Transmitter-based Channel Equalization and MUI Suppression for UWB Systems

Mo Chen and Xiaohua (Edward) Li

Abstract – **In ultra-wide bandwidth (UWB) communication systems, dense multipath and severe multiuser interference (MUI) make the receiver high in hardware complexity and power consumption. It is desirable to shift the computation-demanding work to users with plentiful power and computation resources. We propose two prefiltering schemes which combat the MUI and multipath effects in the transmitter so that the receivers of users without sufficient resources can be simplified to just fixed correlators.**

Keywords – **ultrawideband system, transmitter, multipath equalization, multiuser interference, channel estimation**

I. INTRODUCTION

ULTRA-WIDEBAND (UWB) technology has attracted considerable research interests due to its promising ability to provide high data rate for the short-range indoor data communications. However, the existence of dense multipath propagation (number of channel taps >=40 [2][3]) and multiuser interference limits its performance. Although RAKE receiver is very effective for multipath mitigation, its complexity goes up linearly with the number of RAKE fingers which may be as high as l28 [2].

Considering the typical indoor applications of UWB systems where a number of simple devices (end users) such as TV, PDA, home appliances and toys, communicate with an access point, it is desirable to reduce the complexity of both the transmitter and the receiver of those low cost end users. On the other hand, the complexity of the access point is not very critical because there can be only one of such device in each house to provide networking capability just as IEEE 802.11 wireless LAN. Therefore, it is necessary to shift the complex receiving operations such as channel equalization and MUI suppression from the end users to the access point.

For the uplink transmission from end users to the access point, traditional UWB transmitting and receiving techniques can be directly used because they require only simple transmitters. However, for the downlink transmission from the access point to the end users, we need new techniques to reduce the complexity of the receivers of the end users. Therefore, in this paper, we consider in particular the downlink transmission and try to perform channel equalization and MUI suppression in the transmitter instead of the receiver.

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The desire to keep UWB receiver simple and power efficient makes complex receiving algorithms unattractive. If the receiver structure at the end users is simplified to a conventional matched filter (which only requires the desired user's code), the hardware complexity and power consumption will be greatly relieved.

Although there have not been such work in UWB systems, much work has been done on algorithms that exploit the fact that the channel impulse responses are known to the transmitter so that multiuser detection and equalization can be performed at the transmitter, not at the receiver. Besides simplified receivers, there are other advantages such as bandwidth efficiency due to the remove of training and the optimal scheduling of transmission according to channel conditions.

In [4], transmitter-based filters, called "precode", are designed to minimize the mean-square error between the received signal and a block of symbols. The method performs remarkably well if the receiver includes a RAKE processor, which unfortunately does not reduce receiver's complexity. In addition, there are many algorithms such as [5] that address joint optimization of transmitter and receiver, which, however, only make the receiver more complex.

In contrast, algorithms in [1][8] are proposed to make the receiver as simple as a fixed matched filter. In [8], a common signal for all users is determined in the transmitter before transmitting. After passing through different multipath channels to different users, the signal is just filtered by a matched filter at the receiving end of each user. Unfortunately this algorithm is extremely high in computational complexity. On the other hand, in [1] the transmitter filters the users' signals using "prerake" filters which are in fact the time-inverted complex-conjugated channels of the corresponding users. This prerake method mitigates the multipath problem, but unfortunately does not significantly reduce MUI.

Although transmitter-based prerake or joint transmitter-receiver design techniques have been widely investigated and are promising for significant performance improvement, a practical problem is how the transmitter knows the channels. Many papers rely on channel status feedback from the receivers to the transmitter, which unfortunately wastes too much bandwidth, especially for wireless communications with time-varying channels. Another applicable scenario is fast time division duplex where the transmitter can estimate the channels from previously received signals. However, for most systems, transmitters and receivers may use different frequency bands and thus their channels are different.

For UWB systems, such a problem no longer exists. Because all the transmitters and receivers share the same spectrum and perform transmitting or receiving at the same time, it is possible for the access point to predict the downlink channels from the uplink received signals because the channels from it to the end users will be the same as those from the end users. Therefore, in this paper, we assume that the transmitter knows the channels. Then we focus on transmitter-based MUI and multipath mitigation by two effective zero-forcing schemes used in the transmitters.

This paper is organized as follows. In Section II, the UWB system model is described. In Section III, prerake filter [1] is restated and new methods are proposed. In Section IV, we use simulations to demonstrate the performance of the new methods and to compare them with the prerake filter. Finally, conclusions are given in Section V.

