BLIND CHANNEL ESTIMATION FOR MULTIPLE ACCESS UWB COMMUNICATIONS BASED ON PERIODIC TIME HOPPING AND PULSE-RATE MODELING

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ABSTRACT

In an impulse radio based ultra-wideband (UWB) wireless communication system, transmitted signal propagates through a rich multipath channel. So far, channel estimation has been addressed for a single-user system by a computationally demanding maximum-likelihood technique. For a multiuser system, a pulse-rate channel input/output vector model is first derived in this paper which exhibits a tri-linear structure with a new-defined input, effective channel and hopping-code dependent matrix for each user. Such a structure can be regarded as similar to a code-division multiple access (CDMA) system. Therefore covariance based blind channel estimation methods successful therein can be applied to estimate UWB channels.

1. INTRODUCTION

With recent release of the spectral mask from the Federal Communications Commission, the communication society has witnessed an increasing interest in impulse radio (IR) or ultra-wideband (UWB) technology [1]. A conventional IR system transmits trains of time-hopping (TH) short-duration pulses with a low duty cycle and uses pulse position modulation (PPM). Therefore, multipath down to path delay differentials in nanosecond is resolvable at the receiver, significantly mitigating multipath distortion and providing path diversity [2].

In an UWB system, a receiver consists of waveform correlators [1]. To fully capture the signal energy spread over multiple paths, the receiver needs to know channel parameters when correlation is performed. However, in a dense multipath wireless environment, channel information is not known *a priori*. Although channel parameters can be estimated by maximum likelihood (ML) methods [2], [3], the multiple access interference (MAI) is approximated as a Gaussian process which may not be accurate. Meanwhile, they are computationally expensive. Low complexity channel estimation methods with explicit consideration of MAI need to be developed.

In this paper, we study UWB channel estimation based on the second order statistics (SOS) of the received signal in a multiple access UWB system. For low complexity and easy implementation, some existing estimation and detection techniques also use the SOS, for example in acquisition of the arrival time of the first path of UWB channels [4], linear detection of input symbols [6]. First, an UWB system based on pulse-rate sampling is shown to follow a similar model as a direct sequence (DS) code-division multiple access (CDMA) system, similar to an observation in [5], [6] but clearly in a tri-linear format. Then we clearly define a matrix for each user from its unique time-hopping (TH) sequence. Each matrix can be treated as a code matrix, similar to the code matrix constructed from spreading codes in a CDMA system [7]. But it consists of only zeros and ones, indicating whether there exists a contribution to the received signal during a particular time interval from a multipath channel or not. Locations of zeros and ones are different for different users. Then, using either covariance matching idea [7] or subspace concept [8] aided by unique code matrices, corresponding channel vectors for all users can be estimated.

2. SYSTEM MODEL

Consider a multiple access (MA) TH UWB system with K users. The transmitted baseband UWB signal from user k can be described by [6]

$$\alpha_k(t) = \sqrt{\mathcal{P}_k} \sum_{i=-\infty}^{\infty} w(t - iT_f - c_k(i)T_c - \tau_{I_k(\lfloor i/N_f \rfloor)}) \quad (1)$$

where $\mathcal{P}_k$ is the kth user's transmission power, $w(t)$ is the baseband monopulse, T_f is the frame duration, N_f is the number of frames over which an M-ary PPM symbol repeats, $c_k(i) \in [0, N_c - 1]$ is a periodic hopping sequence with period equal to one symbol period. Each chip has duration T_c. $I_k(\lfloor i/N_f \rfloor) \in [0, M-1]$ is the kth user's information bearing symbol during the ith frame, $\tau_{I_k(\lfloor i/N_f \rfloor)} = I_k(\lfloor i/N_f \rfloor)\sigma$ is the corresponding modulation delay in a multiples of σ seconds. Assume $T_f = N_c T_c$ and $T_c = M\sigma$. If we define $w_m(t) \triangleq w(t - m\sigma)$ where $m = 0, \cdots, M-1$ and $s_{k,m}(\lfloor i/N_f \rfloor) = \delta\Big(I_k(\lfloor i/N_f \rfloor) - m\Big)$, then (1) may be expressed by linear modulation in a chip rate as [6]

$$\alpha_k(t) = \sqrt{\mathcal{P}_k} \sum_{i=-\infty}^{\infty} \sum_{m=0}^{M-1} u_{k,m}(i) w_m(t - iT_c) \quad (2)$$

where chip index has replaced frame index in (1),

$$u_{k,m}(i) = s_{k,m}\Big(\lfloor i/(N_c N_f) \rfloor\Big) \widetilde{c}_k(i),$$

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$$\widetilde{c}_k(i) = \delta\Big(\lfloor i/N_c \rfloor N_c + c_k(\lfloor i/N_c \rfloor) - i\Big).$$

It is clear according to (2) that input $u_{k,m}(i)$ is modulated by waveform $w_m(t)$ at a chip rate. The transmitted signal $\alpha_k(t)$ propagates through a linear channel with channel $\bar{g}_k(t)$. At the receiver, the channel output is first passed through a matched filter matched to the monopulse $w(t)$. We can define a front-end effective channel including effects from modulated pulse at the transmitter, propagation channel and matched filter at the receiver by $g_{k,m}(t) = w_m(t) \star \bar{g}_k(t) \star w(-t)$ where $\star$ denotes convolution. Considering additive white Gaussion noise (AWGN) $v(t)$, the output of the matched filter becomes

$$y(t) = \sum_{k,i_1,m} \sqrt{\mathcal{P}_k} u_{k,m}(i_1) g_{k,m}(t - i_1 T_c) + v(t). \quad (3)$$

