Bandwidth-Efficient OFDM Transmission with Iterative Cyclic Prefix Reconstruction

Jong-Bu Lim, Eung-Sun Kim, Cheol-Jin Park, Hui-Chul Won, Ki-Ho Kim, and Gi-Hong Im

Abstract: **For orthogonal frequency division multiplexing (OFDM), cyclic prefix (CP) should be longer than the length of channel impulse response, resulting in a loss of bandwidth efficiency. In this paper, we describe a new technique to restore the cyclicity of the received signal when the CP is not sufficient for OFDM systems. The proposed technique efficiently restores the cyclicity of the current received symbol by adding the weighted next received symbol to the current received symbol. Iterative CP reconstruction (CPR) procedure, based on the residual intersymbol interference cancellation (RISIC) algorithm, is analyzed and compared to the RISIC. In addition, we apply the CPR method to Alamouti space-time block coded (STBC) OFDM system. It is shown that in the STBC OFDM, tail cancellation as well as cyclic reconstruction of the CPR procedure should be repeated. The computational complexities of the RISIC, the proposed CPR, the RISIC with STBC, and the proposed CPR with STBC are analyzed and their performances are evaluated in multipath fading environments.**

We also propose an iterative channel estimation (CE) method for OFDM with insufficient CP. Further, we discuss the CE method for the STBC OFDM system with the CPR. It is shown that the CPR technique with the proposed CE method minimizes the loss of bandwidth efficiency due to the use of CP, without sacrificing the diversity gain of the STBC OFDM system.

Index Terms: **Channel estimation (CE), cyclic prefix (CP) reconstruction, intercarrier interference (ICI), interference cancellation, intersymbol interference (ISI), orthogonal frequency division multiplexing (OFDM), space-time block code (STBC).**

I. INTRODUCTION

High-speed transmission techniques have been developed to provide duplex operation at data rates ranging from 51.84 Mb/s up to gigabits for wired and wireless applications [1]–[7]. In various fora and organizations for wireless research, such as the Wireless World Research Forum (WWRF) and International Telecommunication Union (ITU), there have been active discussions about 4G or beyond-3G systems to be deployed around 2010 [8], [9]. 4G systems will support up to approximately 100 Mb/s for high mobility and up to approximately 1 Gb/s for low mobility such as nomadic/local wireless access. High bandwidth efficiency is one of the key requirements to achieve such high data throughput targets with limited spectrum availability.

Multicarrier modulation, such as discrete multitone (DMT)/ orthogonal frequency-division multiplexing (OFDM), has been proposed for bandwidth-efficient data communications in both wired and wireless environments [1]–[4]. DMT/OFDM divides a broadband signal into multiple narrowband subcarriers, where each subcarrier is more robust to multipath. In order to maintain orthogonality among subcarriers, a cyclic prefix (CP) is added at the head of each symbol and the length of CP is greater than the expected length of channel impulse response (CIR). Since, however, the use of CP results in a lowering of spectral efficiency, several approaches have been proposed to cope with this problem [10]–[19]. DMT receivers apply a finite impulse response time domain equalizer in order to shorten the length of the overall CIR [10]–[12]. As an alternative to time domain equalization, per tone equalization was proposed in [13], where the equalization is carried out in the frequency domain with a multitap frequency domain equalizer for each tone. A precoder is also used to remove the distortion due to insufficient CP [14]. With only a few linear combinations of redundant (or unused) carrier output samples, intersymbol interference (ISI) and intercarrier interference (ICI) can be totally eliminated even in receiver structures without CP [15], [16]. Another approach is to employ an iterative cancellation method, such as the residual ISI cancellation (RISIC) technique, to eliminate the ISI and ICI due to insufficient CP [17]–[19].

The RISIC technique [17] uses a combination of tail cancellation and cyclic reconstruction like the echo cancellation method in [20]. The echo or crosstalk cancellation is a technique to foster high-speed full-duplex communications [20]–[22]. The RISIC technique removes ISI from the received signal in the tail cancellation step, and restores cyclicity so as to avoid ICI in the cyclic reconstruction step. However, the RISIC technique and most other techniques based on RISIC offer good performance only when the length of CIR is moderate so that the interference power is much smaller than the signal power. In order to overcome the above limitation, various methods have been proposed to utilize the soft-input soft-output (SISO) channel decoder, the optimal detection filtering, and the leaked signal energy spread to the next symbol [23]–[24]. However, the SISO channel decoder itself provides only a slight improvement in the OFDM symbol error rate (SER). The optimal detection filtering at the initial step requires a computationally expensive process such as an inversion of the channel transfer function matrix. The ISI combiner, utilizing the leaked signal energy, has a drawback that the estimation of the next transmitted symbol is required.

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In this paper, we discuss an improved CP reconstruction (CPR) method based on the RISIC algorithm. The proposed method, with a simple one-tap frequency domain equalizer and no estimation of the next transmitted symbol, efficiently restores the cyclicity of the current received symbol by adding the weighted next received symbol to the current received symbol. Through simulation and analytical results, we show that the proposed scheme outperforms the conventional RISIC in the SER even when the CIR is longer than a half of the symbol duration. In addition, we apply the CPR algorithm to Alamouti space-time block coded (STBC) OFDM system [25] with insufficient CP. To be applied on the STBC OFDM system, the CPR procedure for each OFDM symbol in the symbol pair is performed consecutively and is combined with the decoding process of the STBC. We also present simulation results for the performance of the RISIC algorithm, the proposed CPR technique, the RISIC with STBC, and the proposed CPR with STBC.

Accurate channel estimation (CE) is necessary since the channel is frequency selective and time-varying in wideband mobile systems. Based on frequency-domain filtering, a CE method for OFDM system has been proposed in [26]. In [27], a minimum mean-square error (MMSE) CE method for OFDM system has been proposed, which exploits the correlation of the channel frequency responses at different times and frequencies. An iterative CE technique that exploits coding information and the extra observation offered by the CP has recently been suggested in [28]. However, these estimation methods [26]–[28] can not be used with the CPR algorithm because they have been developed under the premise of use of sufficient CP. In this paper, we propose an iterative CE method for OFDM system with insufficient CP. The proposed estimation method is based on the expectation-maximization (EM) algorithm [29] and exploits iteratively the extrinsic probabilities of the SISO channel decoder. The rest of the paper is organized as follows. In Section II, the improved CPR procedure using the proposed method is presented and analyzed, and the receiver structure employing the proposed CPR scheme is described. In Section III, we apply the CPR procedure to the STBC OFDM system. Iterative CE method for OFDM with insufficient CP is presented in Section IV. Computational complexity of various CPR algorithms is analyzed in Section V. Simulation results are given and discussed in Section VI. Section VII gives the conclusions.

II. CYCLIC PREFIX RECONSTRUCTION

A. Data Model

In an OFDM system, the data stream is divided into blocks of length N and modulated by using an N-point inverse fast Fourier transform (IFFT). For the ith transmitted symbol sequence $\{X_{i,n}\}_{n=0}^{N-1}$, the time domain sequence at the output of the IFFT is

$$x_{i,k} = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_{i,n} W_N^{-nk}, \ 0 \le k < N \tag{1}$$

where $W_N^{nk} = e^{-j\frac{2\pi nk}{N}}$. A CP of length G is appended in the head of $\{x_{i,k}\}_{k=0}^{N-1}$, and then the OFDM symbol is transmitted

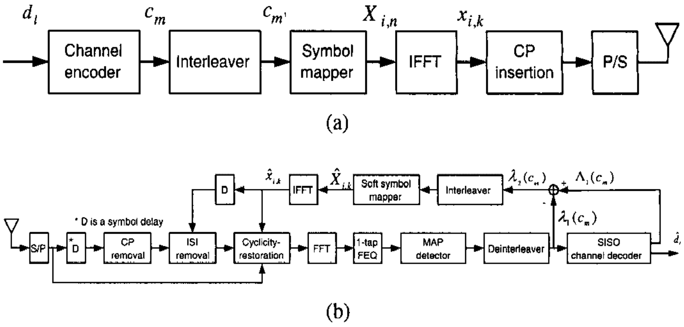

Fig. 1. Block diagram of OFDM system; (a) a coded OFDM transmitter and (b) receiver structure employing the proposed CPR technique.

over a channel, which is modeled as an Lth order FIR filter with additive white Gaussian noise (AWGN). We assume that the channel is known at the receiver by means of an MMSE channel estimator and is static at least during one OFDM symbol. The received sequence for the ith symbol can be represented as

$$r_{i,k} = \begin{cases} \sum_{l=0}^{G+k} h_l x_{i,(k-l)_N} + \sum_{l=G+k+1}^{L} h_l x_{i-1,(k-l+G)_N} \\ +n_{i,k}, \qquad\qquad -G \le k < L-G, \\ \sum_{l=0}^{L} h_l x_{i,(k-l)_N} + n_{i,k}, \quad L-G \le k < N \end{cases} \tag{2}$$

where $(k)_N$ is the residue of k modulo N, h_l is the CIR, and $n_{i,k}$ is AWGN. After removal of the CP, the sequence $\{r_{i,k}\}_{k=0}^{N-1}$ is applied to the next modules including an N-point FFT. As depicted in (2), if $G \geqslant L$, all samples of the sequence $\{r_{i,k}\}_{k=0}^{N-1}$ are free of interference. Otherwise, the first $(L-G)$ samples are contaminated with interference.

In this paper, we consider that the proposed method is applied to a coded OFDM system, employing a convolutional code with interleaving, so as to exploit the benefit of turbo equalization. Fig. 1(a) shows a transmitter structure of the coded OFDM system. At the transmitter, the information bit sequence $\{d_l\}$ is encoded into the coded bit sequence $\{c_m\}$ with a convolutional code. In order to ensure that the codes are terminated at the end of each block, we used a feedforward encoder with an all-zeros tail sequence at the end of each block. Then, the code bit sequence $\{c_m\}$ is interleaved within the block.

