An OFDM/FDMA Indoor Geolocation System

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ABSTRACT: An orthogonal frequency division multiplexing (OFDM) frequency division multiple access (FDMA) indoor geolocation system is discussed in this paper. The principle of operation, transmitter, channel, and receiver are discussed. The system’s theoretical performance and simulation results are presented. This system is an attractive alternative for indoor geolocation for the following reasons: (1) it appears that indoor positioning and navigation are achievable to within a few meters if only pseudorange measurements are used, and (2) the receiver design is far simpler than the design of any conventional GPS receiver.

INTRODUCTION

Ultra-wideband (UWB) systems employ narrow, pulsed signals in the time domain (hence, ultra-wideband signals in the frequency domain). The pulse width can be very narrow, which eliminates multipath effects, thus making UWB systems attractive for indoor applications. The time of arrival (TOA) among pulses can be measured to determine the receiver location.

Signal bandwidth is one of the key factors that affects TOA estimation accuracy in multipath propagation environments [1, 2]. In general, larger bandwidth yields better ranging accuracy. UWB systems, which exploit bandwidths in excess of 1 GHz, have attracted considerable attention as a means of measuring accurate TOA for indoor geolocation applications [3]. The use of high-frequency carriers results in high attenuation; therefore, the frequency band considered for a UWB system is typically focused on licensed operation in the 2–3 GHz band [3]. Results of propagation measurement in a typical modern office building have shown that the UWB signal does not suffer multipath fading [3], which is desirable for accurate TOA estimation in indoor areas.

The actual deployment of UWB systems in the United States is subject to the Federal Communication Commission (FCC) approval of February 14, 2002 [4]. The main concern of the FCC is interference among UWB devices, other licensed services, and GPS systems that operate at the 1176.45, 1227.6, and 1575.42 MHz frequency bands [5–7]. Therefore, the FCC has approved the operation of wall-imaging systems below 960 MHz and in the frequency band of 3.1–10.6 GHz [4]. These systems are designed to detect the location of objects contained within a “wall,” such as a concrete structure, the side of a bridge, or the wall of a mine [4]. Operation of these systems is restricted to law enforcement, fire and rescue organizations, scientific research institutions, commercial mining companies, and construction companies [4]. However, given the weak GPS satellite signals that must be processed by GPS receivers, the noise-like UWB signal is still considered harmful for GPS systems in close proximity [5–7]. Thus we were led to propose the modified or spectralized UWB system presented in this paper.

This paper is organized as follows. First, we describe an orthogonal frequency division multiplexing (OFDM)/frequency division multiple access (FDMA) indoor geolocation system. Second, we discuss the channel model. Third, we propose and discuss the transmitter design. Fourth, we propose and discuss the receiver design. Fifth, we analyze the theoretical performance of a two-dimensional (2D) OFDM/FDMA indoor geolocation system. Sixth, we discuss the navigation performance of a 2D OFDM/FDMA indoor geolocation system. Finally, we present conclusions.

SYSTEM DESCRIPTION

An OFDM/FDMA (or spectralized UWB) indoor geolocation system uses a profile for the configuration of the bandwidth allocation, which determines which portions of the spectrum can and cannot be allocated.
The following steps describe the basic idea behind our spectralized UWB (or OFDM/FDMA) system (see Figure 1). First, we divide the high-end UWB spectrum into I blocks of equal bandwidth W. Second, there is a one-to-one mapping between a frequency block and a transmitter. Hence on the one hand, the modulation scheme of the system is FDMA. On the other hand, each frequency block is composed of tones with consecutive spacing of $\Delta f$; thus the other aspect of the modulation scheme is OFDM.

There are two reasons why we propose an OFDM/FDMA indoor geolocation system: (1) there is a large, available spectrum equal to 7.5 GHz, and (2) the FDMA modulation scheme is well known to achieve the smallest cross-channel interference (an example being GLONASS).

