ULTRA WIDE-BAND COMMUNICATIONS WITH BLIND CHANNEL ESTIMATION BASED ON FIRST-ORDER STATISTICS

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Abstract — Ultra Wide-Band (UWB) communication holds great potential for significantly improved data rate in future wireless systems. Accurate channel estimation and synchronization are critical for successful operation of a UWB system. We propose in this paper a completely blind channel estimation and synchronization algorithm for UWB systems that employ pulse-position modulation. The algorithm exploits the first-order cyclostationarity in the received signal and performs certain circular deconvolution. The complexity is extremely low — only some "overlap-add" operations and FFT operations are needed. The algorithm is capable of simultaneously estimating multiple (say, more than 60) channel taps and there is no ambiguity in either the amplitude or the phase of the estimated channel. It is shown that using estimated channel from 500 information symbols, the performance can approach that with known channel within 2 dB in signal to noise ratio for bit error rate less than 0.01.

Keywords: Ultra Wide-band, blind channel estimation, first-order statistics, pulse position modulation, cyclostationary,

1. INTRODUCTION

Ultra-Wideband (UWB) communication has attracted much attention recently from both academia and industry, partly because of its huge potential for offering short-range broadband wireless services using frequencies currently already allocated to other applications [5, 7, 8].

Channel estimation and synchronization are important tasks for the successful operation of a UWB system because the demodulation of information symbols depends critically on the availability of full or partial channel information. There has been some preliminary work of channel estimation [2] and synchronization [6]. Both data-aided (training based) and non-data-aided (blind) methods are considered in [2], but the methods require multi-dimensional search to maximize the log-likelihood function and therefore have high complexity. Timing acquisition and tracking are considered in [6] based on second-order cyclostationary, but no channel estimation is attempted. Pilot waveform assisted modulation is proposed in [9] to aid the channel estimation task at the receiver.

In this paper, we will consider the blind channel estimation problem for a UWB system employing pulse position modulation (PPM). Our method does not require pilot symbols or pilot waveforms [2, 9]. We will also obtain coarse symbol timing to a precision that is enough for symbol demodulation. The proposed methods is based on the first-order statistics of the received signal and has very low complexity — only some basic signal processing operations such as "overlap add", and Fourier transform and/or filtering are needed.

2. SYSTEM MODELING

Consider a single user in a UWB communication system employing PPM modulation. For simplicity, we consider binary modulation, but the proposed method can also be applied to M-ary modulation with $M > 2$. Let $w(t)$ denote a monocycle, which usually is a twice-differentiated Gaussian pulse of width less than one nano-second. The spectral shaping pulse for binary PPM modulation in a "time-hopping" UWB system [4] can be written as

$$p(t) = \sum_{j=0}^{N_f-1} w(t - jT_f - c_jT_f), \qquad (1)$$

where T_f is the *frame* duration, N_f is the number of frames per symbol, and $(c_0, c_1, \ldots, c_{N_f-1})$ is a sequence of real numbers between 0 and 1 that specifies the user-specific time-hopping code. The code c_j's should be user-dependent and this dependence is not shown here because we will be considering only one user and treat all other users as additive noise. Also notice that a symbol consists of multiple frames rather than vice versa. We assume that the user-specific code $(c_0, c_1, \ldots, c_{N_f-1})$ is known at the receiver and is symbol-periodic, i.e., it does not change from symbol to symbol.

Having specified the spectral shaping pulse $p(t)$, the transmitted signal of the user of interest employing binary PPM can be written as

$$s(t) = \sum_{i=-\infty}^{\infty} p(t - iT_s - b_i\Delta), \qquad (2)$$

where T_s is the symbol duration and is equal to N_fT_f, b_i's are the binary zero-or-one information symbols with equal probability $1/2$, and Δ is a fixed constant, which is a PPM parameter. The parameter Δ is usually slightly larger than the width of the monocycle $w(t)$ and is much smaller than the frame duration T_f (and hence much smaller than the symbol duration T_s).

