An Alternative Approach to Multipath and Near-Far Problem for Indoor Geolocation Systems

Ilir F. Progri, Student Member, ION, and William R. Michalson, Member, ION
Electrical and Computer Engineering (ECE) Department, Worcester Polytechnic Institute (WPI)
100 Institute Road, Worcester, MA 01609
Tel: (508) 831-5297, FAX: (508) 831-5491
Email: {iprogri or wrm @ece.wpi.edu}, URL: http://www.ece.wpi.edu/~iprogri

Biography

Mr. Ilir F. Progri received his Diploma of Engineer Degree in Electrical Engineering (EE) from the Polytechnic University of Tirana, Albania in 1994. He received his MS degree in EE from the ECE Department at WPI in 1997, where he is a Ph.D. candidate. His research interests include carrier phase and signal processing, integrity monitoring, precision approach and landing, adaptive and non-adaptive filter design and implementation. Mr. Progri is the current recipient of the ION Student Sponsorship Paper.

Dr. William R. Michalson is an Associate Professor in the ECE Department at WPI, where he also directs the Satellite Navigation Laboratory. His research focuses on the development, test, and evaluation of GPS integrity monitoring algorithms, with an emphasis on integrity monitoring for sole-means navigation and precision approach.

Introduction

The usage of the pseudolites for local area augmentation and autonomous positioning, navigation and precision landing is well known. A majority of this research, which focuses on the pseudolite’s construction and some of its applications are summarized in [1]. The ionospheric, tropospheric, and ephemeris errors do not apply to most pseudolite applications. Nevertheless, to a large extent the authors address the near-far problem and multipath as the most severe source of error for most pseudolite related applications. Therefore, the signal structure and the receiver design must account for these types of errors.

A possible alternative for eliminating the near-far problem is the technique of pulsing the pseudolite’s signal [1]. This technique suggests that the signal can be turned on during a fraction of the duty cycle and turned off during the remaining fraction of the duty cycle. Duty cycles such as 10-30% are known in the literature and the community [1].

On the other hand, it appears that a higher chipping rate can mitigate the undesired effect of the multipath [2]. In the past we have also suggested that under some circumstances the symmetry can reduce the multipath errors [3].

The research presented in this paper consists of a more sophisticated technique for eliminating both the near-far problem as well as the undesired effect of the multipath. Here the effort consists of designing the system (the pseudolites and the receiver) despite the system geometry, multipath, and dynamics; i.e., we would like to design a system, which will perform well under any of those conditions. Although the main thrust of this application would be indoors, the signal structure and the system design can be applied as well for local area augmentation and all the applications, which are associated with severe multipath and near-far problems.

Previously, we have proposed a pseudolite signal structure, which appears to provide superior auto and cross-correlation signal properties when subject to a hypothetical environment [4]. Next, we performed a preliminary investigation, which appears to suggest that the proposed signal structure can mitigate the near-far problem [5]. The main objective of this work consists of

Abstract

Previously we have considered the navigation characteristics of an autonomous, pseudolite based, indoor geolocation system designed to operate on a single frequency modulated on “orthogonal” or “nearly orthogonal” codes to achieve channel separation. Although this system has the potential to provide accurate positioning under normal conditions; i.e., when low or no multipath is present and when the distance from the receiver is almost equally spaced from the pseudolites. Due to the presence of multiple objects, the multipath effect is significant, which impacts the receiver and total navigation performance. On the other hand, the system we are trying to build requires a random placement of the pseudolites and a random trajectory path of the receiver, which may lead to the near-far problem. Therefore, the impact of the multipath and of the near-far problem when multipath is present is not fully assessed. This paper will discuss the theoretical characteristics of the multi-carrier and multi-code navigation system by looking first at the receiver’s performance due to the multipath, geometry, and dynamics impact. The impact of other important parameters such as carrier frequency, code length and type, chipping rate, data rate, noise and transmitted power are discussed in the paper.
enhancing the investigation of the proposed signal structure and conducting additional analysis, which will assess the performance of the proposed pseudolite’s signal structure when multipath is present.

The paper is organized as follows: A description of the system is given in following section. The receiver model and analysis follows. The theoretical performance is assessed in the section that follows. The paper is concluded with a conclusion section and as always with a useful list of references.

System Description

Consider an indoor application (ex. firefighters in a rescue mission) as shown in Figure 1.

Smoke, fire, different fluids, falling objects (such as ceilings), etc, can be present in the environment in which firefighters operate. Under these conditions, it is conceivable that multipath will be the most severe source of errors.

Assuming an initial placement of the pseudolites, the movements of a firefighter can be described as a random walk; i.e., the firefighter can be very close to one pseudolite and far away from another (see Figure 1). This situation leads to the near-far problem. The initial research on the signal structure has demonstrated that the binary codes do not provide enough separation; hence, a pulsed pseudolite signal can be applied when subject to indoor geolocation systems [1][4].

We intent, however, to explore a different alternative, which is going to employ a system with multi-carriers and multi-codes. Every pseudolite’s transmitted signal power is set to $P_T$.

We will inherit some of the assumptions stated in [5].

A detailed analysis of the received signal model is presented in the following section, with the intent to analyze the impact of the multipath and near-far effect on the receiver’s performance.

The Received Signal Analysis

Consider the $i^{th}$ pseudolite’s signal I and Q components of the $h^{th}$ pseudolite’s group as in [4],

\begin{align}
    s_i^I(t) &= c(n, t) \cos(\omega_h t + \phi_i), \\
    s_i^Q(t) &= d_i(t) p(t) \sin(\omega_h t + \phi_i),
\end{align}

where,

- $c(l, t)$ is the PN code;
- $d_i(t)$ is the data stream coded at $R_d$ data rate;
- $l$ and $n$ correspond to a given $i^{th}$ pseudolite;
- $p(t)$ is an appropriate code [4], which improves the signal cross-correlation properties, resolves the bit timing clock, and reduces the spectral densities. At this stage, our analyses are simplified and do not consider the impact of such a sequence;
- $\omega_h = 2\pi f_h$ with $f_h$ the carrier frequency;
- $\phi_i$ is the initial phase of the signal.

