Multi-Differential Slightly Frequency-Shifted Reference Ultra-wideband (UWB) Radio

Qu Zhang and Dennis L. Goeckel
Department of Electrical and Computer Engineering
University of Massachusetts at Amherst
Amherst, Massachusetts 01003-9292
Email: {quzhang, goeckel}@ecs.umass.edu

***Abstract* - The frequency-shifted reference (FSR) ultra-wideband (UWB) communication scheme has been recently proposed by the authors. The key idea of the FSR-UWB system is to employ a reference signal that is slightly shifted in frequency from the data-bearing signal, and it has been shown that this results in a very simple receiver architecture. In particular, such a scheme obviates the need for the delay element that greatly complicates implementation of the receiver in standard transmitted reference UWB (TR-UWB) systems. In this paper, we propose a multi-differential FSR-UWB system, where multiple data carriers employ a single reference carrier. This modification essential increases the number of (differential) degrees of freedom available for signaling in the system. However, unlike most communication systems that provide such a dimensionality increase, the large ratio between the UWB system bandwidth and the carrier separation allows the multi-differential FSR-UWB to achieve this significant increase in signal space dimensionality over the standard FSR-UWB system with only a negligible increase in bandwidth. After a general performance characterization, applications to parallel data signaling, multi-dimensional signaling, and narrowband interference cancellation are considered that demonstrate the utility of the proposed scheme.**

I. INTRODUCTION

The extremely large bandwidth of ultra-wideband (UWB) systems suggests a number of potential advantages over conventional narrowband communication systems, including low power consumption, high multiple-access capacity and the potential capability to achieve a high transmission rate and diversity against multipath [1]. However, the enormous number of resolvable paths of the fading channel greatly complicates the receiver design. In particular, the conventional rake receiver exhibits significant complexity and difficult channel estimation [2].

Transmitted reference (TR) UWB systems have been proposed to address the timing and channel estimation problems [3]. In a standard TR-UWB system, two interleaved pulse trains are transmitted. One pulse train carries the information, while the other serves as the reference. Since adjacent reference and data pulses pass through nearly the same channel, the reference provides the receiver a template of the channel-distorted data signal. Thus, receiver complexity is greatly reduced by avoiding explicit channel estimation.

However, the receiver implementation of TR-UWB systems is still challenging. In particular, the implied delay element, which must handle a wideband analog signal, is difficult to build in a low power integrated circuit [4]-[6]. To relax this problem, slightly frequency-shifted reference (FSR) UWB has been proposed [7], [8], where the reference is translated slightly in frequency rather than time. The frequency offset between the reference and data signal is only the inverse of the symbol interval. Thus, for low data rate applications, this frequency shift is well below the coherence bandwidth of the channel, and, thus reference and data signals that are orthogonal over a symbol period go through nearly the same channel. Besides the reduction in receiver complexity, FSR-UWB also outperforms the standard TR-UWB system in terms of bit error rate (BER) [7].

In this paper, a multi-differential (MD) FSR-UWB system is proposed. In the proposed scheme, K data signals sharing one common reference are transmitted in parallel. Each data signal is a slightly frequency-shifted version of the reference. The data carrier frequencies are carefully chosen such that the reference signal and all data signals are orthogonal to each other over the symbol period. In low data rate applications, the frequency offsets are still well below the channel coherence bandwidth, so the reference and data signals go through nearly the same channel. Attempts to employ multiple data signals in standard TR-UWB systems were proposed in [9], [10]. However, these improved TR-UWB schemes extend the delay of the original TR-UWB system, thus exacerbating its main implementation difficulty. The additional delays also lead to longer frame lengths and, hence, a lower data rate. In contrast, the proposed scheme provides the same increase in differential degrees of freedom without lengthening the frame period. Moreover, note that for the proposed scheme, the spectra of the reference and the data signals significantly overlap; thus, since the frequency offsets are much smaller than the signal bandwidth, the total bandwidth of the system is just slightly increased compared to the original single-differential (SD) FSR-UWB scheme. It is shown in this paper that the proposed scheme provides a transmission rate that is higher than SD FSR-UWB while outperforming SD FSR-UWB systems in error performance. Moreover, the proposed MD FSR-UWB can also provide a large-dimensional signal space that allows for the mitigation of narrowband interference.

The system model is described in Section II. In Section III, the receiver output of the proposed scheme is analyzed. Applications of the proposed scheme are presented and analyzed in Section IV. Numerical results are presented in Section V, and the conclusions follow.

