Joint Iterative Transmit/Receive Frequency-Domain Equalization & ISI Cancellation for Broadband Single-carrier Block Transmissions

Kazuki Takeda and Fumiyuki Adachi
Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University
6-6-05, Aza-aoba, Aramaki, Aoba-ku, Sendai, 980-8579, JAPAN
+81-22-795-7174
kazuki@mobile.ecei.tohoku.ac.jp adachi@ecei.tohoku.ac.jp

ABSTRACT
In this paper, we propose a new joint transmit/receive equalization technique for single-carrier (SC) block transmissions in a severe frequency-selective fading channel. An iterative frequency-domain inter-symbol interference cancellation (FDIC) is introduced to the joint transmit/receive (Tx/Rx) frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion which was previously proposed by authors. At a receiver, a 1-tap FDE and FDIC are jointly used and they are updated in an iterative manner to achieve larger frequency diversity gain. Since the transmit FDE weight cannot be updated, we introduce a virtual receiver having a receive FDE & FDIC into the transmitter. In the virtual receiver at the transmitter, the same degree of the residual ISI cancellation is assumed to compute the transmit FDE weight, based on the MMSE criterion, matched to the receive FDE & FDIC. The computer simulation results show that the bit error rate (BER) performance is significantly improved by using our proposed scheme.

Categories and Subject Descriptors
C.2.1 [**Computer-Communication Networks**]: Network Architecture and Design – Wireless communication

General Terms
Performance

Keywords
Single-carrier, frequency-domain equalization, interference cancellation

1. INTRODUCTION
High-speed and high-quality data transmissions are demanded in the next generation wireless communication systems [1]. As the data rate increases, the number of resolvable propagation paths tends to increase and the channel frequency-selectivity gets stronger, thereby producing severe inter-symbol interference (ISI) [2]. Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion is able to overcome the SC transmission problem arising from the severe channel frequency-selectivity [3, 4]. The SC signal transmission using FDE is a block transmission. By inserting a cyclic prefix (CP) into a guard interval (GI) of each transmit signal block, the received signal block can be transformed into frequency-domain signals by fast Fourier transform (FFT). One-tap receive FDE weight computed based on the MMSE criterion is multiplied to each component of frequency-domain signal to compensate the spectrum distortion. SC with MMSE-FDE provides almost the same transmission performance as orthogonal frequency division multiplexing (OFDM).

However, SC with MMSE-FDE still suffers from the ISI (residual ISI) which remains after MMSE-FDE. The BER performance of SC with MMSE-FDE is a few dB away from the matched-filter (MF) bound [5]. Reducing the residual ISI can further improve the BER performance. One way to reduce the residual ISI is to introduce an ISI cancellation technique. In [6-8], frequency-domain iterative ISI cancellation (FDIC) (or called block iterative decision feedback equalizer (DFE) in the frequency-domain) was proposed. The residual ISI replica is generated using the log-likelihood ratio (LLR) of the received signal after channel decoding and is subtracted from the received signal after the receive FDE. A series of the receive FDE & FDIC and channel decoding is repeated sufficient number of times. In each iteration, the receive FDE weight is updated by taking the reliability of the residual ISI replica into account.

Another way for reducing the residual ISI is to introduce the transmit FDE into the SC with receive FDE. In [9], we proposed a joint transmit/receive (Tx/Rx) MMSE-FDE for SC transmission. Channel state information (CSI) is shared by the transmitter and the receiver and the FDE is jointly carried out at both the transmitter and receiver. The set of FDE weights is optimized base on the MMSE criterion. We showed by theoretical analysis and computer simulation that the joint Tx/Rx MMSE-FDE provides better BER performance than the receive MMSE-FDE.

In this paper, we introduce the FDIC into the previously proposed joint Tx/Rx MMSE-FDE (called joint iterative Tx/Rx FDE & FDIC in this paper) to further improve the BER performance of broadband SC block transmissions. Similar to [8], the receive FDE weight and the residual ISI replica are updated in each iteration. However, the transmit FDE weight cannot be updated since the receive FDE weight is unknown to the transmitter. To compute the transmit FDE weight, we introduce a virtual receiver having a receive FDE & FDIC into the transmitter. In the virtual receiver at the transmitter, the same degree of the residual ISI cancellation is assumed to compute the transmit FDE weight, based on the MMSE criterion, matched to the receive FDE & FDIC. We evaluate the bit error rate (BER) performance improvement by computer simulation to confirm the effectiveness of our proposed technique.

The rest of the paper is organized as follows. Section 2 presents the system model of SC with joint iterative Tx/Rx FDE & FDIC. The transmit and received signal representations are given in Sect. 3. The set of FDE weights are derived in Sect. 4.

"IWCMC'10, June 28– July 2, 2010, Caen, France.

The simulation results are presented in Sect. 5. Section 6 concludes this paper.

