SIGNAL PROCESSING MODEL AND RECEIVER ALGORITHMS FOR A HIGHER RATE MULTI-USER TR-UWB SYSTEM

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ABSTRACT

Transmit-reference (TR) schemes are commonly used only in low data rate ultra-wideband (UWB) systems because of many restrictions on the pulse spacing, frame and symbol periods (should be longer than the channel length). This paper extends our previous research that tries to remove these restrictions to enable a higher data rate application in a multiuser context. Based on the fact that most UWB channels are highly uncorrelated, we can formulate a CMDA-like signal processing model for an asynchronous multiuser system. Blind and iterative algorithms are derived, of which the performances are compared and verified in simulations.

Index Terms— UWB, multi-user separation, transmit-reference

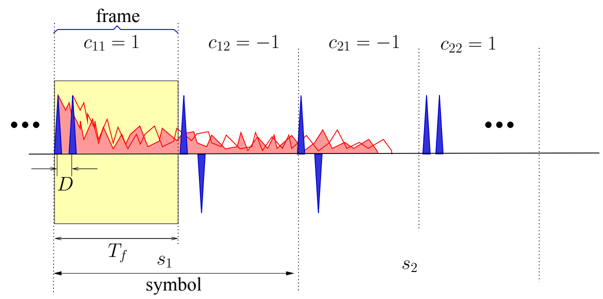

Fig. 1. Pulse sequence structure

1. INTRODUCTION

Since first introduced in [1], the transmit-reference (TR) scheme has received special attention as a low complexity, low data rate candidate for ultra-wideband (UWB) communication systems. The main advantage of a TR scheme is that, by transmitting pulse pairs and using a correlation receiver, we do not have to estimate the individual channel taps but gather the total channel energy to detect a single symbol, which will greatly reduce the receiver's complexity. However, this also implies a very low data rate since no inter-frame interference (IFI) and no inter-pulse interference (IPI) is allowed.

In [2], we have proposed a TR-UWB scheme that allows both IFI and IPI. The delay D between two pulses in a doublet was chosen much smaller than the frame period T_f and the channel length T_h, and $T_f < T_h$. By imposing these conditions, the frame rate can be at least three times higher than other TR-UWB schemes that often assume $D > T_h$ and $T_f > 2D + T_h$ to prevent IPI and IFI. The main idea is that by using integrate and dump, and oversampling to get P samples per frame instead of sliding window, we divide the channel into P unknown parameters called energies of the channel segments. This will help us maintain the low sampling rate nature (much lower than the Nyquist rate) of the original TR-UWB scheme and enable higher data rate applications. In [3], we also dealt with the narrowband interference problem for this scheme.

A signal model based on second-order Volterra systems is developed in [4] that considers IFI in frame differential UWB systems but the algorithm's complexity quickly grows in longer (and more practical) UWB channels and in multiuser context. In [5], the proposed multiuser TR-UWB accepts IPI but no IFI, and assumes perfect frame synchronization. In this paper, we extend our work to a multi-user context that allows both IPI and IFI. The simple data model developed previously for the single user case is now extended to a multi-user data model. Based on this signal processing data model, blind and iterative receiver algorithms are straightforwardly derived, of which the performance comparisons are shown in simulations. Again, by using integrate and dump, and oversampling, users are allowed to transmit signal asynchronously (with known users' offsets at the accuracy of the sampling period) as a CMDA-like system in [6], [7].

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2. DATA MODEL

2.1. Setup

Consider a multi-user TR-UWB system where K asynchronous users transmit blocks of N_s symbols with a known (relative) offset D_k, each symbol consists of N_c frames (similar to "chips" in long code CDMA). The frame period T_f can be smaller than the channel length T_h, which means that there are inter-frame interferences (IFIs). All the users' channels are typical UWB indoor channels, with channel taps that are assumed highly uncorrelated. A pair of pulses (doublet) is transmitted at the start of each frame, in which the first pulse is kept fixed as the reference and the second pulse, delayed by D seconds, carries information (the frame/chip value) in its polarity. Here we choose the delay between two pulses D in a doublet to be very small compared to the frame period and the channel length, i.e. $D \ll T_f < T_h$. The structure of the transmitted pulse sequence by a user is illustrated in Fig. 1.