II. SYSTEM MODEL

Throughout this paper, a baseband low data rate UWB system is assumed. Data is transmitted during a symbol interval of length T_s consisting of $N_f \gg 1$ frames, each of length T_f. Define the regular pulse train

$$u(t) = \sum_{n=0}^{N_f-1} p(t - nT_f), \tag{1}$$

where $p(t)$ is the UWB pulse shape with approximate width T_p, bandwidth W and energy $\frac{1}{N_f}$. It is assumed that $\frac{T_f}{T_p} \gg 1$. In implementation, the pulse train would be dithered to improve the spectral characteristics, but it can be shown that this does not impact the performance analysis.

For an MD FSR-UWB system with K carriers, the transmitted signal over interval $[lT_s, (l+1)T_s]$ is given by

$$x(t) = u(t - lT_s)g(t - lT_s), \tag{2}$$

where

$$g(t) \triangleq \left(\sqrt{E_p} + \sum_{k=0}^{K-1} c_k \sqrt{2E_d^{(k)}} \cos(2\pi f_k t)\right), \tag{3}$$

$\mathbf{c} \triangleq [c_0 \ \ c_1 \ \ \ldots \ \ c_{K-1}]^T$ is the transmitted symbol, the superscript ^{T} denotes matrix transpose, and E_p and $E_d^{(k)}$ are the energy allocated on the reference signal and the k^{th} data signal, respectively.

Define the index set of the carrier frequencies as

$$\mathcal{I} \triangleq \{0, 1, ..., K-1\}.$$

The carrier frequency of the k^{th} data signal is defined as $f_k = (2k+1)\frac{1}{T_s}$, $k \in \mathcal{I}$. To ensure that the reference signal passes through a channel nearly the same as the corresponding data signals, f_{K-1} should be much less than the channel's coherence frequency $(\Delta f)_c$. Since a small or moderate K is assumed in our application, this constraint is easily satisfied.

The original SD FSR-UWB system uses one differential degree of freedom in a bandwidth of $2(W + f_0)$, where $W \gg f_0$. The proposed scheme provides K differential degrees of freedom in a bandwidth of $2(W + (2K+1)f_0)$. As in many applications, it will be shown that these additional degrees of freedom provide the ability to vastly improve system performance. However, unlike many applications, since $W \gg f_0$, the proposed scheme increases the degrees of freedom with only a slight increase in the system bandwidth.

As shown in [7], for a narrow band signal $f(t)$,

$$\int_{lT_s}^{(l+1)T_s} u^2(t - lT_s) f(t) dt \approx \frac{1}{T_s} \int_{lT_s}^{(l+1)T_s} f(t) dt.$$

Then the energy of the transmitted signal per symbol is easily calculated as

$$\begin{aligned} E_s &= \int_{lT_s}^{(l+1)T_s} x^2(t) dt \approx \frac{1}{T_s} \int_{lT_s}^{(l+1)T_s} g^2(t) dt \\ &= E_p + \sum_{k=0}^{K-1} c_k^2 E_d^{(k)}. \end{aligned} \tag{4}$$

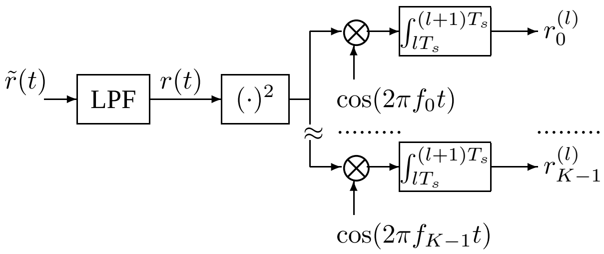

Fig. 1. Receiver structure of a K-carrier FSR-UWB system, where $\tilde{r}(t)$ is the received signal, and $r(t)$ is a lowpass-filtered version of $\tilde{r}(t)$.

The receiver structure is shown in Fig. 1. Note that the receiver complextiy is roughly linear in K, and, hence, the essence of the simplicity of the receiver from [7] is maintained. The received signal is given by $\tilde{r}(t) = s(t) + \tilde{n}(t)$, where $\tilde{n}(t)$ is a zero mean Gaussian process with two-sided power spectral density $S_{\tilde{n}}(f) = \frac{N_0}{2}$, $s(t) \triangleq h(t) * x(t)$, and $h(t)$ is the channel impulse response. Suppose the front-end lowpass filter (LPF) of bandwidth W and frequency response $H(f)$ does not distort $s(t)$; then, the output of the LPF is

$$r(t) = s(t) + n(t),$$

where $n(t)$ is a zero-mean Gaussian process with power spectral density $S_n(f) = \frac{N_0}{2}|H(f)|^2$.

