff}$) impedances as this distance is increased, calculated from z-parameters generated in HFSS®. The simulation model has an outer dome shape slightly modified from Figure 1. Evidently, for common-mode excitation, the two-wire length that yields the lowest mismatch loss is about 0.3λ, while the differential case is best matched at the usual quarter wavelength.
The maximum directivity and realized gain in the upper hemisphere are plotted over ground distance for the common-mode pattern in Figure 4a, where there is a peak in realized gain around the same 0.3λ mark. This is consistent as the input phase is increased up to about 60°. Beyond that, the differential components of the current begin to take over until the system again presents as a spiral load with a two-wire transmission line, Figure 4b. The chosen z-height of 0.28λ represents a trade-off favoring common-mode, optimized at L1. Mismatch losses at L2/L5 were accounted for through simulation and verified with measurements in Section IV-A, Figure 11.

C. Model geometry and simulations

Figure 5 shows the model for the hemispiral, with relevant dimensions annotated or given in the caption. It is a design compromise between the differential and common-mode patterns that yields good performance through all in-between modes. The antenna’s largest extent, its diameter, is 0.34λ at the lowest design frequency L5, comparatively smaller than those of the leaky-wave antennas used in [8], [9] (1λ+), and the 0.5λ-diameter ESPAR used in [7].

Reference equations for the Archimedean spiral may be found in [19]. The dome curvature is given by the following parametric equation, which also defines the distance in z between the arms:

\[ z(\phi) = h \cos^p \left( \frac{\phi}{\phi_0} \right) \]  

where \( h = 48 \text{ mm} \), \( \phi_0 = 11\pi \text{ rad} \), \( p = 0.5 \)

Decreasing the height \( h \) of the curvature generally worsens AR, while increasing it adds z-height and impacts the impedance discussed earlier. The exponent \( p \) defines the rate of curvature, with larger \( p \) resulting in more concavity at the edge. The \( p \) that produces the lowest mean AR over look angle changes with input phase. The variable \( \phi_0 \) determines the number of turns and the outer radius of the spiral. The impact of these variables were assessed using analysis similar to Sections II-A/II-B, and the values shown here are based on performance and size trade-offs over pattern modes.

The polar plots in Figure 6 show simulated 3D power patterns over the upper hemisphere for two modes. Since there are two arms, the null may not be arbitrarily steered to any angle; however, changing the input phase results in a set of distinct patterns in each azimuthal cut-plane, making it possible to estimate DOA. Using four different patterns in our modified MUSIC algorithm [22], we show simulation results in Figure 7 for 2D direction finding in \( \theta \) for each \( \phi \)-plane. Going forward, we define DOA error as the absolute difference between the true source angle and the MUSIC spectrum peak. Gaussian noise is added to the incoming signal, and results are averaged over a broad SNR range that depends on receiver implementation [1]. We include a source near the horizon (\( \theta = 85^\circ \)) to show the expected degradation in performance,
Because single-input MUSIC is modified to compose a received vector over time, to guarantee coherence it must sample at the exact same point of the incoming waveform for each pattern, which can only be done in real time if sampling is at a multiple of the lowest source frequency [7]. For moving sources (or receivers) with changing Doppler shifts, this is impossible because Doppler is determined in the digital domain (after sampling). Noncoherent processing may be used in this case, where received power is time-averaged in each pattern mode, but if more than one source is present the results will be incorrect. To our knowledge, this is not addressed in the open literature for modified MUSIC. Conventional MUSIC for arrays does not have this limitation as long as each input is synchronously received.

The solution we propose is a phase alignment in post-processing, which can be done with an accurate understanding of the multi-source received signal’s characteristics. For clarity, we illustrate with two waveforms in Equation (3), but generalization to more sources is straightforward.

\[
R_m(t) = B_m(\theta, \phi) \cos(\omega_c(t - t_m) + \delta) + C_m(\theta, \phi) \cos(\omega_b(t - t_m)) \\
= D_m \left[ \cos \left( \frac{(\omega_c - \omega_b) t - (\omega_c - \omega_b) t_m + \delta}{2} \right) \right] \\
+ F_m \left[ \cos \left( \frac{(\omega_c - \omega_b) t - (\omega_c - \omega_b) t_m - \delta}{2} \right) \right] \\
+ F_m \left[ \sin \left( \frac{(\omega_c + \omega_b) t - (\omega_c + \omega_b) t_m + \delta}{2} \right) \right] \\
+ F_m \left[ \sin \left( \frac{(\omega_c + \omega_b) t - (\omega_c + \omega_b) t_m - \delta}{2} \right) \right]
\]

where \( D_m = B_m(\theta, \phi) + C_m(\theta, \phi) \)
\( F_m = B_m(\theta, \phi) - C_m(\theta, \phi) \)

