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Xiaojing Huang
University of Wollongong, huang@uow.edu.au

Y. Li
Twincall Pty Ltd, Ryde

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Performance of Impulse-Train-Modulated Ultra-Wideband Systems

Xiaojing Huang, *Member, IEEE*, and Yunxin Li, *Senior Member, IEEE*

***Abstract*—The performance of impulse-train-modulated ultra-wideband (UWB) systems for the ideal additive white Gaussian noise channel is analyzed in this letter. The derived formulae are also used to optimize the modulation parameter of a Gaussian monocycle UWB impulse radio.**

***Index Terms*—Additive white Gaussian noise (AWGN) channel, impulse radio (IR), ultra-wideband (UWB).**

I. Introduction

ULTRA-WIDEBAND (UWB) systems offer unique advantages, such as higher processing gain and multipath resolution, deeper material penetration, and more covert operation, over conventional narrowband systems [1]–[3]. For low-complexity implementation, UWB systems often use impulse-train modulations to carry information, so that they are sometimes called impulse radios (IRs). Considerable studies on UWB signal-propagation characterization [4]–[6], [15]–[17] and UWB system-performance evaluation [7]–[11] have been carried out over the recent years.

In this letter, the performance of three impulse-train-modulation schemes, i.e., biphase modulation (BPM), pulse position modulation (PPM), and hybrid modulation [12], [13], which is a combination of BPM and PPM, is analyzed for the ideal additive white Gaussian noise (AWGN) channel. Simpler closed-form bit-error probability (BEP) formulae are derived, and modulation-parameter selection for system performance optimization is illustrated.

II. Impulse-Train-Modulated UWB Signals and Detection

We consider UWB signals generated by modulating a pseudonoise (PN) impulse train with input data information and an ideal transmission channel with only AWGN interference. The simplified transmission model is shown in Fig. 1. The PN impulse train is generally expressed as $c(t) = \sum_i c_i \delta(t - iT_c - d_i)$, which consists of a series of Dirac delta impulses, with nominal impulse repetition period (called chip time) T_c, modulated by a direct-sequence $\{c_i\}$ and/or time-hopping sequence $\{d_i\}$, $d_i \in [0, T_c)$. Assuming a short direct-sequence and/or time-hopping sequence, so that the duration of $c(t)$ is the same as the data symbol interval T_s, the modulated signal $x(t)$ in the time period $nT_s \le t \le (n+1)T_s$ can be expressed as $a_n c(t - nT_s)$, $c(t - nT_s - ((1-a_n)/2)\Delta)$, or $a_n^{(1)} c(t - nT_s - ((1 - a_n^{(0)})/2)\Delta)$ for BPM, PPM, or hybrid modulation, respectively, where $a_n \in [-1, +1]$ is a binary data symbol, $a_n^{(1)}$ and $a_n^{(0)} \in [-1, +1]$ are the most significant bit (MSB) and the least significant bit (LSB), respectively, of a quaternary data symbol $2a_n^{(1)} + a_n^{(0)} \in [-3, -1, +1, +3]$, and Δ denotes the time shift when a_n (for PPM) or $a_n^{(0)}$ is -1. $y(t)$ denotes the convolution of $x(t)$ with the overall impulse response $w_r(t)$ of the transmitter antenna and the receiver antenna. $z(t)$ is a Gaussian noise with double-sided power spectral density $(1/2)N_0$. The attenuation factor α (a real-valued number) models the propagation of the UWB signal over the channel. The received UWB signal plus interference is expressed as

$$r(t) = \alpha y(t) + z(t). \tag{1}$$

Further denoting

$$g(t) = c(t) * w_r(t) = \int_0^{T_s} c(\tau) w_r(t - \tau) d\tau \tag{2}$$

which represents the received signal waveform to carry one data symbol, $y(t)$ in the time period $nT_s \le t \le (n+1)T_s$ can be expressed as $a_n g(t - nT_s)$, $g(t - nT_s - ((1 - a_n)/2)\Delta)$, or $a_n^{(1)} g(t - nT_s - ((1 - a_n^{(0)})/2)\Delta)$ for BPM, PPM, or hybrid modulation, respectively. Since all these modulations are memoryless, each transmitted data symbol can be detected independently from the received signal $r(t)$ in the time period $nT_s \le t \le (n+1)T_s$. Following a well-defined procedure [14] and defining two decision variables

$$U = \alpha \int_0^{T_s} r(t + nT_s) g(t) dt \tag{3}$$

$$V = \alpha \int_0^{T_s} r(t + nT_s) g(t - \Delta) dt \tag{4}$$

the decision rule for BPM is

$$a_n = \begin{cases} +1, & \text{if } U \ge 0 \\ -1, & \text{if } U < 0. \end{cases}$$