II. SYSTEM DESCRIPTION

Consider the transmission from the access point to K end users with UWB spread spectrum signaling through multipath channels, where each user's binary baseband pulse amplitude modulated (BPAM) information signal can be represented as either

$$s_n^{(k)} = \sum_{m=-\infty}^{\infty} \sum_{j=1}^{N_q} w(n - mN_d - jN_f - (c_w)_j^{(k)} N_c) d_m^{(k)} \tag{1}$$

for time-hopping (TH) PAM [2] or

$$s_n^{(k)} = \sum_{m=-\infty}^{\infty} \sum_{j=1}^{N_q} w(n - mN_d - jN_c)(c_p)_j^{(k)} d_m^{(k)} \tag{2}$$

for DS-CDMA [3], where d_m is the mth data bit, $(c_p)_j$ is the jth chip of the PN code, $(c_w)_j$ is the jth code phase defined by the PN code, and $w(n)$ is the discrete pulse waveform (Gaussian or it's derivative). Parameter N_q represents the number of pulses to be used per data bit, N_c is the chip length, and N_f is the nominal pulse repetition interval. Bit length is $N_d = N_q N_c$ with DS-UWB and $N_d = N_q N_f$ with TH-UWB.

Assume a time-invariant baseband multipath channel for each user, e.g., the kth user's multipath channel is $h_n^{(k)}$. Since the access point simply adds all users' signals together to form the transmitting signal, if without any extra filtering at the transmitter and without noise, the user l receives the signal component of user k as

$$y_n^{(k)} = s_n^{(k)} \otimes h_n^{(l)} , \tag{3}$$

where $\otimes$ denotes convolution. The overall received signal at the user l is then

$$r_n^{(l)} = \sum_{k=1}^{K} y_n^{(k)} + v_n^{(l),I} + v_n^{(l),t} , \tag{4}$$

where $v_n^{(l),I}$ denotes the narrowband interference which is assumed to be white Gaussian noise, and $v_n^{(l),t}$ is the thermal noise. We can combine $v_n^{(l),I}$ and $v_n^{(l),t}$ together as AWGN $v_n^{(l)} = v_n^{(l),I} + v_n^{(l),t}$. Then in order to formulate (4) into the matrix form, we construct $M \times 1$ dimensional received sample vector as

$$\begin{aligned} [r_n^{(l)}, \cdots, r_{n-M+1}^{(l)}]^T &= \sum_{k=1}^{K} \mathsf{H}^{(l)} \mathbf{s}^{(k)} \\ &\triangleq \sum_{k=1}^{K} \begin{bmatrix} h_0^{(l)} & \cdots & h_{L_l-1}^{(l)} & & \\ & \ddots & & \ddots & \\ & & h_0^{(l)} & \cdots & h_{L_l-1}^{(l)} \end{bmatrix} \begin{bmatrix} s_n^{(k)} \\ \vdots \\ s_{n-M-L_l+2}^{(k)} \end{bmatrix} + \begin{bmatrix} v_n^{(l)} \\ \vdots \\ v_{n\text{-}M+1}^{(l)} \end{bmatrix}, \end{aligned} \tag{5}$$

where $(\cdot)^T$ denotes transposition. The channel matrix $\mathsf{H}^{(l)}$ for the user l is with dimension $M \times (M + L_l - 1)$, where L_l is the channel length.

III. ALGORITHM DEVELOPMENT

III.1. Prerake scheme

For convenience, let us first restate the prerake filters of [1], whose block diagram is shown in Fig. 1.

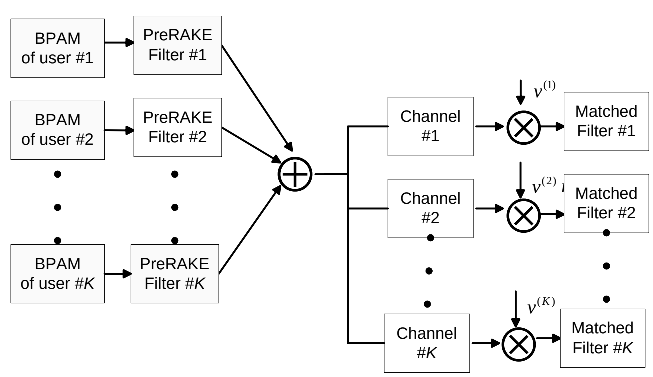

Figure.1 Block diagram of the prerake scheme for the UWB downlink transmission.

Signal of each user k is first filtered by the time-inverted and complex-conjugated channel of this user. The channel effect during transmission is then equivalent to maximal ratio combining, and is the same as applying a traditional rake receiver for each user. Therefore, at the receiver of each user, a simple matched filter consisting of the code sequence can be directly used for dispreading.