Typically, synchronization is achieved using a pseudorandom sequence, as is the case with GPS. This is necessary because the phase information from a single sinusoid is lost as a result of multipath. In the system described below, we employ the characteristics of an OFDM signal to preserve the phase information even in the presence of multipath, thus eliminating the need for a conventional pseudorandom noise (PRN) scheme.

Based on this discussion, we propose an OFDM/FDMA indoor geolocation system that contains I OFDM/FDMA transmitters and one OFDM/FDMA receiver. The channel that enables the signal propagation from a transmitter to the receiver is discussed next.

**CHANNEL MODEL**

The channel is the medium between the transmitting antenna and the receiving antenna. An electromagnetic wave propagating through a channel undergoes a loss of power and dispersion in direction. The ability of the medium to absorb an electromagnetic wave depends on its physical properties, and is known as the channel’s propagation power loss. The power loss in decibels can be inversely proportional with 2, 3, 4, or $n$ times the distance and $m$ times the signal frequency. The path loss, $Q_i(t) = Q_i(R_i(t), f_i)$ in decibels, between the $i$-th transmitter and the receiver can be written as [8]

$$Q_i(R_i(t), f_i)_{\text{dB}} = Q_i(R_0, f_0)_{\text{dB}} + 10 \log_{10} \frac{R_0}{R_i(t)} + 10m \log_{10} \frac{f_0}{f_i} + N(0, \sigma)$$  \hspace{1cm} (1)

where $R_0$ is the reference distance equal to 1 m, $f_0$ is the reference frequency equal to 1575.42 MHz, $R_i$ is the distance between the transmitter and receiver, $f_i$ is the operation frequency, $Q_i(R_0, f_0)_{\text{dB}}$ is the path loss at the reference distance and frequency, and $N(0, \sigma)$ is a zero-mean normally distributed noise with standard deviation of $\sigma$. Typical values of $\sigma$ vary between 0 and 1.

Let $p_i(t)$ represent the ratio of the received power, $P_r(t)$, over the transmitted power, $P_t(t)$, as

$$p_i(t) = \frac{p_i(t)}{p_t(t)} = G_t G_r Q_i(t)$$  \hspace{1cm} (2)

where $G_t$ and $G_r$ denote the transmitter and receiver antenna gain, respectively.

The dispersion of the electromagnetic wave results from nonuniformity of the environment geometry and its absorption properties. The three most common phenomena that perturb an electromagnetic wave are reflection, refraction, and scattering (or diffraction). When there is a direct path between the transmitter and the receiver, the signal is received through the line-of-sight (LOS) path. A signal received through the LOS path loses power because of both the lack of conductivity of the medium and refraction, which occurs as a result of the existence of physical layers with different refractive coefficients. Part of the LOS signal may also be reflected and scattered. When a signal is received through paths different from the LOS path, it is called multipath.

When modeling multipath, we assume that every channel has a fixed number of paths, $L$, and that the properties of every path in the channel are stationary over the longest symbol period. The channel is also assumed to be a slowly varying, frequency-selective, and Rayleigh-fading channel modeled as

$$C_i(t) = \sum_{h=1}^{L} a_{ih} h(t - \tau_{ih}) \exp(-j0_{ih})$$  \hspace{1cm} (3)

where $a_{ih}$, $\tau_{ih}$, and $0_{ih}$ are the $i$-th user $h$-th path gain, time delay, and phase shift, respectively.

![Fig. 1–Spectralized UWB Frequency Allocation](image-url)
The gain $a_i^h$ is Rayleigh distributed, and the phase shift $\theta_i^h$ is uniformly distributed over $[0, 2\pi]$.

**TRANSMITTER DESIGN**

The transmitted signal is a superposition of sinusoids (or tones) equally spaced in the frequency domain. A tone is defined as a narrowband signal whose bandwidth does not exceed 40 kHz. This kind of signal structure is often referred to as OFDM owing to the exact orthogonality of its components over a fundamental period. The generation of an OFDM signal is depicted in Figure 2.