The output of the receiver can be expressed by a K-by-1 vector $\mathbf{r}^{(l)} \triangleq [r_0^{(l)} \ r_1^{(l)} \ \ldots \ r_{K-1}^{(l)}]^T$, where

$$r_k^{(l)} = \int_{lT_s}^{(l+1)T_s} r^2(t) \cos(2\pi f_k t) dt$$

is the output of the k^{th} branch during the l^{th} symbol interval. It is important to note that, although the input signals to the receiver in Fig. 1 are of high bandwidth, $\mathbf{r}$ only needs to be sampled at the symbol rate. Hence, any fairly complex manipulations of $\mathbf{r}$ that are required (e.g. in Section IV-C below) can be performed digitally.

III. RECEIVER ANALYSIS

Without loss of generality, the analysis considers the 0^{th} symbol interval, and the symbol index is suppressed for simplicity. The received vector $\mathbf{r}$ can be rewritten as

$$\mathbf{r} = \mathbf{s} + \mathbf{n}, \tag{5}$$

where the desired signal vector $\mathbf{s} \triangleq [s_0 \ \ s_1 \ \ \ldots \ \ s_{K-1}]^T$ has components given by

$$s_k = \int_0^{T_s} s^2(t) \cos(2\pi f_k t) dt, \tag{6}$$

and the components of the noise vector $\mathbf{n} \triangleq [n_0 \ n_1 \ \ldots \ n_{K-1}]^T$ are given by

$$n_k = n_{k,0} + n_{k,1},$$

where

$$\begin{cases} n_{k,0} \triangleq \int_0^{T_s} 2s(t)n(t) \cos(2\pi f_k t) dt \\ n_{k,1} \triangleq \int_0^{T_s} n^2(t) \cos(2\pi f_k t) dt. \end{cases} \tag{7}$$

A. Additive White Gaussian Noise (AWGN)

On AWGN channels, $s(t) = x(t)$. It is trivial to show that

$$s_k \approx \frac{1}{T_s}\int_0^{T_s} g^2(t)\cos(2\pi f_k t)dt = \sqrt{2E_p E_d^{(k)}}c_k.$$

Following the arguments of previous work [3], the noise term n_k is approximately Gaussian with zero mean. By following the derivation in [8] and applying trigonometric identities, it can be shown that n_k has conditional variance

$$\mathrm{E}\{n_k^2|\mathbf{c}\} = \sigma_0^2 + \sigma_1^2$$

where

$$\sigma_1^2 \triangleq \mathrm{E}\{n_{k,1}^2|\mathbf{c}\} = \frac{1}{2}N_0^2 T_s W,$$

and

$$\begin{aligned}\sigma_0^2 &\triangleq \mathrm{E}\{n_{k,0}^2|\mathbf{c}\}\\ &\approx \frac{2N_0}{T_s}\int_0^{T_s} g^2(t)\cos^2(2\pi f_k t)dt\\ &= N_0\Bigg(E_p + \frac{E_d^{(k)}c_k^2}{2} + \sum_{n=0}^{K-1}E_d^{(n)}c_n^2 + \\ &\quad \frac{1}{2}\sum_{\substack{m=0\\(m,n)\in\mathcal{A}_k}}^{K-1}\sum_{n=0}^{K-1}c_m c_n\sqrt{E_d^{(m)}E_d^{(n)}}\Bigg),\end{aligned}$$

where $\mathcal{A}_k$ is defined as a set of pairs (m,n) as follows:

$$\begin{aligned}\mathcal{A}_k \triangleq\ &\{(m,n)\},\ \text{where } m,n\in\mathcal{I},\ m\neq n, \text{ and}\\ &m+n=2k \text{ or } m-n=2k \text{ or } m-n=-2k.\end{aligned}$$

B. Multipath (MP)

A discrete-path model is assumed in this paper. Express the channel impulse response as

$$h(t) = \sum_{l=0}^{L-1}\alpha_l\delta(t-\tau_l), \tag{8}$$

where L is the number of paths, α_l is the amplitude of the l^{th} path, and τ_l represents the delay of the l^{th} path. For analytic convenience, it is assumed that the path delay between two adjacent paths is no less than T_p so that there is no inter-path interference. Define

$$\begin{cases}\gamma \triangleq \sum_{l=0}^{L-1}\alpha_l^2\\ \gamma_k \triangleq \sum_{l=0}^{L-1}\alpha_l^2\cos(2\pi f_k\tau_l),\ \text{for } k\in\mathcal{I}.\end{cases}$$

Then it is straightforward to show that the signal component is given by

$$s_k = \sqrt{2E_d E_p^{(k)}}c_k\gamma_k,$$

and the noise variance given the MP channel and transmitted symbol is

$$\mathrm{E}\{n_k^2|\mathbf{c},\mathbf{h}\} = \sigma_0^2\gamma + \sigma_1^2.$$

IV. Applications

A. Parallel Transmission

First it is shown that this MD FSR-UWB system can provide significantly higher data rate transmission than the SD FSR-UWB system proposed in [7], while still outperforming SD FSR-UWB in terms of bit error rate.

