Selective Multipath Interference Canceller with Linear Equalization for DS-UWB Systems with Low Spreading Factor

Chunhua Geng, Yukui Pei, and Ning Ge Member, IEEE

Abstract—In high rate DS-UWB systems with low spreading factor, the selective multipath interference canceller with linear equalization (SMPIC-LE) is developed to alleviate severe multipath interferences induced by the poor orthogonality of spreading codes. The SMPIC iteratively mitigates the strongest inter-path interference, inter-chip interference and inter-symbol interference, while the former two are unresolvable in conventional RAKE-decision feedback equalizer (DFE) receivers. The numerical results and complexity analysis demonstrate that SMPIC-LE with proper parameters provides an attractive overall advantage in performance and computational complexity compared with RAKE-DFE. In addition, it approaches the matched filter bound well as the RAKE finger in SMPIC increases.

Index Terms—Equalization, iterative receiver, multipath interference, RAKE, ultra-wideband

I. INTRODUCTION

Ultra-wideband (UWB) is a promising technology for wireless high rate and short range communications [1]. Direct-sequence spreading based UWB (DS-UWB) and multiband orthogonal frequency-division multiplexing UWB (MB-OFDM UWB) are two main physical layer standards for high data rate wireless personal area networks (WPAN) [2], [3]. Because of the fine properties of coherent processing of the occupied bandwidth and the widest contiguous bandwidth, DS-UWB has received considerable attention from both academia and industry [4], [5].

For high data rate DS-UWB systems supporting transmission rate ranging from several Megabit per second to more than one Gigabit per second, most recent research on the receiver design focuses on the RAKE reception with symbol level decision feedback equalizer (DFE) [6]-[8]. In practical high rate DS-UWB systems, limited by state-of-the-art ADC technology, the spreading factor (SF) cannot be large enough to maintain the ideal orthogonality between spreading codes [9]. Therefore, the conventional RAKE-DFE receiver would suffer significant performance loss due to severe inter-path interference (IPI), inter-chip interference (ICI) and inter-symbol interference (ISI) [10]-[13]. The former two kinds of interference can not be mitigated by the RAKE-DFE receiver effectively. Furthermore, in order to combat the severe ISI induced by long channel delay spread, the DFE tap number has to be quite large. The demanding computational complexity of DFE always exceeds that of the RAKE receiver significantly by far and becomes a heavy burden for system design.

To resolve the above problems, the selective multipath interference canceller (SMPIC) with symbol level linear equalization (LE) is proposed in this paper for practical high rate DS-UWB systems with low SF: The SMPIC is capable of mitigating the IPI, ICI and ISI by reconstructing and subtracting the selected strongest multipath interferences from the received signal in an iterative way. Then the symbol level LE is concatenated to alleviate the remaining ISI. In addition, to validate the effectiveness of the SMPIC-LE receiver, we derive the matched filter bound (MFB), which takes into account such practical constrains as the sampling rate and the RAKE diversity order, i.e. the finger number. Simulation results and complexity analysis show that the proposed SMPIC-LE can achieve similar or even better performance with much lower computational complexity compared with the conventional RAKE-DFE receiver in various realistic UWB channels. Moreover, as the RAKE diversity order increases, the performance of SMPIC-LE receivers can approach the derived MFB well.

The remainder of this paper is organized as follows. The DS-UWB system model with the proposed SMPIC-LE receiver is introduced in Section II. In Section III, the computational complexity and performance of the SMPIC-LE receiver are analyzed, and the MFB is also derived. In Section IV, the corroborating simulation results are presented. Section V summarizes the whole paper.

II. SYSTEM MODEL

In this paper, the IEEE802.15.3a UWB indoor channel model is employed [14]. The equivalent complex-valued baseband model with the proposed SMPIC-LE receiver is shown in Fig.1

A. Transmitter

In this paper, we only focus on binary phase-shift keying (BPSK) modulation, which is the mandatory transmission mode for DS-UWB systems. At the transmitter, the random source symbol is spread and modulated with chip pulse $g_T(t)$. For each symbol, the pulse shape is defined as

$$g(t) = \sum_{n=0}^{N-1} c[n]g_T(t - nT_c)$$

(1)
where \( c[n] \) denotes the \( n \)-th chip of the spreading code of length \( N_c \) and \( T_c \) is the chip duration. Assume \( M \) symbols are contained in each frame, and each transmitted frame can be written as

\[
s(t) = \sum_{m=0}^{M-1} b[m]g(t - mT_b)
\]

(2)

where \( b[m] \in \{-1, +1\} \) represents the \( m \)-th symbol of each frame, and \( T_b = NT_c \) is the symbol interval.