Next, we assume that the pseudolite signal is received at the antenna port through $l$ independent paths; hence, the analytical expression of the received signal is given by,

\begin{align}
    r_i^I(t) &= \sum_{k=1}^{l} A_i^k a_i^k c(n, \tilde{r}_k) \cos(\omega_h \tilde{r}_k + \phi_i^k), \\
    r_i^Q(t) &= \sum_{k=1}^{l} A_i^k a_i^k d_i(\tilde{r}_k) \sin(\omega_h \tilde{r}_k + \phi_i^k),
\end{align}
where,

\[ a_i^k \] is the overall reflection factor of the \( k \)th path starting from the \( i \)th pseudolite to the receiver;

\[ \tilde{t}_k = t - \tau_k^i, \] \( t \) is the time difference (delay) between the \( i \)th pseudolite and the receiver if the signal travels through the \( k \)th path, \( R_i^k(t) \) is the \( k \)th path distance (m), and \( c \) is the speed of light (m/s);

\[ A_i^k = \sqrt{P_i^k}, \] \( P_i^k \) is the power received at the antenna from the \( i \)th pseudolite (W/Hz) defined as [1],

\[ dP_i^k = dP_i + dG_a - 20 \log_{10} \left( \frac{4\pi R_i^k(t) f_h}{c} \right) \text{(dBW/Hz)}, \] (5)

\[ P_i^k = 10 \log_{10} \left( \frac{c^2 G_a P_T}{4\omega^2 k^2 R_i^k(t)^2} \right) \text{(W/Hz)}, \] (6)

\( G_a \) is the antenna aperture gain (no units);

\[ \tilde{\omega}_h = \omega_h \left( 1 + \frac{\tilde{R}_i(t)}{c} \right), \] \( \tilde{R}_i(t) \) is the geometric range rate (m/s) between the \( i \)th pseudolite and the receiver;

\[ \phi_i^k = \phi_i - \phi^k, \] \( \phi^k \) is the initial phase of the \( k \)th path’s received signal.

We recognize that,

\[ \tilde{\omega}_h \tilde{t}_k = \omega_h t - \tilde{\omega}_h \tau_k^i. \] (7)

Hence, if we define,

\[ \tilde{\phi}_i^k = \phi_i^k - \tilde{\omega}_h \tau_k^i, \] (8)

then substituting (7) and (8) into (3) and (4) yields,

\[ r_i^I(t) = \sum_{k=1}^{I} A_i^k a_i^k c(n, \tilde{t}_k) \cos(\tilde{\omega}_h t + \tilde{\phi}_i^k), \] (9)

\[ r_i^Q(t) = \sum_{k=1}^{I} A_i^k a_i^k c(n, \tilde{t}_k) \sin(\tilde{\omega}_h t + \tilde{\phi}_i^k). \] (10)

Assume that the adjacent signal is given by,

\[ r_j^I(t) = \sum_{m=1}^{J} A_j^m a_j^m c(f, \tilde{t}_m) \cos(\tilde{\omega}_h t + \tilde{\phi}_j^m), \] (11)

\[ r_j^Q(t) = \sum_{m=1}^{J} A_j^m a_j^m d_i^j \tilde{t}_m \sin(\tilde{\omega}_h t + \tilde{\phi}_j^m). \] (12)

where,

\[ \hat{t}_m = t - \tau_m^j, \] \( \tau_m^j = \frac{R_m^j(t)}{c} \] is the \( m \)th path’s time difference (delay) between the \( j \)th pseudolite and the receiver, \( R_m^j(t) \) is the \( m \)th path’s distance (m);

\[ \hat{\omega}_v = \omega_v \left( 1 + \frac{\tilde{R}_j(t)}{c} \right), \] \( \tilde{R}_j(t) \) is the geometric range rate (m/s) between the \( j \)th pseudolite and the receiver; \( \phi_j^m = \phi_j^m - \hat{\omega}_v \tau_m^j \) and \( J \) denotes the total number of signal’s paths from the \( j \)th pseudolite to the receiver.

The I and Q components of the total received signal can be written as,

\[ r_i^I(t) = r_i^I(t) + \sum_{j \neq i} r_j^I(t) + \sigma_0 v^I(t), \] (13)

\[ r_i^Q(t) = r_i^Q(t) + \sum_{j \neq i} r_j^Q(t) + \sigma_0 v^Q(t), \] (14)

where,

\[ \sum_{j \neq i} r_j(t) \] denotes the total effect of all the adjacent signals;

\[ \sigma_0 = \sqrt{N_0} \] is the noise standard deviation, \( N_0 \) is the thermal noise PSD (W/Hz);

\( v^I(t) \) and \( v^Q(t) \) are uncorrelated white Gaussian processes with zero mean and unit variance or WGN(0,1).