During each symbol interval T_s, K bits are transmitted simultaneously. Suppose the c_k are independently and identically distributed (i.i.d.) symbols, each corresponding to a single data bit with equal probability to be -1 or $+1$. On AWGN channels, the received signal energy at the output of the k^{th} branch of the receiver is

$$E_{r,k} = \sqrt{2E_p E_d^{(k)}}.$$

At low-to-moderate SNRs (which yield the error rates typically of interest), $n_{k,1}$ dominates the receiver performance, and the optimal partitioning of the transmitted symbol energy E_s to maximize the received symbol energy results in

$$E_p = \frac{E_s}{2};\ E_d^{(k)} = \frac{E_s}{2K}\ \text{for all } k\in\mathcal{I}. \tag{9}$$

Thus, the optimal symbol energy splitting results in the identical transmitted bit energy $E_b = \frac{E_s}{K}$ and identical received bit energy $E_{r,k} = \sqrt{\frac{K}{2}}E_b$. For simplicity, we use (9) in later notations.

Using the Gaussian assumption for the variance of the noise given $\mathbf{c}$, the average BER can be calculated as

$$P_b = \mathrm{E}_{\mathbf{c}}\left\{Q\left(\sqrt{\frac{E_{r,k}}{\mathrm{E}\{n_k^2|\mathbf{c}\}}}\right)\right\}, \tag{10}$$

which is complicated by the dependence of the conditional variance on the value of $\mathbf{c}$. However, we have observed that taking the expectation of the variance of the noise over $\mathbf{c}$ (i.e., assuming essentially that the *unconditional* noise is Gaussian) results in an accurate BER characterization. In this case, the identical unconditional variances are given by

$$\sigma^2 = \mathrm{E}_{\mathbf{c}}\{\mathrm{E}\{n_k^2|\mathbf{c}\}\} = \frac{1}{2}N_0^2 T_s W + N_0 E_b(K+\frac{1}{4}).$$

Hence, each bit has an identical BER

$$P_b = Q\left(\frac{E_b\sqrt{K}}{\sqrt{N_0^2 T_s W + 2N_0 E_b(K+\frac{1}{4})}}\right). \tag{11}$$

Over MP channels, it is straightforward to show that the average bit error rate for the k^{th} bit is

$$P_{b_k} = \mathrm{E}_{\mathbf{h}}\left\{Q\left(\frac{E_b\gamma_k\sqrt{K}}{\sqrt{N_0^2 T_s W + 2N_0 E_b(K+\frac{1}{4})\gamma}}\right)\right\}. \tag{12}$$

When the symbol duration T_s is much greater than the longest path delay and K is not large, the γ_k are almost the same for all $k\in\mathcal{I}$, and, thus, each bit achieves nearly identical performance.

The comparison of (11) and (12) with [7] shows that, at low-to-moderate SNRs, which yield the error rates typically of interest, the "noise-cross-noise" term $n_{k,1}$ dominates the error performance. The proposed scheme demonstrate a significant $10\log_{10}\sqrt{K}$dB gain over SD FSR-UWB due to the fact that the variance of $n_{k,1}$ does not depend on K. This gain is significantly more than that obtained by simply amortizing the reference energy over multiple data symbols, as has been done in standard TR-UWB [9]. Note that, although it looks like the gain increases indefinitely in K, for very large K, the "signal-cross-noise" term $n_{k,0}$ will start to dominate the error performance. Moreover, bear in mind that the proposed scheme provides a transmission rate that is K times higher than SD FSR-UWB with only a negligible increase in bandwidth.

B. General M-ary Signaling

A generalized TR-UWB system is proposed in [10], where M-ary signaling is employed to improve performance. However, besides the drawbacks discussed in Section I, the optimal signal design in the generalized TR-UWB of [10] is still an open question. In this section, it is shown that the proposed scheme can also employ M-ary signaling to improve the performance, and, at low-to-moderate SNRs, the optimal decision can be made by choosing the scaled signal point closest in Euclidean distance to the receiver output vector.

In each symbol interval, the transmitted signal $x(t)$ is independently chosen from the M-ary signal set $\{x_m(t) : m = 0, 1, ..., M-1\}$ with equal probability. Each possible transmitted signal is given by $x_m(t) = u(t)g_m(t)$, where

$$g_m(t) = \sqrt{\frac{E_s}{2}} + \sqrt{E_s}\sum_{k=0}^{K-1} c_{m,k}\cos(2\pi f_k t).$$

Hence, the transmitted code for the m^{th} signal is defined as $\mathbf{c}_m = [c_{m,0}\ c_{m,1}\ ...\ c_{m,K-1}]^T$. Again, the received vector can be written in the form of (5). At the receiver, the received vector will be processed jointly to make decisions.

