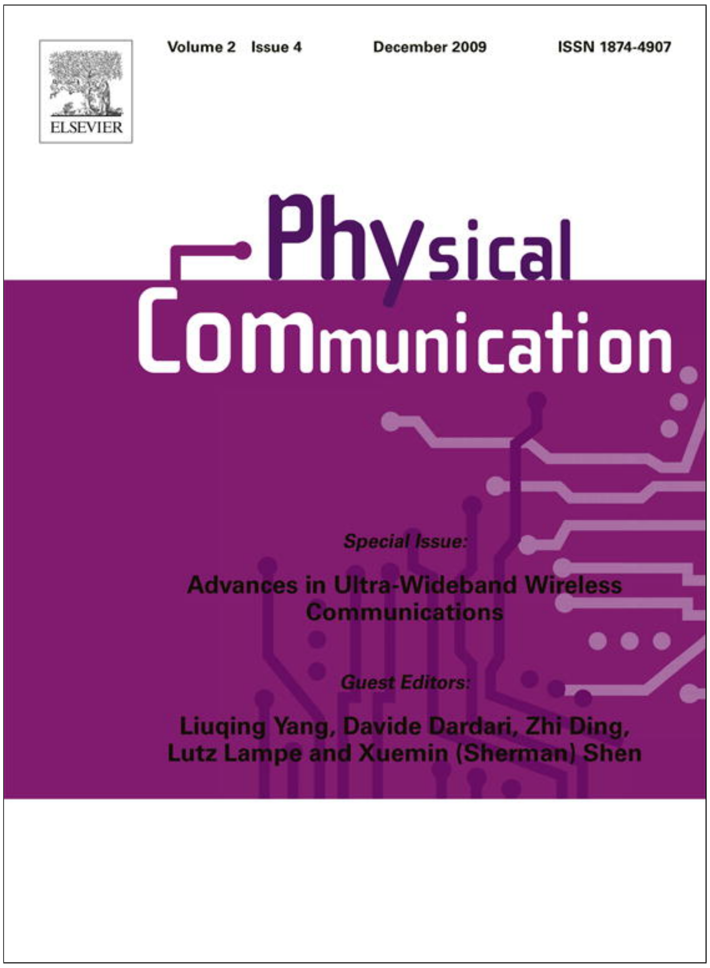
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Hybrid coherent and frequency-shifted-reference ultrawideband radio☆

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ABSTRACT

Ultrawideband communications often occur in heterogeneous networks where different receivers have different complexity and energy consumption requirements. In this case it is desirable to have a modulation scheme that works well with coherent receivers as well as simpler receivers, namely transmitted-reference (TR) receivers. In particular, we consider a TR scheme that employs slightly frequency-shifted-reference (FSR) signals [D.L. Goeckel, Q. Zhang, Slightly frequency-shifted-reference ultrawideband (UWB) radio, IEEE Trans. Commun. (2007)] and thus avoids one of the main drawbacks of conventional TR schemes, namely the need to implement a delay line. We propose and analyze a modulation scheme that works well with both FSR receivers (where it has at least the same performance as conventional TR modulation), and coherent receivers. Coherent receivers receiving conventional TR modulation suffer a 3 dB penalty, because they cannot make use of the energy invested into the reference pulse. Our proposed scheme avoids this drawback by including a data preprocessor that can be viewed as a nonsystematic rate-1/2 convolutional code. These codes give 1.5 dB gain over our previously proposed constraint-length-two systematic codes at a BER of 1×10^{-4} in 802.15.3a CM4 multipath fading channels. We also develop a sliding-window based scheme to derive the template waveform that is needed for coherent rake receivers. This scheme exploits the data preprocessor structure and flexibly uses the received signal over a certain window. The distortion to the self-derived template waveform is a decreasing function of the window length; in the extreme but unrealistic case of a very long window length in a slowly fading channel, the self-derived template waveform is noiseless (i.e., the ideal template without distortion).

1. Introduction

One of the key challenges for pulsed ultrawideband (UWB) systems [1–4] is the construction of low-cost receivers that work well in multipath environments. In impulse radio, each symbol is represented by a series of pulses. In multipath environments, pulses suffer delay dispersion, and their energy has to be collected by rake receivers or equivalent structures [5]. Coherent rake receivers that combine all resolvable multipath components (MPCs) [6] are optimum; however, tracking, estimating, and combining a large number of MPCs (e.g., tens or even hundreds in indoor environments [7,8]), or even a subset of all the resolvable paths [9], result in complex receivers. Consequently, simpler alternatives, in particular transmitted-reference (TR) receivers, have drawn significant attention in recent years [10–14].

In conventional TR systems, the symbol is represented by a series of *pulse doublets*, where each doublet consists

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of two pulses separated by a fixed delay: the first, unmodulated, pulse serves as a reference pulse; the second pulse is data modulated and is referred to as the data pulse. The received signal can simply be demodulated by correlating it with a delayed version of itself, and integrating the resulting signal. The main problem for implementing TR receivers is the delay unit, which must handle wideband analog signals and must be precise, making it difficult to build in low-power integrated fashion [15,16]. In order to avoid the delay unit, an alternative, called slightly frequency-shifted reference (FSR), was recently proposed by one of us [15]. In this scheme, the pulse doublet does not consist of pulses that are offset in the delay domain, but rather in the frequency domain. A receiver thus only has to implement a mixer, and not a delay line, and is thus much easier to build; furthermore such a receiver can actually perform slightly better than the conventional TR receiver over both additive white Gaussian noise (AWGN) and multipath fading channels [15].