If the incoming waves have arbitrary radian frequencies \( \omega_c \) and \( \omega_b \), constant amplitudes \( B_m \) and \( C_m \), and a relative constant phase \( \delta \), we can represent the total received signal \( R_m \) in the time domain as (3) for the \( m \)th sampling period. We may assume that \( \omega_c > \omega_b \) and \( B_m > C_m \) without any loss of generality. \( B_m \) and \( C_m \) are a function of their location and the delay term \( t_m \) can be referenced to any sampling period, as long as it is consistent. The term with amplitude \( D_m \) and only cosine terms includes a spectral component at the average of \( \omega_c \) and \( \omega_b \) multiplied by another component at half the difference of the frequencies. This is simply a lower frequency envelope modulated on a carrier, a basic AM waveform. The second sine-only term is the same waveform in quadrature that serves to adjust the modulation depth.

When \( \omega_c - \omega_b \) is small, the delay term manifests as a larger phase offset in the carrier than the envelope. This would be a realistic scenario for most satellite communications, as carrier frequencies are generally fixed, channels are relatively narrow-band, and the range for expected Doppler shift is very small compared to the carrier. We align the envelopes first by taking the magnitude of \( R(t) \) and its quadrature component, which is periodic at the envelope frequency. Each
local peak may be detected and used as the starting point for aligning the carrier peaks. Conveniently, $t_{\text{m}}$ may also represent the effect of dropped samples, which relaxes any timing synchronization requirements between the data capture and the pattern-switching mechanism.

This simple step is not addressed in previous authors' modifications of MUSIC to accommodate a single RF input instead of a vector from an array. It is critical, however, when adapting the method to a multi-source system, especially when transmitters and/or receivers are not necessarily stationary. In the absence of a highly tunable front end and pre-existing knowledge of the present frequencies, this method may act as a substitute for receiving data in each mode simultaneously.

IV. RESULTS AND DISCUSSION

An early hemispiral proof-of-concept was made by hand and introduced in [24]. The current prototype, shown in Figure 8, is printed in copper C18150 by Stratasys Direct, Inc. It avoids the previous dielectric structure and enjoys higher realized gain, especially on patterns with input phasing away from 180°. Several fabrication flaws include uneven gaps between the two spiral arms (requiring suboptimal two-wire spacing to avoid shorts), a reduced outer radius, and a distorted outer profile. These all impact the resulting radiation in various ways, most notably reducing the realized gain by up to 1-2 dB depending on input phase and introducing pattern asymmetry, but updates were made to the simulation model to reflect the imperfections where possible. The results from this prototype are discussed in the remaining sections.

A. Measurements

Figure 9 is a simplified block diagram of the general data collection setup for DOA estimation, though some measurements discussed in this section require variations of this configuration. The diagram is a top-view, where the $y$-axis points to the chamber ceiling and the origin is referenced to the phase center of the hemispiral, the receiving antenna in this case. Radiation patterns were measured in the $xz$-plane, so for this demonstration direction estimation is limited to one plane. A system capable of accurately taking measurements in the entire upper hemisphere of the antenna would remove the limitation.

In all DOA data collection configurations, the hemispiral was the receiving antenna and dual-feed RHCP standard gain horns were the transmitting antennas. The receiver path included a USRP N210/DBSRX2 front end.

The hemispiral’s gain patterns are shown in Figure 10. Under normal operation with 180° of phase between the arms, the top plot includes measured data (solid lines) for L1, L2, and L5 (black, red, and blue respectively), which are 9.6, 4.9, and 6.8 dB at maximum. The dashed lines represent data from HFSS® simulations. There is a slight tilt in the measured data due to error in the manual $xz$ alignment between the hemispiral and the horn. The most significant sources of error would come from the actual versus specified gain of the “known” calibration antenna at these frequencies. Figure 10 also includes several examples out of the ten total patterns used for direction estimation, labeled by their input phasing. These measurements constitute a calibration step for MUSIC. We used an analog phase shifter on one of the two coupler outputs, so a source of error for the null location is introduced by the difference between the expected and actual input phase, the latter of which may fluctuate with temperature and even slight control voltage error. Additionally, the attenuation in the phase shifter path varies with control voltage, and the compensation in the coupler’s other leg was chosen as an average over phase steps, causing an imbalance that varies with each pattern. In practice, eliminating the phase shifter is straightforward once the desired patterns are chosen. One possibility is to use RF switches to change line lengths on the coupler corresponding to the discrete set of desired phases.