Let $h(t)$ denote the impulse response of the channel that the user of interest's signal experiences. Let $x(t) = p(t) \star h(t)$, where "$\star$" denotes linear convolution. Assume that other than the multipath channel $h(t)$, there is an additional delay τ between the transmitted signal of the user of interest and the received signal; τ is assumed to be a real number between 0 and T_s without loss of generality. The received signal can be written as

$$y(t) = \sum_{i=-\infty}^{\infty} x(t - iT_s - b_i\Delta - \tau) + n(t), \qquad (3)$$

where $n(t)$ is the additive noise plus the multiple-user interference; $n(t)$ is assumed to be zero-mean and white.

We assume that the channel remains constant for a duration of I symbols. The channel estimation will therefore be based on I consecutive received symbols. In addition, we assume that **AS1:** The channel is causal and Finite Impulse Response (FIR), and the duration of the channel is less than a fraction of T_s; i.e., $h(t) = 0$, if $t < 0$ or $t > \epsilon T_s$, where $0 < \epsilon < 1$ and ϵ is known. We require $\epsilon < 1$ for the unique identification of τ.

3. PROPOSED CHANNEL ESTIMATION METHOD

The blind channel estimation problem can be formulated as follows: Given a segment of $y(t)$, say for $0 \le t \le T$, and the spectral shaping pulse $p(t)$, estimate the channel $h(t)$ and the timing information τ.

Our proposed channel estimation method is based on the cyclostationary in the first-order statistics of the received signal $y(t)$. To derive the method, we first evaluate the expectation $\mathrm{E}[y(t)]$ over the information symbols b_i's and the additive noise $n(t)$.

For a fixed channel realization $h(t)$, since b_i has equal probability of being zero or one, and $n(t)$ is zero-mean, we have [cf. (3)]

$$\bar{y}(t) := E[y(t)] = \sum_{i=-\infty}^{\infty} \frac{1}{2}[x(t-iT_s-\tau)+x(t-iT_s-\Delta-\tau)]. \qquad (4)$$

It can be readily seen that $\bar{y}(t)$ is periodic with period T_s and non-zero, thanks to the PPM modulation format. This simple but important observation will provide the basis of our channel estimation and synchronization algorithm.

Since $x(t) = h(t) \star p(t)$, we can re-write $\bar{y}(t)$ in (4) as $\bar{y}(t) = [\frac{1}{2}p(t) + \frac{1}{2}p(t-\Delta)] \star \tilde{h}(t-\tau)$ where $\tilde{h}(t) = \sum_{i=-\infty}^{\infty} h(t-iT_s)$ is a periodically extended "channel". Let us also define the "average pulse" $p_a(t) = \frac{1}{2}[p(t)+p(t-\Delta)]$ and periodically extend it as

$$\tilde{p}_a(t) = \sum_{i=-\infty}^{\infty} p_a(t - iT_s). \qquad (5)$$

In practice, $p(t)$ and $p(t-\Delta)$ can be easily designed to be confined to one symbol period, in which case $\tilde{p}_a(t) = p_a(t)$, for $0 \le t \le T_s$. Denote the circular convolution $z_{12}(t)$ of two periodic signals $z_1(t)$ and $z_2(t)$ both with period T_s as

$$z_{12}(t) = z_1(t) \otimes z_2(t) = \int_0^{T_s} z_1(a) z_2(t-a)da. \qquad (6)$$

With this notation and defining $\tilde{h}_\tau(t) := \tilde{h}(t-\tau)$, we can write $\bar{y}(t)$ in terms of circular convolution as

$$\bar{y}(t) = \tilde{p}_a(t) \otimes \tilde{h}_\tau(t). \qquad (7)$$

Given $\bar{y}(t)$, which can be estimated from the received signal $y(t)$, and $\tilde{p}_a(t)$, which can be constructed from the known spectral shaping pulse $p(t)$, we would like to estimate $h(t)$ and τ under the assumption AS1. We will decompose the problem into two steps. In the first step, we estimate $\tilde{h}_\tau(t)$ and in the second step, we estimate τ from the estimate of $\tilde{h}_\tau(t)$.