Devices in a UWB network often have different cost/complexity/performance requirements. For example, some UWB devices may be battery-operated and thus require a simple, energy-saving receiver, while other devices may be connected to a power supply and thus can use a high-complexity, power hungry receiver that offers better communication range [17]. The resulting heterogeneous network requires a "universal" modulation method compatible with different types of receivers such as coherent rake and FSR or TR receivers. Technically, it is possible to demodulate FSR signals with a coherent receiver, by simply "throwing away" the reference pulses. However, this implies a 3 dB signal energy penalty compared to a system that is designed to use coherent receivers only. On the other hand, signals designed for coherent receivers (i.e., those without reference pulses) obviously cannot be demodulated by a FSR receiver. In [18], we proposed a hybrid modulation scheme that enables efficient reception by both coherent and TR receivers. The key idea is to make the "reference pulse" information bearing, without modifying the phase relationship between the reference pulse and data pulse. This makes sure that the energy in the reference pulse is not "wasted" for the coherent receiver, and recovers the 3 dB loss by "normal" TR signaling. Furthermore, the information in the reference pulse is made dependent on the *previous* information symbol, which introduces memory into the modulation, and leads to further performance gain for the coherent receiver.

In the current paper, we show how our hybrid modulation scheme can be modified to enable coherent rake receivers and *FSR receivers* in the same wireless network, and present several modifications that further improve the performance. In particular, our contributions are:

- we propose a hybrid modulation scheme that works with coherent and FSR receivers.
- we proposed a modified encoding structure. The data preprocessor in the basic hybrid scheme proposed in [18] can be viewed as a rate-1/2 systematic convolutional code with a constraint length two. We show in this paper that the systematic code can be extended into a nonsystematic convolutional code with longer constraint lengths that gives a higher coding gain, while the desired properties between the data pulse and reference pulse are still maintained for FSR receivers.
- we develop a channel estimation scheme that efficiently exploits the proposed signaling structure to obtain nearly noise-free channel coefficient estimates for coherent reception in slowly fading channels.
- we derive analytical bounds for the coding gain of the new preprocessing structure.
- we show that a frequency shift proposed in [15] can be reduced by a factor of two, improving performance, and allowing for the transmission of higher-data rates.
- we analyze the resulting receiver structure and performance for the resulting UWB transmission scheme. We derive closed-form equations for the BER in multipath channels.

The remainder of the paper is organized in the following way. Section 2 explains the improved hybrid FSR-coherent UWB scheme, and coherent and FSR receivers for the proposed hybrid modulation scheme are analyzed in Section 3. Section 4 presents simulation results of the proposed scheme with a coherent receiver. Comparison is made between the proposed improved hybrid FSR-coherent scheme and the basic hybrid TR-coherent scheme described in [18]. Concluding remarks are given in Section 5.

2. Hybrid FSR-coherent system

2.1. Transmitter

2.1.1. Transmitted signal

We consider a time-hopping impulse radio system, where each bit interval T_b is partitioned into N_f frames, each having a duration T_f ($T_b = N_f T_f$). Let us define a basic template waveform for the transmission of a bit

$$u(t) = \sum_{j=0}^{N_f-1} \mathrm{d}_j p(t - jT_f - c_j T_c) \tag{1}$$

where $p(t)$ is the basic UWB pulse shape normalized so that $\int_{-\infty}^{\infty} p^2(t)\mathrm{d}t = 1/N_f$ and $\int_{-\infty}^{\infty} u^2(t)\mathrm{d}t = 1$. The elements of the pseudorandom sequence c_j, which determines the position of the pulse within each frame, are uniformly distributed integers between 0 and T_f/T_c, where T_c is the chip duration. The elements of the pseudorandom sequence d_j are taken from $\{-1, +1\}$, and effect a randomization of the polarities, which is useful for smoothing the transmit spectrum and enhancing the multiuser separation. For ease of notation, we henceforth omit the d_j, since they do not influence the operation of the TR receiver (where their impacts on reference pulse and data pulse cancel out) or the coherent receiver (where the receiver cancels its effect by multiplying the received signal with the sequence d_j).

We furthermore define waveforms for the reference and data signals (assuming equal energies)

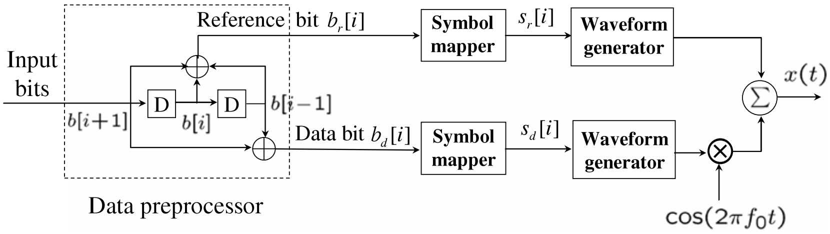

Fig. 1. Block diagram of the slightly frequency-shifted-reference hybrid system transmitter.

$$x_r(t) = \sum_{i=-\infty}^{\infty} \sqrt{E_b/2} s_r[i] u(t - iT_b) \tag{2a}$$

$$x_d(t) = \sum_{i=-\infty}^{\infty} \sqrt{E_b/2} s_d[i] u(t - iT_b). \tag{2b}$$

The data streams $s_r[i]$ and $s_d[i]$ ($s_r[i], s_d[i], \in \{-1, +1\}$) modulating the reference and data signals will be discussed in Section 2.1.2.