**Receiver’s Front End and Baseband Sampling**

The received signal of the \( i \)th pseudolite will be downconverted to the baseband and sampled; thus, resulting in the following expression (see Figure 2),

\[ o_i^I(u) = \sum_{k=1}^{I} A_i^k a_i^k c(n, \tilde{u}_k) \cos(\omega_h u + \tilde{\phi}_i^k), \] (15)

\[ o_i^Q(u) = \sum_{k=1}^{I} A_i^k a_i^k d_i^k \tilde{u}_k \sin(\omega_h u + \tilde{\phi}_i^k), \] (16)

where,

\[ \omega_h = 2\pi f_h \hat{R}_i(u) / c. \] (17)

After downconverting and sampling the adjacent signal results in,

\[ o_j^I(u) = \sum_{m=1}^{J} A_j^m a_j^m c(f, \tilde{u}_m) \cos(\omega_h u + \tilde{\phi}_j^m), \] (18)

\[ o_j^Q(u) = \sum_{m=1}^{J} A_j^m a_j^m d_i^j \tilde{u}_m \sin(\omega_h u + \tilde{\phi}_j^m), \] (19)
where,

$$\omega_i^{D} = (\omega_i - \omega_h) \pm \omega_r \frac{\dot{R}_r(t)}{c}. \quad (20)$$

The I and Q components of the total received signal after downconverting and baseband sampling can be written as,

$$o^I (t) = o^I_i (t) + \sum_{j \neq i} o^I_j (t) + \sigma_i v^I (u), \quad (21)$$

$$o^Q (t) = o^Q_i (t) + \sum_{j \neq i} o^Q_j (t) + \sigma_i v^Q (u). \quad (22)$$

**Doppler Removal and Phase Rotation**

During this process Doppler shift is removed and the phase is rotated, such that the desired, received signal should look like (see Figure 2),

$$u^I_i (t) = A^I_i c(n, \tilde{u}_i) + \sum_{k=2}^{l} A^I_k a^I_k c(n, \tilde{u}_k) \cos(\tilde{\phi}^{k,1}_i), \quad (23)$$

$$u^Q_i (t) = \sum_{k=2}^{l} A^I_k a^Q_k d_j (\tilde{u}_k) c(l, \tilde{u}_k) \sin(\tilde{\phi}^{k,1}_i), \quad (24)$$

where,

$$\tilde{\phi}^{k,1}_i = \phi^k - \phi^I_i. \quad (25)$$

The adjacent signal components are transformed as,

$$u^I_j (t) = \sum_{m=1}^{J} A^I_m a^I_m c(n, \tilde{u}_m) \cos(\omega_{i, m}^{D} u + \phi^m_{j, i}), \quad (26)$$

$$u^Q_j (t) = \sum_{m=1}^{J} A^Q_m a^Q_m d_j (\tilde{u}_m) \sin(\omega_{i, m}^{D} u + \phi^m_{j, i}), \quad (27)$$

where,

$$\omega_{i, m}^{D} = \omega_{i}^{D} - \omega_{m}^{D}$$

$$\phi^m_{j, i} = \phi^m_j - \phi^I_i. \quad (28)$$

At this stage, the I and Q components of the total received signal can be written as,

$$u^I (t) = u^I_i (t) + \sum_{j \neq i} u^I_j (t) + \sigma_i v^I (u), \quad (29)$$

$$u^Q (t) = u^Q_i (t) + \sum_{j \neq i} u^Q_j (t) + \sigma_i v^Q (u). \quad (30)$$

**Correlation**

Assuming full correlation for the $i^{th}$ pseudolite’s code yields (see Figure 2),

$$w^I_i (t) = A^I_i + \sum_{k=2}^{l} A^I_k a^I_k c(n, \tilde{u}_k) \cos(\tilde{\phi}^{k,1}_i), \quad (31)$$

$$w^Q_i (t) = \sum_{k=2}^{l} A^I_k a^Q_k d_j (\tilde{u}_k) c(l, \tilde{u}_k) \sin(\tilde{\phi}^{k,1}_i). \quad (32)$$

Similarly, the adjacent signal’s components are determined from,

$$w^I_j (t) = \sum_{m=1}^{J} A^I_m a^I_m c(n, \tilde{u}_m) \cos(\omega_{i, m}^{D} u + \phi^m_{j, i}), \quad (33)$$

$$w^Q_j (t) = \sum_{m=1}^{J} A^Q_m a^Q_m d_j (\tilde{u}_m) c(l, \tilde{u}_m) \sin(\omega_{i, m}^{D} u + \phi^m_{j, i}). \quad (34)$$

The I and Q components of the total signal are given by,

$$w^I (t) = w^I_i (t) + \sum_{j \neq i} w^I_j (t) + \sigma_i v^I (u), \quad (35)$$

$$w^Q (t) = w^Q_i (t) + \sum_{j \neq i} w^Q_j (t) + \sigma_i v^Q (u). \quad (36)$$

**Integration and Dump**

The $i^{th}$ signal components due to only the effect of the direct path, after $NT_r$ accumulation intervals are as follows (see Figure 2),

$$x^I_i (a) = A_i, \quad (37)$$

$$x^Q_i (a) = 0. \quad (38)$$

The $i^{th}$ signal components due to only the effect of the remaining paths, after $NT_r$ accumulation intervals can be written as,

$$x^I_i (a) = \int_0^{NT_r} \sum_{k=2}^{l} A^I_k a^I_k c(n, \tilde{u}_k) \cos(\tilde{\phi}^{k,1}_i) \, du \quad (39)$$

$$x^Q_i (a) = \int_0^{NT_r} \sum_{k=2}^{l} A^Q_k a^Q_k d_j (\tilde{u}_k) c(l, \tilde{u}_k) \sin(\tilde{\phi}^{k,1}_i) \, du \quad (40)$$

Assuming that the Gold or Kasami sequences are used then the above expressions can be written as,

$$x^I_i (a) \leq \frac{2 \text{seq}(m)}{M} \sum_{k=2}^{l} A^I_k |a^I_k|, \quad (41)$$

$$x^Q_i (a) \leq \frac{2 \text{seq}(m)}{M} \sum_{k=2}^{l} A^Q_k |a^Q_k|, \quad (42)$$

where [6],

$$\text{seq}(m) = \begin{cases} k(m) & \text{Kasami} \\ g(m) & \text{Gold} \end{cases} \quad (43)$$