B. Cyclic Prefix Reconstruction Procedure

In order to recover the contaminated samples in (2), two steps, called tail cancellation and cyclic reconstruction, must be taken [17], [20]. The procedure can be described as

$$\tilde{r}_{i,k} = r_{i,k} - \sum_{l=G+k+1}^{L} h_l x_{i-1,(k-l+G)_N} + \sum_{l=G+k+1}^{L} h_l x_{i,(k-l)_N}, \ 0 \le k < L-G. \tag{3}$$

The first step (tail cancellation) is to remove the ISI from the received symbol by subtracting the second term in (3), and the second step (cyclic reconstruction) is to restore the cyclicity by adding the last term. In practical implementation, the RISIC algorithm uses the $(i-1)$th decoded symbol $\{\hat{x}_{i-1,k}\}_{k=0}^{N-1}$ for the ISI removal. For cyclic reconstruction, the decoding process for $\{\hat{x}_{i,k}\}_{k=0}^{N-1}$ and the cyclic reconstruction are repeated several times. This method offers good SER performance when the difference of $L-G$ is moderate. However, as the difference becomes larger, it cannot obtain a reliable estimation of the transmitted symbol $\{x_{i,k}\}_{k=0}^{N-1}$ at the initial step, which causes a degradation in SER performance.

To solve this problem, we utilize the fact that the required signal for the cyclic reconstruction of the ith received symbol is found in the first $(L-G)$ samples of the $(i+1)$th received symbol. Those samples are given by

$$r_{i+1,k-G} = \sum_{l=0}^{k} h_l x_{i+1,(k-G-l)_N} + \sum_{l=k+1}^{L} h_l x_{i,(k-l)_N} + n_{i+1,k}, 0 \le k < L-G. \quad (4)$$

The second term in (4) contains the last term in (3). From this observation, we can obtain a reliable estimation of the ith transmitted symbol before the first iteration, and the SER performance becomes satisfactory after several iterations. The proposed CPR procedure can be described as follows.

1) Transform the estimated symbol $\{\hat{X}_{i-1,n}\}_{n=0}^{N-1}$ into the time domain symbol $\{\hat{x}_{i-1,k}\}_{k=0}^{N-1}$.
2) Remove ISI from $r_{i,k}$ as

$$\tilde{r}_{i,k}^{(0)} = \begin{cases} r_{i,k} - \sum\limits_{l=G+k+1}^{L} \hat{h}_l \hat{x}_{i-1,(k-l+G)_N}, & 0 \le k < L-G, \\ r_{i,k}, & L-G \le k < N. \end{cases} \quad (5)$$

3) Restore the cyclicity by using the first $(L-G)$ samples of $(i+1)$th symbol as

$$\tilde{r}_{i,k}^{(1)} = \begin{cases} \tilde{r}_{i,k}^{(0)} + \xi_k r_{i+1,k-G}, & 0 \le k < L-G, \\ \tilde{r}_{i,k}^{(0)}, & L-G \le k < N \end{cases} \quad (6)$$

where weighting coefficients ξ_k are given by

$$\xi_k = \frac{\sum\limits_{l=G+k+1}^{L} |\hat{h}_l|^2}{\sum\limits_{l=0}^{L} |\hat{h}_l|^2}, \quad 0 \le k < L-G. \quad (7)$$

Derivation of the optimum weighting coefficients is given in the sequel.

4) Estimate the transmitted symbol $\{\hat{X}_{i,n}\}_{n=0}^{N-1}$ from the $\{\tilde{r}_{i,k}^{(1)}\}_{k=0}^{N-1}$ and transform it into the time domain symbol $\{\hat{x}_{i,k}\}_{k=0}^{N-1}$.
5) Set the iteration index I to one.
6) Restore the cyclicity of the ith received symbol as

$$\tilde{r}_{i,k}^{(I+1)} = \begin{cases} \tilde{r}_{i,k}^{(0)} + \sum\limits_{l=G+k+1}^{L} \hat{h}_l \hat{x}_{i,(k-l)_N}^{(I)}, & 0 \le k < L-G, \\ \tilde{r}_{i,k}^{(0)}, & L-G \le k < N. \end{cases} \quad (8)$$

7) Estimate the transmitted symbol $\{\hat{X}_{i,n}\}_{n=0}^{N-1}$ from the $\{\tilde{r}_{i,k}^{(I+1)}\}_{k=0}^{N-1}$.
8) For more iteration, transform the $\{\hat{X}_{i,n}\}_{n=0}^{N-1}$ into the time domain symbol $\{\hat{x}_{i,k}\}_{k=0}^{N-1}$, and go to step 6) with $I \leftarrow I+1$.

The weighting coefficients in (7) have been obtained to minimize the interference power in $\{\tilde{r}_{i,k}^{(1)}\}_{k=0}^{N-1}$. In order to simplify the problem, we assumed that the samples $x_{i,k}$ are independent, identically distributed with variance P_x, the estimated channel $\hat{h}_l$ is identical to h_l, and the ISI removal in (5) is perfect. In addition, we neglected the noise $n_{i,k}$. Although the assumption is not completely true, (6) can effectively reduce the interference power. Substituting (2), (4), and (5) into (6), the $\{\tilde{r}_{i,k}^{(1)}\}_{k=0}^{L-G-1}$ can be rewritten as

$$\begin{aligned} \tilde{r}_{i,k}^{(1)} &= \sum_{l=0}^{L} h_l x_{i,(k-l)_N} + \xi_k \sum_{l=k+1}^{G+k} h_l x_{i,(k-l)_N} \\ &\quad + (\xi_k - 1) \sum_{l=G+k+1}^{L} h_l x_{i,(k-l)_N} \\ &\quad + \xi_k \sum_{l=0}^{k} h_l x_{i+1,(k-G-l)_N}, \quad 0 \le k < L-G \end{aligned} \quad (9)$$

where the first term is the desired signal, and the rest are interferences. Then, the interference power for the kth sample $\tilde{r}_{i,k}^{(1)}$ is given by

$$\begin{aligned} P_I(k) &= P_x \Big\{ \xi_k^2 \sum_{l=k+1}^{G+k} |h_l|^2 + (1-\xi_k)^2 \sum_{l=G+k+1}^{L} |h_l|^2 \\ &\quad + \xi_k^2 \sum_{l=0}^{k} |h_l|^2 \Big\} \\ &= P_x \Big\{ (1-\xi_k)^2 \sum_{l=G+k+1}^{L} |h_l|^2 + \xi_k^2 \sum_{l=0}^{G+k} |h_l|^2 \Big\}, \\ &\quad 0 \le k < L-G. \end{aligned} \quad (10)$$

The optimum ξ_k is obtained by setting the derivative of (10) with respect to ξ_k equal to zero as

$$\begin{aligned} \frac{\partial P_I(k)}{\partial \xi_k} &= P_x \Big\{ -2(1-\xi_k) \sum_{l=G+k+1}^{L} |h_l|^2 + 2\xi_k \sum_{l=0}^{G+k} |h_l|^2 \Big\} \\ &= 0, \ 0 \le k < L-G. \end{aligned} \quad (11)$$

Substituting $\hat{h}_l$ for h_l, the solution of the equation is obtained as (7).

C. Performance of OFDM with Insufficient CP

We evaluate the SER performance by calculating the signal-to-interference ratio (SIR) for each subcarrier of the OFDM system with insufficient CP. The SER performance of RISIC is well summarized in [17]. In this subsection, it is shown through analysis that by incorporating the proposed CPR technique, the interference power is significantly reduced as compared to the RISIC. All assumptions used in (9) are applied in this subsection except that the ISI from the $(i-1)$th transmitted symbol is considered. In this case, (6) can be expressed as

$$\begin{aligned}\tilde{r}_{i,k}^{(1)} = &\sum_{l=G+1}^{L} h_l x_{i-1,(k-l-G)_N} u(l-G-k-1)\\ &+\sum_{l=0}^{L} h_l x_{i,(k-l)_N} u(k-l+G)\\ &+\xi_k \left(\sum_{l=0}^{L} h_l x_{i+1,(k-G-l)_N} u(k-l) \right.\\ &\left. +\sum_{l=0}^{L} h_l x_{i,(k-l)_N} u(l-k-1) \right) u(L-G-k-1) \quad (12)\end{aligned}$$

where $u(n)$ is the unit step function. The ICI-reduced sample sequence $\{\tilde{r}_{i,k}^{(1)}\}_{k=0}^{N-1}$ is demodulated by taking the N-point FFT as

$$Y_n = \mathrm{FFT}\{\tilde{\mathbf{r}}_i^{(1)}\}(n) = \eta_n X_n + I_n \quad (13)$$

where $\tilde{r}_i^{(1)}$ represents a vector with elements from $\{\tilde{r}_{i,k}^{(1)}\}_{k=0}^{N-1}$,

$$\begin{aligned}\eta_n = &\sum_{l=0}^{G} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\\ &+\sum_{l=G+1}^{L} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\left(1+\frac{G}{N}-\frac{l}{N}\right)\\ &+\frac{1}{N}\sum_{l=0}^{L} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\sum_{k=0}^{L-G-1}\xi_k u(l-k-1) \quad (14)\end{aligned}$$

and

$$I_n = I_{i,n} + I_{i-1,n} + I_{i+1,n}. \quad (15)$$

Here, $I_{i,n}$ is the ICI term, and $I_{i-1,n}$ and $I_{i+1,n}$ are the ISI terms.

The SIR for the nth subcarrier before step 3) is defined by

$$\mathrm{SIR}_b(n) = P_{ub}(n)/P_{Ib}(n) \quad (16)$$

where the subscript "b" represents "*before* step 3)," and

$$P_{Ib}(n) = \frac{1}{2}E|I_n|_b^2 = \frac{1}{2}E|I_{i,n}|_b^2 + \frac{1}{2}E|I_{i-1,n}|_b^2. \quad (17)$$

The signal power (P_{ub}) and the interference power (P_{Ib}) were given in [17].