When signal $x_m(t)$ is transmitted, the received signal vector can be expressed as

$$\mathbf{s} = \frac{E_s\gamma_k}{\sqrt{2K}}\mathbf{c}_m.$$

As argued before, when the UWB symbol duration T_s is much greater than the maximum multipath delay, and the number of carriers K is small or moderate, the γ_k are nearly identical for all $k \in \mathcal{I}$ conditioned on the channel. Therefore, $\mathbf{s} \approx \lambda\mathbf{c}_m$, where the constant λ is given by:

$$\lambda = \frac{E_s\gamma_k}{\sqrt{2K}}. \tag{13}$$

Conditioned on the transmitted signal $x_m(t)$, the received vector $\mathbf{r}$ is jointly Gaussian, with conditional mean

$$\mathrm{E}\{\mathbf{r}|x_m(t)\} = \lambda\mathbf{c}_m.$$

The accurate covariance matrix of $\mathbf{r}$ conditioned on $x_m(t)$ is complicated and not diagonal, which makes optimal signal set design difficult. However, at low-to-moderate SNRs, where $n_{k,1}$ dominates the noise term, it can be shown that the noise components are approximately i.i.d. zero-mean Gaussian with variance $\frac{1}{2}N_0^2T_sW$. Therefore, the optimal decision can be made by simply finding the m for which the Euclidean distance between $\mathbf{r}$ and $\lambda\mathbf{c}_m$ is minimized, and, hence, at low-to-moderate SNR environments, simply designing the vectors $\{\mathbf{c}_m, m = 0, 1, ..., M-1\}$ whose minimum Euclidean distance is maximized achieves the best performance.

C. Narrowband Interference Suppression

One of the drawbacks of UWB systems is their sensitivity to narrowband interference. This is particularly true of TR-UWB and FSR-UWB systems, where the multiplication embedded in the receiver architecture results in interference multiplication. How to mitigate NBI in standard TR-UWB systems is still an open question. In this section, the potential for NBI suppression in the proposed scheme is discussed.

Suppose a narrowband interfering signal $i(t)$ is given by

$$i(t) = A_I u_I(t)\cos(2\pi f_I t + \theta_I),$$

where the data signal $u_I(t)$ of the interferer modulates a carrier of amplitude A_I, frequency f_I and phase θ_I. Because $i(t)$ is narrowband, $u_I(t)$ is assumed to be a constant $u_{I,n}$ over the n^{th} UWB frame.

In the presence of narrowband interference, the received signal becomes

$$r(t) = s(t) + i(t) + n(t),$$

and the output of the k^{th} branch of the receiver yields

$$r_k = s_k + i_k + n_k,$$

where the signal component s_k is the same as (6), the interference component i_k is given by

$$i_k = \int_0^{T_s}\left(2s(t)i(t) + i^2(t)\right)\cos(2\pi f_k t)dt,$$

and the Gaussian noise n_k can be expressed as

$$n_k = n_{k,0} + n_{k,1} + n_{k,I},$$

where $n_{k,0}$ and $n_{k,1}$ are given in (7), and $n_{k,I}$ is the noise-cross-interference component, which is given by

$$n_{k,I} = \int_0^{T_s} 2i(t)n(t)\cos(2\pi f_k t)dt. \tag{14}$$

Furthermore, assume that the system employs a UWB pulse $p(t)$ that satisfies $\int_{-\infty}^{\infty} p(t)dt \approx 0$. Often, $p(t)$ is modeled as the second or higher order derivative of Gaussian [1], which satisfies this assumption. Then,

$$\begin{aligned}
&\int_0^{T_s} 2s(t)i(t)\cos(2\pi f_k t)dt \\
\approx\ & \sum_{l=0}^{L-1}\sum_{n=0}^{N_f-1} 2\alpha_l g(nT_f+\tau_l)i(nT_f+\tau_l)\cdot \\
& \int_0^{T_s} p(t-nT_f-\tau_l)dt \\
=\ & 0.
\end{aligned}$$

Now, it can be shown that

$$\begin{aligned} i_k &\approx \int_0^{T_s} i^2(t)\cos(2\pi f_k t)dt \\ &\approx \frac{A_I^2 \bar{u}_I^2}{T_s}\int_0^{T_s} \cos^2(2\pi f_I t + \theta_I)\cos(2\pi f_k t)dt, \end{aligned}$$

where $\bar{u}_I^2 \triangleq \frac{1}{N_f}\sum_{n=0}^{N_f-1} u_{I,n}^2$. In general, $f_k \neq 2f_I$; therefore,

$$i_k = \frac{A_I^2 \bar{u}_I^2 f_I}{\pi(4f_I^2 - f_k^2)T_s}\cos(2\pi f_I T_s + 2\theta_I)\sin(2\pi f_I T_s). \quad (15)$$

The noise-cross-interference term $n_{k,I}$ can be shown to be approximately Gaussian, with zero mean and variance depending on the representation of $i(t)$ in $\mathbf{r}$.