**B. UWB channels**

In order to compare standardization proposals for high data rate WPANs, IEEE802.15.3a task group developed a standard channel model for UWB systems [14]. This model is based on the Saleh-Valenzuela model [15] with some modification to account for the properties of realistic UWB channels. In this model, multipath arrivals are grouped into two categories: cluster arrivals and ray arrivals within each cluster. The channel impulse response is defined as:

\[
h(t) = X \sum_{l=0}^{L-1} \sum_{k=0}^{K-1} \alpha_{k,l} \delta(t - T_l - \tau_{k,l})
\]

(3)

where \( X \) represents the log-normal shadowing, \( \alpha_{k,l} \) is the multipath gain coefficient, \( T_l \) is the delay of \( l \)-th cluster and \( \tau_{k,l} \) is the delay of the \( k \)-th multipath component relative to the \( l \)-th cluster arrival time (\( T_l \)). By definition, we have \( \tau_{0,l} = 0 \) for \( l \in \{0, 1, ..., L - 1\} \).

**C. SMPIC-LE Receiver**

In the DS-UWB system, the SMPIC-LE receiver is developed to alleviate the severe multipath interferences induced by the poor orthogonality of spreading codes. In the first stage, the SMPIC is employed to specifically mitigate the strongest multipath interference components, including IPI, ICI and ISI. Then, a conventional symbol level LE with small tap number is concatenated to combat the residual ISI. In the sequel of this section, we mainly focus on the proposed SMPIC scheme.

The structure of SMPIC is presented in Fig[2]. The SMPIC works in an iterative manner. Its purpose is to eliminate the interference induced by multipath delay at each RAKE finger. Similar with selective-RAKE (SRAKE) [16], the SMPIC selects the instantaneously strongest \( J \) multipath components and combines them together at first. Then the interference is estimated and subtracted from the received data at each RAKE finger to get more precise input signals for the RAKE reception in the next iteration. In order to reduce the complexity, the interference is reconstituted by using the hard decision of the SRAKE output. Through this iterative process, the precision of the correlation result in each RAKE finger is improved, so is the output of the receiver.

In DS-UWB systems, the received signal of each frame is given by

\[
r(t) = \sum_{l=0}^{L-1} \sum_{k=0}^{K-1} \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} a_{k,j} b[m] c[n] \delta(t - T_l - \tau_{k,j} - nT_c - mT_b) + z(t)
\]

(4)

where \( z(t) \) is additive white Gaussian noise (AWGN) with mean being zero and power spectral density being \( N_0/2 \) W/Hz.

We assume the receiver can get the perfect channel knowledge. Received data \( r(t) \) is first fed into maximal ratio combining (MRC) SRAKE in SMPIC. After conventional RAKE processing, the output is sent to hard-decision module. The estimation of \( m \)-th bit of each frame at the output of hard-decision module is denoted as \( b^{(0)}[m] \). This estimated sequence is then spread, modulated and processed by a very simple multipath interference regenerator (MIR). In this sub-module, the modulated sequence is multiplied by the amplitude of selected \( J \) paths and delayed by corresponding time. So the reconstituted \( j \)-th path signal can be expressed as

\[
r_j^{(0)}(t) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} a_{k,j} b^{(0)}[m] c[n] \delta(t - T_l - \tau_{k,j} - nT_c - mT_b)
\]

(5)

where \( j \in \{1, 2, ..., J\} \), and \( \alpha_{k,j} \) is the multipath gain coefficient corresponding to the \( j \)-th RAKE finger. \( T_l \) and \( \tau_{k,j} \) denote the delays.