Assume each effective channel has maximum delay spread $q\sigma$. Then $y(t)$ is sampled very σ seconds to yield a discrete-time output $y(n) = y(t)|_{t=n\sigma}$. Using the discrete-time version of the effective channel and invoking $T_c = M\sigma$, we obtain a pulse-rate model [5]

$$y(n) = \sum_{k,m} \sum_{i_2=0}^{q} \sqrt{\mathcal{P}_k} u_{k,m}(\lfloor \frac{n - i_2}{M} \rfloor) g_{k,m}(i_2) + v(n). \quad (4)$$

Consider P symbol intervals of data samples with coppresponding time instants $nMN_cN_f + p$ for $p = 1, \cdots, MPN_cN_f$ and collect them in a big vector $\boldsymbol{y}_n$ of length $\nu = MPN_cN_f$, a vector form model follows

$$\boldsymbol{y}_n = \sum_{k,m,l} \sqrt{\mathcal{P}_k} \boldsymbol{C}_{k,l} \boldsymbol{T}_m \boldsymbol{g}_k s_{k,m}(n + l) + \boldsymbol{v}_n \quad (5)$$

where symbol index l takes all integers $-\lceil q/(MN_cN_f) \rceil$, $\cdots, P-1$, $\boldsymbol{g}_k$ is an unknown channel vector for user k, $\boldsymbol{T}_m = [\boldsymbol{0}, \boldsymbol{I}, \boldsymbol{0}]^T$ is a tall selection matrix in order to obtain the mth subchannel for user k, $\boldsymbol{C}_{k,l}$ is a matrix constructed from corresponding $\widetilde{c}_k(i)$ and is uniquely determined by the TH sequence. It only contains zeros and ones and repeats from symbol to symbol because the TH sequence has period equal to one symbol interval. By employing this structure, all channels can be estimated based on the covariance/correlation of $\boldsymbol{y}_n$.

3. BLIND CHANNEL ESTIMATION

There exist two typical covariance based methods to be employed. The direct one is to apply covariance matching. Since $\boldsymbol{y}_n$ has non-zero mean because $E\{s_{k,m}(n+l)\} = 1/M$, correlation of $\boldsymbol{y}_n$ involves cross terms among different channel vectors. Therefore, covariance is more convenient for channel estimation. Denote the mean of $\boldsymbol{y}_n$ by $\bar{\boldsymbol{y}}$, the correlation by $\boldsymbol{R}$ and covariance by $\boldsymbol{A}$. It can be easily verified that the cross variance of $s_{k,m_1}(n+l_1)$ and $s_{k,m_2}(n+l_2)$ in (5) is $\sigma^2_{s,m_1,m_2} = 1/M\delta(m_1 - m_2)\delta(l_1 - l_2) + 1/M^2(1 - \delta(l_1 - l_2))$. Then from (5), we obtain

$$\boldsymbol{A} = \sum_{i,k,m_1,m_2} \sigma^2_{s,m_1,m_2} \boldsymbol{C}_{k,i} \boldsymbol{T}_{m_1} \boldsymbol{G}_k \boldsymbol{T}^H_{m_2} \boldsymbol{C}^H_{k,i} + \sigma^2_v \boldsymbol{I}$$

where $\boldsymbol{G}_k = \mathcal{P}_k \boldsymbol{g}_k \boldsymbol{g}_k^H$, σ^2_v is the noise power. Since an estimate of $\boldsymbol{A}$ can be easily obtained from N data vectors as

$$\widehat{\boldsymbol{A}} = \frac{1}{N} \sum_{n=1}^{N} \boldsymbol{y}_n \boldsymbol{y}_n^H - \bar{\boldsymbol{y}} \bar{\boldsymbol{y}}^H, \;\; \bar{\boldsymbol{y}} = \frac{1}{N} \sum_{n=1}^{N} \boldsymbol{y}_n,$$

all $\boldsymbol{G}_k$ together with σ^2_v can be estimated first by minimizing the squared Frobenius norm of an error matrix [7]

$$(\widehat{\sigma}^2_v, \widehat{\boldsymbol{G}}_1, \cdots, \widehat{\boldsymbol{G}}_K) = \arg \min_{\sigma^2_v, \boldsymbol{G}_1, \cdots, \boldsymbol{G}_K} ||\boldsymbol{A} - \widehat{\boldsymbol{A}}||^2_F. \quad (6)$$

Singular value decomposition (SVD) on each $\widehat{\boldsymbol{G}}_k$ yields a singular vector corresponding to its maximum singular value. That singular vector becomes an estimate of $\boldsymbol{g}_k$ up to a multiplicative scalar.

Subspace method is another powerful technique. If the noise subspace $\boldsymbol{U}_n$ is obtained from eigenvalue decomposition of $\boldsymbol{R}$, then it is observed from (5) that the signature waveform of $s_{k,m}(n+i)$ which is $\boldsymbol{C}_{k,i} \boldsymbol{T}_m \boldsymbol{g}_k$ is orthogonal to $\boldsymbol{U}_n$ for all possible m. Therefore, a criterion can be developed as follows [8]

$$\widehat{\boldsymbol{g}}_k = \arg \min_{||\boldsymbol{x}||=1} \sum_{i,m=0} ||\boldsymbol{U}_n^H \boldsymbol{C}_{k,i} \boldsymbol{T}_m \boldsymbol{x}||^2 \quad (7)$$

Under some conditions, (7) gives a unique channel vector up to a multiplicative scalar.

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