Theoretically, the problem of estimating $\tilde{h}_\tau(t)$ from $\bar{y}(t)$ can be solved by computing the Fourier series of both sides of (7). The Fourier series $Z[k]$ of a periodic signal $z(t)$ of period T_s is defined as $Z[k] = \int_0^{T_s} z(t)e^{-2\pi kt}dt$. The Fourier series of both sides of (7) is then

$$\bar{Y}[k] = \tilde{P}_a[k]\tilde{H}[k], \qquad (8)$$

where $\bar{Y}[k]$, $\tilde{P}_a[k]$, and $\tilde{H}[k]$ are the Fourier series of $\bar{y}(t)$, $\tilde{p}_a(t)$, and $\tilde{h}(t)$, respectively. It follows that $\tilde{H}[k]$ can be recovered as $\bar{Y}[k]/\tilde{P}_a[k]$, provided that $\tilde{P}_a[k] \neq 0$. This simple division is *ad hoc* in nature. But in AWGN, if we ignore the constraints on the channel $h(t)$ (e.g., its finite duration ϵT_s), then this per-frequency division minimizes the least-square distance between $\bar{Y}[\cdot]$ and $\tilde{H}[\cdot]$, and hence is the maximum likelihood estimator for $\tilde{H}[\cdot]$ and its time-domain counterpart $\tilde{h}_\tau(t)$.

Once $\tilde{H}[k]$ is found, we can obtain an estimate $\hat{\tilde{h}}_\tau(t)$ of $\tilde{h}_\tau(t)$ by performing the inverse Fourier transform. We propose to estimate τ by maximizing the energy of the multipath channel estimate:

$$\hat{\tau} = \arg\max_{0 \le \tau \le T_s} \int_\tau^{\tau+T_s} [\hat{\tilde{h}}_\tau(t)]^2 dt. \qquad (9)$$

We develop the algorithmic steps for channel estimation and synchronization as follows.

<u>Steps for channel estimation and synchronization</u>

- S1. Construct $\tilde{p}_a(t)$ from $p(t)$ according to (5) and its Fourier series $\tilde{P}_a[k]$, $k = 0, 1, \ldots$.
- S2. Receive $y(t)$ for $0 \le t \le (I-1)T_s$;
- S3. Obtain an estimate $\hat{\bar{y}}(t)$ of $\bar{y}(t) = E[y(t)]$ by performing the following "overlap-add" operation: $\hat{\bar{y}}(t) = \frac{1}{I}\sum_{i=0}^{I-1} y(t+iT_s)$, $0 \le t \le T_s$.

S4. Compute the Fourier series $\hat{\tilde{Y}}[k]$ of $\hat{\tilde{y}}(t)$ and obtain $\hat{\tilde{H}}[k]$ as $\hat{\tilde{H}}[k] = \hat{\tilde{Y}}[k]/\tilde{P}_a[k]$ if $\tilde{P}_a[k] \neq 0$, and 0 otherwise.

S5. Do inverse Fourier transform of $\hat{\tilde{H}}[k]$ to obtain $\hat{\tilde{h}}_\tau(t)$.

S6. Estimate τ according to (9) to obtain $\hat{\tau}$.

S7. Estimate the channel $h(t)$ as $\hat{h}(t) = \hat{\tilde{h}}_\tau(t+\hat{\tau})$ if $0 \leq t \leq \epsilon T_s$, and 0 otherwise.

Detection algorithm

S8. Based on the estimates $\hat{h}(t)$ and $\hat{\tau}$, construct the "difference" signal $x_d(t) = [p(t) - p(t-\Delta)] \star \hat{h}(t-\hat{\tau})$.

S9. Obtain the decision statistics $d_i = \int_{-\infty}^{\infty} y(t) x_d(t - iT_s)dt$, and decide that $\hat{b}_i = 0$ if $d_i \geq 0$, and 1 otherwise.

