Techniques for Reducing the Near-Far Problem in Indoor Geolocation Systems

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BIOGRAPHY

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ABSTRACT

When using pseudolites to provide navigation in an indoor or underground environment there is a concern that nearby pseudolites will negatively impact the ability of a receiver to track all of the available pseudolites. This problem, known as the near-far problem is a consequence of limited dynamic range. Some techniques for solving this problem, such as modifying the pseudolite duty cycle or synchronizing pseudolite transmissions have been previously proposed. In contrast, this paper explores the dynamic range issue directly by investigating the behavior of a receiver’s tracking loops. By subjecting the tracking loop to various high dynamic range scenarios, the effect of signal and loop characteristics on dynamic range is directly visible. Based on this approach, proposals for increasing receiver dynamic range will be presented and validated using both analytically and simulated results.

INTRODUCTION

We start with an example similar to that presented by Dixon[2,p.374]. In Figure 1 transmitter 1 and transmitter 2 are intentionally broadcasting a message to the receiver. We assume CDMA channels, each transmitter uses a different PRN code and the receiver is normally capable of simultaneously receiving these signals. We further assume that the transmitters broadcast with equal power and are attenuated according to the square law propagation characteristic.

Figure 1: Example of a near-far problem

Because the distance from transmitter 1 is ten times that from transmitter 2, there is a 40dB disparity between the received signals. To receive both signals, the interference rejection capability must exceed 40dB and the receiver must accommodate at least a 40dB dynamic range.

Dixon proposes that the transmitters be coordinated in some way to reduce the cross correlation properties or employing time division multiplexing. In considering this
approach, we first consider the dynamic range that a receiver can accommodate.

In receiving such a weak signal, it is well known that the process of despreading provides an associated processor gain. In addition, given that the interference is highly structured, Verdu[11] discusses how an understanding of this mutual interference can lead to an optimal detector. Unfortunately, as in Dixon’s example, the roles of synchronous demodulation and that of detection are not clearly elaborated. This paper is not concerned with the problem of detection, but rather outlines our initial progress in understanding the more fundamental problem of coherent demodulation. In particular, how the near-far problem is manifested at the level of the tracking loops.

Blanchard[1,p.318] points out that the notion of operating threshold is subjective, based on the circumstances where phase-lock is needed. Blanchard as well as Gardner[3] point out that as the receive signal to noise ratio (SNR) decreases, phase variance of the digital counter oscillator (DCO) first increases to a point where performance rapidly degrades. At such a low SNR, the DCO occasionally slips cycles. At lower SNR, phase-lock is lost entirely. To regain phase-lock the SNR must first be increased several decibels.

For the purposes of this paper, we are not concerned with measuring the exact operating threshold for a tracking loop. Due to the empirical nature of this paper, we adjust SNR using 3dB or 6dB steps and rely on the fact that at the threshold, performance tends to degrade rapidly.

**RECEIVER MODEL**

The results presented in this paper were produced using a computer program written in ‘C’ that is similar to the Plessy Builder chipset, designed for Global Positioning System (GPS) C/A code only receivers. The model was selected only as a starting point for future research. The GPS uses direct sequence spread spectrum (DSSS) signaling and employs the familiar code division multiple access (CDMA) technique to differentiate between satellites.

Figure 2 is a conceptual model of such a multi channel GPS receiver. The AGC block is provided to avoid overloading analog components and to maximize the dynamic range of the analog to digital (A/D) converter.

The f0 multiplier and IF stage represent the converter used to move the desired signals to the intermediate frequency (IF) band. Collectively, the AGC, multiplier, IF and A/D block are referred to as the front-end. An individual channel composed of a multiplier and (CORR) correlator block may despread at most one broadcast signal.

Using Gardner’s[3,p.157] distinction we refer to the AGC block as wideband AGC as the control voltage is derived by rectifying the output of an IF amplifier in some way. Realize that the control signal is proportional to the total output of the amplifier, which includes the desired signal and noise (interference). If the noise exceeds the desired signal, as is typical, the control signal and hence the gain is determined by the noise.

The Plessey GP1010[8] serves as an example front-end circuit. The GP1010 includes an on-chip frequency synthesizer, mixers, AGC, and a four level A/D converter. The A/D magnitude is used to control the AGC loop so that the output is ±3 thirty percent of the time and is ±1 seventy percent of the time.

Figure 3 illustrates the hardware provided for each channel by a Plessey GP1020[9] correlator channel. The A/D converter described earlier provides the sampled IF signal. The PRN block is a pseudo-random signal generator and the ADM blocks are accumulate-and-dump modules, used to perform correlation.

![Figure 2: Receiver Model](image1)

![Figure 3: Channel Hardware](image2)
The Plessy PRN code generator is unusual in comparison to other generators in that pairs of successive locations in the shift registers are used to represent PRN code chips. In this way, two such updates are required to advance to the next PRN code chip. In generating the PRN code in this way, timing of the early and late codes is ½ chip apart.

After estimating the envelope of the early and late correlator outputs by forming the square root of the sum of squares, the code phase discriminator for a delay-locked loop may be realized[4]. Spilker[10] also serves as a useful reference.