In an FSR scheme, the total transmitted waveform $x(t)$ is

$$\begin{aligned} x(t) &= x_r(t) + x_d(t)\sqrt{2}\cos(2\pi f_0 t) \\ &= \sum_{i=-\infty}^{\infty} \sqrt{\frac{E_b}{2}} \left\{ s_r[i] + \sqrt{2} s_d[i] \cos(2\pi f_0 t) \right\} \\ &\quad \times u(t - iT_b) \end{aligned} \tag{3}$$

where f_0 is the shift of the center frequency of the reference pulses relative to that of the data pulses.

In order to allow simple demodulation, it is essential that f_0 be chosen such that the first and second term of the r.h.s. of Eq. (3) are orthogonal or at least approximately orthogonal. At the same time, f_0 should be chosen to be as small as possible; this ensures that the reference signal and the (frequency-shifted) data signal "see" the same propagation channel even in frequency-selective channels. In the following, we derive the minimum frequency spacing. Without loss of generality, let us focus on the first bit interval, i.e., $i = 0$:

$$\begin{aligned} &\int_0^{T_b} x_r(t) x_d(t) \cos(2\pi f_0 t) \mathrm{d}t \\ &= \frac{E_b}{2} s_r[0] s_d[0] \int_0^{T_b} \sum_{j=0}^{N_f-1} p^2(t - jT_f - c_j T_f) \\ &\quad \times \cos(2\pi f_0 t) \mathrm{d}t \\ &\approx \frac{E_b}{2} s_r[0] s_d[0] \sum_{j=0}^{N_f-1} \cos(2\pi f_0 j T_f) \int_0^{T_b} p^2(t - jT_f) \mathrm{d}t \\ &= \frac{E_b}{2N_f} s_r[0] s_d[0] \sum_{j=0}^{N_f-1} \cos(2\pi f_0 j T_f) = 0 \end{aligned} \tag{4}$$

where the approximation is obtained by considering that $p(t - jT_f)$ is a very narrow pulse, compared with a bit interval T_b, located at $t = jT_f$. When N_f is large and the N_f pulses are equally spaced over T_b, choosing $f_0 = \frac{1}{2T_b}$ will imply that the summands $\cos(2\pi f_0 j T_f)$ in the last line are the uniform samples of the sinusoidal signal $\cos(2\pi f_0 t)$ over the first half of its fundamental period from 0 to T_b, which is approximately equal to zero.

This choice of the frequency-shift reduces the difference in the center frequencies of the data pulse and reference pulse by a half compared with the suggested value of $f_0 = \frac{1}{T_b}$ in [15].

2.1.2. Data preprocessor

In the conventional FSR scheme, the reference bits are fixed, i.e., $s_r[i] = 1$. However, coherent receivers can achieve better performance if the reference signal also carries information; of course, it is essential to retain a signaling structure that allows demodulation by an FSR receiver.

The proposed hybrid scheme is shown in Fig. 1, where the 'reference' bit $b_r[i]$ and 'data' bit $b_d[i]$ can be considered as the outputs of a rate-1/2 convolutional encoder. In the related scheme of [18], we proposed a convolutional encoder with constraint length two, resulting in a minimum free distance of $\sqrt{6}$. We now investigate a new encoding scheme with a constraint length three and code generator polynomial $[1 \oplus D \oplus D^2, 1 \oplus D^2]$, where $\oplus$ denotes modulo-two addition. Note that the 'reference' bit is not really a reference bit as in a conventional TR system; rather, it is one of the two output bits of the data preprocessor, i.e., the rate-1/2 convolutional encoder. Obviously, a coherent receiver can detect the transmitted bits by demodulating both the reference bit and the data bit, and then applying a Viterbi decoder to recover the original information bits.

This encoder structure must satisfy certain conditions so that FSR receivers can detect the transmitted signal as well. Table 1 shows the polarities of the modulated reference and data waveforms, and their phase/polarity difference for all eight combinations of three consecutive input bits. As seen from the table, the phase difference of the pair of pulses – the reference and data pulses – solely depends on input information bit $b[i]$; therefore, an FSR receiver can also demodulate the received signals.

The ith data bit and the ith reference bit are expressed as

$$b_d[i] = b[i+1] \oplus b[i-1] \tag{5a}$$

$$b_r[i] = b[i+1] \oplus b[i] \oplus b[i-1]. \tag{5b}$$

For bipolar signaling, the symbol mapper simply performs the following mapping: $b_r[i] = 0/1 \to s_r[i] = -1/1$; $b_d[i] = 0/1 \to s_d[i] = -1/1$.

Table 1
Input–output combinations of data preprocessor designed as a rate-1/2 convolutional code with constraint length three.

$b[i+1]$	$b[i]$	$b[i-1]$	$s_r[i]$	$s_d[i]$	Phase difference between $s_r[i]$ and $s_d[i]$ (°)
0	0	0	−1	−1	0
0	0	1	+1	+1	0
0	1	0	+1	−1	180
0	1	1	−1	+1	180
1	0	0	+1	+1	0
1	0	1	−1	−1	0
1	1	0	−1	+1	180
1	1	1	+1	−1	180

For coherent rake receivers, the data preprocessor is effectively a rate-1/2 convolutional encoder with a constraint length three. The minimum Euclidean distance is $\sqrt{10\varepsilon}$, resulting in a 4 dB coding gain, which is 2.22 dB higher than the basic hybrid TR-coherent scheme in [18].