There are $K$ OFDM frequencies (or tones) generated using the same clock; therefore, all OFDM tones have the same initial phase. These tones are combined with the help of an analog combiner. Next, the combined signal is sampled and digitized using an analog-to-digital converter. At this stage, the signal may go through additional filtering and processing, which is not shown in the figure because it is not relevant to the discussion here. Next, the signal is converted back to the analog domain using a digital-to-analog converter and modulated on a radio frequency (RF) carrier. The mathematics of the signal structure is discussed next.

Let $S_i(t)$ be the complex OFDM transmitted signal from the $i$-th transmitter before upconversion, which is defined as

$$S_i(t) = \sum_{k=1}^{K} \exp(j[\omega_k t + \phi_i])$$  \tag{4}$$

where $\omega$ is the signal radian frequency, $\phi_i$ is the initial phase, and the subscript $k$ denotes the frequency component of the OFDM signal (see Figure 2).

The transmitted signal (see equation (4)) is sampled at the sampling frequency, $f_s$, producing the discrete-time signal

$$S_i[n] = \sum_{k=1}^{K} \exp\left(j \left(\frac{f_k}{f_s} n + \phi_i\right)\right)$$  \tag{5}$$

The cross-correlation function, $\gamma(m, \tau)$, of the discrete-time signal $S_i[n]$ with a delayed and conjugated version of itself can be defined as

$$\gamma(m, \tau) = \frac{1}{T} \int_{0}^{T} S_i(t)S_i(t + \tau)^* \, dt$$

$$= \frac{1}{N} \sum_{n=1}^{N-|m|} S_i[n]S_i[n + m + \tau]^*$$  \tag{6}$$

To evaluate equation (6), we compute the expression inside the sum as follows:

$$S_i[n]S_i[n + m + \tau]^* = \sum_{k=1}^{K} \exp\left(j \left(\frac{f_k}{f_s} n + \phi_i\right)\right)$$

$$\times \exp\left(-j \frac{f_k(m + \tau) + \phi_i}{f_s}\right)$$

$$\times \sum_{k=1}^{K} \exp\left(j \left(\frac{(k-1)\Delta}{f_s} n\right)\right)$$  \tag{7}$$

where $\Delta$ is the difference between two sinusoids (or tones).

Substituting the results of equation (7) into equation (6) yields

$$\gamma(m, \tau) = \sum_{h=1}^{K} \exp\left(-j \frac{f_h(m + \tau)}{f_s}\right)$$

$$\times \exp\left(j \left(\frac{(k-1)\Delta}{f_s} n\right)\right)$$  \tag{8}$$

Let $\beta_h$ denote the sum

$$\beta_h = \sum_{k=1}^{K} \left[\frac{1}{N} \sum_{n=1}^{N-|m|} \exp\left(j \left(\frac{(k-1)\Delta}{f_s} n\right)\right)\right]$$  \tag{9}$$

These coefficients $\beta_h$, $h \in \{1, \ldots, K\}$ are independent of the delay $\tau$; therefore, they will be the same for every $\tau$. Henceforth, equation (8) can be written as

$$\gamma(m, \tau) = \sum_{h=1}^{K} \beta_h \exp\left(-j \frac{f_h(\tau + m)}{f_s}\right)$$  \tag{10}$$

We seek a relation between the spacing parameter, $\Delta$, and the sampling frequency, $f_s$, such that the

Fig. 2—An OFDM/FDMA Transmitter [8]
coefficient, $\beta_h$, has the largest possible value and is independent of h. If we assume that

$$\frac{\Delta}{f_s} << \frac{2\pi}{NK} \quad (11)$$

then

$$\beta_h = K \frac{N - |m|}{N}, \forall h \epsilon \{1, \ldots, K\} \quad (12)$$

The above equality makes it possible to write the cross-correlation function (equation (10)) as follows:

$$\gamma(m, \tau) = K \frac{N - |m|}{N} \sum_{h=1}^{K} \exp \left( j \frac{f_h(\tau + m)}{f_s} \right),$$

for $\frac{\Delta}{f_s} << \frac{2\pi}{NK} \quad (13)$

We also seek a relation among the OFDM frequencies, $f_h$, $\forall h \epsilon \{1, \ldots, K\}$; the sampling frequency, $f_s$; and the number of samples, N. If we assume

$$\frac{f_h}{f_s} = \frac{2\pi}{N} \quad (14)$$

then the magnitude of the sum of all the complex exponents in equation (13) will have the shape of a sinc function with maximum occurring for $m = -\tau$. Thus, the magnitude of this cross-correlation function serves as a detector statistic.