In designing the loop filters for the Costas loop and the delay-locked loop, our choice in loop filters is motivated by the need to eliminate the steady state phase error. In both cases the loop filter is the simple integrator plus proportional filter \( L(z) \), expressed in equation (1). The values \( K_j \) and \( K_i \) correspond to the placement of the loop filter zero and the loop filter gain constant.

\[
L(z) = K_j \frac{z - K_i}{z - 1}
\] (1)

**NOISE MODEL**

To introduce noise into our simulator, a discrete value random number generator is used that produces the values ±1 and ±3, using the same probabilities as described earlier for the GP1010 device. The carrier DCO values are listed in Table 1. For a moment consider the statistics of the correlator outputs in the absence of signal. During one C/A code period each ADM accumulates 5714 samples. Thus, we apply the central limit theorem[7] to calculate the mean \( \mu_y \) and variance \( \sigma_y^2 \).

\[
\mu_y = 0
\]
\[
\sigma_y^2 = 48571
\] (2)

Thus the standard deviation of the correlator noise output is \( \sigma_y = 220 \). With the probability density of the ADM output being approximately Gaussian, the square root of the sum of squares of pairs of correlator outputs has Rayleigh density with mean \( \mu_z \) and variance \( \sigma_z^2 \). The mean value \( \mu_z \) serves as a useful indicator.

\[
\mu_z = \sigma_y \sqrt{\frac{1}{2} \pi} = 276
\]
\[
\sigma_z^2 = (2 - \frac{1}{2} \pi) \sigma_y^2 = 20847
\] (3)

**SIGNAL TO NOISE RATIO**

For this paper we assume that the wideband AGC is dominated by noise so that gain applied by the front end is independent of the desired signal energy. In our simulator we simply add in the received signal. We define a standard input reference level such that in the presence of the noise process described above, the signal to noise ratio at \( y_{ce} \) is +20dB. Relative to the reference level, the broadcast signal amplitude is changed in 6dB steps.

**SIGNAL ACCOMMODATION**

First, we consider the ability of a receiver to accommodate a wide dynamic range for a single received signal in the face of white Gaussian noise. In such a model the operating threshold sets a lower limit on the received signal to noise ratio that may be tracked. Stability of the tracking loops sets an upper limit. Hill and Michalson[5] as well as Natali[6] show that a discrete time Costas loop becomes unstable as loop gain and hence receive signal energy is increased.

We suspect that simply modifying the loop parameters affects the upper and lower limits in a similar fashion, causing the overall dynamic range to remain approximately constant.

In the design of each tracking loop we have the option of modifying either the loop filter gain term \( K_f \), or placement of the loop filter zero \( K_z \). Figure 4 from [5] is the root locus for the Costas loop, drawn for three values of \( K_z \). The root locus illustrates the trajectory of the system poles as the loop gain is varied. To be stable, the system poles must remain inside the unit circle. For this
research, $K_z$ was selected to be 0.875. The placement of the zero was chosen to control damping, leaving us only the loop filter gain $K_r$ to change. It was observed that increasing the loop filter gain lowers the operating threshold.

![Root Locus](image)

Figure 4: Root locus for Costas loop

An intuitive explanation why threshold decreases as loop filter gain is increased is that the overall loop gain is dependent on the received signal amplitude. For the case of a modified Costas loop, Hill and Michalson[5] show that loop gain is directly proportional to the received signal amplitude. It appears that increasing the loop filter gain by 6dB partially makes up for a 6dB loss in SNR.

THE ACCOMMODATION TEST

Figure 1 illustrates how the simulator was configured to perform this test. The initial carrier phase error and code phase error is set to zero, so that the tracking loops start in the locked condition. An additional ramp increase in phase is applied, for the Costas Loop the ramp corresponds to a shift of 4.2575 Hz and for the delay-locked loop this corresponds to an increase by 4.2575 chips per second.

![Test Configuration](image)

Figure 5: Test configuration

Rather than relying on a lock detector, the results are manually inspected. Three guidelines are used to determine if the tracking loops are working.

- With the delay-locked loop tracking the desired signal, the envelope of the ADM outputs will have an average value larger than $\mu_z$.
- Our Costas loop is designed so that when locked, signal energy is routed to $y_{ce}$ rather than $y_{se}$. With the Costas loop tracking the desired signal, the magnitude of $y_{ce}$ will approximately equal the envelope of $y_{ce}$ and $y_{se}$.
- The carrier and code rate shift of the transmitter is selected so that when tracking, both the Costas loop and delay-locked loop feedback have an average value of 100.

The amplitude of the broadcast signal is first adjusted so that the SNR of $y_{ce}$ is +20dB. We refer to this level as our starting reference (REF) level. Next, the broadcast amplitude is adjusted in 6dB steps, identifying when the receiver successfully tracks the desired signal. For this test the loop filter zeros were placed at $K_z = 7/8$. The values for the loop filter gain term $K_r$ is listed in Table 2.