At the receiver, we use a simple correlator followed by an "integrate and dump" operator as shown in Fig.2. The oversampling rate $P = T_f/T_{sam}$ is chosen such that there is at least one sample per frame, where T_{sam} is the sampling period, much smaller than the frame period T_f and the symbol period T_s.

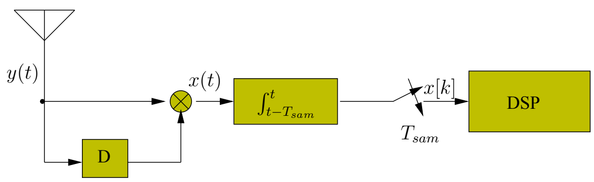

Fig. 2. Autocorrelation receiver

2.2. Single user

For simplicity and clarity, we first derive the data model for the single user case as in [2]. The received signal at the antenna output for a block of N_s symbol $\mathbf{s}$ is

$$y(t) = \sum_{i=1}^{N_s} \sum_{j=1}^{N_f} [h(t-(ij-1)T_f) + s_i c_{ij} h(t-(ij-1)T_f - D)] \quad (1)$$

where $h(t) = h_p(t) * g(t) * a(t)$ is the convolutional product of the physical channel $h_p(t)$, the UWB pulse shape $g(t)$ and the antenna template $a(t)$; $\mathbf{s} = [s_1 \cdots s_{N_s}]^T$ is the source symbol vector, $s_i \in \{+1, -1\}$; and $\mathbf{c}_i = [c_{i1} \cdots c_{iN_f}]^T$ is the code vector for the i-th symbol s_i, $c_{ij} \in \{+1, -1\}$.

At the multiplier output, the signal $x(t) = y(t)y(t-D)$ will be integrated and dumped at the oversampling rate $P = T_f/T_{sam}$. Due to uncorrelated channels, the cross terms can be ignored, and the data model in [2] can be easily extended to include the "code" terms c_{ij}. The resulting discrete samples $x[n] = \int_{(n-1)T_{sam}}^{nT_{sam}} x(t)dt$, $n = 1, \cdots, (NsNf-1)P + T_h/T_{sam}$ are stacked into a column vector $\mathbf{x}$, which can be expressed as (see Fig. 3)

$$\mathbf{x} = \mathbf{H}\text{diag}\{\mathbf{c}_1, \cdots, \mathbf{c}_{N_s}\}\mathbf{s} + \text{noise} \quad (2)$$

where $\mathbf{H}$ contains shifted versions of a "new" channel vector $\mathbf{h}$, with entries

$$h[n] = \int_{(n-1)T_{sam}}^{nT_{sam}} h^2(t)dt \qquad k = 1, \cdots, T_h/T_{sam}$$

So $h[n]$ is actually the energy of the n-th segment of the channel $h(t)$. Here we replace the continuous channel with thousands of channel taps, by only T_h/T_{sam} new channel parameters in vector $\mathbf{h}$.

The data model in (2) can also be rewritten in another form

$$\mathbf{x} = \mathcal{C}(\mathbf{I} \otimes \mathbf{h})\mathbf{s} + \text{noise} \quad (3)$$

where $\mathcal{C}$ is the code matrix of size $((N_f N_s - 1)T_f + T_h)/T_{sam} \times (T_h N_s)/T_{sam}$, with entries taken from $\mathbf{c}_i$ and structure illustrated in Fig. 3.

2.3. Multiple delays

In the previous section, we used a single delay to simplify the mathematical expressions and the receiver structure. However, this will cause spikes at $1/D$ frequency intervals in the spectrum of the received UWB signal, which may conflict with the FCC spectrum mask. To avoid this problem, the delay between two pulses in a doublet can be made to vary from frame to frame, of which the pattern is known. To match all delays, multiple banks of correlation receivers are used. The signals at the outputs are added together before being processed in the DSP block.

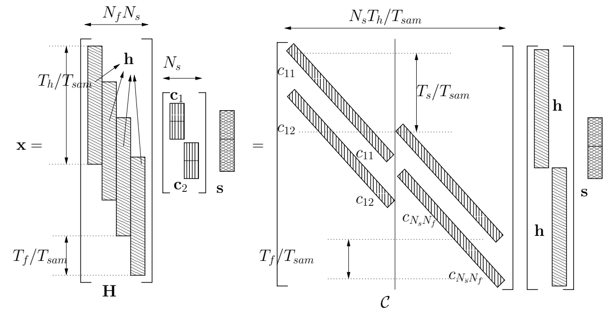

Fig. 3. The data model

Since the channels are uncorrelated, we have the same resulting data model as in (2) and (3), which is independent of the delay value(s). In this case, the role of multiple delays is merely to smooth out the signal's spectrum.