The proposed data preprocessor can be further extended to a rate-1/2 convolutional encoder with a longer constraint length. For example, the convolutional code with generator polynomial $[1 \oplus D \oplus D^2 \oplus D^3, 1 \oplus D^2 \oplus D^3]$ is a good choice with a constraint length four. The relationship among the input bits, the reference bit, and the data bit for this preprocessor is shown in Table 2. The minimum Euclidean distance with this preprocessor becomes $\sqrt{12\varepsilon}$, which results in an additional 0.8 dB coding gain over the preprocessor shown in Fig. 1, or 3 dB coding gain over the scheme in [18].

For even higher constraint lengths, it becomes difficult to find good codes with the maximum, or near maximum, free distance while the phase relationship between the reference pulse and data pulse required by FSR receivers is still maintained. A longer constraint length would generally result in a higher coding gain at the expense of a higher decoding complexity. Thus, system complexity-performance tradeoff can be made flexible by choosing an appropriate preprocessor structure.

2.2. Receiver

2.2.1. Received signal

The standardized 802.15.3a channel model [7] describes the channel impulse response as

$$h(t) = \sum_{l=0}^{L-1} h_l \delta(t - \tau_l) \tag{6}$$

where L is the total number of multipath components, h_l is the channel fading coefficient for the lth path, τ_l is the arrival time of the lth path relative to the first path ($\tau_0 = 0$ assumed), and $\delta(t)$ is the Dirac delta function. The channel gain h_l is modeled as $h_l = \lambda_l \beta_l$, where λ_l takes on the values of -1 or 1 with equal probability and β_l has a lognormal distribution [7]. Since multipath components tend to arrive in clusters [7], τ_l in (6) is expressed as $\tau_l = \mu_c + \nu_{m,c}$, where μ_c is the delay of the cth cluster that the lth path falls in, $\nu_{m,c}$ is the delay (relative to μ_c) of the mth multipath component in the cth cluster. The relative power of the lth path to the first path can be expressed as $E\{|h_l|^2\} = E\{|h_0|^2\} \mathrm{e}^{-\mu_c/\Gamma} \mathrm{e}^{-\nu_{m,c}/\gamma}$, where $E\{\cdot\}$ denotes

Table 2
Input–output combinations of data preprocessor designed as a rate-1/2 convolutional code with constraint length four.

$b[i+1]$	$b[i]$	$b[i-1]$	$b[i-2]$	$s_r[i]$	$s_d[i]$	Phase difference between $s_r[i]$ & $s_d[i]$ (°)
0	0	0	0	−1	−1	0
0	0	0	1	+1	+1	0
0	0	1	0	+1	+1	0
0	0	1	1	−1	−1	0
0	1	0	0	+1	−1	180
0	1	0	1	−1	+1	180
0	1	1	0	−1	+1	180
0	1	1	1	+1	−1	180
1	0	0	0	+1	+1	0
1	0	0	1	−1	−1	0
1	0	1	0	−1	−1	0
1	0	1	1	+1	+1	0
1	1	0	0	−1	+1	180
1	1	0	1	+1	−1	180
1	1	1	0	+1	−1	180
1	1	1	1	−1	+1	180

expectation, Γ is the cluster decay factor, and γ is the ray decay factor. Note that, different from common baseband models of narrow-band systems, h_l is real-valued.

Given the transmitted signal $x(t)$ in (3) and the channel impulse response $h(t)$ given by (6), the received signal is expressed as

$$\begin{aligned} r(t) &= \sum_{l=0}^{L-1} h_l x(t - \tau_l) \\ &= \sum_{i=-\infty}^{\infty} \sqrt{\frac{E_b}{2}} \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l)\{s_r[i] \\ &\quad + \sqrt{2} s_d[i] \cos(2\pi f_0 (t - \tau_l))\} + n_0(t) \end{aligned} \tag{7}$$

where $n_0(t)$ is the additive white Gaussian noise with the two-sided power spectral density $N_0/2$.

2.2.2. FSR receiver

The FSR receiver is similar to the one described in [15] and is shown in Fig. 2 for convenience. The received signal is bandpass-filtered,[1] squared, multiplied with $\sqrt{2}\cos(2\pi f_0 t)$, and then integrated over each bit interval.

Using an analysis similar to [15], we can show that for the ith bit, the integrator output r_i in the absence of noise and multipath fading can be expressed as

$$\begin{aligned} r_i &= \int_{iT_b}^{(i+1)T_b} \left(\sqrt{\frac{E_b}{2}} \{s_r[i] + \sqrt{2} s_d[i] \cos(2\pi f_0 t)\} \right. \\ &\quad \left. \times\ u(t - iT_b) \right)^2 \sqrt{2}\cos(2\pi f_0 t) \mathrm{d}t \\ &= E_b s_r[i] s_d[i] \end{aligned} \tag{8}$$

[1] The bandpass filter (BPF) has bandwidth B, which is chosen to be equal to or larger than the signal's 10 dB bandwidth, to avoid severe, filter-induced distortion to the signal. For simplicity of analysis, it is also assumed that the BPF is ideal and its bandwidth B is an integer multiple of $1/(2T_b)$. Let the noise at the output of the BPF be $n(t)$. The filtered noise is no longer white, but its autocorrelation can be very narrow in time due to the ultrawide filter bandwidth B. The BPF does not change the received signal component. This assumption applies to the coherent receiver to be discussed in the next section.

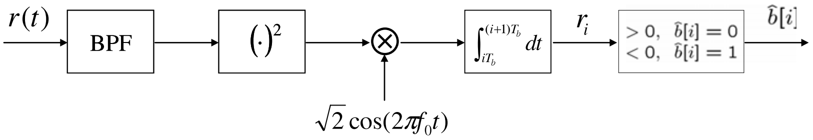

Fig. 2. FSR receiver for the proposed hybrid FSR-coherent scheme.

where the second equation can be derived using the relationship given in (4). The analysis for multipath channels follows the approach in [15], which also leads to the product term $s_r[i]s_d[i]$ in the decision variable given by (8). From Tables 1 and 2, we observe that the product term $s_r[i]s_d[i]$ maps exactly to the symbol-mapped input bits. Therefore, if the relationship between the reference and data bits as described in Section 2.1.2 is maintained, FSR receivers can demodulate the transmitted signal.