In summary, the output vector can be written by

$$\mathbf{r} = \mathbf{s} + \mathbf{i} + \mathbf{n}.$$

The signal vector $\mathbf{s} = \lambda \mathbf{c}_m$ is proportional to the transmitted symbol vector $\mathbf{c}_m$, where λ is a constant given in (13). The interference vector $\mathbf{i}$ can be rewritten as $\mathbf{i} = \eta\tilde{\mathbf{i}}$, where

$$\eta = \frac{A_I^2 \bar{u}_I^2 f_I}{\pi T_s}\cos(2\pi f_I T_s + 2\theta_I)\sin(2\pi f_I T_s) \quad (16)$$

is a constant across k, and the k^{th} element of the K-by-1 vector $\tilde{\mathbf{i}}$ is defined as $\tilde{i}_k = \frac{1}{4f_I^2 - f_k^2}$, which is completely characterized by the interfering frequency f_I and the data carrying frequency f_k. Therefore, if f_I can be estimated at the receiver, the narrowband interference can be mitigated by projecting the received vector $\mathbf{r}$ onto the null space of the space spanned by $\tilde{\mathbf{i}}$ [11].

Suppose the $(K-1)$-by-K matrix $\mathbf{M}$ is made up of the orthogonal basis for the null space of that spanned by $\tilde{\mathbf{i}}$ such that $\mathbf{M}\tilde{\mathbf{i}} = 0$. Define

$$\hat{\mathbf{r}} = \mathbf{M}\mathbf{r} = \hat{\mathbf{s}} + \hat{\mathbf{n}},$$

where $\hat{\mathbf{s}} = \mathbf{M}\mathbf{s}$, and $\hat{\mathbf{n}} = \mathbf{M}\mathbf{n}$. Denote the projection of $\mathbf{c}_m$ as $\hat{\mathbf{c}}_m = \mathbf{M}\mathbf{c}_m$. To make a decision on the space spanned by $\mathbf{M}$, measuring the Euclidean distance between $\hat{\mathbf{r}}$ and $\lambda\hat{\mathbf{c}}_m$ is a simple but effective way. The vector $\hat{\mathbf{n}}$ is still zero-mean Gaussian; however, its components are not necessarily i.i.d., so the solution obtained via comparing minimum Euclidean distances is suboptimal.

In the case of persistent narrowband interference, the transmitter can steer away from such. Under a transmitted signal energy constraint, vectors $\hat{\mathbf{c}}_m$ can be designed to have maximum minimum Euclidean distance. Then the symbol $\mathbf{c}_m$ can be obtained by employing the concept of the generalized inverse:

$$\mathbf{c}_m = \rho_m(\mathbf{M}^T\mathbf{M})^{-1}\mathbf{M}^T\hat{\mathbf{c}}_m,$$

where the factor ρ_m is set to ensure that $\mathbf{c}_m^T\mathbf{c}_m = K$, so that the transmitted signal energy is E_s.

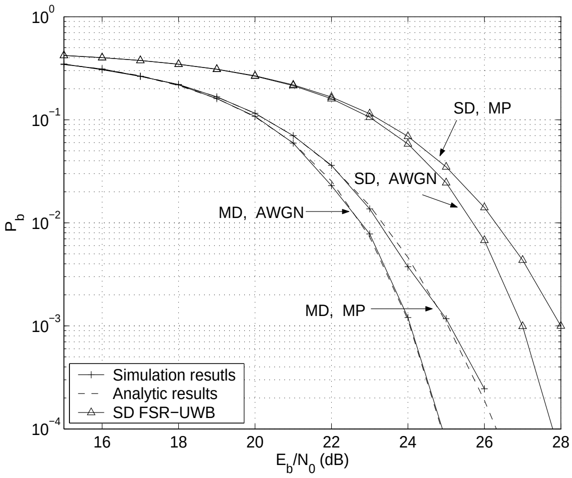

Fig. 2. BER versus average SNR, for a 4-differential FSR-UWB system employing parallel transmission and an SD FSR-UWB system. The analytic results are obtained from (11) and (12), respectively.