Remark 1 (parameterization) We do not parameterize the channel in terms of a number of taps and their delays and amplitudes. There are therefore essentially an infinite number of parameters to estimate: the function value of $h(t)$ for $t \in [0, \epsilon T_s]$. This problem as it stands can be ill-posed because it is impossible to estimate infinite unrelated parameters from the finite data. Therefore, in practice, in order to produce reliable estimation of the channel, some more constraints on it, in addition to AS1, have to be made. These constraints can be on the bandwidth of the channel frequency response, the number of discrete taps, or based on some other finite parameterization schemes. Also, in practice, we will work in the discrete-time domain and use Fast Fourier Transform (FFT) in steps 4 and 5. In general, we need to sample the received signal at a rate much higher than the symbol rate, which can be a big challenge in practice. This drawback is a disadvantage of the proposed blind channel estimation algorithm as compared with other non-blind transmitted-reference methods in e.g., [9]. Recently, it has been shown that for synchronization and channel estimation purposes, it is possible to sample that received signal at a rate lower than the Nyquist rate [1, 3].

Remark 2 (Identifiability) If there are zeros in the Fourier transform of $\tilde{p}_a(t)$, then consistency in the estimation of $h(t)$ may not be achievable. Some frequency component of the channel impulse response may not be recoverable due to the lack of excitation at the corresponding frequency in the signal $\tilde{p}_a(t)$. The consistency for estimating $h(t)$ can be achieved if and only if the Fourier series $\tilde{H}[k]$ of $\sum_{i=-\infty}^{\infty} h(t - iT_s)$ is zero for every k such that $\tilde{P}_a[k]$ is zero. Even if the channel cannot be consistently estimated due to lack of excitation at certain frequencies of $\tilde{p}_a(t)$, the effect of missing certain frequency components in the channel estimation is not severe: it is as if the channel has filtered out those frequencies. However, it should be noticed that due to the division involved in Step 4 of the algorithm, noise will be enhanced when $\tilde{P}_a[k]$ is small.

Remark 3 (Blindness) Our algorithm is completely blind, i.e., no training is needed. For phase-shift keying (PSK) modulation, blind algorithms usually cannot resolve some phase ambiguity due to the inherent rotational invariance in the transmitted signal. Here, thanks to the PPM based transmission, our blind algorithm does not have ambiguity.

4. SIMULATION RESULTS

In this section, we present some simulation results for the proposed algorithms. We will assume that the channel is FIR and follows a tap-delay-line model. The simulations followed the steps described in Section 3, with the continuous-time Fourier series and inverse Fourier transform replaced by FFT and IFFT operations.

The system parameters are: $T_f = 11ns$, $N_f = 12$, $T_s = 132ns$. The parameter Δ is set $1ns$ or $2ns$ and it will be indicted which one is used. A discrete-time equivalent model of the UWB system is used. The channels we simulated had 32 or 64 equally spaced taps with spacing $1ns$. The channel taps were assumed to be independent, zero-mean, and Gaussian distributed with equal variance. The average channel energy is normalized to one. We remark that the UWB channel estimator developed in [2] does not seem possible for this setup because the parameter space (32 or 64 taps) is too large to search using the methods proposed therein.

In our simulations, we had modified the algorithm presented in Section 3 to avoid large noise enhancement caused by small elements of $\tilde{P}_a[k]$. Specifically, the conditions $\tilde{P}_a[k] \neq 0$ and $\tilde{P}_a[k] = 0$ were modified to $|\tilde{P}_a[k]| \geq \gamma$ and $|\tilde{P}_a[k]| < \gamma$, respectively, where the *tolerance threshold* γ is a small positive number. In our simulations, we have chosen γ in an *ad hoc* way.

We report in Fig. 1 the MSE as a function of the data record lengths I for fixed SNR's (10 and 15 dB). Larger data record length provided better performance in both MSE, as can be expected. We used $\gamma = 0.06$ in this simulation.

For fixed data length $I = 500$, which corresponds to a delay of 66 μs, we report in Fig. 2 and Fig. 3 the MSE and BER, respectively, as functions of the SNR. We have considered four combinations of choices for PPM parameter Δ and the tolerance threshold γ: i) $\gamma = 0.1$, $\Delta = 1ns$; ii) $\gamma = 0.1$, $\Delta = 2ns$; iii) $\gamma = 0$, $\Delta = 1ns$; and iv) $\gamma = 0$, $\Delta = 2ns$. We chose different choices for Δ to see its effect on the system performance. The performance (both MSE and BER) was better if we use $\gamma = 0.1$ to avoid large noise enhancement, as compared with the case when $\gamma = 0$. There was little performance difference between $\Delta = 1ns$ and $\Delta = 2ns$. In Fig. 3, we also plot the performance when the channel was known. It was noted that the performance with estimated channel came close to that with known channel within 2 dB in SNR for BER less than 10^{-2}.