The measured antenna VSWR for the common-mode input (worst case) is shown in Figure 11 over the frequency range between L5 and L1, with the actual GPS carriers marked. Civil GPS signals inhabit about 2 MHz centered around L1, L2 (L2C), and about 20 MHz at L5. Military codes are allocated 24 MHz at L1 and L2 [1]. This element is evidently well-matched over the entire range and not just at the GPS bands. Future work may include extending the measurements beyond the bordering GPS frequencies and possibly to other applications.

The MUSIC spectrums for $xz$-plane sources placed at various angles relative to the hemispiral are shown in Figure 12. During the calibration step, the turntable used to measure the radiation patterns has a tolerance of roughly 3°, so this is expected to be the limiting factor of our direction estimation accuracy. The next section discusses and quantifies how the estimation error varies with other factors, like SNR and the patterns used.
FIG. 10. RHCP gain (dB) for the differential at all bands (a) and other pattern modes by input phase at L1 (b). Both simulated and measured data are taken in the $xz$-plane.

Fig. 11. Measured common-mode VSWR: 1.16 (L1), 1.93 (L2), 1.32 (L5)

B. Further Analysis

To quantify the system’s limits, we used the measured pattern data to predict DOA performance in various simulation environments. The most pressing concern for GPS is the robustness of this system when signals of interest have low SNR, as is the case with satellite reception. As a rough benchmark, our algorithm could expect post-correlation SNR levels around 20 dB or more, depending on the receiver’s integration time [1]. We found that above 26 dB of SNR, just two patterns are needed to produce results within 1° of accuracy after factoring in the tolerance of 3° noted in Section IV-A. Below 26 dB, it was observed that increasing the number of patterns yields higher estimation accuracy, with calibration-limited performance down to 12 dB when using up to eight patterns. However, not all additional patterns improve the estimate by the same amount, or at all. To be perfect substitutes for the mode vectors described in Section III, the matrix $A$ must be full column rank, which is not necessarily the case for an arbitrary set of patterns. The Pearson correlation coefficient, a measure of linear dependence, may be used for evaluation [25]. Its calculation for a pair of variables $X$ and $Y$ over $T$ samples with mean $\mu$ and standard deviation $\sigma$ is defined as [25]

$$\rho(X,Y) = \frac{1}{T-1} \sum_{i=1}^{T} \left( \frac{X_i - \mu_X}{\sigma_X} \right) \left( \frac{Y_i - \mu_Y}{\sigma_Y} \right)$$

(4)

where $\rho = 1$ indicates perfect correlation. We thus calculated a $\rho$-matrix of pairwise coefficients between all the patterns. For each pair of patterns, DOA estimation was performed for a sweep of incoming signals across $\theta$, with error averaged over increasing noise levels. The median over $\theta$ is shown in Figure 13, against increasing correlation coefficients on the $x$-axis. While the error does trend higher for more correlated patterns, the relationship is obviously not perfect. Using only
two patterns potentially leads to ambiguities at certain angles that may result in high error, especially in the presence of significant noise. Additionally, the correlation metric can only measure linear independence, so higher-order effects are not captured.

We also quantified system performance using incoming sources across all $\theta$ in the $xz$-plane, again averaging over a range of noise levels at each step. The hemispiral’s expected wide FOV was confirmed with as few as three patterns, shown in Figure 14 for several example pattern combinations with at least one low-correlation pair. The dashed line represents the calibration tolerance of the turntable.

Finally, we assessed the system’s performance with respect to incoming signals with a potential frequency offset. For this experiment, we placed two incoming sources at $\theta = -35^\circ$ and $70^\circ$ in the $xz$-plane. With one source kept at L1, the second source was swept in frequency over the range of possible Doppler shifts for GPS. The results are reported in Figure 15, again with the calibration tolerance shown.

V. CONCLUSION

In this study, a hemispiral antenna has been presented and evaluated for DOA with a modified MUSIC algorithm. Its design is particularly well-suited for wide-angle satellite reception, as required by GPS systems, but it can easily be adapted to other applications. Unlike prior work on single-antenna MUSIC, we also present methodology and results for sensing the arrival angles of multiple sources with frequency offsets. While we demonstrated 2D direction finding over one angle, $\theta$, the method supports azimuthal angle detection. The processing change for 3D DOA includes expanding the $A$-matrix to capture pattern variation over $(\theta, \phi)$. Future investigations will involve comparisons of performance against known antennas, design revisions with better manufacturability, and alternate methods of implementation.

REFERENCES


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