We now return to the discussion of the transmitter design. The signal $S_i[n]$ is passed through a digital-to-analog converter; upconverted at the RF frequency, $f_{RF(i)}$; and amplified with gain $\sqrt{2P}$. This procedure yields the RF signal at the transmitter antenna (or at the output of the transmitter)

$$\tilde{S}_i(t) = \sqrt{2P} \sum_{k=1}^{K} \exp[j(\omega_k t + \phi_{RF(i)} + \phi_i)] \quad (15)$$

where P is the signal power, and $\phi_{RF(i)} = 2\pi f_{RF(i)}$ is the RF radian frequency of the i-th transmitter (see Figure 2).

Equation (15) enables us to compute the transmitted signal amplitude and power density in a free-space environment at any point in time and at a distance $R_i(t)$ from the i-th transmitter based only on the transmitter characteristics.

For example, assume that the transmitter has 10 frequencies in the OFDM band, spaced 1 MHz apart. Assume a 500 MHz sampling frequency. Assume that we use N = 256 samples to digitize the signal. Figure 3 illustrates the autocorrelation function $\gamma$ for delay $\tau = (-435.2 \text{ (dashed o)}, 0 \text{ (solid +)}, 435.2 \text{ (dashed ^)})$ ns. The maximum detectable range without ambiguity for this transmitter is

$$d_{max} = \frac{(N - 1)e}{f_s} = T_c$$

$$= 153 \text{ m and } T = \frac{(N - 1)}{f_s} \quad (16)$$

As is shown in Figure 3, when the delay between two sequences is zero, we have the maximum cross-correlation peak. When $\tau = 0.85T = 435.2$ ns, the maximum cross-correlation peak occurs slightly earlier. Also, when $\tau = -0.85T = 435.2$ ns, the maximum cross-correlation peak occurs slightly later. However, the closer the maximum cross-correlation peak is to the center, the smaller this time difference becomes.

**RECEIVER DESIGN**

The block diagram of a generic receiver is presented in Figure 4. As shown, the incoming signal is received through an antenna element, and is amplified, filtered, and downconverted to the intermediate frequency (IF). (In actual implementation, there may be several downconversion stages.) A local oscillator drives the locally generated reference frequency of every mixer. It is conceivable that this frequency will differ both in absolute value and in rate of change from the actual RF signal frequency; therefore, the frequency of the signal after the downconversion will exhibit an error with statistics depending on the local receiver clock and the propagation models. This frequency error is expected to have an impact on the following stages of the receiver. The IF signal is sampled and is either amplified or attenuated, depending on the gain from the automatic gain controller (AGC) (not shown in the figure). The signal is then ready for digital signal processing (DSP) at the digital signal processor. The digital signal processor utilizes a locally generated reference OFDM signal (LGS) to produce an estimate of the time delay, pseudorange, Doppler, carrier phase, and finally navigation solution. The phase of the LGS is expected to be different from the phase of the OFDM signal from the i-th transmitter. Once the time delay has been estimated, a coherent phase lock loop can be used to
track the phase of the OFDM signal from the i-th transmitter, thus improving the time delay estimation and navigation solution accuracy.

Taking propagation effects into account (see equations (1) and 2)), the received noiseless signal is

\[
S_j(t) = \sqrt{2P_f} \sum_{k=1}^{K} d_k(t) \exp(j(\omega_k t + \phi_k)) \tag{17}
\]

\[
P_r(t) = p(t)P_t = Q_i(t)G_jG_p \tag{18}
\]

\[
\hat{R}_i(t) = R_i(1 \pm R_i(t)/c) \tag{19}
\]

where \( \hat{R}_i(t) \) is the geometric range rate in meters/second between the i-th transmitter and the receiver.

