Sliding Window and Successive Cancellation Channel Estimation Schemes based on Pilot Spread Code in DS-UWB System

Yupeng Wang*, JungJu Kim* *Associate member*, KyungHi Chang* *Lifelong member*

ABSTRACT

In this paper, the performances of a single-user DS-UWB system applying two simple proposed channel estimation schemes are introduced, according to the newly updated DS-UWB PHY Layer standard from IEEE P802.15.3a. The performances of error control coding, different combining schemes in selective Rake receiver for DS-UWB system are analyzed. Both of the two channel estimation schemes using data-independent structure work well in DS-UWB system with few pilot bits. For the purpose of channel estimation and reduces the number of pilot bits, we apply a pilot symbol spreaded with 2^8-1 or 2^9-1 periods of m-sequence for different channel estimation schemes.

Key Words : DS-UWB, Error-Control Coding, Selective Rake Receiver, Sliding Window, Successive Cancellation

Ⅰ. Introduction

The Ultra Wide Band(UWB) technology is one available candidate for future short-range indoor radio communication systems providing very high bit rate services, lower power consumption and accurate positioning capability.

Conventional UWB radio systems employ the transmission of very short impulses of radio energy whose characteristic spectrum signature extends across a wide range of radio frequencies. To improve its Multiple-Access(MA) capability, UWB impulse radio technology can be combined with traditional Spread-Spectrum(SS) techniques. Direct-Sequence (DS) SS technique is a well known and powerful MA technology. Antipodal signaling, such as Binary Phase-Shift Keying(BPSK), is employed in DS-UWB system[1]. The newly updated DS-UWB PHY standard can provide a wireless PAN services with payload capabilities of 28, 55, 110, 220, 500, 660, 1000 and 1320 Mbps.

In conventional DS-CDMA system, there is pilot channel to provide time reference, and to enable coherent detection, power estimation and power control. The unmodulated pilot channel spread spectrum signal is transmitted at all times by the base station with higher power level than the other channels. While in UWB system, there is no dedicated channel for pilot, and all the control information and data are transmitted in the same channel. So we should consider the energy loss due to the pilot symbols and efficient channel estimation schemes.

In this paper, the DS-UWB data link performance with selective Rake receiver and error control coding is presented. Based on the autocorrelation function of the pilot spread code, we use sliding window and successive cancellation schemes to estimate the channel impulse response with corresponding delay. Besides, we compare the performance of two different channel estimation schemes with two different

* 인하대학교 정보통신대학원 이동통신연구실 (khchang@inha.ac.kr)
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m-sequence as a pilot spread code.

The organization of this paper is as follows. In Section II, the DS-UWB system model using anti-podal signaling and the realistic UWB channel are introduced. The theory and structure of the selective Rake receiver are given in section III. We will describe the proposed channel estimation schemes in Section IV. Section V presents the structures of transmitter and receiver, simulation assumptions, parameter settings and numerical simulation results with corresponding explanations. Finally, conclusions are made in section VI.

II. DS-UW B System Model

2.1 Definition of DS-UWB Signals

We consider a single-user DS-UWB system. Assuming that the user has a pseudorandom sequence with N_c chips per message symbol period T_s such that $N_cT_c=T_s$. The signal transmitted can be expressed as

$$s(t)=\sum_i b(i)p(t-iT_s) \qquad (1)$$

where $\{b(i)\}$ is the BPSK modulated data stream which takes values of {+1, -1}, and $p(t)$ is the spread waveform, which is

$$p(t)=\sum_{n=0}^{N_c-1} c(n)g(t-nT_c) \qquad (2)$$

where $c(n)$ is the signature sequence, and $g(t)$ is the chip pulse[2].

2.2 IEEE 802.15.SG3a UWB Channel Model

The IEEE 802.15.SG3a channel model for UWB transmission was derived from the Saleh-Valenzuela model, where multipath components arrive in clusters [3]. In addition, independent fading is assumed for each cluster as well as for each ray within a cluster.