2. SYSTEM MODEL

The system model of SC using the joint iterative Tx/Rx FDE & FDIC is illustrated in Fig. 1. We assume perfect knowledge of CSI is shared by the transmitter and the receiver. For channel coding, we apply turbo coding with rate-1/3 having two recursive systematic convolutional (RSC) encoders [10, 11]. At the transmitter, two parity bit sequences generated by encoders are punctured to form the codeword with the length K. The coding rate is represented by R. The codeword is then data-modulated. The data-modulated sequence is divided into N_c-symbol blocks and each block is transformed into the frequency-domain signal by N_c-point FFT. After multiplying with the transmit FDE weight, frequency-domain signal is transformed back into the time-domain signal by N_c-point inverse FFT (IFFT). Then, the N_g-sample CP is added to each N_c-symbol block and the resultant signal blocks are transmitted over the frequency-selective channel.

At the receiver, after removing the CP, a series of receive FDE and FDIC is carried out in an iterative fashion. Let us consider the ith iteration stage ($0<i\leq I$). The received signal block is transformed into the frequency-domain signal by FFT. Each component of the frequency-domain signal is multiplied with the receive FDE weight and then, subtracted the residual ISI replica generated after the turbo decoding in the (i–1)th iteration stage. The frequency-domain signal is transformed into the time-domain signal by IFFT. After obtaining all the decision variables associated with the data-modulated symbols in a block, bit LLRs are computed for turbo decoding. The turbo decoder output will be used for generating an updated residual ISI replica.

The joint iterative Tx/Rx FDE & FDIC is performed block-by-block and hence, we consider the transmission of one N_c-symbol block in this section.

2.1 Channel Model

In this paper, we use a symbol-spaced discrete-time signal representation. The propagation channel is assumed to be an L-path frequency-selective block fading channel having its impulse response as

$$h(\tau)=\sum_{l=0}^{L-1}h_l\delta(\tau-\tau_l), \tag{1}$$

where h_l and τ_l are the complex-valued path gain and the delay time of the lth path, respectively. We assume $\sum_{l=0}^{L-1}E[|h_l|^2]=1$ and $\tau_l=l$.

2.2 Transmit Signal

The data symbol block is represented using a vector form as $\mathbf{d}=[d(0),\ \ldots,\ d(t),\ \ldots,\ d(N_c-1)]^T$. N_c-point FFT is carried out on $\mathbf{d}$ to obtain the frequency-domain transmit signal $\mathbf{D}=[D(0),\ \ldots,\ D(k),\ \ldots,\ D(N_c-1)]^T$ as

$$\mathbf{D}=\mathbf{Fd} \tag{2}$$

with $\mathbf{F}$ being an $N_c\times N_c$ FFT matrix given as

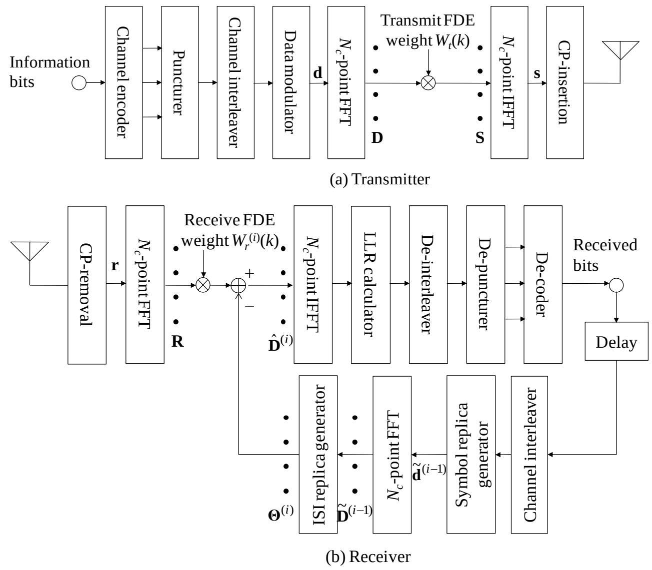

Figure 1. System model.

$$\mathbf{F}=\frac{1}{\sqrt{N_c}}\begin{bmatrix}1 & 1 & \cdots & 1\\ 1 & e^{-j2\pi\frac{(1\times1)}{N_c}} & \cdots & e^{-j2\pi\frac{(1\times(N_c-1))}{N_c}}\\ \vdots & \vdots & \ddots & \vdots\\ 1 & e^{-j2\pi\frac{((N_c-1)\times1)}{N_c}} & \cdots & e^{-j2\pi\frac{((N_c-1)\times(N_c-1))}{N_c}}\end{bmatrix}. \tag{3}$$

The transmit FDE weight $\mathbf{W}_t=diag\{W_t(0),\ \ldots,\ W_t(k),\ \ldots,\ W_t(N_c-1)\}$ is multiplied to $\mathbf{D}$ as

$$\mathbf{S}=[S(0),\ \ \ldots,\ \ S(k),\ \ \ldots,\ \ S(N_c-1)]^T=\mathbf{W}_t\mathbf{D} \tag{4}$$

with

$$tr[\mathbf{W}_t\mathbf{W}_t^H]=N_c. \tag{5}$$

Eq. (5) is satisfied to keep the transmit signal power unchanged.