2.4. Multiple users

Now we derive the data model for an asynchronous multi-user system where each user is characterized by a code matrix $\mathbf{C}_k$, channel vector $\mathbf{h}_k$ and a known offset D_k. Since each user goes through a different channel, we can safely assume that two different channels are uncorrelated, which means that all the cross-terms between two users' channels are noise-like. Therefore, the received signal will be modeled as

$$\begin{aligned} \mathbf{x} &= \sum_{k=1}^{K} \mathbf{H}_k \text{diag}\{\mathbf{c}_{k1}, \cdots, \mathbf{c}_{kN_s}\}\mathbf{s}_k + \text{noise} \\ &= \sum_{k=1}^{K} \mathcal{C}_k(\mathbf{I} \otimes \mathbf{h}_k)\mathbf{s}_k + \text{noise} \end{aligned}$$

where $\mathbf{H}_k$, $\mathcal{C}_k$ are the channel matrix and code matrix for the k-th user, both of which include the offset D_k. The multi-user data model can be straightforwardly derived as

$$\mathbf{x} = \mathcal{C}\mathcal{H}\mathbf{s} + \text{noise} \quad (4)$$

where $\mathcal{C} = [\mathcal{C}_1 \cdots \mathcal{C}_K]$ is the known code matrix; $\mathcal{H} = \text{diag}\{\mathbf{I} \otimes \mathbf{h}_1, \cdots, \mathbf{I} \otimes \mathbf{h}_K\}$ is the unknown channel matrix, in which $\mathbf{h}_k$ contains the unknown channel coefficients; and $\mathbf{s} = [\mathbf{s}_1^T \cdots \mathbf{s}_K^T]^T$ contains the unknown source symbols.

3. RECEIVER ALGORITHMS

3.1. Single user algorithm

In certain cases, we only need to focus on a single user while treating all other users' signal as interference (MUI). We can use the same iterative algorithm proposed in [2], which can converge rapidly with a good initial channel estimate. Here blind algorithms are introduced as the initial estimate by exploiting the "code" structure in data model (3), assuming that the code of the user of interest is known.

If the symbol period is chosen longer than the channel length, i.e. $T_s = N_f T_f > T_h$, we can implement a blind algorithm by multiplying both sides with the left-inverse of the code matrix $\mathcal{C}$, giving

$$\mathbf{x}' := \mathcal{C}^\dagger \mathbf{x} \approx (\mathbf{I} \otimes \mathbf{h})\mathbf{s} \tag{5}$$

Restack vector $\mathbf{x}'$ into a matrix $\mathbf{X}'$ of size $(T_h/T_{sam}) \times N_s$, we have

$$\mathbf{X}' \approx \mathbf{h}\mathbf{s}^T$$

Subsequently, the channel vector $\mathbf{h}$ and the source symbols $\mathbf{s}$ are found, up to an unknown scaling, by taking the rank-1 decomposition (SVD) of $\mathbf{X}'$.

Alternatively, a "matched filter" algorithm can be similarly implemented by replacing $\mathcal{C}^\dagger$ with $\mathcal{C}^T$ in (5).

3.2. Multi-user algorithms

The data model in (4) suggests that we can apply existing results in CDMA literature in the receiver of this TR-UWB scheme. The only difference is that the "channel matrix" $\mathcal{H}$ consists of energies of channel segments, instead of the channel fingers in CDMA.

The algorithms can be straightforwardly derived as in [7], [6]. Some constraints are needed for the code matrix $\mathcal{C}$ or $\mathcal{CH}$ to be left-invertible. If orthogonal codes are used, the simple "matched filter" algorithm can achieve a good performance. Other blind techniques proposed in [8, 9] for long-code CDMA can be applied. However, we only focus on some simple blind and iterative algorithms to reduce the receiver's complexity.