Therefore, the multipath model consists of the following discrete time impulse response

$$h_i(t)=X_i\sum_{l=0}^{L}\sum_{k=0}^{K}\alpha^i_{k,l}\delta(t-T^i_l-\tau^i_{k,l}) \qquad (3)$$

Table. 1. IEEE 802.15.SG3a UWB channel characteristics and corresponding model parameters.

Channel Characteristic	CM1 LOS (0-4 m)	CM2 NLOS (0-4 m)	CM3 NLOS (4-10 m)	CM4 NLOS
Mean excess Delay(nsec)(τ_m)	5.05	10.38	14.18	
RMS delay (nsec)	5.28	8.03	14.28	25
NP_{10dB}			35	
NP(85%)	24	36.1	61.54	
Model Parameter				
Λ(1/nsec)	0.0233	0.4	0.0667	0.0667
λ(1/nsec)	2.5	0.5	2.1	2.1
γ	4.3	6.7	7.9	12
Γ	7.1	5.5	14.00	24.00
σ_1(dB)	3.3941	3.3941	3.3941	3.3941
σ_2(dB)	3.3941	3.3941	3.3941	3.3941
σ_x(dB)	3	3	3	3

where $\{\alpha^i_{k,l}\}$ are the multipath gain coefficients, $\{T^i_l\}$ is the delay of the l^{th} cluster, $\{\tau^i_{k,l}\}$ is the delay of the k^{th} multipath component relative to the l^{th} cluster arrival time $\{T^i_l\}$, $\{X_i\}$ represents the log-normal shadowing, and i refers to the i^{th} realization. By definition, we have $\tau_{0,l}=0$.

The amplitude $|\alpha_{k,l}|$ has a log-normal distribution and the phase $\angle\alpha_{k,l}$ is uniformly distributed in $\{0, \pi\}$. The expected value $E(|\alpha_{k,l}|^2)$ is proportional to $\exp(-T_1/\Gamma-\tau_{k,l}/\gamma)$, where Γ and γ denote a cluster- and a ray-decay factor, respectively. The interval time between two clusters $T_{l+1}-T_l$ or two rays within one cluster $\tau_{k+1,l}-\tau_{k,l}$ is exponentially distributed. Log-normal shadowing is modeled with $X=10^{n/20}$ where n has a normal distribution with mean $\mu=0$ dB and standard deviation $\sigma=3$ dB. There are four different models, CM 1, CM 2, CM 3 and CM 4 for different channel characteristics, shown in Table 1. NP_{10dB} is the number of paths within 10 dB of the strongest path and $NP_{85\%}$ gives the number of paths containing 85 percents of the energy. The above channel model is assumed quasi-static or flat during the data transmission of one packet. The impulse response realizations of CM 3 is presented in Fig. 1, and the

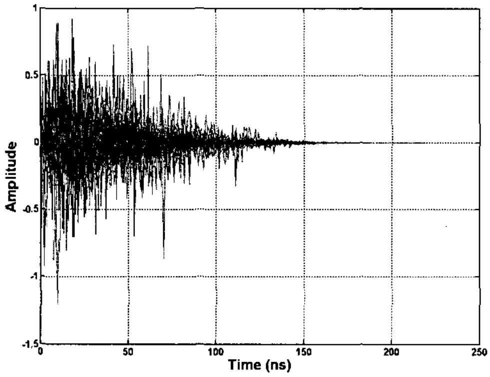

Fig. 1. Impulse response realization (CM 3).

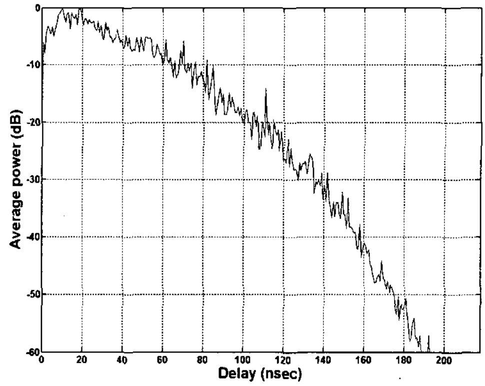

Fig. 2. Average power decay profile (CM 3).

average power decay profile of CM 3 is shown in Fig. 2[4].