We have assumed a one-to-one correspondence between channels and transmitters. Hence, the channel index is denoted by the subscript j, which means that the j-th receiver channel is assigned to track the i-th transmitted signal. The medium-free received signal (equation (17)) going through channel (3) and the receiver’s front end (FE) section produces the received signal, \( r_j(t) \), determined from

\[
r_j(t) = \sum_{i=1}^{I} \int_{-\tau}^{\tau} C_{ik}(\tau) S_i(t-\tau) d\tau + n(t) \]

\[
= \sqrt{2P_f} \sum_{i=1}^{K} \sum_{k=1}^{K} \sum_{h=1}^{K} a_i^h d_i(t-\tau) \exp(j\beta_i^h(t) + n(t) \tag{20}
\]

where

\[
\beta_i^h(t) = \omega_i(t - \tau_i^h) + \phi_i^h + \theta_i^h \tag{21}
\]

The received signal, given by equation (17), goes through several processing stages, which are considered separately. It is important to emphasize that the received signal model is independent of the processing performed by the receiver; that is, the same signal model can potentially lead to a variety of different receivers or receiver designs. Nevertheless, the model discussed here can be used as a baseline for future receiver designs.

**THEORETICAL PERFORMANCE**

The communication performance evaluation includes the receiver’s FE and IF sampling and the DSP. The receiver’s FE and IF sampling consists of downconverting the RF signal into the IF band, then sampling. The DSP consists of cross-correlation estimation in the digital signal processor, which leads to time delay estimation and finally to the navigation or localization solution.

**Receiver’s FE and IF Sampling**

The receiver’s FE section has a two-fold impact on the received signal: a desired impact and an undesired impact. The desired impact consists of downconverting the received signal to the desired IF frequency for further signal processing. The undesired effect consists of introducing thermal or Johnson noise into the received signal [8].

The received signal, \( r_{j;\tau}(t) = r(t) \), is downconverted to the IF frequency and sampled and then digitized (see Figure 5), and the subscript j is no longer used in our analysis. Therefore, the expression for the received signal at the IF band, \( o(t) \), can be written as

\[
o(t) = \sqrt{2PG_a} \sum_{i=1}^{I} \sum_{k=1}^{K} \sum_{h=1}^{K} a_i^h d_i(t-\tau) \exp(j\phi_{i,k,h}) + n_{out}(t) \tag{22}
\]

where

\[
\phi_{i,k,h} = (\omega_i - \hat{\omega}_i)(t - \tau_i^h) + \omega_k(t - \tau_i^h) + \phi_i + \theta_i^h \tag{23}
\]

![Fig. 4–OFDM/FDMA Receiver Block Diagram [8]](image)

![Fig. 5–Block Diagram of the Receiver’s Front End and IF Sampling [8]](image)
and \( G_a \) is the FE amplifier gain. Note that the overhead hat (\(^\hat{\}\)) symbol denotes a locally generated or estimated parameter by the receiver, and \( n_{\text{out}}(t) \) is the output noise with average power given by

\[
N_{\text{out}} = G_F F_a N_{\text{in}} \tag{24}
\]

where \( F_s \) is the noise factor, and \( N_{\text{in}} \) is the average power of the input noise.

To perform some analysis, we must simplify our notation for the received signal given by equation (22). Therefore, we rewrite equation (22) as follows:

\[
o(t) = \sqrt{2PG_a} \sum_{k=1}^{K} a_i d_i(t - \tau_i) \exp j\varphi_{i,k,l} + \sqrt{2PG_a} \sum_{k=1}^{K} \sum_{k=1}^{L} a_h d_h(t - \tau_h) \exp j\varphi_{h,k,h} + \sqrt{2PG_a} n_{\text{out}}(t) \tag{25}
\]

Based on equation (25), we recognize four components (or terms) of the received signal \( o(t) \): the desired component, the intersymbol interference (ISI) term, the mutually accessed interference (MAI) term, and the noise component. For notation simplicity, the above expression can be written as

\[
o(t) = A \exp j[\omega_o(t - \tau) + \phi + 0] + i(t) \tag{26}
\]

where

\[
\phi = \phi_i, \quad \theta = \theta_i, \quad \tau = \tau_i, \quad \omega_o = \omega_{o_i} = \omega - \omega_i,
\]

and \( A = \sqrt{2PG_a} a_i d_i(t - \tau) \tag{27} \)

and \( i(t) \) is the interference signal, which is assumed to be normally (or Gaussian) distributed, keeping in mind the central limit theorem and the filtering that occurs after the mixer (or downconversion).