Similarly, the adjacent signal’s components can be written as,

$$x^I_j (a) \leq \frac{2 \text{seq}(m)}{M} \sum_{m=1}^{J} A^I_j \mu^I_m, \quad (44)$$

$$x^Q_j (a) \leq \frac{2 \text{seq}(m)}{M} \sum_{m=1}^{J} A^Q_j \mu^Q_m. \quad (45)$$
\[
\begin{align*}
x^0_j(a) & \leq \frac{2\text{seq}(m)}{M} \sum_{m=1}^{J} A_m^n \left\vert a_m^n \right\vert, \\
x^1_j(a) & = x^1_j(a) + \sum_{j \neq i} x^1_j(a) + \sigma_{NT} \eta^1(a), \\
x^0(a) & = x^0(a) + \sum_{j \neq i} x^0_j(a) + \sigma_{NT} \eta^0(a),
\end{align*}
\]

Hence, yields the total signal’s I and Q components,

\[
y^1(a) = y^1_i(a) + \sum_{j \neq i} y^1_j(a) + \sigma_r \eta^1(a), \\
y^0(a) = y^0_i(a) + \sum_{j \neq i} y^0_j(a) + \sigma_r \eta^0(a).
\]

The average power of the above signal is given by,

\[
\bar{P}_r \equiv E\left\{y^1_i(a)^2 + y^0_i(a)^2 \right\} = \bar{P}_r^1 + \bar{P}_r^0 + 2E\left\{y^1_i(a) \sum_{j \neq i} y^1_j(a) + y^0_i(a) \sum_{j \neq i} y^0_j(a) \right\} + \sigma_r^2,
\]

where,

\[
\begin{align*}
\bar{P}_r^1 & \equiv E\left\{y^1_i(a)^2 + y^0_i(a)^2 \right\}, \\
\bar{P}_r^0 & \equiv E\left\{y^0_i(a)^2 + y^0_i(a)^2 \right\}.
\end{align*}
\]

The expression of the desired, average signal power can be further written as,

\[
\bar{P}_r = \bar{P}_r^1 + \bar{P}_r^0 + 2E\left\{y^1_i(a) y^1_i(a) + y^0_i(a) y^0_i(a) \right\}.
\]

Define the detector’s power degradation ratio as,

\[
dP_{b,i} = 10\log_{10}\left(\frac{\bar{P}_y - \bar{P}_y^i}{\bar{P}_y^i}\right) \leq 10\log_{10}\left[\max_{b} \left\{P_{b,i}^i\right\}\right] (dB).
\]

An upper bound of the above expression can be achieved by applying the upper bounds of the desired signal multipath component and the adjacent signal’s components into the average power expression.

**Tracking Model**

Once, the data bit transition is determined then we can expand the acquisition model to account for the tracking case. Assuming that the data rate, \(R_d\), is 10 times the code repetition rate produces the following desired signal’s components (see Figure 2),

\[
\begin{align*}
z^1_i(a) & = A_i, \\
z^0_i(a) & = 0.
\end{align*}
\]

The expressions for the adjacent signal’s components are determined from,

\[
\begin{align*}
y^1_{ir}(a) & \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{J} A_k^n \left\vert a_k^n \right\vert, \\
y^0_{ir}(a) & \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{J} A_k^n \left\vert a_k^n \right\vert.
\end{align*}
\]
\[ z_{\text{avg}}^O(a) = \frac{\sum_{k=2}^{I} A_k^i a_k^i d_l(u_k) \cos(\phi_k^i)}{T_i}. \]  

(66)

The above signal’s components may reach the following upper-bound,

\[ z_{\text{avg}}^l(a) \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{I} A_k^i |a_k^i|. \]  

(67)

\[ z_{\text{avg}}^O(a) \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{I} A_k^i |a_k^i|. \]  

(68)

The same procedure can be followed to determine the adjacent signal components during the tracking phase in accordance with,

\[ z_{\text{adj}}^l(a) \leq \frac{2\text{seq}(m)}{M} \sum_{m=1}^{J} \sum_{k=2}^{I} A_{m,k}^i |a_{m,k}^i|. \]  

(69)

\[ z_{\text{adj}}^O(a) \leq \frac{2\text{seq}(m)}{M} \sum_{m=1}^{J} \sum_{k=2}^{I} A_{m,k}^i |a_{m,k}^i|. \]  

(70)

Note that only the noise power is reduced during this phase because of the longer averages compared to the acquisition phase. Hence, yields the total signal’s I and Q components,

\[ z^l(a) = z^l(a) + \sum_{j \neq i} z^l_j(a) + \sigma_T \eta^l(a), \]  

(71)

\[ z^O(a) = z^O(a) + \sum_{j \neq i} z^O_j(a) + \sigma_T \eta^O(a). \]  

(72)

The average power of the above signal is given by,

\[ \bar{P}_z \equiv E\left[z^l(a)^2 + z^O(a)^2\right] = \bar{P}_z^l + \bar{P}_z^O + 2E\left[z^l(a)z^O(a)\right] + \sigma_T^2, \]  

(73)

where,

\[ \bar{P}_z^l \equiv E\left[z^l(a)^2\right], \]  

(74)

\[ \bar{P}_z^O \equiv E\left[z^O(a)^2\right]. \]  

(75)

The expression of the desired, average signal power can be further written as,

\[ \bar{P}_z^l = \bar{P}_z^{ll} + \bar{P}_z^{lr} + 2E\left[z^l_1(a)z^l_2(a) + z^O_1(a)z^O_2(a)\right]. \]  

(76)

The transmitted signal power can be further written as,

\[ P_z^{lh} = \frac{\bar{P}_z^{ll} - \bar{P}_z^{lr}}{\bar{P}_z^{ll}} \leq \max\left(P_z^{lh}\right) \] (no units), 

(77)

\[ dP_z^{bh} = 10 \log_{10}\left(P_z^{bh}\right) \leq 10 \log_{10}\left[\max\left(P_z^{bh}\right)\right] \text{ (dB)}. \]  

(78)

An upper bound of the above expression can be achieved by employing the upper bounds of the desired signal’s multipath component and the adjacent signal’s components into the average power expression. In general due to the complexity of calculating the detector’s power degradation ratio (expressions (77) and (78)), in our theoretical performance results we are going to evaluate the upper bound of expressions (77) and (78).