For PPM, the decision rule is

$$a_n = \begin{cases} +1, & \text{if } U \ge V \\ -1, & \text{if } U < V \end{cases}$$

with coherent detection (the sign of α is known) or

$$a_n = \begin{cases} +1, & \text{if } |U| \ge |V| \\ -1, & \text{if } |U| < |V| \end{cases}$$

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X. Huang is with the School of Electrical, Computer and Telecommunications Engineering, Faculty of Informatics, University of Wollongong, Wollongong, NSW 2522, Australia (e-mail: huang@uow.edu.au).

Y. Li is with the Twincall Education Center, Twincall Pty Ltd, Ryde, NSW 2112, Australia (e-mail: jeff@twincall.com).

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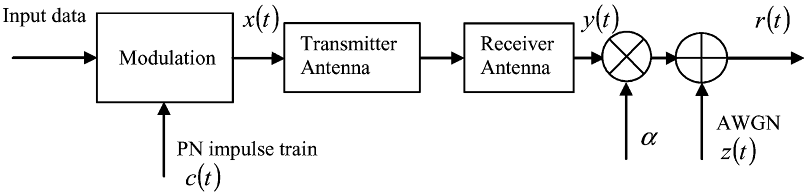

Fig. 1. Simplified UWB transmission model.

with noncoherent energy detection (no knowledge of α is required). For hybrid modulation, the decision rules are

$$a_n^{(1)} = \begin{cases} +1, & \text{if } U \geq |V| \text{ or } V \geq |U| \\ -1, & \text{if } U \leq -|V| \text{ or } V \leq -|U| \end{cases}$$

for MSB, which can be further simplified as

$$(a_n^{(1)} = \begin{cases} +1, & \text{if } U \geq -V \\ -1, & \text{if } U < -V \end{cases} \tag{5}$$

$$\text{and} \quad a_n^{(0)} = \begin{cases} +1, & \text{if } |U| \geq |V| \\ -1, & \text{if } |U| < |V| \end{cases}$$

for LSB.

III. The Bit-Error Probability Analyses

When $a_n = +1$ for BPM and PPM or $a_n^{(1)} = a_n^{(0)} = +1$ for hybrid modulation are transmitted, U is a Gaussian variable with mean $E_b = \alpha^2 \int_0^{T_s} g^2(t)dt$, which represents the received signal energy per bit, and variance $\sigma^2 = N_0 E_b/2$, and V is a Gaussian variable with mean $E_b\rho$ and the same variance σ^2, where

$$\rho = \frac{\int_0^{T_s} g(t)g(t-\Delta)dt}{\int_0^{T_s} g^2(t)dt} \tag{6}$$

is the normalized autocorrelation coefficient of $g(t)$ at offset Δ. Note that U and V are correlated if $\rho \neq 0$ with joint central moment $\mu = \mathbf{E}[(U - E_b)(V - E_b\rho)] = (N_0/2)E_b\rho$. The joint probability density function of them can be expressed as

$$p(u,v) = \frac{1}{\pi N_0 E_b \sqrt{1-\rho^2}} e^{-\frac{(u-E_b)^2 - 2\rho(u-E_b)(v-E_b\rho) + (v-E_b\rho)^2}{N_0 E_b (1-\rho^2)}}. \tag{7}$$

Since BPM is a binary modulation with antipodal signals, its BEP is well known as $Q(\sqrt{2\gamma_b})$ [14], where $\gamma_b = E_b/N_0$ is the normalized signal-to-noise ratio (SNR) per bit, and $Q(x) = (1/\sqrt{2\pi})\int_x^{+\infty} e^{-(t^2/2)}dt$. The BEP of the PPM with coherent detection is also easily found to be $1 - \int_{-\infty}^{+\infty}\int_{-\infty}^{u} p(u,v)dvdu = Q(\sqrt{(1-\rho)\gamma_b})$, which is the same as that of the binary modulation with correlated signals [14]. For PPM with noncoherent energy detection, the closed-form formula of its BEP can be found in the literature [14], but it is complicated, since the Marcum Q function and the modified Bessel function are involved. According to the decision rule provided in the previous section, we can alternatively evaluate this BEP as $P_e(\gamma_b) = 1 - \int_{-\infty}^{0}\int_{u}^{-u} p(u,v)dvdu - \int_0^{\infty}\int_{-u}^{u} p(u,v) dvdu \approx Q(\sqrt{(1+\rho)\gamma_b}) + Q(\sqrt{(1-\rho)\gamma_b})$ for $\gamma_b > 6.3$ (8 dB), at which, a reasonable BEP of less than 10^{-2} is secured.