The SIR for the nth subcarrier after step 3) in the CPR procedure is defined by

$$\mathrm{SIR}_a(n) = P_{ua}(n)/P_{Ia}(n) \quad (18)$$

where the subscript "a" stands for "*after* step 3)." The signal power is given by

$$\begin{aligned}P_{ua}(n) = &\frac{1}{2}E|\eta_n X_n|_a^2\\ = &P_X\left|\sum_{l=0}^{G} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\right.\\ &+\sum_{l=G+1}^{L} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\left(1+\frac{G}{N}-\frac{l}{N}\right)\\ &\left.+\frac{1}{N}\sum_{l=0}^{L} h_l \exp\left(-j\frac{2\pi nl}{N}\right)\sum_{k=0}^{L-G-1}\xi_k u(l-k-1)\right|^2 \quad (19)\end{aligned}$$

where the transmitted symbol power $P_X = \frac{1}{2}E|X_n|^2$. The interference power can be expressed as

$$\begin{aligned}P_{Ia}(n) = &\frac{1}{2}E|I_n|_a^2\\ = &\frac{1}{2}E|I_{i-1,n}|_a^2 + \frac{1}{2}E|I_{i,n}|_a^2 + \frac{1}{2}E|I_{i+1,n}|_a^2 \quad (20)\end{aligned}$$

where

$$\frac{1}{2}E|I_{i-1,n}|_a^2 = \frac{1}{2}E|I_{i-1,n}|_b^2, \quad (21)$$

$$\begin{aligned}\frac{1}{2}E|I_{i,n}|_a^2 = &\frac{1}{2}E|I_{i,n}|_b^2 - \frac{P_X}{N^2}\left[2\mathrm{Re}\left\{\sum_{l=G+1}^{L}\sum_{l'=0}^{L} h_l h_{l'}^*\right.\right.\\ &\cdot\sum_{k=0}^{l-G-1}\sum_{k'=0}^{L-G-1}\xi_{k'} u(l'-k'-1)\exp\left(-j\frac{2\pi n(k-k')}{N}\right)\\ &\left.\cdot\sum_{\substack{m=0\\m\neq n}}^{N-1}\exp\left(-j\frac{2\pi m(l-l'-k+k')}{N}\right)\right\}\\ &+\sum_{l=0}^{L}\sum_{l'=0}^{L} h_l h_{l'}^* \sum_{k=0}^{L-G-1}\sum_{k'=0}^{L-G-1}\xi_k\xi_{k'} u(l-k-1)\\ &\cdot u(l'-k'-1)\exp\left(-j\frac{2\pi n(k-k')}{N}\right)\\ &\left.\cdot\sum_{\substack{m=0\\m\neq n}}^{N-1}\exp\left(-j\frac{2\pi m(l-l'-k+k')}{N}\right)\right], \quad (22)\end{aligned}$$

and

$$\begin{aligned}\frac{1}{2}E|I_{i+1,n}|_a^2 = &\frac{P_X}{N}\sum_{l=0}^{L}\sum_{l'=0}^{L} h_l h_{l'}^* \sum_{k=l}^{L-G-1}\sum_{k'=l'}^{L-G-1}\xi_k\xi_{k'} u(k-l)\\ &\cdot u(k'-l')\exp\left(-j\frac{2\pi n(k-k')}{N}\right)\\ &\cdot\delta(l-l'-k+k'). \quad (23)\end{aligned}$$

Comparing the interference power of the proposed CPR (see (20)) and the RISIC (see (17)), we see that the ISI is increased for the proposed CPR, due to the addition of the weighted $(i+1)$th symbol. However, the operation of the addition of the weighted $(i+1)$th symbol leads to a significant reduction in the

ICI power of the proposed CPR, thus resulting in a reduced total interference power. Using the signal and interference powers, the SER performance for different block sizes can be obtained as in Fig. 2. A guard interval is not used and the channel impulse response is assumed to be known, whose delay profile is represented as the first column in Table 3. Note from Fig. 2 that by adding the weighted next received symbol, a reliable estimation of the current received symbol is obtained before the first iteration. As such, the proposed CPR technique outperforms the RISIC before the first iteration, which yields a significantly improved SER performance after several iterations.

D. Receiver Structure

Fig. 1(b) represents receiver structure employing the proposed technique with turbo equalization. Unlike the RISIC, the proposed receiver structure inserts a delay of one symbol before the CP removal component. Thus, the proposed structure can utilize the next received symbol to restore the cyclicity of the current received symbol as depicted in Fig. 1(b). The ISI removal component performs the step 2) in the proposed CPR procedure. The cyclicity-restoration component performs the step 3) during the first iteration and the step 6) during the other iterations. The cyclicity-restored sequence $\{\tilde{r}_{i,k}^{(I)}\}_{k=0}^{N-1}$ is transformed into the $\{\tilde{R}_{i,n}^{(I)}\}_{n=0}^{N-1}$ using the FFT and is fed into the one-tap frequency domain equalizer. The tap coefficients of the one-tap frequency domain equalizer can be obtained as

$$E_n = \frac{H_n^*}{|H_n|^2 + \Omega_n/P_X}, \quad 0 \leq n < N-1 \tag{24}$$

where the channel gain H_n in the frequency domain is given by

$$H_n = \sum_{l=0}^{L} h_l W_N^{nl}, \quad 0 \leq n < N-1$$

and Ω_n is the noise power of the nth subcarrier.

The maximum *a posteriori* (MAP) detector calculates the log-likelihood ratio (LLR) $\lambda_1(c_{m'})$ of the coded bit $c_{m'}$ from the equalized symbol $Y_n = E_n \tilde{R}_{i,n}^{(I)}$. For instance, in case of 4-QAM, $\lambda_1(c_{m'})$ can be expressed as

$$\lambda_1(c_{m'}) = \begin{cases} 2\frac{\mathrm{Re}(Y_n)\mu_n}{|E_n|^2\Omega_n}, & m' = 2n, \\ 2\frac{\mathrm{Im}(Y_n)\mu_n}{|E_n|^2\Omega_n}, & m' = 2n+1 \end{cases} \tag{25}$$

where $\mu_n = E_n H_n$, and $\mathrm{Re}(\cdot)$ and $\mathrm{Im}(\cdot)$ represent the real and imaginary part of the input, respectively. Then, the $\lambda_1(c_{m'})$ is deinterleaved and fed into the SISO channel decoder. The SISO channel decoder, using the max-log-map algorithm, computes the *a posteriori* LLR $\Lambda_1(c_m)$ of the coded bit, and subtracts the *a priori* LLR $\lambda_1(c_m)$ from it. Next, the resultant extrinsic information $\lambda_2(c_m)$ is interleaved and converted into $\hat{X}_{i,n}$ by the soft symbol mapper. In case of 4-QAM, $\hat{X}_{i,n}$ can be expressed as

$$\hat{X}_{i,n} = \frac{1}{\sqrt{2}} \tanh\left[\frac{\lambda_2(c_{2n})}{2}\right] + j\frac{1}{\sqrt{2}} \tanh\left[\frac{\lambda_2(c_{2n+1})}{2}\right]. \tag{26}$$

Then, the estimate $\hat{X}_{i,n}$ is transformed into the time domain and is used in ISI removal and cyclicity restoration.

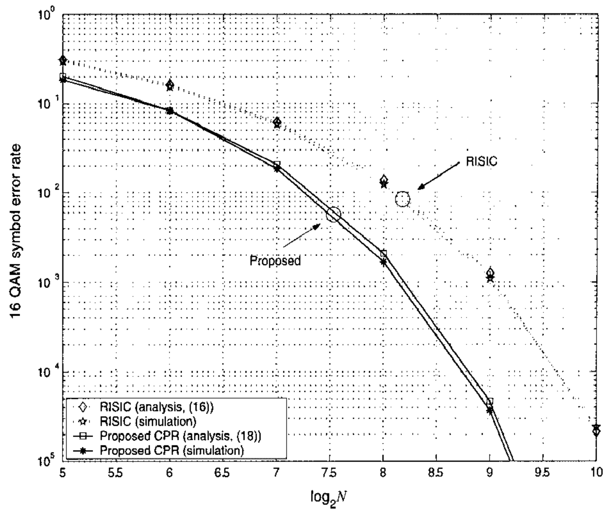

Fig. 2. Performance of OFDM with no CP for different block sizes.