The signal degradation is the inverse measure of the signal to noise ratio. Spilker suggests a simplified estimate of the code and phase error based on the signal to noise ratio [6]. Employing the detector’s power degradation ratio we provide first, the expression for the code error standard deviation (m) as,

\[ \sigma_c = cT_c \sqrt{\frac{P_z^{bh}}{z_{cc}}} \leq cT_c \sqrt{\max\left(P_z^{bh}\right)} = \max\left(\sigma_c\right), \]  

(79)

and second the similar expression for the phase error (m),

\[ \sigma_\phi = \frac{c}{\omega_h} \sqrt{\frac{P_z^{bh}}{z_{cc}}} \leq \frac{c}{\omega_h} \sqrt{\max\left(P_z^{bh}\right)} = \max\left(\sigma_\phi\right). \]  

(80)

Similarly, we have provided a simplified expression for the upper bound of the receiver’s phase and code error, which can be computed relatively easily as opposed to the correct expression of the receiver’s phase and code error. If the upper bounds for the receiver’s phase and code error can achieve good performance then it is conceivable that the receiver can achieve better performance under normal circumstances.

Next, we proceed with the theoretical assessment of the receiver performance when subject to server multipath and near-far problem.

**Theoretical Performance**

This section describes several scenarios, which are designed to depict the theoretical performance of the pseudolite’s signal structure for a noisy environment under severe multipath, geometry, and dynamics constraints.

Two sets of results are created to assess the theoretical performance of the model. For the first set of results the assumptions including the initial assumptions summarized in [5] are provided below:

- The thermal noise PSD = –200 dBW/Hz.
- The carrier frequency of the adjacent signal = 1227.6 MHz.
- Geometric distance (or LOS path) of the desired signal = 100 m and of the adjacent signal = 1 m.
- Doppler of the desired signal = 1 m/s and of the adjacent signal = 3 m/s.
- The transmitted signal power = –100 dBW/Hz.
- Antenna aperture gain = 10 dB.
- Number of paths including the LOS path = 100.
- The reflection coefficients are selected as a random number between (0,1).
- The distance of every path excluding the LOS path is selected as a random variable greater than the LOS distance.

Since the results obtained for this set of assumptions are almost identical with the results presented in [5], we will only comment on them here.

For the second set of results, the assumptions are identical, except that the carrier frequency of the adjacent signal is the same as the carrier frequency of the desired signal. This represents a worst case scenario that both confirms the validity of our signal model and which demonstrates the advantages of using a multiple carrier system. These results are presented and commented on in this paper.

The results that are presented include the effects changing the carrier frequency, code type and length, chipping and data rate, geometry and dynamics, equivalent noise power and the transmitted signal power.

**Carrier Frequency**

The impact of the desired signal’s carrier frequency on the detector’s power degradation ratio and receiver’s phase and code error is presented in Figure 3 and Figure 4 respectively. It appears that the system with single carrier frequency experiences a higher detector power degradation ratio (see Figure 3 of [5]).

For a signal carrier frequency, the receiver’s phase and code error depend on both the code and the carrier frequency (see Figure 4). A 3 mm phase error is observed for a 1575.42 MHz frequency and 1023 Gold code length, which is well known in the navigation community. For a multicarrier system as observed previously, the receiver’s phase remains unchanged at (0.2mm) and the receiver’s code error increases proportionally from 2 cm to 8 cm with the increase of the carrier frequency from 1575.42 MHz to 6598.35 MHz (see Figure 4 of [5]).

**Code Length and Type**

The impact of the code length (M) and type (Kasami or Gold) on the detector’s power degradation ratio and the receiver’s phase and code error (mm and m) for a single frequency system is shown in Figure 3 and Figure 5 respectively. For this system, for the same carrier frequency, the higher the code length is the smaller the phase and code error becomes and vice versa (see Figure 5).

For a multifrequency system the impact of the code length and type is almost unnoticeable (see Figure 5 of [5]).

Table 1: The detector’s power degradation ratio (dB) vs. carrier frequency (MHz) and code length (M) and type (Gold or Kasami).

Figure 3: The detector’s power degradation ratio (dB) vs. carrier frequency (MHz) and code length (M) and type (Gold or Kasami).

Figure 4: The receiver’s phase and code error (mm and m) vs. carrier frequency (MHz).

Figure 5: The receiver’s phase and code error (mm and m) vs. code length (M) and type (Gold or Kasami).
implies the major contribution in the signal separation results from the carrier frequency [5].

A preliminary conclusion reads:

The performance of a multicarrier system can be up to 10 times better than the performance of a single carrier system under the same multipath, geometry, and dynamics conditions. The presence of multipath appears to be insignificant to the receiver’s performance in a multicarrier system.

Figure 6: The detector’s power degradation ratio (dB) vs. chipping rate (Mbps) and data rate (kbps).

Figure 7: The receiver’s phase and code error (mm and m) vs. chipping rate (Mbps).

Chipping Rate

The impact of the chipping rate (Mbps) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single frequency system is presented in Figure 6 and Figure 7 respectively. For the same data rate, the higher the chipping rate the smaller the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) and vice versa.

For a multicarrier system, even in the presence of multipath, the detector’s power degradation ratio seems to remain constant at –43.75 dB during the change of the chipping rate. Under the same circumstances the receiver’s phase error (2 mm) during the change of the chipping rate from 1.023 Mbps to 1.023 Gbps. The receiver’s code error changes inverse proportionally from 20 cm down to 2 mm with the change of the chipping rate from 1.023 Mbps to 1.023 Gbps (see Figure 6 and Figure 7 of [5]).