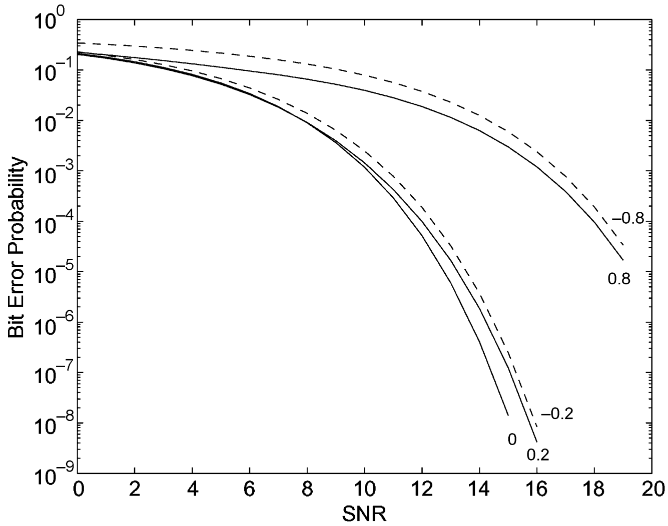

Fig. 2. Average BEP for hybrid modulation with different values of ρ as a function of SNR per symbol.

Finally, let us evaluate the BEP for the detection of a hybrid-modulated signal. Assuming that $a_n^{(1)} = +1$ is transmitted, the MSB error probability is derived from (5) as $P_e^{(1)}(\gamma_s) = 1 - \int_{-\infty}^{\infty}\int_{-u}^{\infty} p(u,v)dvdu = Q(\sqrt{(1+\rho)\gamma_s})$ where $\gamma_s = \alpha^2 \int_{-\infty}^{+\infty} g^2(t)dt/N_0$ denotes the SNR per symbol. The LSB error probability has the same expression as that for the PPM noncoherent energy detection except that γ_b should be replaced by γ_s, i.e., $P_e^{(0)}(\gamma_s) = P_e(\gamma_s)$. Therefore, the averaged BEP is $(1/2)P_e^{(1)}(\gamma_s) + (1/2)P_e^{(0)}(\gamma_s) \approx Q(\sqrt{(1+\rho)\gamma_s}) + (1/2)Q(\sqrt{(1-\rho)\gamma_s})$, which is plotted in Fig. 2.

Table I summarizes the above analytical results. We see that the BPM has the best performance, whereas the PPM with coherent detection offers better performance than the PPM with noncoherent energy detection. The hybrid modulation offers similar performance as the PPM with noncoherent detection at the same SNR per symbol, but the bit rate is doubled. If transmitted at the same bit rate, the required SNR per bit for the quaternary hybrid-modulated UWB system is 3 dB less than that of the binary PPM UWB system in order to achieve similar BEP.

TABLE I
SUMMARY OF BEPS OF DIFFERENT IMPULSE-TRAIN-MODULATED UWB SYSTEMS

Biphase Modulation	$Q\left(\sqrt{2\gamma_b}\right)$	γ_b : SNR per bit
Pulse Position Modulation	$Q\left(\sqrt{(1-\rho)\gamma_b}\right)$ (coherent) $Q\left(\sqrt{(1+\rho)\gamma_b}\right)+Q\left(\sqrt{(1-\rho)\gamma_b}\right)$(noncoherent)	 For γ_b > 8dB
Hybrid Modulation	$Q\left(\sqrt{(1+\rho)\gamma_s}\right)$ (MSB) $Q\left(\sqrt{(1+\rho)\gamma_s}\right)+Q\left(\sqrt{(1-\rho)\gamma_s}\right)$(LSB) $Q\left(\sqrt{(1+\rho)\gamma_s}\right)+\frac{1}{2}Q\left(\sqrt{(1-\rho)\gamma_s}\right)$(average)	γ_s : SNR per symbol for γ_s > 8dB for γ_s > 8dB

IV. SYSTEM OPTIMIZATION

From the above analysis, we see that the performance of the impulse-train-modulated UWB system depends not only on the modulation scheme and the detection method, but also on the normalized autocorrelation function of the received signal waveform (except for BPM). It is obvious that for a given modulation scheme and a chosen detection method, the system performance can be optimized through appropriate signal waveform design and modulation-parameter selection. Assuming that the received signal impulse $w_r(t)$ has already been decided, we illustrate in this section how to minimize the BEP by selecting the time shift Δ used in PPM and hybrid modulation.