V. NUMERICAL RESULTS

The UWB pulse employed is the second derivative of a Gaussian pulse with zero-to-zero width 1.2ns. The noise bandwidth W is 2.5GHz. The symbol interval T_s consists of $N_f = 100$ frames, each of length $T_f = 100$ns. The multipath fading channel model considered is a discrete-path model given in (8), with uniform path delays (i.e. $\tau_l = l\tau_1$). The path gains are zero mean Gaussian random variables with variances given by an exponentially decaying MP intensity profile $\mathrm{E}\{\alpha_l^2\} = \frac{1}{D}\exp -\frac{l\tau_1}{D}$, where D is the decay factor, and the average aggregate power of the channel model is normalized to one. Throughout this section, an MD FSR-UWB system with $K = 4$ and MP channels with parameters $\tau_1 = 2$ns, $L = 20$, and $D = 40$ns are employed.

First, parallel transmission MD FSR-UWB is compared with SD FSR-UWB in Fig. 2. The analytic results are obtained from (11) and (12) respectively. As predicted, when $K = 4$, parallel transmission demonstrates a 3dB gain over an SD FSR-UWB system at low-to-moderate SNRs. Note that MD FSR-UWB also provides a transmission rate that is four times higher than SD FSR-UWB.

Next, a 4-ary signal set is employed. Suppose the transmitted codes for bit pairs $(0,0)$, $(0,1)$, $(1,0)$ and $(1,1)$ are $[1\ 1\ 1\ 1]^T$, $[1\ -1\ 1\ -1]^T$, $[-1\ -1\ -1\ -1]^T$ and $[-1\ 1\ -1\ 1]^T$, respectively. If the received noise vector $\mathbf{n}$ has covariance matrix $\frac{1}{2}N_0^2 T_s W\mathbf{E}$, where $\mathbf{E}$ is a K-dimensional identity matrix, then the receiver that makes decisions via minimum Euclidean distance is optimal; the BERs for this idealized case are shown in dashed-lines in Fig.3. As argued in Section IV, strictly speaking the noise components in the received vector are not i.i.d., therefore making decisions via minimized Euclidean distance not optimal. However, at low-to-moderate SNRs, the noise components are approximately i.i.d.. In Fig. 3, as predicted, the solid-line curves show the BERs of this 4-ary signaling MD FSR-UWB system are very close to those for which the i.i.d. assumption is artificially forced at the receiver. As Fig. 3 shows, this 4-ary signaling

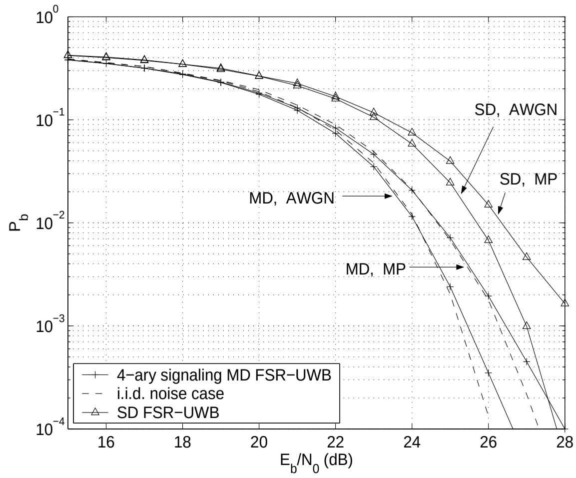

Fig. 3. BER versus average SNR, for a 4-differential FSR-UWB employing 4-ary signaling and an SD FSR-UWB system. The i.i.d. noise cases are obtained by artificially inserting white noise on the receiver output vector, in which case the receiver is optimal.

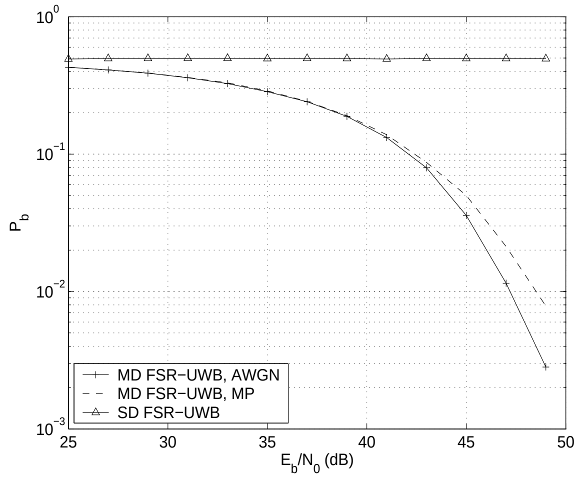

Fig. 4. BER versus average SNR, for a 4-differential FSR-UWB system employing 2-ary signaling and an SD FSR-UWB system with normalized transmitted energy in the presence of narrowband interference. The NBI signal has an energy of around 5000, with $A_I = 1$, $\theta_I = 0.345\pi$ and $f_I = 1.33$GHz. The interferer data signal $u_I(t)$ is assumed to be ± 1 during each frame.