Again for the same chipping rate, the multicarrier system’s performance can be up to 10 times better than the performance of a single carrier system when subject to the same conditions.

Figure 8: The receiver’s phase and code error (mm and m) vs. data rate (kbps).

Figure 9: The detector’s power degradation ratio (dB) vs. geometric range (distance) (m) and Doppler (m/s).

Data Rate

The impact of the data rate (kbps) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single carrier system is presented in Figure 6 and Figure 8 respectively. For the same chipping rate, the higher the data rate is the higher the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) are.

For a multicarrier system, even in the presence of multipath, the detector’s power degradation ratio seems to remain constant at –43.75 dB during the change of the data rate. Under the same circumstances the receiver’s phase error (2 mm) during the change of the data rate
from 10 bps to 100 kbps (see Figure 6 and Figure 7 of [5]). The receiver’s code error changes only as a result of the change of the chipping rate.

Again for the same data rate, the multicarrier system’s performance can be up to 10 times better than the performance of a single carrier system when subject to the same conditions.

Figure 10: The receiver’s phase and code error (mm and m) vs. geometric distance (m).

Figure 11: The receiver’s phase and code error (mm and m) vs. Doppler (m/s).

Geometry (Distance)

The impact of the adjacent signal’s geometric distance (m) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single carrier system is presented in Figure 9 and Figure 10 respectively. These figures demonstrate the near far problem. For the same Doppler, the higher the difference between the desired signal and receiver distance with the adjacent signal and the receiver distance is the higher the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) are and vice versa (see Figure 9 and Figure 10).

For a multicarrier and multicode system however, it appears that the near-far problem is resolved with or without the multipath. Although the multipath model is included, we note that almost identical receiver’s performance can be achieved (see Figures 9 and 10 of [5]).

The performance of a multicarrier and multicode system can be up to 50 times better than the performance of a single carrier system when subject to the same geometry conditions as shown in Figures 9 and 10.

Figure 12: The detector’s power degradation ratio (dB) vs. equivalent noise power (dBW/Hz) and transmitted signal’s power (dBW/Hz).

Figure 13: The receiver’s phase and code error (mm and m) vs. equivalent noise power (dBW/Hz).

Dynamics (Doppler)

The impact of adjacent signal’s Doppler (m) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single carrier system is presented in Figure 9 and Figure 11 respectively. For the same relative difference distance, the smaller the relative difference Doppler the higher the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) are.

For a multicarrier and multicode system the reader may refer to Figure 9 and Figure 11 of [5].
**Equivalent Noise Power**

The impact of the equivalent noise power (dBW/Hz) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single carrier and multicarrier system is displayed in Figure 12 and Figure 13 respectively.

For a multicarrier and multicarrier system, the receiver’s performance due to the change of equivalent noise power achieved when multipath is present (see Figure 12 and Figure 13 of [5]) changes almost at the same rate compared to the receiver performance of a single carrier system.

**Transmitted Power**

The impact of the pseudolite’s transmitted power (dBW/Hz) on the detector’s power degradation ratio (dB) and receiver’s phase and code error (mm and m) for a single carrier multicarrier system is presented in Figure 12 and Figure 14 respectively.

For a multicarrier and multicarrier system, the receiver’s performance due to the change of transmitted power achieved when multipath is present (see Figure 12 and Figure 13 of [5]) changes almost at the same rate compared to the receiver performance of a single carrier system; however this change is in opposite direction with the change of the equivalent noise power.

![Graph](image)

Figure 14: The receiver’s phase and code error (mm and m) vs. transmitted signal’s power (dBW/Hz).

**Conclusion**

We have analytically investigated the receiver’s performance model by including the multipath effect.

For a multi-carrier and multi-code system, we have provided theoretical performance results that are an order of magnitude better than the performance results for a single carrier system. The first ones can be summarized as follows:

- With and/or without multipath the receiver’s phase error appears to remain constant for any changes of the carrier phase. On the other hand the code error seems to increase proportionally with the increase of the carrier frequency.
- Code length and type have almost no impact on the receiver’s phase and code error.
- Chipping rate appears to impact only the receiver’s code error and having no effect on the receiver’s phase error. Data rate appears to impact neither the receiver’s phase nor the code error.
- Even when multipath is present, the near-far problem imposes no threat to a multi-carrier and multi-code system.
- The increase of the equivalent noise power impacts the receiver’s performance at the same rate as the decrease of the pseudolite’s transmitted signal power and vice versa.

Although the multipath model is included for a multi-carrier and multi-code system, the receiver’s performance appears to be identical to the case when the multipath model was not included.

**References**


An Alternative Approach to Multipath and Near-Far Problem for Indoor Geolocation Systems

I. F. Progri and W. R. Michalson
ECE-WPI, Worcester, MA 01609

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Salt Lake City, UT

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Introduction

- Single frequency Pseudolite applications [1] are limited from near-far and multipath.
- An indoor geolocation system [2] employs 6 wideband receivers with a signal structure different from the GPS.
- Near-far problem to some extent can be eliminated by pulsing the signal [1].
- A higher chipping rate can mitigate the multipath [3] and symmetry may reduce the multipath errors [4].
- Our approach is a more sophisticated technique to deal with multipath and near-far effect.
Previously we have proposed a Pseudolite signal structure [5].

Next, we performed a preliminary investigation, which appears to suggest that it can eliminate the near-far effect [6].

In this paper we will enhance our previous investigation when subject to multipath:

- Theoretical assessment of this system (Pseudolite and receiver)
- Numerical results
- Summary
System Description

- Three groups of Pseudolites with as many as four Pseudolites/group.
- Within the group the frequency is the same and the PRN code is different.
- Every Pseudolite’s transmitted signal power is set to $P_t$.
- Noise, geometry, dynamics and multipath are included.
System Description Cont.