III. CPR FOR STBC OFDM

A. Data Model

Let $\{X_{m,n}^p\}_{n=0}^{N-1}$, $p = 1, 2$, $m = 2i, 2i+1$, denote the transmitted symbol sequence of the pth transmit antenna at the mth OFDM symbol duration. Here, N is the number of subcarriers. The STBC OFDM encoder generates the coded symbol $X_{m,n}^p$ of nth subcarrier as follows [25]:

$$\begin{bmatrix} X_{2i,n}^1 & X_{2i+1,n}^1 \\ X_{2i,n}^2 & X_{2i+1,n}^2 \end{bmatrix} = \begin{bmatrix} D_{i,n}^1 & -D_{i,n}^{2*} \\ D_{i,n}^2 & D_{i,n}^{1*} \end{bmatrix} \tag{27}$$

where $\boldsymbol{D}_i = [D_{i,0}^1, D_{i,1}^1, \cdots, D_{i,N-1}^1, D_{i,0}^2, D_{i,1}^2, \cdots, D_{i,N-1}^2]$ is the modulated symbol vector for the $(2i)$th and $(2i+1)$th OFDM symbol durations and $(\cdot)^*$ denotes the complex conjugate operation. By IFFT, the transmitted symbol sequence $\{X_{m,n}^p\}_{n=0}^{N-1}$ is converted into the time domain sequence $\{x_{m,k}^p\}_{k=0}^{N-1}$, which is given by

$$x_{m,k}^p = \frac{1}{N} \sum_{n=0}^{N-1} X_{m,n}^p W_N^{-nk}, \quad k = 0, 1, \cdots, N-1 \tag{28}$$

where $p = 1, 2$, $m = 2i, 2i+1$. The received sequence for the $(2i)$th symbol can be represented as

$$r_{2i,k} = \begin{cases} \sum_{p=1}^{2} \sum\limits_{l=0}^{G+k} h_l^p x_{2i,(k-l)_N}^p \\ \quad + \sum_{p=1}^{2} \sum\limits_{l=G+k+1}^{L} h_l^p x_{2i-1,(k-l+G)_N}^p + n_{2i,k}, \\ \qquad\qquad -G \leq k < L-G, \\ \sum\limits_{p=1}^{2} \sum\limits_{l=0}^{L} h_l^p x_{2i,(k-l)_N}^p + n_{2i,k}, \\ \qquad\qquad L-G \leq k < N \end{cases} \tag{29}$$

where h_l^p is the lth tap of CIR between the receive antenna and pth transmit antenna. G is the length of CP and L is the length of the CIR. Similarly, for the $(2i+1)$th symbol duration, the received sequence can be written as

$$r_{2i+1,k} = \begin{cases} \sum_{p=1}^{2}\sum_{l=0}^{G+k} h_l^p x^p_{2i+1,(k-l)_N} \\ \quad + \sum_{p=1}^{2}\sum_{l=G+k+1}^{L} h_l^p x^p_{2i,(k-l+G)_N} + n_{2i+1,k}, \\ \qquad\qquad -G \le k < L-G, \\ \sum_{p=1}^{2}\sum_{l=0}^{L} h_l^p x^p_{2i+1,(k-l)_N} + n_{2i+1,k}, \\ \qquad\qquad L-G \le k < N. \end{cases} \quad (30)$$

For the STBC OFDM system, we assume that the channel is static during two consecutive OFDM symbol intervals.

B. The CPR Procedure for STBC OFDM

As depicted in (29) and (30), if $G \ge L$, all samples of the sequences $\{r_{2i,k}\}_{k=0}^{N-1}$ and $\{r_{2i+1,k}\}_{k=0}^{N-1}$ are free of interference. Otherwise, the first $(L-G)$ samples are contaminated with interference. In order to recover the contaminated samples, the tail cancellation and the cyclic reconstruction must be taken. The CPR procedure for STBC OFDM can be described as

$$\tilde{r}_{2i,k} = r_{2i,k} - \sum_{p=1}^{2}\sum_{l=G+k+1}^{L} h_l^p x^p_{2i-1,(k-l+G)_N} + \sum_{p=1}^{2}\sum_{l=G+k+1}^{L} h_l^p x^p_{2i,(k-l)_N}, \quad 0 \le k < L-G \quad (31)$$

and

$$\tilde{r}_{2i+1,k} = r_{2i+1,k} - \sum_{p=1}^{2}\sum_{l=G+k+1}^{L} h_l^p x^p_{2i,(k-l+G)_N} + \sum_{p=1}^{2}\sum_{l=G+k+1}^{L} h_l^p x^p_{2i+1,(k-l)_N}, \quad 0 \le k < L-G. \quad (32)$$

The first step is to remove the ISI by subtracting the second term in (31) and (32), and the second step is to restore the cyclicity by adding the last term. We observe from (31) and (32) that the ISI removal process for $\{r_{2i+1,k}\}_{k=0}^{N-1}$ is not the same as that for $\{r_{2i,k}\}_{k=0}^{N-1}$. For the ISI removal in $\{r_{2i,k}\}_{k=0}^{N-1}$, the $(2i-1)$th decoded symbols $\{\hat{x}^p_{2i-1,k}\}_{k=0}^{N-1}$, $p=1,2$, are used. However, in order to remove the ISI in $\{r_{2i+1,k}\}_{k=0}^{N-1}$, the decoding process for $\{\hat{x}^p_{2i,k}\}_{k=0}^{N-1}$, $p=1,2$, and the ISI removal should be repeated several times. The detailed CPR procedure for the STBC OFDM combined with CE will be discussed in Section IV. Unlike the single transmit antenna case, the ISI removal for $\{r_{2i+1,k}\}_{k=0}^{N-1}$ as well as the cyclicity restoration for $\{r_{2i,k}\}_{k=0}^{N-1}$ and $\{r_{2i+1,k}\}_{k=0}^{N-1}$ should be repeated in the STBC OFDM.

C. STBC OFDM Transceiver Structure Employing the CPR

Fig. 3(a) shows the STBC OFDM transmitter employing an interleaved convolutional code. As in the single antenna transmitter, the information bit sequence $\{d_l\}$ is converted into the

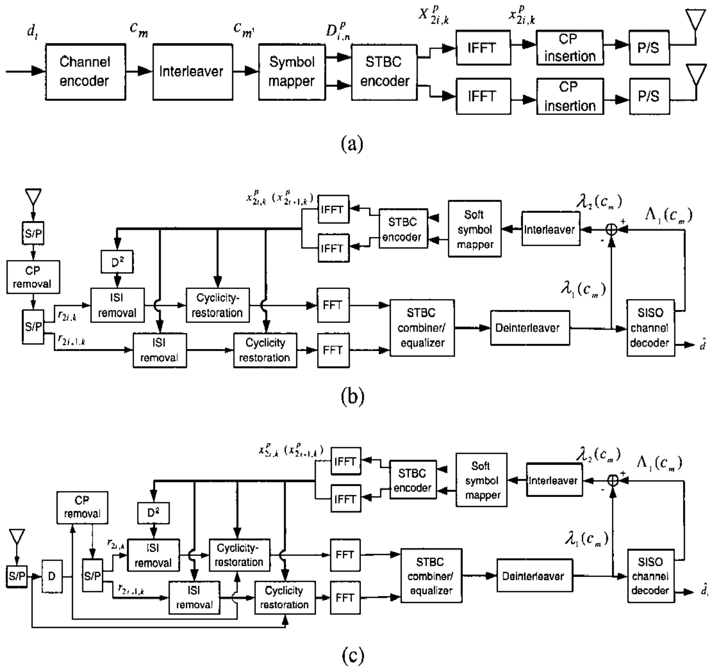

Fig. 3. Block diagram of STBC OFDM system; (a) STBC OFDM transmitter structure. STBC OFDM receiver structure employing (b) the RISIC technique or (c) the proposed CPR technique.

interleaved code bit sequence $\{c_{m'}\}$ with the interleaved convolutional code. The coded bits are mapped into QAM symbol sequence, encoded using the STBC encoder, and transmitted from the two antennas. Figs. 3(b) and 3(c) represent receiver structures employing the RISIC procedure and the proposed CPR procedure, respectively. The difference between the two receivers is that a delay of one symbol is inserted before the CP removal operation so as to enable the proposed cyclicity-restoration operation that is step 3) in the CPR procedure. The cyclicity-restored sequence $\{\tilde{r}^{(I)}_{m,k}\}_{k=0}^{N-1}$, $m=2i, 2i+1$, is transformed into the $\{\tilde{R}^{(I)}_{m,n}\}_{n=0}^{N-1}$, $m=2i, 2i+1$, using the FFT and is fed into the STBC combiner, which combines the received signals as follows:

$$S^1_{i,n} = \tilde{R}^{(I)}_{2i,n} H_n^{1*} + \tilde{R}^{(I)*}_{2i+1,n} H_n^2, \quad (33a)$$

$$S^2_{i,n} = \tilde{R}^{(I)}_{2i,n} H_n^{2*} - \tilde{R}^{(I)*}_{2i+1,n} H_n^1 \quad (33b)$$

where H_n^p is the nth subcarrier channel between the receive antenna and pth transmit antenna. The combined signals are sent to the one-tap frequency domain equalizer, whose coefficient values are computed as

$$E_n^p = \frac{\mathcal{H}_n}{|\mathcal{H}_n|^2 + \Omega_n^p / P_X}, \quad p = 1,2 \quad (34)$$

where $\mathcal{H}_n = |H_n^1|^2 + |H_n^2|^2$, and Ω_n^p is the noise power of the nth subcarrier at the pth transmit antenna.

Then, the LLR $\lambda_1(c_{m'})$ of the coded bit $c_{m'}$ is obtained from the equalized symbol $Y^p_{i,n} = E_n^p S^p_{i,n}$, $p=1,2$. The $\lambda_1(c_{m'})$ is deinterleaved and fed into the SISO channel decoder as the *a priori* information. Using the same soft symbol mapper as for the case of single antenna, we can obtain the estimate of information, $\{\hat{D}^p_{i,n}\}_{n=0}^{N-1}$, $p=1,2$, and then the estimate of coded symbol, $\{\hat{X}^p_{m,n}\}_{n=0}^{N-1}$, $p = 1, 2$, $m = 2i, 2i+1$, is obtained by the STBC encoder.

IV. ITERATIVE CHANNEL ESTIMATION METHOD FOR THE CPR

In this section, we first propose an iterative CE method for the OFDM system with the CPR algorithm and a single transmit antenna, based on the assumption that the channel is varied across the OFDM symbols. In this case, h_l in (2) can be represented as $h_{i,l}$ which is the lth tap of CIR at the ith OFDM symbol duration. Accurate channel information should be used for the CPR since error propagation due to imperfect channel information may occur (see (3)). The proposed CE method is based on the EM algorithm and exploits iteratively the extrinsic probabilities of the SISO decoder, which are the ISI removed and cyclic restored values by the CPR. We also extend the proposed iterative CE method to the STBC OFDM system with the CPR algorithm and two transmit antennas.