Ⅲ. Coherent Selective Rake Receiver

A receiver structure that is usually used for CDMA systems is Rake receiver, depicted in Fig. 3. In UWB channel model, there is abundant multipath components and the several stronger components usually don't come first. If we use the fixed spaced Rake receiver such as chip-spaced or symbol spaced, it is possible to miss such stronger impulse reponses which contain stronger signal energy than the other channel impulse responses. If we locate the finger at the moments when the strong channel impulse response occurs, we will get better signal-to-noise (SNR) than conventional Rake receiver. So we call it selective Rake Receiver and fingers are located at the N largest impulse response moments, N is the finger number.

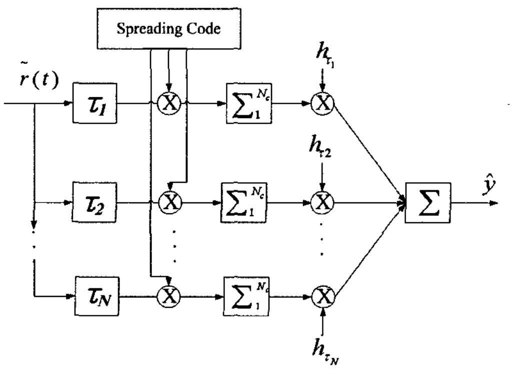

Fig. 3. Rake receiver structure.

We assume that the channel impulse response is time invariant. The output of the i^{th} finger can be written as

$$y_i = \int_{-iT_c}^{T_f - iT_c} \tilde{r}(t + iT_c) p^*(t) dt \tag{4}$$

where $\tilde{r}(t + iT_c)$ is the received complex baseband signal.

Let's represent the baseband signal as

$$\tilde{r}(t) = h\tilde{s}(t) + \tilde{w}(t) \tag{5}$$

where $\{h\}$ is the propagation loss due to multipath channel and $\tilde{s}(t)$ represents the spreaded data, $\tilde{w}(t)$ is the complex envelope representation of white Gaussian noise with zero mean and variance of one. We can write $\tilde{s}(t)$ as below

$$\tilde{s}(t) = b(t)\sqrt{E_b}p(t) \tag{6}$$

In equation (6), E_b is the bit energy. We can use equation (5) and (6) to rewrite the output of i^{th} finger as

$$y_i = h_i b\sqrt{E_b} + b\sqrt{E_b}\sum_{l \neq i} h_l R_{gg}(l - i) + \eta_i \tag{7}$$

where $R_{gg}(t)$ is the autocorrelation of the spread sequence and η_i is the additional white Gaussian noise sample for the i^{th} finger[5].

In maximal-ratio-combining(MRC) RAKE receiver, the signals of the selected fingers are despreaded, coherently added, and then weighted by their respective amplitudes, before a bit decision is made. The output of the RAKE receiver is given as below

$$\begin{aligned} y &= \sum_{l=1}^{N} h^*_l y_l \\ &\approx \sqrt{E_b}\sum_{l=1}^{N} |h_l|^2 b + \sum_{l=1}^{N} h^*_l \eta_l \end{aligned} \tag{8}$$

where $\{h^*\}$ is the conjugate of the channel response, and y_i is the output of the i^{th} finger. The resulting output SNR is described in the following equation (9).

$$\begin{aligned} SNR &= \frac{(E[y]^2)}{\sigma^2_Y} \\ &= \frac{E_b \sum_{l=1}^{N} |h_l|^2}{N_0} \end{aligned} \tag{9}$$

The MRC RAKE receiver produces an instantaneous output SNR that is the sum of the instantaneous SNR of individual fingers.