The composite impulse response of the prerake filter of the user k and the channel of the user l is

$$u_n^{(k,l)} = h_{L_k-n}^{*(k)} \otimes h_n^{(l)}, \tag{6}$$

where $1 \le k,l \leqslant K$ and each $\{u_n^{(k,l)}\}$ has length $L_k + L_l - 1$. For $k=l$, the composite channel $\{u_n^{(k,k)}\}$ has a unique strong peak whereas for $k \neq l$ such a peak may not exist. This provides the basis for maximal-ratio combining utilizing the channels along.

Although the pre-rake method mitigates multipath, there is still leftover ISI, and thus the error rates at the receivers will not be small enough because the matched filters in the receivers can not mitigate such residual ISI.

III.2. Pre-equalization scheme

To further remove the multipath effect, one way is to replace the prerake filters in Fig.1 with zero-forcing (ZF) pre-equalization filters $f_n^{(k)}$, which change the composite impulse response $u_n^{(k,k)} = f_n^{(k)} \otimes h_n^{(k)}$ in (6) into an ideal

$$u_n^{(k,k)} = \delta_{n-\Delta_k}, \tag{7}$$

where δ_n is the delta function, and Δ_k is a certain number of delay. In the matrix form, it becomes

$$\begin{aligned}(\mathbf{f}^{(k)})^T \mathrm{H}^{(k)} &= \begin{bmatrix} f_0^{(k)} & \cdots & f_{M-1}^{(k)} \end{bmatrix} \begin{bmatrix} h_0^{(k)} & \cdots & h_{L_k-1}^{(k)} & & \\ & \ddots & & \ddots & \\ & & h_0^{(k)} & \cdots & h_{L_k-1}^{(k)} \end{bmatrix} \\ &= \begin{bmatrix} 0 & \cdots & 1 & \cdots & 0 \end{bmatrix}_{1\times(M+L_k-1)} = (\mathbf{e}^{(k)})^T,\end{aligned} \tag{8}$$

where $\mathbf{e}^{(k)}$ is a unit vector with a unique non-zero value in the $\Delta_k th$ entry. The solution to the $M \times 1$ filter $\mathbf{f}^{(k)}$ can be found to be the multiplication of $(\mathbf{e}^{(k)})^T$ and the pseudo-inverse of $\mathrm{H}^{(k)}$.

III.3. MUI pre-mitigating scheme

As we mentioned above, although the prerake and the pre-equalization schemes mitigate the multipath effects, they do not significantly reduce MUI. The multiuser interference becomes especially strong when the channels of users are correlated.

To decorrelate the channels and to reduce the multiuser interference, we add an MUI mitigating filter at the transmitter to filter the composite (summation) signal from all users after prerake filters. The proposed diagram is shown in Fig.2.

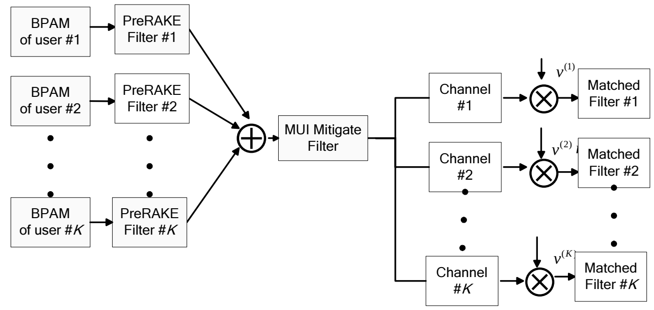

Figure 2. Proposed scheme with MUI pre-mitigating filter for UWB downlink transmission.

The impulse response of the MUI mitigating filter, g_n, has the following property

$$\begin{cases} g_n \otimes u_n^{(k,k)} = \delta_{n-\Delta_k}; \\ g_n \otimes u_n^{(k,l)} = 0; \quad for\ k \neq l \end{cases}, \tag{9}$$

where $1 \leqslant k, l \leqslant K$. Similarly, we can formulate (9) in matrix form. Let $\mathbf{g}$ be the $M \times 1$ vector to represent the MUI pre-mitigating filter and the matrix $\mathbf{U}^{(k,l)}$ with dimension $M \times (L_k + L_l + M - 2)$ to represent the composite channel matrix obtained from the convolution of the prerake filter of user k and the channel of user l.

$$\mathbf{U}^{(k,l)} = \begin{bmatrix} u_0^{(k,l)} & \cdots & u_{L_k+L_l-2}^{(k,l)} & & \\ & \ddots & & \ddots & \\ & & u_0^{(k,l)} & \cdots & u_{L_k+L_l-2}^{(k,l)} \end{bmatrix}. \tag{10}$$

The vector $\mathbf{d}^{(k,l)}$ of size $(L_k + L_l + M - 2) \times 1$ is defined as

$$\mathbf{d}^{(k,l)} = \begin{cases} [0, \ \cdots, \ 1, \ \cdots \ 0]^T, & k = l \\ [0, \ \cdots, \ 0, \ \cdots \ 0]^T, & k \neq l \end{cases} \tag{11}$$

where 1 is in the $\Delta_k th$ entry.