MD FSR-UWB system outperforms SD FSR-UWB in error performance, and note that the data rate is also doubled. One might infer from Fig. 2 and Fig. 3 that the parallel transmission scheme outperforms the M-ary signaling scheme. However, recognizing the former as a particular type of the latter, reveals that the difference in performance observed is because the higher K in Fig. 2 allows more gain.

Per Section IV-C , M-ary signaling MC FSR-UWB provides a high-dimensional signal space that can be used to combat NBI. The BERs of the systems in the presence of NBI are shown in Fig. 4. The NBI signal considered has an amplitude $A_I = 1$, frequency $f_I = 1.33$GHz, and phase $\theta_I = 0.345\pi$. The data bearing signal $u_I(t)$ of the NBI signal is assumed to be constant during the frame, with equal probability to be $+1$ or -1. Compared to the transmitted UWB signal with a normalized energy of unity, the very strong NBI signal has an energy of around 5000. The 2-ary MD FSR-UWB system employs the NBI suppression scheme with $\hat{\mathbf{c}}_0$ and $\hat{\mathbf{c}}_1$ proportional to $\pm[1\ 1\ 1]^T$ respectively, while the SD FSR-UWB system does not use any NBI suppression scheme. As shown, for the SD FSR-UWB system, due to the huge energy of the NBI signal compared to that of the UWB signal, the BER is always 0.5 - even at high SNRs. For an MD FSR-UWB system, on the other hand, the performance is greatly improved. Note from Fig. 4 that very high SNRs are required even for the MD FSR-UWB systems, because of the high variance of the noise-cross-interference term $n_{k,I}$. However, note that unlike the SD FSR-UWB system, the proposed system does not exhibit an error floor at high SNRs.

VI. Conclusion

A multi-differential FSR-UWB system is proposed in this paper, in which several data dimensions share a common reference. Compared to SD FSR-UWB, the proposed scheme provides more degrees of freedom with only a slight increase of the system bandwidth. It is shown that the proposed scheme outperforms SD FSR-UWB in error performance, while providing a higher data rate. Unlike similar enhancements to standard TR-UWB systems, the gain is larger than that obtained by simply amortizing the reference signal energy over multiple data symbols. Moreover, by employing M-ary signaling, the proposed scheme can exploit the large-dimensional signal space to perform effective narrowband interference suppression.

References

[1] M. Z. Win and R. A. Scholtz, "Impulse Radio: How it works,"*IEEE Communication Letters*, Vol. 2: pp. 10–12, Jan. 1998.

[2] M. Z. Win and R. A. Scholtz, "On the energy capture of ultra-wide bandwidth signals in dense multipath environments,"*IEEE Commununication Letters*, vol. 2, pp. 245–247, Sept. 1998.

[3] R. Hoctor and H. Tomlinson, *An Overview of Delay-Hopped, Transmitted-Reference RF Communications,* General Electric Technical Report 2001CRD198, Jan. 2002.

[4] M. Casu and G. Durisi, "Implementation Aspects of a Transmitted-Reference UWB Receiver," *Wireless Communications and Mobile Computing,* Vol. 5: pp. 537-549, May 2005.

[5] L. Feng and W. Namgoong, "An Oversampled Channelized UWB Receiver with Transmitted Reference Modulation," to appear in the *IEEE Transactions on Wireless Communications*.

[6] S. Mitra, *Digital Signal Processing: A Computer-Based Approach,* McGraw-Hill, 1998.

[7] D. Goeckel and Q. Zhang, "Slightly frequency-shifted reference ultra-wideband (UWB) radio: TR-UWB without the delay element," *Proceedings of the Military Communication Conference,* Oct., 2005.

[8] D. Goeckel and Q. Zhang, "Slightly Frequency-Shifted Reference Ultra-Wideband (UWB) Radio", submitted to the *IEEE Transactions on Communications,* June 2005. Available from: "http://www-unix.ecs.umass.edu/∼goeckel/uwb.html".

[9] J. Choi and W. Stark, "Performance of ultra-widebnad communications with suboptimal receivers in multipath channels," *IEEE Journal on Selected Areas in Communications*, Vol. 20: pp. 1754-1766, Dec. 2002.

[10] H. Zhang and D. Goeckel, "Generalized Transmitted-Reference UWB Systems," *Proceedings of the Conference on Ultra-Wideband Systems and Technologies (UWBST),* Nov. 2003.

[11] R. Lupas and S. Verdu, "Linear multiuser detectors for synchronous code-division multiple-access channels Lupas", *IEEE Transactions on Information Theory,* Vol. 35: pp. 123-136, Jan. 1989.