A. OFDM System without Transmit Diversity

A.1 Proposed Channel Estimation Method for OFDM System with Insufficient CP

If $G < L$, orthogonality of the subcarriers at the receiver is not preserved, resulting in ICI. Therefore, conventional CE methods [26]–[28], which have been developed with sufficient CP, cause degradation of the performance of OFDM system with insufficient CP. For the ith OFDM symbol with insufficient CP, we can rewrite (2) using such matrices as (35) shown on the top of the next page.

From (35), the ML estimate of $\boldsymbol{h}_i$ in the training symbol block can be obtained as

$$\hat{\boldsymbol{h}}_i = \arg\min_{\boldsymbol{h}_i} ||\boldsymbol{r}_i - \boldsymbol{\Upsilon}_i \boldsymbol{h}_i||^2. \quad (36)$$

In the data symbol block, since the input $\boldsymbol{\Upsilon}_i$ is not directly observable, we can obtain the estimate of $\boldsymbol{h}_i$ by minimizing an averaged form of the cost function [30], which is the essence of the EM algorithm [29]:

$$\begin{aligned}\hat{\boldsymbol{h}}_i &= \arg\min_{\boldsymbol{h}_i} E\left[||\boldsymbol{r}_i - \boldsymbol{\Upsilon}_i \boldsymbol{h}_i||^2\right] \\ &= \left(E\left[\boldsymbol{\Upsilon}_i^H \boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \boldsymbol{h}_i\right]\right)^{-1} E\left[\boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \boldsymbol{h}_i\right]^H \boldsymbol{r}_i. \quad (37)\end{aligned}$$

As shown in Fig. 4, the SISO decoder computes the *a posteriori* LLR $\Lambda_1(c_m)$ of the coded bit, and the extrinsic information $\lambda_2(c_m)$ is obtained by subtracting the *a priori* LLR $\lambda_1(c_m)$ from the *a posteriori* LLR $\Lambda_1(c_m)$. The terms $E\left[\boldsymbol{\Upsilon}_i^H \boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \boldsymbol{h}_i\right]$ and $E\left[\boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \boldsymbol{h}_i\right]$ in (37) can be calculated using the extrinsic probabilities of the SISO channel decoder and the $(i-1)$th decoded symbol $\{\hat{x}_{i-1,k}\}_{k=0}^{N-1}$.

A.2 Cyclic Prefix Reconstruction Procedure with Channel Estimation

Fig. 4 shows an OFDM receiver structure with CPR and CE. To restore the cyclicity, the cyclic reconstruction is repeated N_I times through the inner loop (solid line). After the CPR,

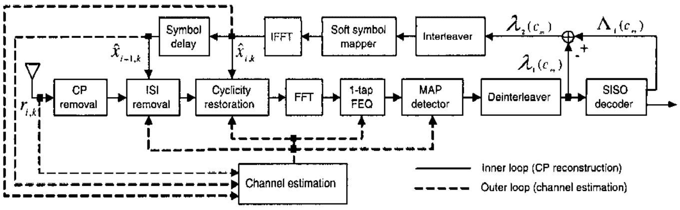

Fig. 4. OFDM receiver with CP reconstruction and channel estimation.

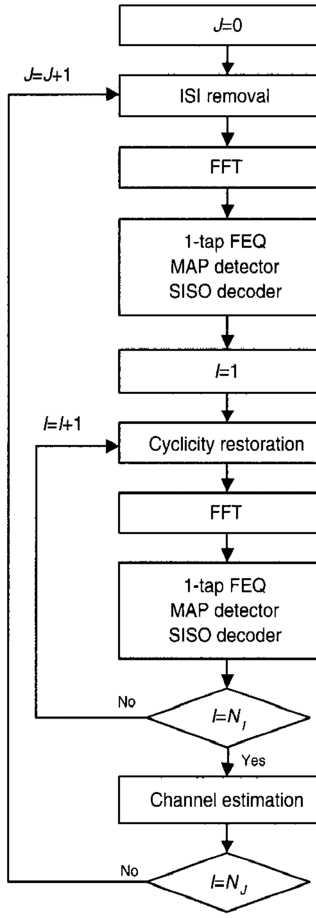

Fig. 5. CP reconstruction and channel estimation procedure for OFDM system (N_I and N_J are the numbers of iteration for CP reconstruction and channel estimation, respectively).

the channel is estimated with the extrinsic probabilities of the decoder, which are fed through the outer loop (dashed line). Whenever the channel is estimated, the ISI removal and the cyclicity restoration should be repeated through the inner loop. Since the CE is repeated N_J times, the equalization and the SISO channel decoding are executed $N_I N_J$ times. Fig. 5 shows the flow diagram for the CPR and the CE procedures. Detailed procedure of the CPR with the proposed iterative CE method will be discussed in the next subsection.

B. OFDM System with Transmit diversity

B.1 Proposed Channel Estimation Method for STBC OFDM System with Insufficient CP

Using (29) and (30), two ($(2i)$th and $(2i+1)$th) consecutive OFDM symbols with insufficient CP can be obtained as follows

$$\boldsymbol{r}_i = \boldsymbol{\Upsilon}_i \boldsymbol{h}_i + \boldsymbol{n}_i \Longleftrightarrow$$

$$\boldsymbol{r}_i = \boldsymbol{\Upsilon}_i \boldsymbol{h}_i + \boldsymbol{n}_i \Longleftrightarrow$$

$$\begin{bmatrix} r_{i,-G} \\ r_{i,-G-1} \\ \vdots \\ r_{i,L-G-1} \\ r_{i,L-G} \\ \vdots \\ r_{i,N-1} \end{bmatrix} = \begin{bmatrix} x_{i,N-G} & x_{i-1,N-1} & x_{i-1,N-2} & \cdots & x_{i-1,N-L} \\ x_{i,N-G+1} & x_{i,N-G} & x_{i-1,N-1} & \cdots & x_{i-1,N-L+1} \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ x_{i,N+L-G-1} & \cdots & x_{i,N-G+1} & x_{i,N-G} & x_{i-1,N-1} \\ x_{i,N+L-G} & x_{i,N+L-G-1} & \cdots & x_{i,N-G+1} & x_{i,N-G} \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ x_{i,N-1} & x_{i,N-2} & \cdots & x_{i,N-L} & x_{i,N-L-1} \end{bmatrix} \begin{bmatrix} h_0 \\ h_1 \\ \vdots \\ h_L \end{bmatrix} + \begin{bmatrix} n_{i,-G} \\ n_{i,-G-1} \\ \vdots \\ n_{i,L-G-1} \\ n_{i,L-G} \\ \vdots \\ n_{i,N-1} \end{bmatrix} \quad (35)$$

$$\begin{bmatrix} r_{2i,-G} \\ r_{2i,-G-1} \\ \vdots \\ r_{2i,N-1} \\ r_{2i+1,-G} \\ r_{2i+1,-G-1} \\ \vdots \\ r_{2i+1,N-1} \end{bmatrix} = \begin{bmatrix} \boldsymbol{\Upsilon}_{2i,1} & \boldsymbol{\Upsilon}_{2i,2} \\ \boldsymbol{\Upsilon}_{2i+1,1} & \boldsymbol{\Upsilon}_{2i+1,2} \end{bmatrix} \begin{bmatrix} h_0^1 \\ h_1^1 \\ \vdots \\ h_L^1 \\ h_0^2 \\ h_1^2 \\ \vdots \\ h_L^2 \end{bmatrix} + \begin{bmatrix} n_{2i,-G} \\ n_{2i,-G-1} \\ \vdots \\ n_{2i,N-1} \\ n_{2i+1,-G} \\ n_{2i+1,-G-1} \\ \vdots \\ n_{2i+1,N-1} \end{bmatrix} \quad (38)$$

where $\boldsymbol{\Upsilon}_{m,p} \triangleq$

$$\begin{bmatrix} x^p_{m,N-G} & x^p_{m-1,N-1} & x^p_{m-1,N-2} & \cdots & x^p_{m-1,N-L} \\ x^p_{m,N-G+1} & x^p_{m,N-G} & x^p_{m-1,N-1} & \cdots & x^p_{m-1,N-L+1} \\ & & \ddots & & \\ x^p_{m,N+L-G-1} & \cdots & x^p_{m,N-G+1} & x^p_{m,N-G} & x^p_{m-1,N-1} \\ x^p_{m,N+L-G} & x^p_{m,N+L-G-1} & \cdots & x^p_{m,N-G+1} & x^p_{m,N-G} \\ & & \ddots & & \\ x^p_{m,N-1} & x^p_{m,N-2} & \cdots & x^p_{m,N-L} & x^p_{m,N-L-1} \end{bmatrix}$$

and $p = 1, 2$, $m = 2i, 2i+1$. From (38), the estimate of $\boldsymbol{h}_i$ in the training symbol block and in the data symbol block can be obtained in a similar way to the case of single transmit antenna (see (36) and (37)).

With the proposed CE algorithm, the CPR procedure for STBC OFDM is shown in Fig. 6, and is briefly described in the steps below.

1) Set an initial channel estimate of $(2i)$th and $(2i+1)$th symbols as $\hat{\boldsymbol{h}}_i^{(J=0)} = \hat{\boldsymbol{h}}_{i-1}^{(J=N_J)}$. Set the iteration index J for the CE to zero.
2) Transform the estimated symbols $\{\hat{X}^p_{2i-1,n}\}_{n=0}^{N-1}$, $p = 1, 2$, into the time domain symbols $\{\hat{x}^p_{2i-1,k}\}_{k=0}^{N-1}$.
3) Remove ISI from $r_{2i,k}$ as

$$\tilde{r}^{(0)}_{2i,k} = \begin{cases} r_{2i,k} - \sum\limits_{l=G+k+1}^{L} \hat{h}_l^{(J)} \hat{x}^1_{2i-1,(k-l+G)_N} \\ - \sum\limits_{l=G+k+1}^{L} \hat{h}^{(J)}_{L+1+l} \hat{x}^2_{2i-1,(k-l+G)_N}, & 0 \le k < L-G, \\ r_{2i,k}, & L-G \le k < N \end{cases} \quad (39)$$

where $\hat{h}_l^{(J)}$ is the lth element of $\hat{\boldsymbol{h}}_i^{(J)}$.