In the next iteration, the input signal to the \( j \)-th finger is represented as

\[
r_j^{(1)}(t) = r(t) - w \sum_{j' \neq j} r_j^{(0)}(t)
\]

(6)

where \( w \in [0, 1] \) is a constant named as interference rejection weight, which allows to reduce the impact of possible errors presented in the estimated multipath interference replicas. Then \( r_j^{(1)}(t) \) is delivered to the MRC SRAKE receiver in the next iteration. The SRAKE output can be used as either the input of MIR for the following iterations, or the output of the SMPIC receiver if the pre-defined iteration time \( p \) is achieved. Finally, the output of the SMPIC is sent to the LE to reduce the remaining ISI.
III. PERFORMANCE AND COMPUTATIONAL COMPLEXITY ANALYSIS

A. Performance Analysis and the Matched Filter Bound

In this subsection, the effect of multipath components (MPCs) on conventional SRAKE receivers and the validity of SMPIC are analyzed first.

The energy of \( g_T(t) \) is defined as \( E_g \).

\[
E_g = \int_{-\infty}^{+\infty} |g_T(t)|^2 dt
\]

The normalized autocorrelation function of \( g_T(t) \) expresses as

\[
R_g(\Delta t) = \frac{1}{E_g} \int_{-\infty}^{+\infty} g(t)g(t+\Delta t)dt
\]

The channel is assumed perfectly known at the receiver. The local template of the \( \tilde{m} \)-th bit in the \( j \)-th finger of SRAKE is given by

\[
v_j(t) = \sum_{n=0}^{N-1} \alpha^{*}_{k,j,n} c[n] g_T(t-T_{ij} - \tau_{kj,\tilde{m}} - \tilde{m}T_b - \tilde{n}T_c)
\]

where \( (.)^* \) denotes complex conjugation. The correlation output of the \( j \)-th finger is

\[
R_j(t) = \int_{\tilde{m}T_b}^{(\tilde{m}+1)T_b} r(t)v_j(t)dt = b[\tilde{m}] (S + I_1 + I_2) + I_3 + Z
\]

where \( Z \) represents the effect of noise, and

\[
S = NE_g |\alpha_{k,j,\tilde{m}}|^2
\]

is the signal component. \( I_1, I_2 \) and \( I_3 \) are the IPI, ICI and ISI respectively.

\[
I_1 = NE_g \sum_{l=0}^{L-1} \sum_{k=0}^{K-1} \alpha_{k,j,\tilde{m}} \alpha_{k,j,\tilde{n}} R_g(T_l - T_{ij} + \tau_{kj,\tilde{m}} - \tau_{kj,\tilde{n}})
\]

where \( l \neq l_j \) or \( k \neq k_j \),

\[
I_2 = E_g \sum_{l=0}^{L-1} \sum_{k=0}^{K-1} \sum_{n=0}^{N-1} \sum_{\tilde{n}=0}^{N-1} \alpha_{k,j,n} \alpha_{k,j,\tilde{n}} c[n] c[\tilde{n}]
\]

\[
R_g(T_l - T_{ij} + \tau_{kj,\tilde{m}} - \tau_{kj,\tilde{n}} + (n-\tilde{n})T_c)
\]

where \( n \neq \tilde{n} \),

\[
I_3 = E_g \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \sum_{k=0}^{K-1} \sum_{n=0}^{N-1} \sum_{\tilde{n}=0}^{N-1} \alpha_{k,j,n} \alpha_{k,j,\tilde{n}} b[m] c[n] c[\tilde{n}]
\]

\[
R_g(T_l - T_{ij} + \tau_{kj,\tilde{m}} - \tau_{kj,\tilde{n}} + (m-\tilde{m})T_c)
\]

where \( m \neq \tilde{m} \).

The accuracy of the RAKE output is closely related to the statistical properties of IPI, ICI, and ISI, which follow an impulsive distribution \( Q(\sqrt{2}\gamma) \). The conventional symbol level equalizer can only combat long ISI at the cost of high computational complexity, but fails to mitigate IPI and ICI effectively. When the SF is small, which means that the autocorrelation property of the spreading code is poor, the multipath interferences degrade the performance of the RAKE-DFE receiver dramatically. From (5) and (6), we can see that the proposed SMPIC can subtract the J-1 strongest interference components in every finger before the correlation and combining at each iteration, hence the strongest interferences, including \( I_1, I_2, \) and \( I_3, \) in \( Q(\sqrt{2}\gamma) \) can be mitigated effectively.