In addition, we can construct a matrix $\mathbf{V}$ of dimension $M \times K^2 (L_k + L_l + M - 2)$ that includes all matrices in (10) as $\mathbf{V} = [\mathbf{U}^{(1,1)}, \mathbf{U}^{(1,2)}, \cdots, \mathbf{U}^{(k,l)}, \cdots, \mathbf{U}^{(K,K)}]$. Similarly we construct the vector $\mathbf{d}$ of dimension $K^2 (L_k + L_l + M - 2) \times 1$ from (11) as $\mathbf{d}^T = [(\mathbf{d}^{(1,1)})^T, (\mathbf{d}^{(1,2)})^T, \cdots, (\mathbf{d}^{(k,l)})^T, \cdots, (\mathbf{d}^{(K,K)})^T]$. Then the equation (9) becomes

$$\mathbf{g}^T \mathbf{V} = \mathbf{d}^T. \tag{12}$$

Similar to the pre-equalization method, the solution to the $M \times 1$ filter $\mathbf{g}^{\mathrm{T}}$ can be found to be the multiplication of $\mathbf{d}^{\mathrm{T}}$ and the pseudo-inverse of $\mathbf{V}$.

IV. SIMULATIONS

The performance of the proposed two zero-forcing methods has been assessed by simulation. Following [6], we choose the Gaussian monocycle whose time domain representation is

$$w(t) = A\left[1-\left(\frac{t}{\sigma}-3.5\right)^2\right]\exp\left[-0.5\left(\frac{t}{\sigma}-3.5\right)^2\right]$$

where $0 < t < 1ns$, $\sigma = 1/7\,ns$. Sampling it at chip rate, we obtain $w(n)$ and then the DS-UWB waveform (2) is simulated with the chip rate $N_c = 121$. A total of 5 users are considered, each experiencing a channel of $L_k = 41$ taps to simulate indoor dense multipath environment. Path gains vary from user to user and are modeled as independent random variables with a Rayleigh distribution [7]. The narrowband interference and thermal noise are simulated as AWGN. The signal to noise ratio (SNR) ranges from –20 dB to 5 dB.

We use 200 Monte Carlo runs, during each of which 100 bits are transmitted for each user, to evaluate the average BER. The final results are shown in Fig. 3.

From Fig. 3, we can see that the pre-equalization scheme and the MUI pre-mitigating scheme both outperform the prerake scheme. The improved performance of the pre-equalization scheme is due to the mitigating of most of the multipath effect (more than the prerake scheme). The MUI pre-mitigating scheme has the best performance because it removes not only the multipath effect, but also the MUI that the prerake and the pre-equalization schemes can not mitigate.

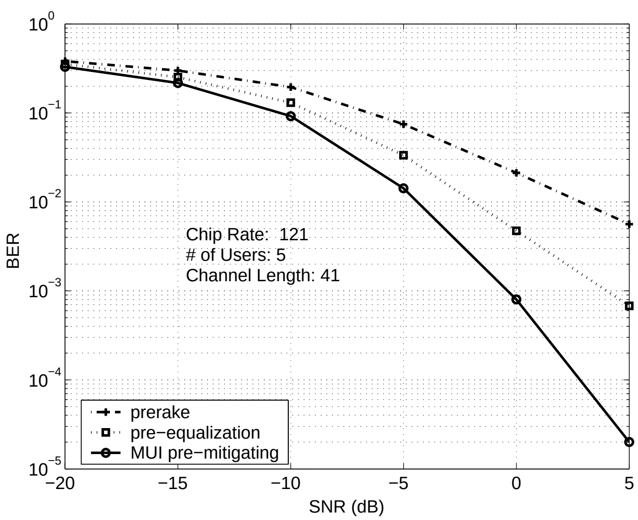

Figure.3. BER of the prerake scheme, the pre-equalization scheme and the MUI pre-mitigating scheme.

VI. CONCLUSION

In this paper, we consider to relieve the hardware complexity and power consumption of the receivers of those users without sufficient power and computation resources in the UWB systems. We move the functionalities of equalization and multiuser interference mitigation from the receivers to the transmitter under the assumption that the transmitter has the knowledge of all channels. The receivers can thus be as simple as conventional matched filters. We propose two effective zero-forcing schemes whose superior performance is studied by simulations.

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