<table>
<thead>
<tr>
<th>Rec.Power</th>
<th>$K_r = \frac{1}{4}$</th>
<th>$K_r = \frac{1}{2}$</th>
<th>$K_r = 1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>REF+18dB</td>
<td>NO</td>
<td>NO</td>
<td>NO</td>
</tr>
<tr>
<td>REF+12dB</td>
<td>YES</td>
<td>NO</td>
<td>NO</td>
</tr>
<tr>
<td>REF+6dB</td>
<td>YES</td>
<td>YES</td>
<td>NO</td>
</tr>
<tr>
<td>REF</td>
<td>YES</td>
<td>YES</td>
<td>YES</td>
</tr>
<tr>
<td>REF-6dB</td>
<td>YES</td>
<td>YES</td>
<td>YES</td>
</tr>
<tr>
<td>REF-12dB</td>
<td>NO</td>
<td>YES</td>
<td>YES</td>
</tr>
<tr>
<td>REF-18dB</td>
<td>NO</td>
<td>NO</td>
<td>NO</td>
</tr>
</tbody>
</table>

In examining Table 2, it first appears there is approximately a 24dB range over which the receiver can track the desired signal. For GPS applications, such a range is adequate. In all the cases listed here, the Costas loop was the first to lose lock. The idea of a coherent attenuator is presented to extend the upper range of signal energies that may be tracked.
COHERENT AGC

In contrast to the notion of wideband AGC, we use Gardner’s [3] distinction in reference to coherent AGC, which obtains the control signal from a synchronous detector. Any noise contribution is easily filtered out so that ideally, coherent AGC is independent of the received noise. Such an AGC may simply attenuate an excessively large correlation peak.

Unfortunately, since coherent AGC is dependent on synchronous detection, it cannot be applied in place of the wideband AGC. As illustrated in Figure 6, each channel in the receiver must provide its own coherent AGC mechanism (CAGC).

![Figure 6: Receiver with coherent AGC](image)

For our simulator, scaling the discriminator values by a constant value models the behavior of a coherent AGC. We decided that the AGC only provides attenuation, becoming active for signals having amplitude larger than the reference signal. Since the broadcast amplitude is known, we use equation (4) to produce the scaling factor, when the received signal amplitude exceeds the reference amplitude.

\[
\text{SCALE} = \frac{\text{REF\_AMPLITUDE}}{\text{AMPLITUDE}} \quad (4)
\]

Table 3: Summary with coherent AGC

<table>
<thead>
<tr>
<th>Rec.Power</th>
<th>( K_f = 1 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>REF+18dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF+12dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF+6dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF</td>
<td>YES</td>
</tr>
<tr>
<td>REF-6dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF-12dB</td>
<td>NO</td>
</tr>
<tr>
<td>REF-18dB</td>
<td>NO</td>
</tr>
</tbody>
</table>

In examining Table 3, it is clear that the dynamic range over which the tracking loops function has been extended. Of course, there are practical limits that prevent us from having infinite dynamic range.

AN INTERFERING SIGNAL

Next, we consider how an interfering signal elevates the ambient noise level. Therefore, a second carrier modulated PRN signal is added. The GPS PRN code 1 serves as the desired signal (produced with the frequency shift described earlier) and the GPS PRN code 2 depicts the interference signal (adjusted to 18dB above the reference level). For the tracking loops we select \( K_f \) to be 1.0. Coherent AGC is applied using the technique described in the previous section. The results are summarized in Table 4. For this test, failure is indicated when the percent difference between the average value of \( y_o \) and the envelope differs by more than 5% (the second guideline).

Table 4: Interference (18dB) PRN2

<table>
<thead>
<tr>
<th>Rec.Power</th>
<th>( K_f = 1 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>REF+18dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF+12dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF+6dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF</td>
<td>YES</td>
</tr>
<tr>
<td>REF-6dB</td>
<td>YES</td>
</tr>
<tr>
<td>REF-12dB</td>
<td>NO</td>
</tr>
<tr>
<td>REF-18dB</td>
<td>NO</td>
</tr>
</tbody>
</table>

From these results we can only glean that the interfering signal has apparently increased the noise level. In continuing this research it will be necessary to define a statistical means to indicate whether or not the performance level of the tracking loops is adequate.

CONSIDERING CODE LENGTH

From the outset we considered that the length of the PRN sequence used would provide a means to mitigate effects of the near-far problem. On one hand, processor gain is linearly related to code length \( L \). On the other hand, Dixon [2, p.88] outlines how for long codes, leakage attributable to cross-correlation decreases approximately by the square root of the code length. This gives a ratio of processor gain to leakage that is proportional to \( L^{\frac{1}{2}} \). Clearly, increasing the code length provides a means of reducing the lower limit illustrated in Table 4.

CONCLUSION

In this paper we first presented the problem of accommodating the dynamic range of a single received signal. For receivers requiring GPS signal compatibility,
coherent AGC provides a means to extend the dynamic range. The upper bound on this improvement has not yet been determined, but is worthy of further investigation. If GPS signal compatibility is not required, then extending the code length should provide a means for extending the lower bound component of the dynamic range. Since this research is based on the use of a Costas loop and since the fundamental limitation is based on the Costas loop to maintain lock, an investigation of other tracking structures is warranted.

REFERENCES