In the equal-gain-combining(EGC) RAKE receiver, we don't consider the magnitude of each finger and only compensate the phase distortion for individual fingers due to the multipath fading channel. The output of the EGC RAKE receiver is given in equation (10).

$$\begin{aligned} y &= \sum_{l=1}^{N} e^{-j\theta_l} y_l \\ &\approx \sqrt{E_b}\sum_{l=1}^{N} h_l b + \sum_{l=1}^{N} e^{-j\theta_l}\eta_l \end{aligned} \tag{10}$$

For EGC RAKE receiver, we just set the magnitude of each finger equal to 1 and compensate the phase distortion only.

Ⅳ. Proposed Channel Estimation Schemes

In the newly updated IEEE P802.15.3a DS-UWB proposal, there is no reference about the channel estimation scheme and the proposed system is a short-code spread-spectrum system whose spread code period is shorter than the channel time dispersion. If we use data-aided(DA) approach, due to the autocorrelation function of the spread code, the multipath components with time delay larger than the period of data spread code will be added to the multipath components with time delay smaller than the period of data spread code in corresponding position. For example, if the period of data spread code is N_c, the multipath component with time delay N_c+1 chips will be added to the multipath components with time delay 1 chip and we can't get the multipath component with time delay N_c+1 chips. So we use a data-independent pilot approach that is based on the known pilot bits to estimate the channel impulse response[6]. The known pilot bits are transmitted before the user data. Here pilot bits are spreaded using a pilot spread code. The transmitted symbol corresponding to the data-independent pilot scheme can be represented as follows

$$s(t) = \begin{cases} \sum_{i=0}^{N_p-1} b_p(t) p_s(t-iT_{sp}) & \textit{before data} \\ \sum_j b(t)p(t-jT_s) & \textit{data symbol period} \end{cases} \tag{11}$$

and

$$p_s(t) = \sum_{n=0}^{N_{pc}-1} c_p(n) g(t-nT_c) \tag{12}$$

where p_s is pilot spread waveform with the same chip duration and chip pulse as that of data spread code, $\{b_p(t)\}$ is the known pilot bits whose length is N_p. The period of the pilot spread code is N_{pc} and the duration of the pilot symbol is T_{sp} such that $N_{pc}T_c=T_{sp}$. Assuming that a packet starts with N_p known pilot bits, the received signal can be represented as

$$r(t) = \sum_{l=1}^{L} \alpha_l s(t-\tau_l) + \omega(t) \tag{13}$$

where α_l and τ_l are the channel gain and delay of the l^{th} path, $\omega(t)$ is a Gaussian noise, L is the number of multipath components.

From (13) we have

$$\mathbf{r} = \mathbf{s}(\boldsymbol{\tau})\boldsymbol{\alpha} + \boldsymbol{\eta} \tag{14}$$

where we take η as a random vector with zero mean and covariance matrix $\mathbf{C}_\eta = E\{\boldsymbol{\eta}\boldsymbol{\eta}^H\}$, and $\boldsymbol{\alpha} = [\alpha_1,$

$\alpha_2,...,\alpha_L]^T$, for pilot symbol duration, $\mathbf{s}(\boldsymbol{\tau})$ is a matrix with entries

$$[\mathbf{s}(\boldsymbol{\tau})]_{m,l} = s(mT_{sp} - \tau_l) \qquad 1 \le l \le L \tag{15}$$

We note that, for a given pair $(\boldsymbol{\alpha}, \boldsymbol{\tau})$, vector $\mathbf{r}$ is a Gaussian, with mean $\mathbf{s}(\boldsymbol{\tau})\boldsymbol{\alpha}$ and covariance matrix $\mathbf{C}_\eta$. The covariance matrix $\mathbf{C}_\eta$ depends on the interfering users. For single user case, it is an identity matrix. Thus the likelihood function of the unknown parameters takes form

$$\Lambda(\tilde{\boldsymbol{\alpha}}, \tilde{\boldsymbol{\tau}}) = 2\,\mathrm{Re}\{\mathbf{r}^H \mathbf{C}_\eta^{-1} \mathbf{s}(\tilde{\boldsymbol{\tau}})\tilde{\boldsymbol{\alpha}}\} - \tilde{\boldsymbol{\alpha}}^H \mathbf{s}^H(\tilde{\boldsymbol{\tau}}) \mathbf{C}_\eta^{-1} \mathbf{s}(\tilde{\boldsymbol{\tau}})\tilde{\boldsymbol{\alpha}} \tag{16}$$

where $\tilde{\boldsymbol{\alpha}}$ and $\tilde{\boldsymbol{\tau}}$ are trial values of $\boldsymbol{\alpha}$ and $\boldsymbol{\tau}$.