To validate the effectiveness of the SMPIC-LE receiver, in the following simulation part, the performance of SMPIC-LE is compared with the MFB of DS-UWB systems, which yields the absolute performance limit for equalization schemes. In order to obtain expressions for the bit error rate (BER) of the MFB, we define the signal-to-noise ratio (SNR) as

\[
\gamma_r = \frac{E_b(r)}{N_0} \tag{15}
\]

where \( E_b(r) \) is the received energy per bit for the \( r \)-th UWB channel realization. The corresponding BER(\( \gamma_r \)) for BPSK in one particular channel realization can be written as

\[
BER(\gamma_r) = Q(\sqrt{2}\gamma_r) \tag{16}
\]

where \( Q(\cdot) \) stands for \( Q \) function. The average BER is obtained semi-analytically by averaging over \( R \) channel realizations

\[
BER_{MFB} = \frac{1}{R} \sum_{r=1}^{R} BER(\gamma_r) \tag{17}
\]

As for the DS-UWB systems employing \( J \)-finger RAKE receiver, where \( J \) is much smaller than the total number of resolvable multipath components for complexity reasons, we obtain

\[
E_b(r) = \sum_{j=1}^{J} |h(t_j)|^2 \tag{18}
\]

where \( t_j (j \in 1, 2, ..., J) \) denotes the positions of the strongest \( J \) multipath components in one channel realization. The resolution of \( t_j \) equals to the sampling rate at the receiver. In this paper, the derived MFB takes into account the sampling rate at
the receiver and the effect of selective RAKE combining with a limited number of RAKE fingers. Therefore, it demonstrates an accurate performance bound for practical receivers in DS-UWB systems.

B. Computational Complexity Analysis

In this paper, the computational complexity of SMPIC-LE and SRAKE-DFE receiver is calculated in terms of multiplications and divisions per output symbol (MADPOS) [17].

The proposed SMPIC is comprised of two kinds of basic sub-modules: one is the MRC SRAKE, and the other is the MIR. When the SF is small, correlators, the main part of SRAKE, are quite simple. The MIR can be seen as the inverse of RAKE processing from Section II. Therefore, the computational complexity of SRAKE and SMPIC is given by

$$C_{SRAKE} = 2 \times J$$
$$C_{SMPIC} = 2 \times (p + 1) \times J + p \times 3J$$

where $J$ stands for the RAKE finger number and $p$ denotes the iteration time in SMPIC.

For equalization, the widely used adaptive Kalman recursive least-square (K-RLS) algorithm is employed for adjusting tap coefficients to ensure fast convergence and lower steady-state mean square error (MSE), and hence a favorable detection performance in UWB systems [18]. The adaptive equalizer works in two stages: in training stage, a training sequence is employed to initially adjust the tap weights; in decision directed stage, the decisions at the output of the equalizer are used to continue the coefficients adaption process. The computational complexity of equalizers based on K-RLS is approximately given by [17]

$$C_{K-RLS} = 2.5 \times N^2 + 4.5 \times N$$

where $N$ represents the total tap number in equalizers.

IV. NUMERICAL RESULTS AND DISCUSSION

A. System Parameters

Monte Carlo simulations have been run to access the performance of the proposed receiver in high rate DS-UWB systems with low SF. The spreading code is set as [-1 +1] with SF being an extreme of 2. The sampling rate is $T_s/4$. At the receiver, the value of interference rejection weight $w$ in SMPIC is chosen as 0.9 by investigating the effect of different weights on the system BER performance. The iteration time $p$ is set as 2, which can guarantee the performance convergence in most cases through our simulations. The forgetting factor in K-RLS algorithm is 0.99999. As for equalization, without notable instructions, the lengths of LE tap $L$, DFE feedforward tap $FF$ and feedback tap $FB$ are set as 15, 25 and 20, respectively. The IEEE 802.15.3a CM1 line-of-sight (LOS) and CM4 extreme non-LOS (NLOS) UWB indoor channel models are considered here. According to the recommended instructions in [14], the numerical results are averaged over the best 90 out of 100 channel realizations.