The signal \( o(t) \) is sampled at the rate (or sampling frequency) \( f_s \), thus producing a digital signal

\[
o[n] = A \exp j \left[ \frac{f_s}{f_s} (n - \tau) + \phi + \theta \right] + i[n] \tag{28}
\]

The signal \( o[n] \) can be written more simply as

\[
o[n] = A \exp j \left[ \frac{f_s}{f_s} (n - \tau) + \theta \right] S[n - \tau] + i[n] = \hat{S}[n - \tau] + i[n] \tag{29}
\]

The signal given by equation (29) is the first input into the digital signal processor, and the locally generated signal is the second input. These two inputs are processed to form the desired signal detection function as shown in Figure 4, the output of which is the time delay estimation.

**Digital Signal Processing**

The concept behind the processing that occurs in the digital signal processor is similar to that discussed for the transmitter. The detection statistics on the receiver are based on the cross-correlation function, \( \gamma(m, \tau) \), between the received signal, \( o[n] \), and the LGS signal, \( S[n + m] \):

\[
\gamma(m, \tau) = \frac{1}{N} \sum_{n=1}^{N-|m|} o[n] S[n + m]^* = \frac{1}{N} \sum_{n=1}^{N-|m|} i[n] S[n + m]^* \tag{30}
\]

The cross-correlation function at the receiver contains both desired and undesired components. The desired component results from considering the multiplication

\[
\hat{S}[n - \tau] S[n + m]^* = A \exp j\theta \exp j \left[ \frac{f_{ER}}{f_s} \tau \right] \exp j \left[ \frac{f_{ER}}{f_s} n \right] \sum_{k=1}^{K} \exp \left[ -j \frac{f_k(m + \tau)}{f_s} \right]
\]

The undesired component is obtained from the multiplication between \( i[n] \) and \( S[n + m] \) as

\[
v[m] = \frac{1}{N} \sum_{n=1}^{N-|m|} i[n] S[n + m]^* \tag{32}
\]

Substituting the results of equations (31) and (32) into equation (30) yields

\[
\gamma(m, \tau) = A \exp j\theta \exp j \left[ \frac{f_{ER}}{f_s} \tau \right] \sum_{k=1}^{K} \exp \left[ -j \frac{f_k(m + \tau)}{f_s} \right] \sum_{k=1}^{K} \exp \left( j \frac{(k - 1)\Delta + f_k}{f_s} n \right)
\]

We observe three major distortions in the cross-correlation function, \( \gamma(m, \tau) \): (1) multipath, (2) frequency error, and (3) receiver noise and interference. Under assumptions similar to those applied in the transmitter design, the receiver detection function \( \gamma(m, \tau) \) can be approximated as

\[
\gamma(m, \tau) \approx K \frac{N-|m|}{N} A \exp j\theta \exp j \left[ \frac{f_{s}}{f_s} \tau \right] \sum_{k=1}^{K} \exp \left( j \frac{f_k(m + \tau)}{f_s} \right) + v[m],
\]

for \( \Delta + \frac{f_k}{f_s} \leq \frac{2\pi}{NK10^2} \tag{34} \)
QUANTITATIVE ASSESSMENT

Although, as an example, we consider a 2D OFDM/FDMA indoor geolocation system depicted in Figure 6, 3D OFDM/FDMA indoor geolocation systems can be considered as well, as we plan to do in future publications. The 2D system consists of three transmitters and one receiver, all positioned in the same line. The receiver location is defined as the origin of the system. The coordinates of the first, second, and third transmitters are \(-130.47\) m, 0 m, and 130.47 m, respectively. The receiver is, however, located at 0 m.