- Since, the proposed system requires more than one carrier to achieve a higher signal separation; hence, minor receiver modifications are required.
- Each receiver is capable of tracking 12 simultaneous signals (three groups of four, four signals of the same group).
The Received Signal Analysis

- The \(i\)\(^{th}\) transmitted signal of the \(h\)\(^{th}\) Pseudolite group

\[
s_i^I(t) = c(n, t) \cos(\omega_h t + \phi_i) \quad s_i^O(t) = d_i(t) p(t) c(l, t) \sin(\omega_h t + \phi_i)
\]

- The received \(i\)\(^{th}\) transmitted signal of the \(h\)\(^{th}\) Pseudolite group and the signal power

\[
r_i^I(t) = \sum_{k=1}^{I} A_i^k a_k^i c(n, \tilde{t}_k) \cos(\tilde{\omega}_h \tilde{t}_k + \phi_i^k) \quad dP_i^k = dP_T + dG_a - 20 \log_{10} \left( \frac{4\pi R_i^k(t) f_h}{c} \right)
\]

\[
r_i^O(t) = \sum_{k=1}^{I} A_i^k a_k^i d_i(\tilde{t}_k) c(l, \tilde{t}_k) \sin(\tilde{\omega}_h \tilde{t}_k + \phi_i^k) \quad P_i^k = 10^{\frac{dP_i^k}{10}} = \frac{c^2 G_a P_T}{4\omega_h^2 R_i^k(t)^2}
\]
The received adjacent signal of $j^{th}$ transmitted signal of the $v^{th}$ Pseudolite group

$$r_j^I(t) = \sum_{m=1}^{J} A_j^m a_j^m c(f, \hat{\phi}_m) \cos(\hat{\omega}_v t + \hat{\phi}_j^m)$$

$$r_j^O(t) = \sum_{m=1}^{J} A_j^m a_j^m d_j^m c(g, \hat{\phi}_m) \sin(\hat{\omega}_v t + \hat{\phi}_j^m)$$

The total signal’s component are

$$r^I(t) = r_i^I(t) + \sum_{j \neq i} r_j^I(t) + \sigma_0 v^I(t)$$

$$r^O(t) = r_i^O(t) + \sum_{j \neq i} r_j^O(t) + \sigma_0 v^O(t)$$

Receiver’s front end and baseband sampling

$$o_i^I(u) = \sum_{k=1}^{I} A_i^k a_k^i c(n, \tilde{\phi}_k^i) \cos(\omega_i^D u + \tilde{\phi}_k^i)$$

$$o_i^O(u) = \sum_{k=1}^{I} A_i^k a_k^i d_k^i c(l, \tilde{\phi}_k^i) \sin(\omega_i^D u + \tilde{\phi}_k^i)$$

$$o_j^I(u) = \sum_{m=1}^{J} A_j^m a_j^m c(f, \tilde{\phi}_m^j) \cos(\omega_j^D u + \tilde{\phi}_m^j)$$

$$o_j^O(u) = \sum_{m=1}^{J} A_j^m a_j^m d_j^m c(g, \tilde{\phi}_m^j) \sin(\omega_j^D u + \tilde{\phi}_m^j)$$

$$o^I(u) = o_i^I(u) + \sum_{j \neq i} o_j^I(u) + \sigma_0 v^I(u)$$

$$o^O(u) = o_i^O(u) + \sum_{j \neq i} o_j^O(u) + \sigma_0 v^O(u)$$
The Received Signal Analysis
Cont.

- Doppler removal and phase rotation

\[ u_i^I(u) = A_i^k c(n, \tilde{u}_I) + \sum_{k=2}^{J} A_i^k a_k^i c(n, \tilde{u}_k) \cos(\phi_i^{k,1}) \quad u_i^Q(u) = \sum_{k=2}^{J} A_i^k a_k^i d_1(\tilde{u}_k) c(l, \tilde{u}_k) \sin(\tilde{\phi}_i^{k,1}) \]

\[ u_j^I(u) = \sum_{m=1}^{J} A_j^m a_m^j c(f, \tilde{u}_m) \cos(\omega_{v,h}^D u + \phi_{j,i}^{m,1}) \quad u_j^Q(u) = \sum_{m=1}^{J} A_j^m a_m^j d_1(\tilde{u}) c(g, \tilde{u}_m) \sin(\omega_{v,h}^D u + \phi_{j,i}^{m,1}) \]

\[ u^I(u) = u_i^I(u) + \sum_{j \neq i} u_j^I(u) + \sigma_0 v^I(u) \quad u^Q(u) = u_i^Q(u) + \sum_{j \neq i} u_j^Q(u) + \sigma_0 v^Q(u) \]

- Correlation

\[ w_i^I(u) = A_i^k + \sum_{k=2}^{J} A_i^k a_k^i c(n, \tilde{u}_I) c(n, \tilde{u}_k) \cos(\phi_i^{k,1}) \quad w_i^Q(u) = \sum_{k=2}^{J} A_i^k a_k^i d_1(\tilde{u}_k) c(l, \tilde{u}_I) c(l, \tilde{u}_k) \sin(\tilde{\phi}_i^{k,1}) \]

\[ w_j^I(u) = \sum_{m=1}^{J} A_j^m a_m^j c(f, \tilde{u}_I) c(f, \tilde{u}_m) \cos(\omega_{v,h}^D u + \phi_{j,i}^{m,1}) \quad w_j^Q(u) = \sum_{m=1}^{J} A_j^m a_m^j d_1(\tilde{u}) c(g, \tilde{u}_I) c(g, \tilde{u}_m) \sin(\omega_{v,h}^D u + \phi_{j,i}^{m,1}) \]

\[ w^I(u) = w_i^I(u) + \sum_{j \neq i} w_j^I(u) + \sigma_0 v^I(u) \quad w^Q(u) = w_i^Q(u) + \sum_{j \neq i} w_j^Q(u) + \sigma_0 v^Q(u) \]
The Received Signal Analysis
Cont.