4) Estimate the transmitted symbols $\{\hat{X}^p_{m,n}\}_{n=0}^{N-1}$, $p = 1, 2$, $m = 2i, 2i+1$, from the sequences $\{\tilde{r}^{(0)}_{2i,k}\}_{k=0}^{N-1}$ and $\{r_{2i+1,k}\}_{k=0}^{N-1}$, and transform them into the time domain symbols $\{\hat{x}^p_{m,k}\}_{k=0}^{N-1}$.

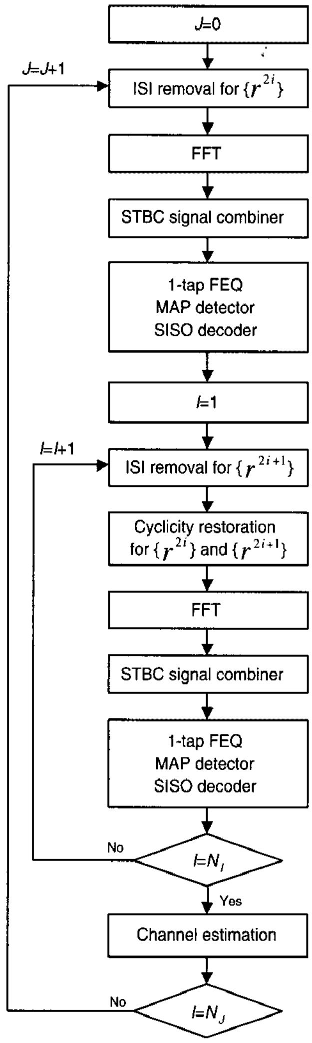

Fig. 6. CP reconstruction and channel estimation procedure for STBC OFDM system.

5) Set the iteration index I for CPR to one.
6) Remove ISI from $r_{2i+1,k}$ as

$$\tilde{r}^{(I-1)}_{2i+1,k} = \begin{cases} r_{2i+1,k} - \sum\limits_{l=G+k+1}^{L} \hat{h}_l^{(J)} \hat{x}^1_{2i,(k-l+G)_N} \\ - \sum\limits_{l=G+k+1}^{L} \hat{h}^{(J)}_{L+1+l} \hat{x}^2_{2i,(k-l+G)_N}, & 0 \le k < L-G, \\ r_{2i+1,k}, & L-G \le k < N. \end{cases} \quad (40)$$

7) Restore the cyclicity of the $(2i)$th and $(2i+1)$th received

Table 1. Computational complexity of the RISIC algorithm and the proposed CPR technique.

	Number of complex multiplication	Number of real multiplication
Tail cancellation (ISI removal)	$(N^\dagger/2)\log N + (L^\dagger - G^\dagger)(L-G+1)/2$	–
Cyclicity restoration & turbo decoding	$(N\log N + (L-G)(L-G+1)/2 + 3N)I^*$	$NI(10S^* + 16)$
Computing weight coefficients‡	–	$3(L-G)+2L$
Total	$(N/2+NI)\log N + (L-G)$ $\cdot(L-G+1)(I+1)/2 + 3NI$	$NI(10S+16)$ $+3(L-G)+2L$

‡ This operation of computing weight coefficients is needed only for the proposed CPR technique.
† N, L, and G represent the number of subcarriers, the length of CIR, and the length of CP, respectively.
* S and I stand for the number of states of the convolution code and the number of iterations, respectively.

Table 2. Total number of multiplications for the RISIC, the proposed CPR, the RISIC with STBC, the proposed CPR with STBC, the proposed CE, and the proposed CE with STBC.

	Total number of complex multiplications	Total number of real multiplications
Coded RISIC	$(N/2+NI)\log N + (L-G)$ $\cdot(L-G+1)(I+1)/2 + 3NI$	$NI(10S+16)$
Proposed CPR	$(N/2+NI)\log N + (L-G)$ $\cdot(L-G+1)(I+1)/2 + 3NI$	$NI(10S+16)$ $3(L-G)+2L$
Coded RISIC with STBC†	$(N(2I+1)\log N + (L-G)$ $\cdot(L-G+1)(3I+1) + 10NI)/2$	$NI(10S+16)+5N/2$
Proposed CPR with STBC†	$(N(2I+1)\log N + (L-G)$ $\cdot(L-G+1)(3I+1) + 10NI)/2$	$NI(10S+16)+5N/2$ $3(L-G)+2L$
Proposed CE	$(L+1)(L+2)(L+N+G+1)$	–
Proposed CE with STBC†	$2(L+1)(2L+3)(L+N+G+1)$	–

† The total number of multiplications per symbol is considered for complexity comparison.

symbols as

$$\tilde{r}_{2i,k}^{(I)} = \begin{cases} \tilde{r}_{2i,k}^{(0)} + \sum\limits_{l=G+k+1}^{L} \hat{h}_l^{(J)} \hat{x}_{2i,(k-l)_N}^1 \\ + \sum\limits_{l=G+k+1}^{L} \hat{h}_{L+1+l}^{(J)} \hat{x}_{2i,(k-l)_N}^2, 0 \le k < L-G, \\ \tilde{r}_{2i,k}^{(0)}, \quad L-G \le k < N \end{cases} \tag{41}$$

and

$$\tilde{r}_{2i+1,k}^{(I)} = \begin{cases} \tilde{r}_{2i+1,k}^{(I-1)} + \sum\limits_{l=G+k+1}^{L} \hat{h}_l^{(J)} \hat{x}_{2i+1,(k-l)_N}^1 \\ + \sum\limits_{l=G+k+1}^{L} \hat{h}_{L+1+l}^{(J)} \hat{x}_{2i+1,(k-l)_N}^2, 0 \le k < L-G, \\ \tilde{r}_{2i+1,k}^{(I-1)}, \quad L-G \le k < N. \end{cases} \tag{42}$$

8) Estimate the transmitted symbols $\{\hat{X}_{m,n}^p\}_{n=0}^{N-1}$, $p=1,2$, $m=2i, 2i+1$, from the sequences $\{\tilde{r}_{2i,k}^{(I)}\}_{k=0}^{N-1}$ and $\{\tilde{r}_{2i+1,k}^{(I)}\}_{k=0}^{N-1}$, and transform them into the time domain symbols $\{\hat{x}_{m,k}^p\}_{k=0}^{N-1}$.
9) If $I < N_I$, go to step 6) with $I \leftarrow I+1$. Otherwise, go to step 10).
10) Calculate the estimate of $\boldsymbol{h}_i$ with the time domain symbols $\{\hat{x}_{m,k}^p\}_{k=0}^{N-1}$, $p=1,2$, $m=2i,2i+1$, as

$$\hat{\boldsymbol{h}}_i^{(J+1)} = \left(E\left[\boldsymbol{\Upsilon}_i^H \boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \hat{\boldsymbol{h}}_i^{(J)}\right]\right)^{-1} E\left[\boldsymbol{\Upsilon}_i | \boldsymbol{r}_i, \hat{\boldsymbol{h}}_i^{(J)}\right]^H \boldsymbol{r}_i. \tag{43}$$

11) If $J < N_J$, go to step 3) with $J \leftarrow J+1$.

V. COMPUTATIONAL COMPLEXITY

Table 1 compares the computational complexity of the RISIC algorithm and the proposed CPR technique for the QPSK modulation, summarizing the number of complex and real multiplications for each operation in the procedure. The operation of computing weight coefficients in Table 1 is needed only for the proposed CPR technique, which is about 0.97% of the total real multiplications with $N=64, L=7, G=0, S=4$, and $I=1$. Table 2 summarizes the total number of multiplications for the RISIC, the proposed CPR, the RISIC with STBC, the proposed CPR with STBC, the proposed CE, and the proposed CE with STBC.

The number of IFFT operations can be reduced from four to two in the STBC OFDM receiver with insufficient CP. From (28), (31), (32), and (37), we observe that four separate IFFT operations are required in order to generate the four time-domain estimated sequences $\{\hat{x}_{m,k}^p\}_{k=0}^{N-1}$, $p=1,2$, $m=2i,2i+1$, which are used for the CPR and the CE. From (27), we note that the following relations for $n=0,1,\cdots,N-1$ are enforced:

$$X_{2i+1,n}^1 = -X_{2i,n}^{2*}, \quad X_{2i+1,n}^2 = X_{2i,n}^{1*}. \tag{44}$$

Furthermore, the DFT operation shows the following well-known symmetry property [31]:

$$x_{(-k)_N}^* \Leftrightarrow X_n^*, \quad n,k = 0,1,\cdots,N-1. \tag{45}$$

From (44) and (45), we can notice that $\{\hat{x}_{2i+1,k}^p\}_{k=0}^{N-1}$, $p=1,2$, are easily obtained from $\{\hat{x}_{2i,k}^p\}_{k=0}^{N-1}$, $p=1,2$, via sign inversion, complex conjugating, and reordering. That is,

$$\hat{x}_{2i+1,k}^1 = -\hat{x}_{2i,(-k)_N}^{2^*}, \quad \hat{x}_{2i+1,k}^2 = \hat{x}_{2i,(-k)_N}^{1^*}, \quad k=0,\cdots,N-1. \tag{46}$$

Thus, two of the four IFFT operations can be replaced by the complex conjugate operation and reordering. The computational complexity of the STBC OFDM in Table II was obtained using the above relations and property.

Table 3. Four-tap, reduced TU, and ad BU channel power delay profiles.