3.2.1. Blind algorithm

When the codes and the offsets are known, the $\mathcal{C}$ matrix in (4) is completely known. By premultiplying both sides of (4) with the left-inverse $\mathcal{C}^\dagger$ or with $\mathcal{C}^T$, we reduce the multiple users problem into K separate single user problems. Similar to section 3.1, we can blindly estimate both the source symbols and the channel coefficients of all users. This can be used as the initial estimate for the iterative algorithms discussed next.

3.2.2. Iterative join source-channel estimation

The data model in (4) is expressed in the "code" by "channel" by "data" form: $\mathbf{x} = \mathcal{CH}\mathbf{s}$, where $\mathcal{H} = \text{diag}\{\mathbf{I} \otimes \mathbf{h}_1, \cdots, \mathbf{I} \otimes \mathbf{h}_K\}$ and $\mathbf{s}$ is the stacking of all users' source symbols $\mathbf{s}_i$. By using a simple property of the Kronecker product, we can rewrite the model in the "code" by "data" by "channel" form: $\mathbf{x} = \mathcal{CS}\mathbf{h}$, where $\mathcal{S} = \text{diag}\{\mathbf{s}_1 \otimes \mathbf{I}, \cdots, \mathbf{s}_K \otimes \mathbf{I}\}$ and $\mathbf{h}$ is the stacking of all users' channel coefficient vectors $\mathbf{h}_i$. This allows us to implement an alternating LS algorithm to iteratively estimate the source symbols and the channel coefficients of all users.

With an initial channel estimate $\mathbf{h}^{(0)}$, for iteration index $k = 1, 2, \cdots$ until convergence,

- keep the channel $\mathbf{h}^{(k-1)}$ fixed, construct $\mathcal{H}$ matrix, estimate the source symbols

$$\mathbf{s}^{(k)} = (\mathcal{CH})^\dagger \mathbf{x}$$

- keep the source symbols $\mathbf{s}^{(k)}$ fixed, construct $\mathcal{S}$ matrix, estimate the channel coefficients

$$\mathbf{h}^{(k)} = (\mathcal{CS})^\dagger \mathbf{x}$$

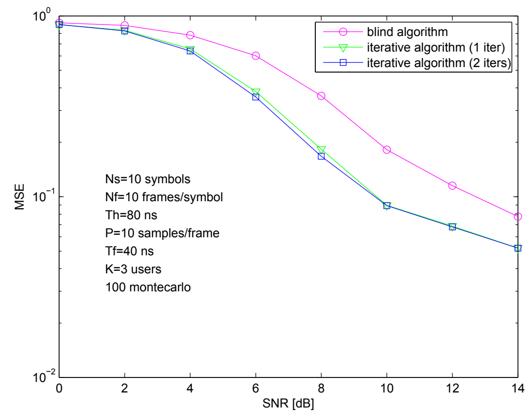

Fig. 4. Channel MSE vs. SNR performance for CM1

After iterations, step 1 is repeated once more to get the final estimate of the source symbols. Hard decision can be used in step 1 to further improve the performance.

Although this is an iterative algorithm that uses matrix inversion operations $(\mathcal{CH})^\dagger$ and $(\mathcal{CS})^\dagger$, it was shown in [7] that, by exploiting the sparse structures of these matrices, we can efficiently implement these operations. In the next section, we will show how much improvement can be achieved from the blind algorithm to the iterative algorithm, with just one or two iterations.

4. SIMULATION RESULTS

We simulate a multi-user TR-UWB system with $K = 3$ equal-powered users transmitting pulse pairs (Gaussian monocycles of width 0.2 ns, spacing $D = 0.5$ ns). These doublets are arranged into blocks of $N_s = 10$ symbols, each symbol consists of $N_f = 10$ frames of duration $T_f = 40$ ns. We use the IEEE channel models (CM1, CM2) which are always longer than the frame period, implying that inter-frame interference (IFI) does exist. The non-ideal antenna effect is also included, i.e. a measured non-ideal antenna response is convolved with the channel.

The received signal is integrated and dumped at oversampling rate of $P = 10$ samples per frame, which means the sampling period is $T_{sam} = T_f/P = 4$ ns. Since oversampling is used to resolve IFI problem, it is obvious that the more samples we get (per frame) the better resolution and performance the algorithms can achieve. However, due to the limits of practical ADCs, the sampling rate can not be arbitrarily high. The choice of sampling rate is the system's tradeoff between performance and complexity. The fact that we can change the sampling rate at will is another advantage of this scheme.