We have assumed that the transmitted signal power is 0 dB. Each signal contains 10 sinusoids with 1 MHz spacing between 2 consecutive sinusoids. Therefore, the bandwidth of each IF OFDM signal is 10 MHz. The first signal is delayed in time by 435.2 ns, corresponding to a distance of \(-130.47\) m. The second signal is generated in true time, corresponding to a distance of 0 m. The third signal is advanced by 435.2 ns, corresponding to a distance of 130.47 m. In reality, there is neither a negative distance nor a negative time. However, to distinguish between the first and second transmitters, we have made such assumptions. For the true principle of operation of such a system, the reader should refer to [8]. The purpose of this quantitative assessment is only to show how an observable can be obtained and what its accuracy would be.

If we were to cross-correlate the three IF signals with an OFDM signal, normalize the cross-correlation function, and plot the cross-correlation function versus time, we would obtain the result shown in Figure 3. As shown in the figure, each cross-correlation function peaks at \(\tau = \{-435.2\) (dashed o), 0 (solid +), 435.2 (dashed ^)\}. Therefore, the cross-correlation peak would indicate the “distance” between each transmitter and the receiver.

Figure 7 illustrates the total RF signal in the time and frequency domains. Each IF OFDM signal is unconverted at 3.105 GHz, 3.125 GHz, and 3.145 GHz, respectively. As shown in Figure 7, two consecutive RF frequencies are 20 MHz apart; hence, there is little overlap between the waveforms of two separate transmitters. As indicated above in the system description, this is known as the OFDM/FDMA modulation scheme, and this waveform is crucial for achieving channelization between a transmitter and a channel of the receiver. Next, we consider the effect of channel losses on this transmitted signal.

We assume that the power spectrum density of the normally distributed interference is 10 dB above the signal power, which is 0 dB. Hence, the received signal at the receiver’s antenna port looks like the waveform shown in Figure 8.

A snapshot of the cross-correlation \(\gamma(m, \tau)\) at the receiver side is shown in Figure 9. As a result of the multipath error, the frequency error, and the receiver noise, the cross-correlation function at the receiver is distorted (see Figure 9) compared with the ideal cross-correlation function at the transmitter (see Figure 3). How accurately we estimate the cross-correlation peak from the center is discussed below.

Navigation Performance Evaluation

Our system is very different from the conventional GPS-like signal structure because it does not use a pseudorandom sequence. Therefore, we should look more carefully at the concept of measurement generation and navigation solution using an OFDM/FDMA indoor geolocation system. Three topics are...
discussed below: (1) time delay estimation, (2) pseudo-range error, and (3) navigation performance.

**TIME DELAY ESTIMATION**

The time delay, \( \tau \), is defined as the sum of the signal time of travel (or arrival), \( \tau_o \), and the transmitter clock bias, \( \tau_t \) (or offset from the reference time) and receiver clock bias, \( \tau_r \) (or offset from the reference time), which reads analytically

\[
\tau = \tau_t + \tau_o + \tau_r
\]  

(35)

If we assume that we can achieve some means of synchronization between transmitters (using either a master GPS receiver, a local area network [LAN], or some other means), then the transmitter clock bias, \( \tau_t \), is estimated separately to enable computation of the navigation solution.

Consider the numerical example presented in the discussion of the transmitter. The time delay can be measured without ambiguity if the maximum distance between the transmitter and receiver is less than 153.6 m (see equation (17)). Comparing the results of Figure 9 with those of Figure 3, we conclude that the location of the cross-correlation peaks is preserved to a large extent. We also observe that the actual distance (in time) of the cross-correlation peak from the center does not reflect the true delay because of the nonlinear nature of the cross-correlation function.