Four-tap static ISI		TU		BU	
Delay (μs)	Fractional power	Delay (μs)	Fractional power	Delay (μs)	Fractional power
0.0	0.15	0.0	0.189	0.0	0.164
0.2	0.65	0.2	0.379	0.3	0.293
0.4	0.15	0.5	0.239	1.0	0.147
0.6	0.05	1.6	0.095	1.6	0.094
		2.3	0.061	5.0	0.185
		5.0	0.037	6.6	0.117

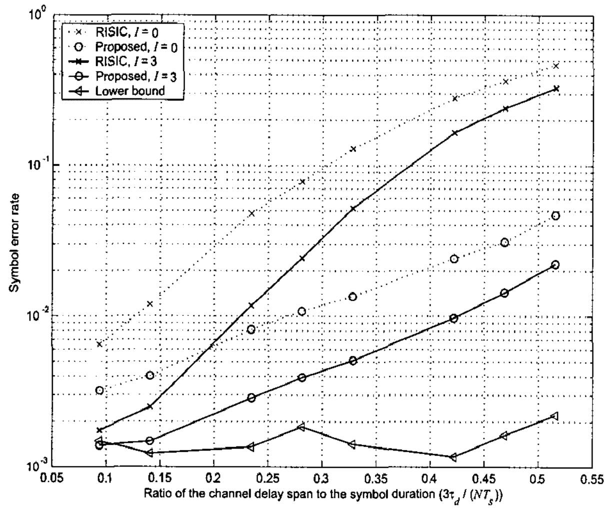

Fig. 7. SER performance of the OFDM system with no CP (G=0) for different delay spans.

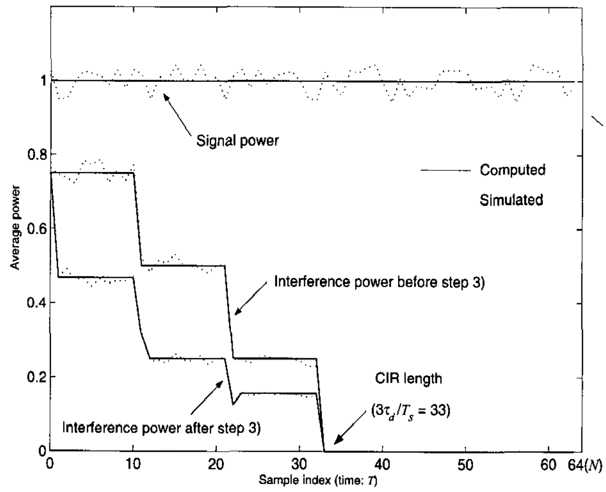

Fig. 8. Interference power before and after step 3) in the CPR procedure.

VI. SIMULATION RESULTS

A. Coded OFDM System with Single Antenna

In the simulations, we consider a coded OFDM system with 64 subcarriers ($N = 64$), no CP ($G = 0$), QPSK modulation, and a 1/2-rate convolutional code with constraint length of 7. In order to show the SER performance for different delay spans of the CIR, we used a four-path Rayleigh fading channel, which is defined as $c(t,\tau) = \sum_{l=0}^{3} c_l(t)\delta(\tau - l\tau_d)$, where $\delta(t)$ is the Dirac-delta function, and τ_d is delay span between adjacent paths. The path gains $c_l(t)$ are independent Gaussian random processes with zero mean and variance 0.25. We used the normalized Doppler frequency $f_D NT_s = 0.001$, where f_D is the maximum Doppler frequency and T_s is the sample period of the OFDM signal. Although we assumed perfect CE at the receiver, a practical CE does not sacrifice much performance for static or slowly fading channels.

Fig. 7 gives the SER performance with no iteration ($I = 0$) and three iterations ($I = 3$) for different delay spans at the average SNR = 9.5 dB, where the interference is not considered in the lower bound. For the channel delay span longer than 20% of the symbol duration, the SER performance of the proposed scheme with no iteration is better than that of the conventional RISIC with three iterations. In order to explain the effect of the proposed scheme more clearly, the interference powers before and after step 3) in the CPR procedure for a static channel with $3\tau_d/NT_s = 0.516$ are shown in Fig. 8. The interference powers before and after step 3) are computed using (10) with zero ξ_k's and the optimum ξ_k's given as (7), respectively. The total interference power is proportional to the area under the interference power curve in Fig. 8, and it can be computed that, through step 3), the total interference power is reduced by 2.2 dB. The reduction of the interference power leads to a reliable estimation of the transmitted symbols, which facilitates the iterative CPR process.

We also compare the SER performance on the six-tap reduced typical urban (TU) and bad urban (BU) channels, which have been defined in [32]. Power delay profiles of TU and BU channels are shown in Table 3. Fig. 9 gives the SER performance on the TU channel with $f_D NT_s = 0.001$. We used $NT_s = 8.0$ μs so that the channel delay span is 62.5% of the OFDM symbol duration. In this case, the use of sufficient CP costs a 38.5% reduction in bandwidth utilization efficiency and a 2.11 dB loss in the SNR. For three iterations, the proposed scheme is about 2.0 dB better than the conventional RISIC at the SER of 10^{-3}, with the SNR loss from the lower bound of only about 0.6 dB. Fig. 10 gives the SER performance on the BU channel with $f_D NT_s = 0.001$ and $NT_s = 12.8$ μs. In this case, the use of sufficient CP costs a 34.0% reduction in bandwidth utilization efficiency, which is equivalent to a 1.8 dB loss in the SNR. Since the interference power due to insufficient CP is very large, the conventional RISIC technique fails to provide an acceptable SER performance. The SNR gap between the conventional scheme and the lower bound is larger than 5 dB at the SER of 10^{-2}. Note also that the gap increases as the SER decreases. In contrast, the SNR loss of the proposed scheme from the lower bound is about 1.6 dB at the SER of 10^{-3}, which is smaller than the SNR loss of 1.8 dB due to the use of sufficient CP.

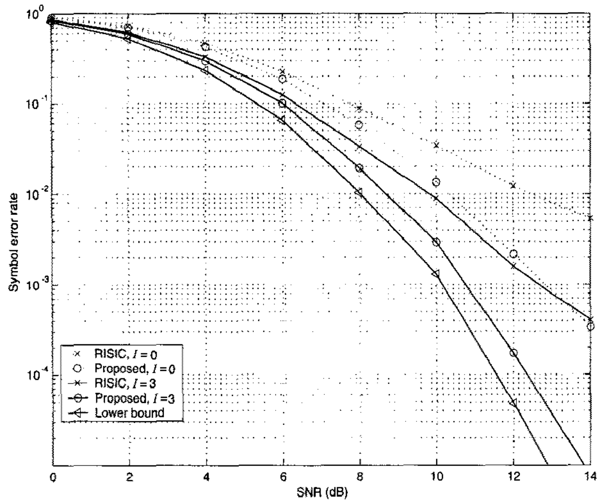

Fig. 9. SER performance of the OFDM system with no CP (G=0) on a TU channel ($L = 40$).

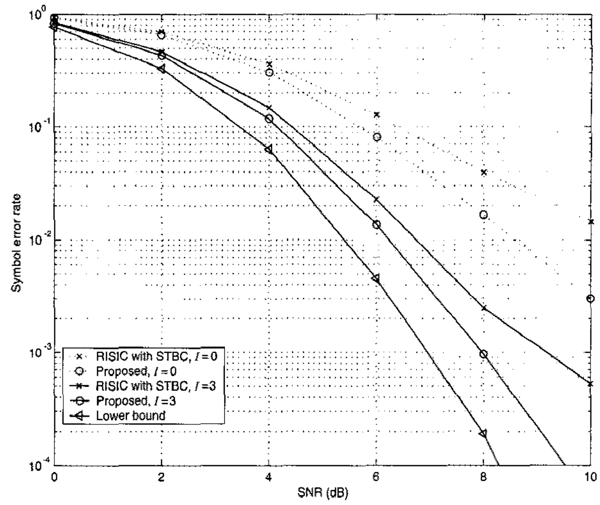

Fig. 11. SER performance of the STBC OFDM system (G=0) on a TU channel ($L = 40$).

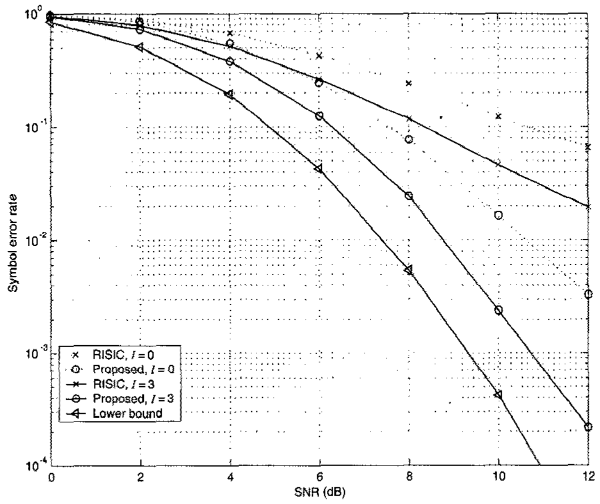

Fig. 10. SER performance of the OFDM system with no CP (G=0) on a BU channel ($L = 33$).

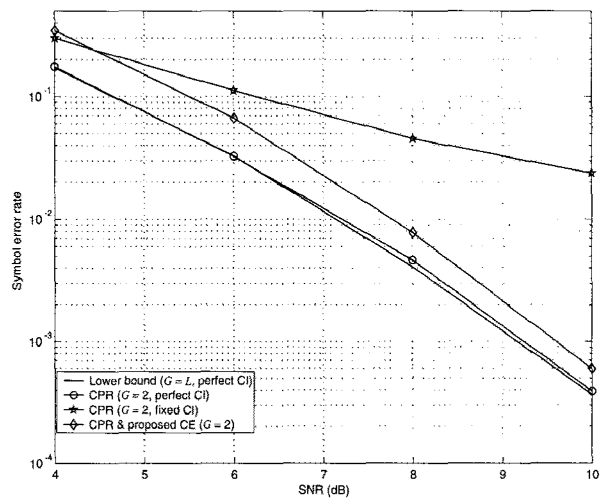

Fig. 12. SER performance of the OFDM system with the CPR and the proposed CE ($L = 7$).