All the users' symbols and codes are generated randomly. A random offset (but known) at the T_{sam} level is assigned to each user. 100 Monte Carlo runs are used to compare the BER vs. SNR and channel MSE vs. SNR plots between the blind algorithm and the iterative algorithms that use the blind algorithm as the initial channel estimate. The reference curve is the zero-forcing receiver when the channel coefficients are completely known. Here, SNR is defined as the pulse energy over the noise spectral density, and channel MSE is defined as the mean square error of the channel coefficient vectors (not the individual channel taps).

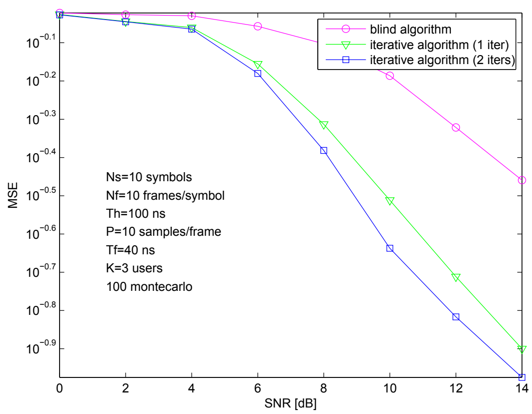

Fig. 5. Channel MSE vs. SNR performance for CM2

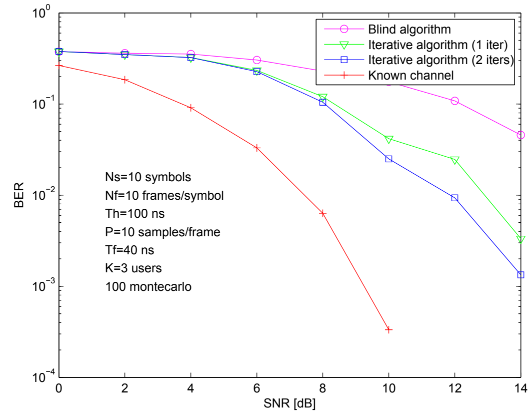

Fig. 7. BER vs. SNR performance for CM2

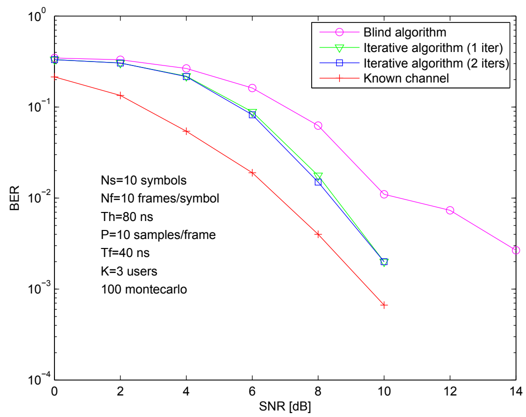

Fig. 6. BER vs. SNR performance for CM1

In Fig. 4 and Fig. 5, it is shown that the iterative algorithms have about 2-3dB gain for CM1 and 4-5dB for CM2 over the initial blind algorithm, which will help improve the BER performance.

From Fig. 6 and Fig. 7, we can see that although the blind algorithm can work well under relaxed conditions, e.g. longer symbol period or higher sampling rate, it seems to have a flooring effect in the high SNR region in higher data rate applications. However, the iterative algorithms do eliminate the effect. The gain increases as SNR increases. The reason is that, in the iterative algorithms, we apply the pseudo-inverse of a much taller matrix $(\mathcal{CH})^\dagger$ instead of $\mathcal{C}^\dagger$. However, the iterative algorithms still have about 2-4 dB gap compared to the "optimal" curve when the channel vectors are known.

5. CONCLUSIONS

Taking into account the practical and the standard (IEEE) channel data, i.e. highly uncorrelated channels, we have established an approximate and simple data model that allows IFI and can be easily extended to the multi-user case. Based on the multi-user data model, receiver algorithms are derived in a CDMA-like context, in which the iterative algorithms can achieve a good performance with only a few iterations. Existing techniques can be used to invert such sparse matrices to reduce the receiver's complexity.

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