2.2.3. Coherent receiver and channel estimation

The coherent receiver for the hybrid FSR scheme is shown in Fig. 3. The received signal for the ith bit interval is first passed through a BPF (see Section 2.2.2) and then correlated with $w_i(t) = \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l)$. This correlator requires the knowledge of $\{h_l\}$ and $\{\tau_l\}$; thus, it is equivalently a coherent rake receiver. The correlator output is processed further to generate the estimate of both the reference bit and the data bit. For the data bit, the correlator output is integrated over each bit interval; for the reference bit, the the correlator output is first multiplied by $\sqrt{2}\cos(2\pi f_0 t)$ and then integrated over each bit interval. Finally, both estimates for the data bits $\{s_d[i]\}$ and reference bits $\{s_r[i]\}$ are applied for Viterbi decoding, yielding the estimated input bits $\{\hat{b}[i]\}$. This is illustrated in Fig. 3.

Note that in practice it is usually impossible to combine all received paths considering receiver complexity; a partial rake can be realized simply by letting the coefficients $\{h_l\}$ that the receiver does not combine equal to zero.

For low- to medium-data rates, it is reasonable to assume that there is no inter-symbol interference (ISI). The derivation of the decision variables for $s_r[i]$ and $s_d[i]$, $r_r[i]$ and $r_d[i]$, is similar; thus, let us focus on deriving $r_d[i]$:

$$
\begin{aligned}
r_d[i] = & \int_{iT_b}^{(i+1)T_b} \sqrt{E_b} s_r[i] \sum_{l=0}^{L-1} h_l^2 u^2(t - iT_b - \tau_l) \\
& \times \cos(2\pi f_0 t)\mathrm{d}t \\
& + \int_{iT_b}^{(i+1)T_b} \sqrt{\frac{E_b}{2}} s_d[i] \sum_{l=0}^{L-1} h_l^2 u^2(t - iT_b - \tau_l) \\
& \times 2\cos(2\pi f_0 t - 2\pi f_0 \tau_l)\cos(2\pi f_0 t)\mathrm{d}t \\
& + \int_{iT_b}^{(i+1)T_b} n(t) \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l)\sqrt{2} \\
& \times \cos(2\pi f_0 t)\mathrm{d}t. \qquad (9)
\end{aligned}
$$

For low-data-rate applications for which $\max\{\tau_l\} \ll T_b$, the first term of the r.h.s. of (9) is approximately equal to zero for the same reason as explained below Eq. (4). Thus

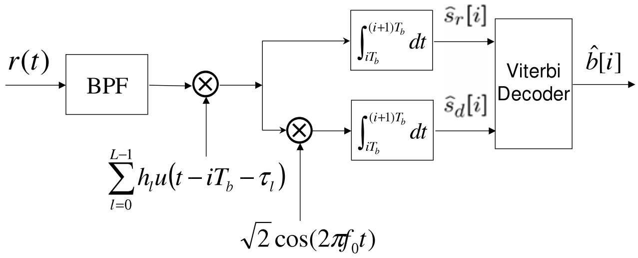

Fig. 3. Coherent receiver for the proposed hybrid FSR-coherent scheme.

we have

$$
\begin{aligned}
r_d[i] \approx & \sqrt{\frac{E_b}{2}} s_d[i] \sum_{l=0}^{L-1} h_l^2 \int_{iT_b}^{(i+1)T_b} u^2(t - iT_b - \tau_l) \\
& \times \cos(2\pi f_0 \tau_l)\mathrm{d}t \\
& + \sqrt{\frac{E_b}{2}} s_d[i] \sum_{l=0}^{L-1} h_l^2 \int_{iT_b}^{(i+1)T_b} u^2(t - iT_b - \tau_l) \\
& \times \cos(4\pi f_0 t - 2\pi f_0 \tau_l)\mathrm{d}t + \xi[i] \\
\approx & \sqrt{\frac{E_b}{2}} s_d[i] \sum_{l=0}^{L-1} h_l^2 + \xi[i] \qquad (10)
\end{aligned}
$$

where we have applied the approximation $\int_{iT_b}^{(i+1)T_b} u^2(t - iT_b - \tau_l)\cos(2\pi f_0 \tau_l)\mathrm{d}t \approx 1$ because $\max\{\tau_l\} \ll T_b$ or $f_0 \tau_l \approx 0$, and the second term on the r.h.s. is approximately equal to zero for the same reason given in deriving Eq. (4).

In the analysis so far, we have assumed that $w_i(t) = \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l)$ is available for detection of the ith bit. This requires the receiver to estimate $\{h_l\}$ and $\{\tau_l\}$ and then construct the receiver template waveform $w_i(t)$ based on the transmitted template waveform given in (1). We analyze a scheme that efficiently exploits the proposed signaling structure to provide an estimate of $w_i(t)$ for the coherent receiver. In a slowly fading channel, $\{h_l\}$ and $\{\tau_l\}$ changes very little over a block of many bit intervals. When a coherent rake receiver is applied to detect the conventional TR signals, the reference bits could be exploited for channel estimation. A disadvantage of making the reference bits data dependent in the hybrid scheme is that the reference bits cannot be exploited directly to estimate $\{h_l\}$. Without an explicit reference signal, a decision-directed approach could be employed [14]. However, the decision-directed approach could potentially suffer from error propagation [14].