Even though the search for the optimum can be performed separately for $\boldsymbol{\tau}$ and $\boldsymbol{\alpha}$, the complexity of the search for delays is prohibitive, since it involves an *L*-dimensional maximization.

In our system, to reduce the complexity we use two kinds of suboptimal channel estimation algorithms called sliding window algorithm(SW) and successive cancellation(SC) algorithm.

The sliding window algorithm calculates the cross-correlation between the received pilot symbol sequence and the known transmitted pilot symbol sequence and to get gains $\{\widehat{\alpha}_l\}$ and delays $\{\widehat{\tau}_l\}$ for the *N* largest amplitudes of the cross-correlated sequence. If the shifted versions of the received signal are mutually orthogonal or the channel is one-tap, the sliding window algorithm is maximum likelihood optimum. But neither of these two cases are met in our system, which results in sub-optimality of the sliding window approach. The sliding window channel estimation scheme is performed as in equation (17) and (18).

$$\hat{\boldsymbol{\tau}} = \arg\max_{\tilde{\tau}} \left\{ \frac{\left|\mathbf{s}^H(\tilde{\boldsymbol{\tau}})\mathbf{C}_\eta^{-1}\mathbf{r}\right|}{\mathbf{s}^H(\tilde{\boldsymbol{\tau}})\mathbf{C}_\eta^{-1}\mathbf{s}(\tilde{\boldsymbol{\tau}})} \right\} \tag{17}$$

and

$$\hat{\boldsymbol{\alpha}} = \frac{\mathbf{s}^H(\hat{\boldsymbol{\tau}})\mathbf{C}_\eta^{-1}\mathbf{r}}{\mathbf{s}^H(\hat{\boldsymbol{\tau}})\mathbf{C}_\eta^{-1}\mathbf{s}(\hat{\boldsymbol{\tau}})} \tag{18}$$

where $\hat{\boldsymbol{\tau}}$ and $\hat{\boldsymbol{\alpha}}$ are estimated channel delay and impulse response, respectively.

The successive cancellation channel estimation scheme improves the performance of sliding window algorithm by an iterative manner. The algorithm begins by using the sliding window algorithm to find the parameters of the strongest tap. The delayed version of the transmitted signal corresponding to the estimated tap is then subtracted from the received sequence and the algorithm is repeated. The main drawback of SC channel estimation algorithm compared with SW algorithm lies in the longer delay than SW channel estimation method before the channel is fully estimated[7]. In DS-UWB system, the pilot symbols and data transmit sequently, besides the pilot symbols take only a little portion of the packet, so the time delay induced by SC has little effect on the system throughout. Based on equation (17) and (18), the algorithm can be performed in the following steps

1. Set $l = 1$ and $\mathbf{r}^{(l)} = \mathbf{r}$.
2. Performing search for the strongest tap τ_l and α_l using the above equation (17) and (18).
3. Subtract the contribution of the estimated path from the received signal as in equation (19).

$$\mathbf{r}^{(l+1)} = \mathbf{r}^{(l)} - \hat{\alpha}_l \mathbf{s}(\hat{\tau}_l) \tag{19}$$

4. If l is lower than the number of the RAKE fingers repeat from step 2 with matrix $\mathbf{r}^{(l+1)}$ instead of $\mathbf{r}$ in (17) and (18), l steps one.

Due to the good autocorrelation property and simple generation method, we choose m-sequence as a pilot spread code, with period longer than the channel maximum time dispersion. The performance of this method will be presented in Section V.