The MSE tended to some error floor as SNR increases (not shown in Fig. 2), which was due to the missing frequency components caused by the introduced threshold γ. But for the SNR range that is shown, namely below 18 dB, introducing γ reduced the MSE because it limited the noise enhancement. Error floor in BER for high SNR was not observed.

5. CONCLUSIONS

We developed a fully blind channel estimation and symbol synchronization algorithm for ultra-wideband communications employing pulse position modulation. The algorithm was based on first-order cyclostationarity of the received signal. Based on FFT operations, the algorithm has low complexity and can deal with channels with a large number of taps that are difficult to estimate using existing parameterized searching algorithms. Simulations show that the blind algorithm together with a simple matched-filter based detector performs within 2 dB of a clairvoyant receiver.

In future work, we will investigate possibilities of utilizing decision-feedback type and iterative type of algorithms that iterate between channel estimation and symbol detection, which will further refine the channel estimation accuracy and improve the system performance in bit error rate.

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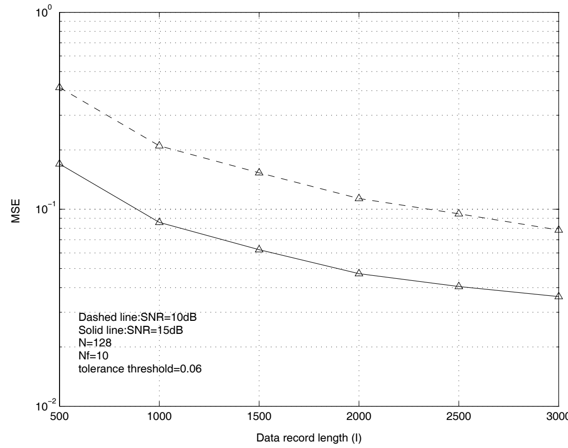

Figure 1: MSE versus I for fixed SNR

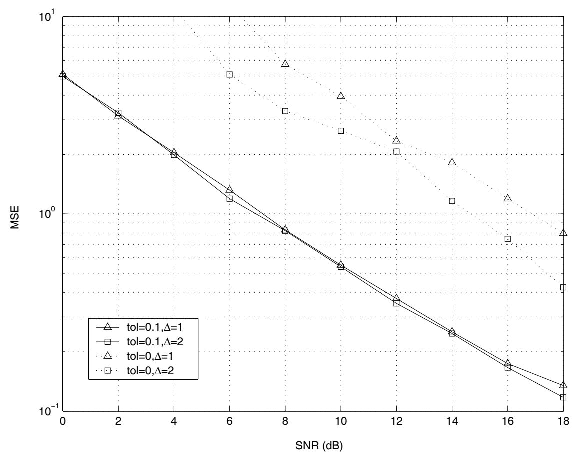

Figure 2: MSE versus SNR for fixed data record length I

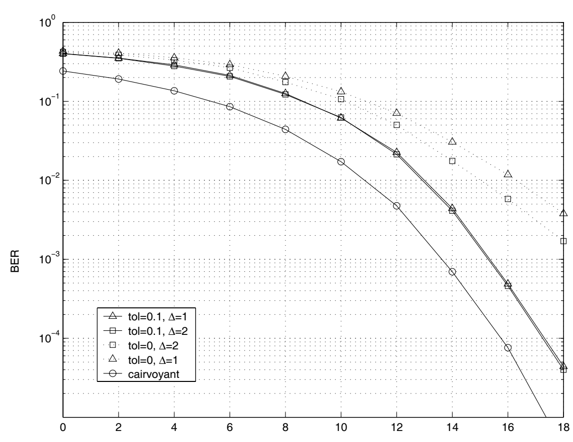

Figure 3: BER versus SNR for fixed data record length I. The performance with known channel is also depicted.