We investigate a sliding-window approach which exploits the FSR receiver structure to estimate the correlation-template $w_i(t)$ for detecting the ith symbol. This approach does not suffer from the error propagation effects. Note that a coherent receiver has all the main components that an FSR requires such as a BPF, correlator, integrator, oscillator, etc., except a squaring device. For clarity of description, let us write the received signal at the BPF output in the ith bit interval as

$$r_i(t) = \sqrt{\frac{E_b}{2}} s_r[i] \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l) + \sqrt{E_b} \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l) s_d[i] \times \cos(2\pi f_0 (t - \tau_l)) + n_i(t) \quad (11)$$

where the first term of the r.h.s. is exactly the correlation template subject to data-dependent scaling factor $\sqrt{\frac{E_b}{2}} s_r[i]$.

Let $\{\tilde{b}[i]\}$ be the bit decisions by the FSR receiver and $\tilde{b}[i-j], i = 1, \ldots, N$, be the temporary FSR bit decisions over a block of N bit intervals preceding the ith bit. The temporary estimates of the reference and data symbols over the same block of N bit intervals, $\tilde{s}_r[i-j]$ and $\tilde{s}_d[i-j], j = 1, \ldots, N$, can be obtained by passing $\tilde{b}[i]$ through the data preprocessor. With the FSR bit decisions, we can "data-demodulate" the received signal in each bit interval given in (11), forming the estimate of $w_i(t)$ as

$$\hat{w}_i(t) = \sqrt{\frac{2}{E_b}} \frac{1}{N} \sum_{j=1}^{N} \tilde{s}_r[i-j] r_{i-j}(t). \quad (12)$$

The estimated correlation template is distorted by one multiplicative distortion and two forms of additive distortions: (a) due to the slightly frequency-shifted data component and (b) due to noise. The multiplicative distortion is due to the FSR bit decision errors, which causes a constant scaling factor to the template. This distortion term is written as $\frac{1}{N}\sum_{j=1}^{N} \tilde{s}_r[i-j] s_r[i-j]$. In the limiting case when $N \to \infty$, it becomes $(1 - P_{b,s_r})$, where P_{b,s_r} is the error rate of the re-encoded reference bits from $\tilde{b}[i]$. This distortion does not degrade the performance of the coherent rake. The two additive distortions are expressed as

$$I_{s_d}(t) = \left(\frac{1}{N} \sum_{j=1}^{N} \tilde{s}_r[i-j] s_d[i-j] \right) \times \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l) \sqrt{2} \cos(2\pi f_0 (t - \tau_l)) \quad (13a)$$

$$I_n(t) = \frac{2}{N\sqrt{E_b}} \sum_{j=1}^{N} \tilde{s}_r[i-j] n_{i-j}(t) \quad (13b)$$

where in (13a) we have applied the fact that over the N different bit intervals, all quantities except $\tilde{s}_r[i]$ and $s_d[i]$ remain the same. When N is large (e.g., 20), the term $\sum_{j=1}^{N} \tilde{s}_r[i-j] s_d[i-j]/N$ in (13a) approaches Gaussian with a zero mean and variance $1/N$. $I_n(t)$ is also zero-mean Gaussian with a power spectral density $\frac{2}{N}\frac{1}{E_b/N_0}$, which is inversely proportional to E_b/N_0 and block size N. In the limiting but unrealistic case when $N \to \infty$, the estimated template is expressed as

$$\lim_{N\to\infty} w_i(t) = (1 - P_{b,s_r}) \sum_{l=0}^{L-1} h_l u(t - iT_b - \tau_l) \quad (14)$$

which is precisely the ideal template waveform, since the constant scaling factor $(1 - P_{b,s_r})$ does not affect the detection performance. For the realistic case when N is finite (e.g., 10), the self-derived template for coherent receivers is a slightly distorted version of the ideal template.

3. Performance analysis

3.1. FSR receiver in multipath

As described in Section 2.2.1, the received signal at the BPF output is expressed as $r(t) = x(t) * h(t) + n(t)$, where $h(t)$ is given in (6), $x(t)$ is given in (3), $*$ denotes convolution, and $n(t)$ is the zero-mean Gaussian noise with a one-sided spectral density N_0. Again using an analysis similar to [15], the BER over multipath channels can be written as

$$P_{\text{FSR}} = E_h \left[Q \left(\frac{E_b \sum_{l=0}^{L-1} h_l^2 \cos(2\pi f_0 \tau_l)}{\sqrt{\frac{5}{2} E_b N_0 \sum_{l=0}^{L-1} h_l^2 + T_b N_0^2 B}} \right) \right] \quad (15)$$

where $E_h[\cdot]$ denotes the expectation over the set of channel coefficients $\{h_l\}$. While a numerical evaluation of the expectation operator has been used in the past, we derive in the following a closed-form expression.