First, we assume that the chip time is large enough so that any impulse delayed by Δ does not overlap with the next impulse, which is the case for most UWB systems using impulse train with low duty cycle, so that the normalized autocorrelation function of $g(t)$ is simply the normalized autocorrelation function of $w_r(t)$, i.e., $\rho = (\int_0^{T_s} g(t)g(t-\Delta)dt)/(\int_0^{T_s} g^2(t)dt) = (\int_{-\infty}^{+\infty} w_r(t)w_r(t-\Delta)dt)/(\int_{-\infty}^{+\infty}[w_r(t)]^2dt)$. Then we choose the impulse response of the transmitter antenna as the ideal Gaussian monocycle pulse, i.e., $w_t(t) = 2\sqrt{\pi e}(t/\tau)e^{-2\pi(t/\tau)^2}$, where τ is a time constant related to the pulse width, which can be defined as 2τ. We also assume that the effect of the receiver antenna on the transmitted impulse is ideally modeled as a derivation operation [10], [11], so that the normalized received impulse $w_r(t)$ at the output of the receiver antenna is $w_r(t) = [1-4\pi(t/\tau)^2]e^{-2\pi(t/\tau)^2}$. The normalized autocorrelation coefficient at time delay Δ is then derived to be $\rho = [1-4\pi(\Delta/\tau)^2 + (4\pi^2/3)(\Delta/\tau)^4]e^{-\pi(\Delta/\tau)^2}$. $w_t(t)$ and $w_r(t)$ as a function of the normalized time t/τ, and ρ as a function of the normalized time delay Δ/τ are plotted in Fig. 3. We see that at $\Delta/\tau = \sqrt{(5-\sqrt{10})/2\pi} \approx 0.54$, ρ takes the minimum value $(4/3)(2-\sqrt{10})e^{-((5-\sqrt{10})/2)} \approx -0.618$. At $\Delta/\tau = \sqrt{(3\pm\sqrt{6})/2\pi} \approx 0.93$ or 0.296, ρ becomes zero. For $(\Delta/\tau) > 2$, ρ also approaches zero. Therefore, for PPM with coherent detection, the optimum time shift is 0.54τ, at which, the system gives optimum performance. For PPM with noncoherent energy detection or for hybrid modulation, the time shift should be chosen as 0.93τ or 0.296τ. If these systems can accommodate large time delay, Δ should be larger than 2τ.

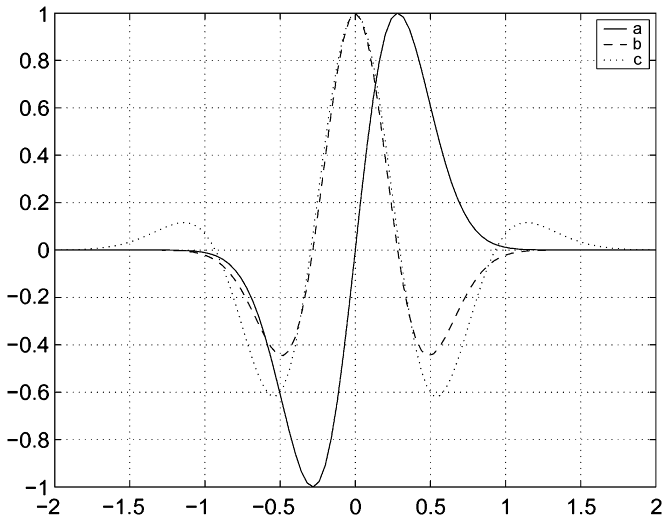

Fig. 3. (a) Transmitted Gaussian monocycle pulse $w_t(t)$ with pulse width of 2τ. (b) Received pulse $w_r(t)$ at output of receiver antenna. (c) Normalized autocorrelation coefficient ρ as a function of normalized time delay Δ/τ.

V. CONCLUSIONS

We have shown that the performance of impulse-train-modulated UWB systems depends not only on the modulation scheme and detection method, but also on the normalized autocorrelation function of the received signal impulse. This observation gives rise to the issue of optimal UWB signal design and modulation-parameter selection for system performance optimization. As a design example, the optimal time shifts for the PPM and the hybrid modulation are determined when a Gaussian monocycle UWB impulse is used.

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