B. Bit Error Rate Performance

As a function of $E_b/N_0$ at the input of receivers, the BER performance of SRAKE-DFE and SMPIC-LE is evaluated and compared with MFB.

First, we present BER curves of SMPIC-LE and SRAKE-DFE receivers with different RAKE finger numbers and transmission data rates. Fig.3 shows the system performance in the CM1 channel model. As seen in this figure, the SMPIC-LE outperforms the conventional SRAKE-DFE receiver. When the transmission data rate equals to 250Mbps, the J=32 SMPIC-LE gets a performance gain about 0.2dB over J = 32 SRAKE-DFE at a BER of $10^{-4}$, and the loss in power efficiency...
compared with the derived MFB is within 1dB. As the data rate increases to 1.5Gbps, the advantage of SMPIC-LE over SRAKE-DFE gets more significant. It is shown when J equals to 32, the performance gain is up to about 1dB, and the performance of SMPIC-LE approaches the MFB well. From Fig.4, it is observed that in the case of CM4 channels, when data rate is 250Mbps, the proposed SMPIC-LE receiver only suffers negligible performance loss compare with SRAKE-DFE with the same RAKE fingers. As the data rate increases to 1.5Gbps, the SMPIC-LE receiver can lower the error floor. From the above two figures, we can conclude that as the data rate increases, the performance gain of SMPIC-LE over SRAKE-DFE improves. This can be attributed to the fact that with the data rate increasing, more resolvable strong interferences, which degrade the system performance dramatically, occur at the receiver, and the proposed SMPIC-LE receiver can alleviate these interferences in a more effective iterative way compared with the conventional SRAKE-DFE receiver.

Then the effect of the LE tap number on the BER performance of the SMPIC-LE receiver is also investigated. From Fig.5, it shows that as LE tap number L increases, the system performance improves as well. In addition, as the RAKE finger J increases, the SMPIC-LE receiver yields a close-to-optimum performance and the performance gain by increasing LE taps become unobvious. For instance, when the finger number J is equal to 16, with L increasing from 15 to 25, the SMPIC-LE receiver can obtain a performance gain of more than 1dB. When J increases to 32, the performance improvement is only about 0.4dB. This is due to the fact that as the RAKE finger number gets larger, the strong ISIs are mitigated by SMPIC effectively, hence increasing LE taps cannot get additional significant performance gain. This fact provides a useful pointer for system designers when specifying system parameters. Our findings also suggest that the SMPIC-LE receiver with more RAKE fingers outperforms the receiver with less RAKE fingers but more equalizer taps, which demonstrates the key role of the proposed SMPIC scheme to mitigate severe multipath interferences in UWB channels.

C. Computational Complexity Comparison

Finally, the computational complexity of SMPIC-LE and SRAKE-SFE receivers adopted in the simulations are compared. The MADPOS of both SMPIC-LE and SRAKE-DFE is shown in Table I. As seen in this table, when J equals to 16, the MADPOS in the SMPIC-LE receiver with L = 15 is 822, which is only 15.5% of that in the SRAKE-DFE with FF = 25 and FB = 20. When J increases to 32, the SMPIC-LE can still save 81% MADPOS than SRAKE-DFE. These results demonstrate that the computational complexity of SMPIC-LE scheme is much less than that of conventional SRAKE-DFE receivers.

V. CONCLUSIONS

The scheme presented in this paper offers a low computational complexity alternative to the conventional SRAKE-DFE receiver, which provides a more efficient way for UWB signal detection by mitigating significant multipath interference components specifically. In this proposed SMPIC-LE scheme, the receiver can alleviate the strongest IPL, ICI and ISI, while the former two interferences are unresolvable in conventional RAKE-DFE receivers. The MFB, which takes into account the effects of sampling rate and the number of RAKE fingers at the receiver, is also derived. Numerical results and complexity analysis show that compared with SRAKE-DFE, the SMPIC-LE receiver, with much lower computational complexity, can achieve similar or even better performance in high rate DS-UWB systems with low SF for various UWB propagation scenarios. In addition, as the RAKE finger number increases, the low-complexity SMPIC-LE receiver approaches the derived MFB limit well.

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