In the following, we again assume that $f_0 \tau_l \ll 1$, which is fulfilled for signaling at low- to medium-data rates. Let $\gamma = \sum_{l=0}^{L-1} h_l^2 = \sum_{l=0}^{L-1} \beta_l^2$. Note that $\{\beta_l\}$ are independent of one another. Since β_l is a lognormal random variable (RV), β_l^2 is also a lognormal RV. Thus, γ is a sum of independent lognormal RVs. Let $\beta_l = e^{v_l}$, where v_l is a normal RV, i.e., $v_l \sim N(\mu_{v_l}, \sigma_{v_l}^2)$. The kth moment of β_l is given as

$$E\{\beta_l^k\} = e^{k\mu_{v_l} + k^2 \sigma_{v_l}^2 / 2}. \quad (16)$$

Although an exact closed-form expression of the pdf of a sum of independent lognormal RVs does not exist, such a sum can be approximated by another lognormal RV [20,21]. The parameters μ and σ of this random variable γ can be obtained from the following set of nonlinear equations [19]:

$$\sum_{i=1}^{N_G} \frac{w_i}{\sqrt{\pi}} \exp\left[-s_m \exp\left(\frac{\sqrt{2}\sigma a_i + \mu}{\xi} \right) \right] = \prod_{i=1}^{K} \widehat{\Psi}_X(s_m; \mu_i, \sigma_i) \quad (17)$$

where $m = 1, 2$ and

$$M(-s) = \int_0^\infty \left[1 + \frac{s\Omega}{m} \right]^{-m} \frac{10/\ln(10)}{\sqrt{2\pi\sigma^2}\Omega} \times e^{-\frac{(10\log_{10}\Omega - \mu)^2}{2\sigma^2}} \mathrm{d}\Omega \quad (18)$$

and the weights, w_i, and abscissas, a_i, of the Gaussian quadrature for different orders, N_G, are tabulated in standard mathematical references. The parameters s_1 and s_2 are the arguments of the moment-generating function at which the exact distribution of the sum of lognormal variables should match the distribution of the equivalent lognormal variable.

The approximated pdf of γ is given as

$$f(\gamma) = \frac{1}{\gamma\sqrt{2\pi\sigma^2}} \exp\left[-\frac{(\ln(\gamma)-\mu)^2}{2\sigma^2}\right]. \tag{19}$$

The average BER can be calculated by averaging the conditional BER $P_{\mathrm{FSR}}(\gamma) = Q\left(\frac{E_b\gamma}{\sqrt{5E_bN_0\gamma/2+T_bN_0^2B}}\right)$ over $f(\gamma)$ as

$$P_{\mathrm{FSR}} = \int_0^\infty P_{\mathrm{FSR}}(\gamma) f(\gamma)\mathrm{d}\gamma. \tag{20}$$

3.2. Coherent receiver

Because of the convolutional decoding involved in the coherent receiver for the hybrid scheme, the exact BER expression is difficult to obtain. However, a union bound can be derived in the following manner. First, let $T(D,I) = \sum_{i,j} n(i,j)I^iD^j$ be the convolutional code transfer function, where $n(i,j)$ is the number of paths with i input 1's and j output 1's that diverge from the all zero's path and re-merge with it later. Assume that $h(t)$ does not vary appreciably between such a divergence and re-merge for the shorter paths that dominate the error performance (i.e., no temporal diversity is achieved). Then, note that: (1) the proposed codes are linear, and (2) the noise in one branch of the receiver is independent of the noise in the other branch of the receiver (since they are uncorrelated and jointly Gaussian). Then, applying a union bound:

$$P_b \le \sum_{i,j} i\, n(i,j) P(\underline{0}_j \to \underline{1}_j) \tag{21}$$

where $\underline{0}_j$ and $\underline{1}_j$ are vectors of j 0's and j 1's, respectively, and $P(\underline{0}_j \to \underline{1}_j)$ is the probability of deciding the latter when the former was sent. Continuing, P_b is upper bounded as:

$$\begin{aligned}
&\le \sum_{i,j} i\, n(i,j) E_h[P(\underline{0}_j \to \underline{1}_j|h)] \\
&= \sum_{i,j} i\, n(i,j) E_h\left[Q\left(\sqrt{\frac{E_bj}{N_0}\sum_{l-0}^{L-1} h_l^2}\right)\right] \\
&= \sum_{i,j} i\, n(i,j) \int_0^\infty Q\left(\sqrt{\frac{E_bj}{N_0}\gamma}\right) f(\gamma)\mathrm{d}\gamma \\
&= \int_0^\infty \left[\sum_{i,j} i\, n(i,j) \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \mathrm{e}^{-\frac{E_b}{2N_0}\frac{j\gamma}{\sin^2\theta}}\mathrm{d}\theta\right] f(\gamma)\mathrm{d}\gamma \\
&= \frac{1}{\pi}\int_0^\infty \int_0^{\frac{\pi}{2}} \sum_{i,j} i\, n(i,j)\left[\mathrm{e}^{-\frac{E_b}{2N_0}\frac{\gamma}{\sin^2\theta}}\right]^j f(\gamma)\mathrm{d}\theta\mathrm{d}\gamma \\
&= \frac{1}{\pi}\int_0^\infty \int_0^{\frac{\pi}{2}} \frac{\partial}{\partial I}T(D,I)\Big|_{I=1,D=\exp\left(-\frac{E_b}{2N_0}\frac{\gamma}{\sin^2\theta}\right)} \\
&\quad \times f(\gamma)\mathrm{d}\theta\mathrm{d}\gamma.
\end{aligned}$$

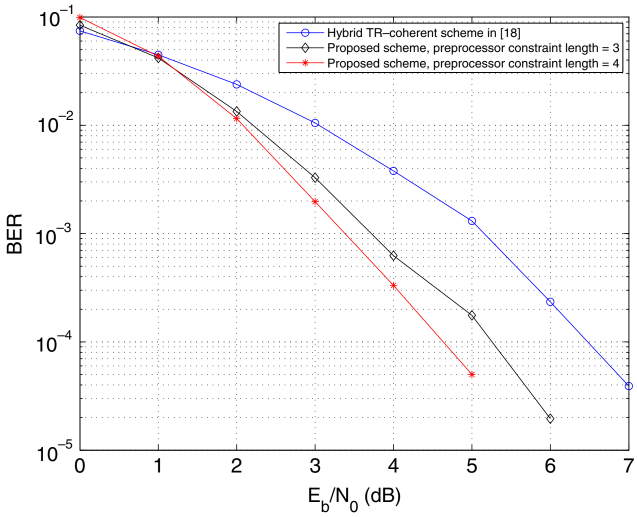

Fig. 4. BER versus E_b/N_0 curves of the proposed hybrid system with a coherent rake receiver operating over an AWGN channel. Performance of the hybrid TR scheme proposed in [18] is provided for comparison.