V. Simulation Conditions and Numerical Results

5.1 Simulation Assumptions and Parameter Sets

The proposed transmitter and receiver structures for

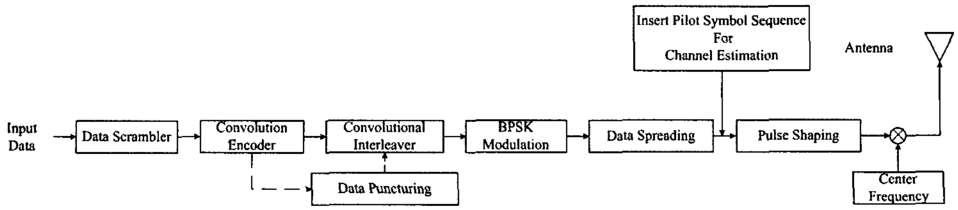

Fig. 4. Structure of DS-UWB transmitter

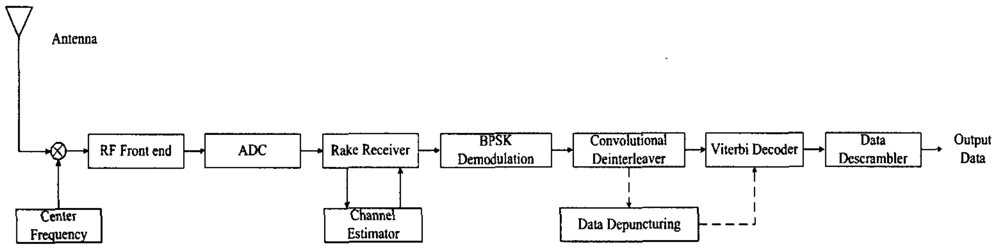

Fig. 5. Structure of DS-UWB receiver

DS-UWB system are shown in Fig. 4. and Fig. 5.

In our simulations, we consider a single user and perfectly synchronized DS-UWB system. No multi-user interference is considered in this paper.

All the simulations in this paper were conducted using the parameter set of 55 Mbps data rate in the lower operating band, according to the newly updated IEEE P802.15. 3a DS-UWB proposal[8].

The explicit setting of parameters is described below

- The uncoded data rate R_b is 55 Mbps.
- We use convolutional coding as error control coding, whose code rates are 1/2 and 3/4, constraint length is 7.
- The transmitted pulses are modulated using BPSK.
- The processing gain is 10.79 dB or 12 chips/symbol. We simulate a short code spread-spectrum system.
- The distortions from the antenna and nonlinear hardware are neglected.
- The m-sequences used for channel estimation are generated by linear feedback shift registers.
- The length 255 m-sequence generation polynomial is

$$c_p(x) = 1 + x^2 + x^3 + x^4 + x^8 \tag{20}$$

The length 511 m-sequence generation polynomial is

$$c_p(x) = 1 + x^4 + x^9 \tag{21}$$

- IEEE P802.15.SG3a UWB channel model is adopted. Here, the number of multipath components is equal to NP_{10dB}.

All of the UWB channel models were simulated using at least 100 channel realizations for each E_b/N_0.

5.2 Numerical Simulation Results

In this section, we present the numerical simulation results in DS-UWB system. In Fig. 7, Fig. 8 and Fig. 10, we use 1/2 rate convolutional code as error control coding and compare the performance difference at target BER of 10^{-4}. In either channel estimation scheme, the number of pilot bits is 4 and 8 for the cases of length 511 and 255 m-sequence respectively. So, we get similar total number of chips for channel estimation in the cases of different pilot spread codes.

Fig. 6. presents the uncoded BER results performances of DS-UWB system in different channels. We simulate with perfect knowledge of the IEEE UWB channel characteristics and using 1-finger selective MRC Rake receiver for comparison with AWGN channel and 1-path Rayleigh fading channel. It is clear that data suffers from larger inter-symbol interference(ISI) in IEEE UWB channel.