Integration and dump

\[ x_{i_1}^i(a) = A_i \quad x_{i_r}^i(a) \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{I} A_i^k |a_k|^i \]

\[ x_j^j(a) \leq \frac{2\text{seq}(m)}{M} \omega_{\text{v},h}^{D} \sum_{m=1}^{J} A_j^m |a_m|^j \]

seq(m) = \begin{cases} k(m) & \text{Kasami} \\ g(m) & \text{Gold} \end{cases}

\[ x^i(a) = x_{i_1}^i(a) + \sum_{j \neq i} x_j^j(a) + \sigma_{NT_r} \eta^i(a) \]

\[ x^0(a) = x_{i_1}^0(a) + \sum_{j \neq i} x_j^0(a) + \sigma_{NT_r} \eta^0(a) \]

Acquisition model

\[ y_{i_1}^i(a) = A_i \quad y_{i_r}^i(a) \leq \frac{2\text{seq}(m)}{M} \sum_{k=2}^{I} A_i^k |a_k|^i \]

\[ y_j^j(a) \leq \frac{2\text{seq}(m)}{M} \omega_{\text{v},h}^{D} \sum_{m=1}^{J} A_j^m |a_m|^j \]

\[ y^i(a) = y_{i_1}^i(a) + \sum_{j \neq i} y_j^j(a) + \sigma_{T_r} \eta^i(a) \]

\[ y^0(a) = y_{i_1}^0(a) + \sum_{j \neq i} y_j^0(a) + \sigma_{T_r} \eta^0(a) \]

\[ dP_y^{h,i} = 10 \log_{10} \left( \frac{\bar{P}_y - \bar{P}_y^{ii}}{\bar{P}_y^{ii}} \right) \leq 10 \log_{10} \left( \max[P_y^{h,i}] \right) \]
The Received Signal Analysis
Cont.

- Tracking model

\[ z_{i1}^I(a) = A_i \]
\[ z_{ir}^I(a) \leq \frac{2 \text{seq}(m)}{M} \sum_{k=2}^{I} A_i^k |a_k^i| \]
\[ z_{ir}^O(a) \leq \frac{2 \text{seq}(m)}{M} \sum_{k=2}^{I} A_i^k |a_k^i| \]

\[ z_{j}^I(a) \leq \frac{2 \text{seq}(m)}{M} \sum_{m=1}^{J} A_j^m |a_m^j| \]
\[ z_{j}^O(a) \leq \frac{2 \text{seq}(m)}{M} \sum_{m=1}^{J} A_j^m |a_m^j| \]

\[ z_{i}^I(a) = z_{i1}^I(a) + \sum_{j \neq i} z_{j}^I(a) + \sigma_T \eta_{i}^I(a) \quad z_{i}^O(a) = z_{i1}^O(a) + \sum_{j \neq i} z_{j}^O(a) + \sigma_T \eta_{i}^O(a) \]

\[ P_{z}^{h,i} = \frac{P_z - P_{z1}^{i}}{P_{z1}^{i}} \leq \max(P_{z}^{h,i}) \quad dP_{z}^{h,i} = 10 \log_{10} (P_{z}^{h,i}) \leq 10 \log_{10} \left[ \max(P_{z}^{h,i}) \right] \]

\[ \sigma_c = cT_c \sqrt{P_{z}^{h,i}} \leq cT_c \sqrt{\max(P_{z}^{h,i})} = \max(\sigma_c) \]

\[ \sigma_\phi = \frac{c}{\omega_h} \sqrt{P_{z}^{h,i}} \leq \frac{c}{\omega_h} \sqrt{\max(P_{z}^{h,i})} = \max(\sigma_\phi) \]
Theoretical Performance

- The thermal noise PSD = –200 dBW/Hz.
- The carrier frequency of the adjacent signal = 1575.42 MHz.
- Geometric distance = 100 m.
- Doppler = 1 m/s.
- The transmitted signal power = –100 dBW/Hz.
- Antenna aperture gain = 10 dBW/Hz.
- Number of paths = 100. The reflection coefficients random (0,1). The distance of every path excluding the LOS path is selected as a random variable greater than the LOS distance.
Carrier Frequency

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Code Type and Length

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Chipping Rate

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Data Rate

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Geometric Distance
Doppler

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Equivalent Noise PSD

![Graph showing Equivalent Noise PSD with two plots: one for Equivalent Noise Power (dBW/Hz) and another for Transmitted Signal Power (dBW/Hz). The graphs illustrate the relationship between the variables with different lines representing different conditions or groups.](image)

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Transmitted Power
Conclusions

- For a multi-carrier and multi-code system, we have provided theoretical performance results that are an order of magnitude better than the performance results for a single carrier system.

- With and/or without multipath the receiver’s phase error appears to remain constant for any changes of the carrier phase. On the other hand the code error seems to increase proportionally with the increase of the carrier frequency.

- Code length and type have almost no impact on the receiver’s phase and code error.
Conclusions Cont.

- Chipping rate appears to impact only the receiver’s code error and having no effect on the receiver’s phase error. Data rate appears to impact neither the receiver’s phase nor the code error.
- Even when multipath is present, the near-far problem imposes no threat to a multi-carrier and multi-code system.
- The increase of the equivalent noise power impacts the receiver’s performance at the same rate as the decrease of the pseudolite’s transmitted signal power and vice versa.
Conclusions Cont.

- A higher chipping rate results in a lower code error.
- A combination of code and carrier can overcome the near far problem, which is very disturbing for indoor applications.
- The increase of the noise power detriments the signal to noise ratio at the same rate as the decrease of the transmitted signal power and vise versa.
- By employing more than one carrier frequency, this system is more immune from intentional or non-intentional interference.
References


References Cont.