For small numbers of states, $T(D,I)$ can be found by hand, but this becomes unwieldy for larger numbers of states. In such cases, Appendix 6A1 of [22] gives an efficient method that can be extended to the case here.

4. Numerical results

Since the performance of the FSR receiver for the proposed hybrid scheme is the same as the FSR scheme in [15], we only focus on simulating the performance with the coherent rake receiver. In all simulations, a carrier-modulated, truncated root-raised-cosine pulse with a roll-off factor 0.25 is used as the UWB pulse shape $p(t)$. The 10-dB signal bandwidth is 1 GHz. We adopt channel models from the IEEE 802.15.3a [23] with a large delay spread (CM4) and consider a low-data-rate system operating at 1 Mbps and the number of pulses to represent a bit is $N_f = 10$. We assume perfect symbol timing and simulate the performances with ideal and estimated template waveforms. For each bit, the receiver combines ten strongest paths using maximal ratio combining (MRC). This is achieved by setting the delays τ_l of the correlator to those of the ten strongest paths. The integrator outputs are fed into a Viterbi decoder, which approximates the maximum likelihood sequence detector. The simulation also assumes a quasi-static fading model, i.e., the channel coefficients and delays do not change over each packet duration but change independently over different packets. The total number of packets is 2000, each with 256 bits; thus, the BER values are effectively the average over 2000 realizations of the channel coefficients.

Fig. 4 shows the performance of the proposed hybrid FSR-coherent system with coherent detection and soft input Viterbi decoding over AWGN channels. BER performances of the proposed hybrid scheme with the data pre-processor of constraint length three and constraint length four over multipath fading channels modeled by IEEE 802.15.3a CM4 are shown in Fig. 5. For comparison, the performance of the hybrid TR-coherent system with a coherent receiver in [18] in the same environment is included in

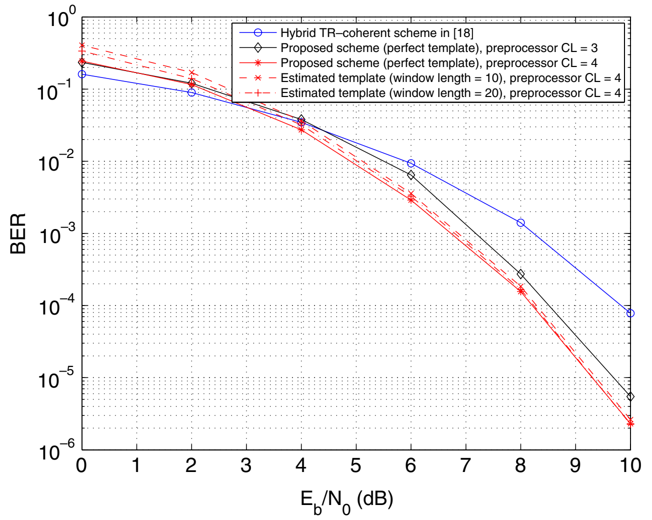

Fig. 5. BER versus E_B/N_0 curves of the proposed hybrid system with different preprocessor constraint lengths (CL) and a coherent rake receiver operating over an IEEE 802.15.3a multipath fading channel modeled by CM4. Performance of the hybrid TR scheme proposed in [18] is provided for comparison.

the two figures. Compared with the case of a perfect template waveform for coherent rake, with an estimated template derived using a window length of $N = 20$, the BER performance is degraded mainly in the low-E_b/N_0 region. It is also observed that with the data preprocessor of constraint length four, the proposed scheme achieves a gain of about 1.5 dB at a BER of 1×10^{-4} over the scheme in [18]; with the data preprocessor of constraint length three, the proposed scheme achieves a gain of about 1.2 dB over the scheme in [18].

5. Conclusion

We have proposed a hybrid UWB modulation scheme that allows reception by both coherent and FSR receivers, and derived closed-form equations and bounds for its performance. For coherent receivers, we introduce a data preprocessing scheme where both the reference signal and data signal in the proposed hybrid FSR-coherent system are generated from the input bits by the data preprocessor. This allows the coherent receiver to exploit not only the energy invested in the reference pulses, but also the memory introduced by the preprocessor for improved performance. The proposed scheme extends the preprocessor in [18] from an effective convolutional encoder with a constraint length two to three or more, improving the performance by more than 2 dB. Compared with coherent detection of the conventional uncoded TR scheme, the proposed scheme not only recovers the 3 dB loss due to the energy wasted in the reference signals, but also achieves an additional 4 dB coding gain. We also proposed a channel estimation scheme that exploits the FSR receiver structure for the coherent rake receiver and derived analytical bounds for the performance of the decoder in multipath environments.

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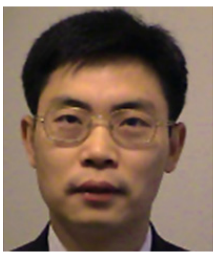

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