In Fig. 7. the performances of sliding window

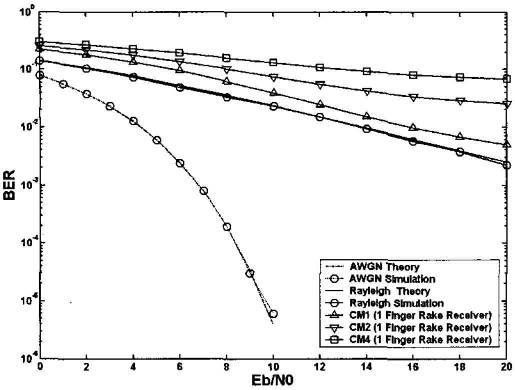

Fig. 6. Performances of uncoded DS-UWB system in different channel models.

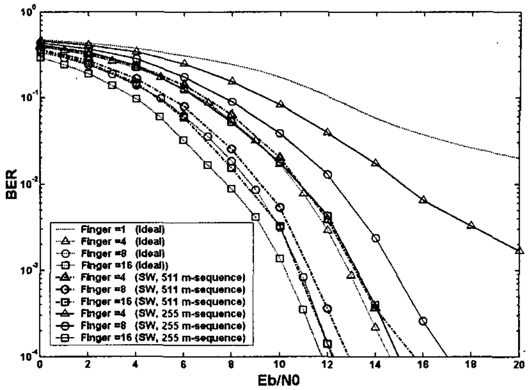

Fig. 7. BER of DS-UWB system with perfect and sliding window channel estimation applying two kinds of pilot spread code under CM 3.

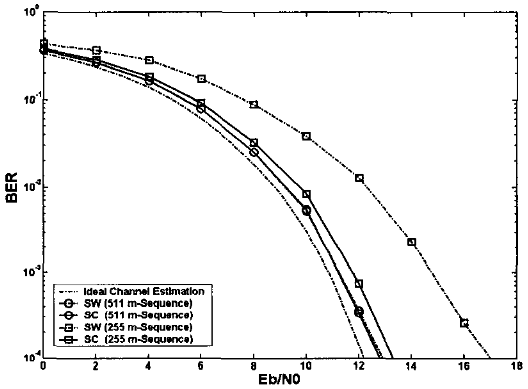

Fig. 8. BER comparison between sliding window and successive cancellation channel estimation schemes under CM 3.

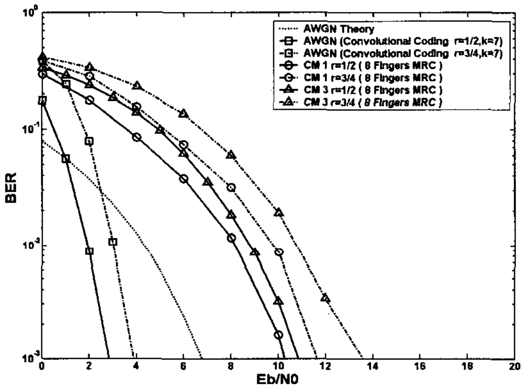

Fig. 9. BER of different convolutional code rates with perfect channel estimation under CM 1 and 3.

channel estimation algorithm with two different length pilot spread codes are presented under channel model 3. We can see that the performance difference between 16 and 8 fingers selective Rake receivers with perfect channel estimation is about 0.5 dB, but the complexity of 16 fingers Rake receiver is twice of the 8 fingers Rake receiver case. So if considering the tradeoff between complexity and performance, we may select 8 fingers selective Rake receiver as our choice. We use two kinds of m-sequence to estimate the channel. We can see from the simulation results that Rake receiver with 8 fingers, length 511 m-sequence provides the performance loss of only 0.6 dB compared with the perfect channel estimation, while in the length 255 m-sequence case the performance loss is more than 4 dB.

Fig. 8. shows the comparison between two different kinds of channel estimation schemes with two different pilot spread codes and 8 fingers MRC Rake receiver under CM 3. We can see that the performances of SW and SC channel estimation schemes with length 511 m-sequence are similar. When length 255 m-sequence is used, the performance of successive cancellation channel estimation scheme is about 0.5 dB worse than that of the sliding window scheme with length 511 m-sequence and about 1 dB worse than that of perfect channel estimation scheme. But the sliding window scheme with length 255 m-sequence induces about 4.8 dB performance loss than the case of perfect channel estimation scheme.