B. STBC OFDM System with Two Transmit Antennas

In this simulation, we used 64 subcarriers, no CP, QPSK modulation, and a 1/2-rate interleaved convolutional code as in the single antenna simulation. The two channels between antenna pairs, which are fading independently, are assumed to have the same statistical properties, i.e., the same delay power profile. The SER performance over the TU channel with $f_D N T_s = 0.001$ ($NT_s = 8.0\ \mu s$) is shown in Fig. 11. For three iterations, the STBC OFDM system employing the proposed CPR algorithm is about 1.2 dB better than the conventional RISIC scheme at the SER of 10^{-3}, with the SNR loss from the lower bound of about 1.0 dB.

C. Iterative Channel Estimation

In the simulations, we used 64 subcarriers with QPSK modulation. For a coded OFDM, a rate-1/2 convolutional code with constraint length of 7. Fig. 12 shows the SER performance at the normalized Doppler frequency $f_D N T_s$ of 0.01. Every frame consists of a training symbol period and nine data symbol periods. The length of CIR is equal to 7 ($L = 7$) and the exponentially power-decaying eight-path Rayleigh fading channel model, which is defined as $h_l = \frac{1}{S_h} e^{-l/\tau_{\max}}$, $l = 0, 1, \cdots, 7$, is used in the system. Here, $\tau_{\max} = 8/\sqrt{3}$ and S_h is the normalization factor ($S_h = \sum_{l=0}^{7} e^{-l/\tau_{\max}}$). The iteration parameters for CPR and CE are set to be equal to 2 ($N_I = 2, N_J = 2$). In the figure, the solid line shows the lower bound, which is obtained with sufficient CP ($G = L$) and perfect channel information. The circle line shows the SER performance of CPR with perfect channel information and insufficient CP ($G = 2$). We observe that the SER performance of CPR with insufficient CP ($G = 2$) is almost identical as the lower bound. The star line shows the SER performance of CPR without CE. In this case, the channel information of training symbol period is used for decoding of data symbol during the data symbol period. Since the channel is fast time-varying ($f_D N T_s = 0.01$), an error floor occurs. Finally, the diamond line shows the SER performance of CPR with the proposed CE method. It is noted from Fig. 12 that the proposed CE method with CPR improves significantly

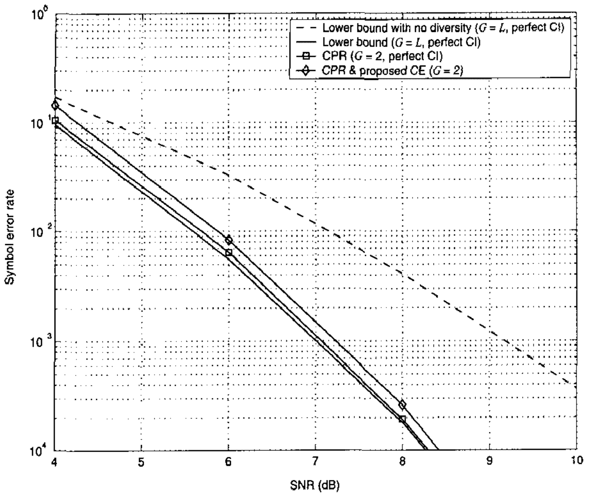

Fig. 13. SER performance of the STBC OFDM system with the CPR and the proposed CE on a exponentially power-decaying Rayleigh fading channel ($L = 7$).

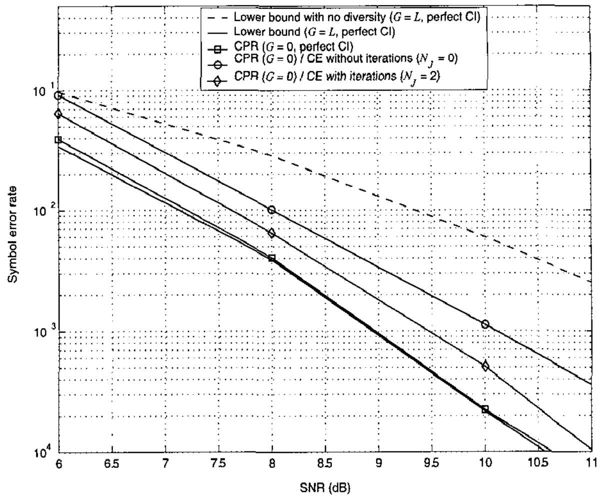

Fig. 14. SER performance of the STBC OFDM system with the CPR and the proposed CE on a TU channel ($L = 8$).

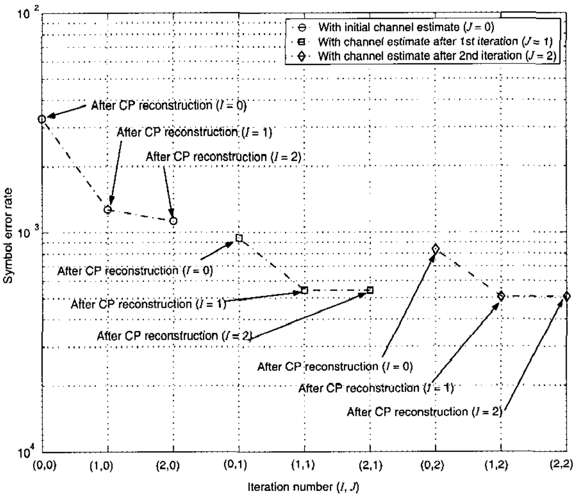

Fig. 15. SER performance for different number of iterations of the CPR and the CE (SNR = 10 dB, $L = 8$, $G = 0$).

the SER performance, compared to the case without CE.

Fig. 13 shows the SER performance in STBC OFDM system, at $f_D NT_s = 0.001$. In the STBC OFDM, every frame consists of two training symbol periods (one STBC symbol pair) and 18 data symbol periods (nine STBC symbol pairs). In the figure, the dashed and the solid lines show the lower bounds with sufficient CP ($G = L$) and perfect channel information, which are obtained in OFDM with no diversity and with the STBC, respectively. We can observe from these lines that the STBC OFDM system has full diversity by introducing temporal and spatial correlation into the signals transmitted from two transmit antennas. Note from Fig. 13 that the performance of CPR in STBC OFDM with insufficient CP ($G = 2$) is almost the same as the lower bound. Therefore, we can improve the spectral efficiency of the OFDM system by using the CPR with $G = 2$, instead of $G = L$. In the figure, the diamond line shows the performance of CPR and CE in STBC OFDM system. We notice from Fig. 13 that the CPR with the proposed CE method in STBC OFDM system has a slight SER performance degradation (about 0.2 dB at the SER of 10^{-3}), compared with the lower bound.

Fig. 14 shows the SER performance in STBC OFDM system on a six-tap reduced TU channel with $f_D NT_s = 0.001$. We used $NT_s = 40.0$ μs so that the length of the CIR is equal to 8 ($L = 8$). Note from Fig. 14 that the performance of the CPR with $G = 0$ is almost the same as the lower bound, and thus we improved the spectral efficiency of the STBC OFDM system by 11.11%. In the figure, the circle and the diamond lines show the performances of CPR with initial channel estimate ($J = 0$) and with the channel estimate obtained after 2nd iteration for the CE ($J = 2$), respectively. We observe that the performance degradation for the proposed iterative CE method is about 0.5 dB, compared to the STBC OFDM with perfect channel information at the SER of 10^{-3}. The figure also shows that using the proposed iterative estimation method ($J = 2$) results in about 0.6 dB gain compared with the case with no iterations ($J = 0$) at the SER of 10^{-3}. The performance with respect to the number of iterations is given in Fig. 15. In Fig. 15, the SNR is set to 10 dB and no CP ($G = 0$) is used. The circle line shows the performance of CPR with initial channel estimate ($J = 0$). By the ISI removal and iterative cyclicity restoration, the SER performance is improved from 3.296×10^{-3} to 1.123×10^{-3}. As such, the extrinsic probabilities of the SISO decoder become more reliable, which is then used for accurate CE. The square and the diamond lines show the performances of CPR with channel estimate after 1st iteration ($J = 1$) and after 2nd iteration ($J = 2$), respectively. It is observed from Fig. 15 that the SER performance of 5.070×10^{-4} is obtained after $N_I N_J$ iterations, which is about six times better than the initial SER of 3.296×10^{-3}.

VII. CONCLUSIONS

An efficient CPR procedure was proposed for the coded OFDM system and was applied to the STBC OFDM system with two transmit antennas. The proposed method restores the cyclicity of the current received symbol, without the aid of the transmitted symbol estimate, by adding the weighted next received symbol to the current received symbol. Simulation results show that the OFDM system employing the new CPR procedure achieves an SER performance close to the lower bound, without the loss of spectral efficiency, on channels with the delay span longer than half of the symbol duration.

This paper also proposed an iterative CE method for the OFDM system with insufficient CP. The proposed method is based on the EM algorithm and exploits the extrinsic probabilities of the channel decoder, which are the ISI removed and cyclic restored values by the CPR. Further, we have applied the RISIC CPR method to Alamouti STBC OFDM system. In the STBC OFDM, ISI removal as well as cyclic restoration should be repeated for the CPR. Finally, an iterative CE method for the STBC OFDM system with CPR has been proposed. The computational complexity of the CPR and the CE with and without transmit diversity was analyzed and their performances were investigated in multipath fading environments. The spectral efficiency gain over OFDM systems using sufficient CP can be converted into a higher coding gain, which exceeds the SNR loss. By applying the proposed CPR algorithm to STBC OFDM system, we can obtain both the benefits of diversity gain and spectral efficiency gain.

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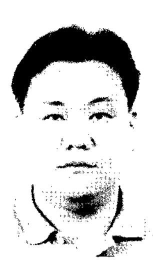

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