The algorithm for computing the time delay is provided in the following steps:

1. Initialize the transmitter and channel models.
2. Compute the LGS at its own local time \( t \).
3. Set \( m = 0 \) and \( \tau = 0 \).
4. Set (step index) index = 0.
5. While \( (m - N - 1) > 1 \) or index < 3:
   6. Compute the cross-correlation function \( \gamma(m, \tau) \) as indicated by equation (34).
   7. Find the maximum value, \( \gamma_{max} \) and its corresponding index, \( m \).
   8. Compute the intermediate time delay, \( d = (m - N - 1)/f_s \).
   9. Compute the actual time delay, \( \tau = \tau + d \).
10. Update the local time, \( t = t + \tau \).
11. Compute the LGS using the new time \( t \).
12. Increment the step index, index = index + 1.
13. End.

Output the time delay.

This algorithm is supposed to converge in no more than two or three steps. If it does not converge by the third step, the computational loop of the algorithm should be interrupted.

Since the time delay estimates are statistical parameters, we conducted the experiment 1,000 times and plotted the cumulative distribution functions in Figure 11. We also computed the sample mean

\[
\mu_\tau = [-435.53, 0.4, 435.03] \text{ns}
\]  

(36) and standard deviation. This result appears to be consistent with those of previous publications indicating that the root-mean-square (RMS) delay
spread for indoor wireless systems is in the range of
tens of nanoseconds [9–13].

Time delay estimation is the first important mea-
surement for producing the navigation solution. As
discussed in the following section, this measurement
can be interpreted as a pseudorange measurement.

\[ \sigma_r = \{8.4, 0.8, 0.3\}\text{ns} \quad (37) \]

**PSEUDORANGE ESTIMATION AND
PSEUDORANGE ERROR**

By definition, the pseudorange measurement, \( \rho \), is the
quantity that results from multiplication of the
time delay estimation (see equation (35)) by the
speed of light, \( c \), i.e.,

\[ \rho = c \cdot \tau = c \cdot \tau_0 + c \cdot \tau_r \quad (38) \]

On the other hand, if we write the time delay as

\[ \tau = \mu_r \pm \sigma_r \quad (39) \]

then the pseudorange error based on equation (38) reads

\[ \rho = c \cdot \tau = c \cdot \mu_r \pm c \cdot \sigma_r = \mu_p \pm \sigma_p \quad (40) \]

Therefore, the pseudorange sample mean, \( \mu_p \), and
standard deviation, \( \sigma_p \), for the experiment under
investigation are

\[ \mu_p = \{-130.6, 0.12, 130.9\}\text{m} \]

and

\[ \sigma_p = \{2.5, 0.24, 0.08\}\text{m} \quad (41) \]

Note that the pseudorange error is independent of the
time delay estimation; i.e., whether the time
delay estimation is positive, zero, or negative, the
pseudorange error is almost 2–3 m. Nevertheless,
the pseudorange error is very large. Therefore, we
should seek alternative measurements to ensure
precise indoor geolocation. One such measurement
results from estimating the phase of the carrier,
which would be the subject of future work.

The cumulative distribution function of the pseudo-
range error is plotted in Figure 11. We observe more
error associated with the first observable than with
the second and third.

**CONCLUSIONS**

Several lessons have been learned from this
investigation. First, we have shown that it is pos-
sible to design an OFDM/FDMA indoor geolocation
system. Second, the advantages of using this system
are as follows:

- There is a 7.5 GHz bandwidth available for
  such systems.
- The design of an OFDM/FDMA indoor geolocation
  system may appear comparable to the design of a
  GPS-like indoor geolocation system. An implementa-
  tion of this system will validate this speculation.

- The pseudorange error of an OFDM/FDMA
  indoor geolocation system is about 2.3 m (1 \( \sigma \)),
  which is much better than the accuracy a GPS-
  like indoor geolocation system.

Nevertheless, there are also some disadvantages to
an OFDM/FDMA indoor geolocation system. In par-
icular, the theoretical maximum detectable range
depends on the number of OFDM tones and on the
sampling frequency (see equation (16)).

Finally, such systems are envisioned as both 3D
and 2D. Further development and implementation of
3D OFDM/FDMA indoor systems will be presented
in future publications.

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