Fig. 9 shows the BER results of DS-UWB system with different convolutional code rates. As the number of multipath components increases, the 3/4 rate convolutional code losses more performance than the 1/2 rate convolutional code.

As shown in Fig. 10, the MRC selective Rake re-

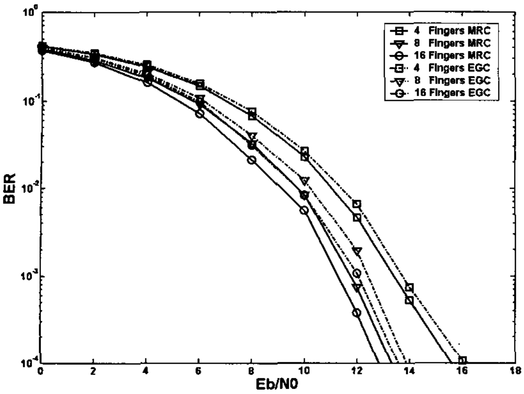

Fig. 10. BER Performances of MRC and EGC RAKE receivers with successive cancellation channel estimation scheme under CM 3.

ceiver performs about 0.52 dB, 0.58 dB and 0.75 dB better than EGC Rake receiver for the case of 4 fingers, 8 fingers, and 16 fingers respectively in coded DS-UWB system using successive cancellation channel estimation scheme with length 255 m-sequence as a pilot spread code. From the above result, we can conclude that as the number of the finger increases, EGC Rake receiver losses more performance than the case of MRC Rake receiver. In MRC case, we need to estimate the magnitude and phase of the channel impulse response, which adds only a little complexity than EGC.

Ⅵ. Conclusions

Even though the data transmitted in the realistic UWB channel environment suffers from very large ISI, we can use error control coding to achieve the target BER with less E_b/N_0 than in the uncoded system. The selective MRC Rake receiver performs well in DS-UWB system with two practical channel estimation algorithms. Successive cancellation channel estimation scheme works better than sliding window channel estimation when short length m-sequence is used as a pilot spread code. From Fig. 8 we see that when we use SC channel estimation scheme with length 255 m-sequence as a pilot spread code, at BER 10^{-4} with 8 fingers selective MRC Rake receiver, the performance is about 0.5 dB worse than the case of sliding window channel estimation scheme with length 511 m-sequence, but the complexity of the former channel estimator is only half of the latter one. According to the newly updated IEEE P802.15.3a DS-UWB PHY specifications, we propose to use selective Rake receiver (MRC) with 8 fingers and successive cancellation channel estimation algorithm with m-sequence as a pilot spread code considering the above simulation results and the discussion on complexity. The period of m-sequence is 2^8-1 in this paper.

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왕 우 붕 (Yupeng Wang) 준회원

2004년 7월 동북대학교 통신공학과(공학사)

2004년 9월~현재 인하대학교 정보통신대학원 석사과정

<관심분야> 4세대 이동통신 시스템, UWB 시스템 무선 전송규격

김 정 주 (JungJu Kim) 준회원

2004년 2월 인하대학교 전자공학과 (공학사)

2004년 3월~현재 인하대학교 정보통신대학원 석사과정

<관심분야> 4세대 이동통신 시스템, Cognitive Radio, UWB 및 WiBro 시스템 무선 전송규격

장 경 희 (KyungHi Chang) 종신회원

1985년 2월 연세대학교 전자공학과(공학사)

1987년 2월 연세대학교 전자공학과(공학석사)

1992년 8월 Texas A & M Univ., EE Dept. (Ph.D.)

1989년~1990년 삼성종합기술원 주임연구원

1992년~2003년 한국전자통신연구원, 이동통신연구소 무선전송방식연구팀장 (책임연구원)

2003년~현재 인하대학교 정보통신대학원 부교수

<관심분야> 3G LTE & 4세대 이동통신 무선전송방식, WMAN 시스템 무선전송기술, Cognitive Radio, Cross-layer Design