An N_c-point IFFT is applied to $\mathbf{S}$ to obtain a time-domain transmit signal block $\mathbf{s}=[s(0),\ \ldots,\ s(t),\ \ldots,\ s(N_c-1)]^T=\mathbf{F}^H\mathbf{S}$. After the insertion of CP, the signal block is transmitted.

2.3 Received Signal

The CP-length is assumed to be equal or longer than the maximum channel time delay. The received signal block $\mathbf{r}=[r(0),\ \ldots,r(t),\ \ldots,\ r(N_c-1)]^T$ after the CP-removal can be expressed as

$$\mathbf{r}=\sqrt{\frac{2E_s}{T_s}}\mathbf{hs}+\mathbf{n}, \tag{6}$$

where E_s and T_s are the average transmit symbol energy and symbol duration, respectively, $\mathbf{h}$ is an $N_c\times N_c$ circulant channel matrix given by

$$\mathbf{h}=\begin{bmatrix}h_0 & & & h_{L-1} & \cdots & h_1\\ h_1 & \ddots & & & \ddots & \vdots\\ \vdots & & h_0 & \mathbf{0} & & h_{L-1}\\ h_{L-1} & & h_1 & \ddots & & \\ & \ddots & \vdots & & \ddots & \\ \mathbf{0} & & h_{L-1} & \cdots & \cdots & h_0\end{bmatrix} \tag{7}$$

and $\mathbf{n}$=[$n(0)$, …, $n(t)$, …, $n(N_c-1)]^T$ is the noise vector with $n(t)$ being a zero-mean additive white Gaussian noise (AWGN) having variance $2N_0/T_c$ (N_0 is the one-sided noise power spectrum density).

An N_c-point FFT is carried out on $\mathbf{r}$ to obtain the frequency-domain received signal block $\mathbf{R}=[R(0), …, R(k), …, R(N_c-1)]^T$ as

$$\mathbf{R}=\mathbf{F}\mathbf{r}=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}\mathbf{S}+\mathbf{N}=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}\mathbf{W}_t\mathbf{D}+\mathbf{N}, \tag{8}$$

where $\mathbf{N}=\mathbf{F}\mathbf{n}$ and $\mathbf{H}=\mathbf{F}\mathbf{h}\mathbf{F}^H$. Due to the circulant property of $\mathbf{h}$, the channel gain matrix $\mathbf{H}$ of size $N_c\times N_c$ is diagonal. The kth diagonal element of $\mathbf{H}$ is given by

$$H(k)=\sum_{l=0}^{L-1}h_l\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right). \tag{9}$$

The receive FDE & FDIC is iteratively performed on the frequency-domain received signal. At the ith iteration stage, the frequency-domain received signal block after the receive FDE & FDIC, $\hat{\mathbf{D}}^{(i)}=[\hat{D}^{(i)}(0), …, \hat{D}^{(i)}(k), …, \hat{D}^{(i)}(N_c-1)]^T$, can be written as

$$\hat{\mathbf{D}}^{(i)}=\mathbf{W}_r^{(i)}\mathbf{R}-\mathbf{\Theta}^{(i)}=\sqrt{\frac{2E_s}{T_s}}{}_s\,\mathbf{W}_r^{(i)}\mathbf{H}\mathbf{W}_t\mathbf{D}-\mathbf{\Theta}^{(i)}+\mathbf{W}_r^{(i)}\mathbf{N}, \tag{10}$$

where $\mathbf{W}_r^{(i)}=diag\{W_r^{(i)}(0), …, W_r^{(i)}(k), …, W_r^{(i)}(N_c-1)\}$ is an $N_c\times N_c$ diagonal receive FDE weight matrix for the ith iteration. $\mathbf{\Theta}^{(i)}$ is an $N_c\times 1$ vector representing the residual ISI replica block, given as

$$\mathbf{\Theta}^{(i)}=\sqrt{\frac{2E_s}{T_s}}\{\mathbf{W}_r^{(i)}\mathbf{H}\mathbf{W}_t-\mathbf{I}\}\tilde{\mathbf{D}}^{(i-1)} \tag{11}$$

with $\tilde{\mathbf{D}}^{(i-1)}=[\tilde{D}^{(i-1)}(0), …, \tilde{D}^{(i-1)}(k), …, \tilde{D}^{(i-1)}(N_c-1)]^T$ being the frequency-domain data symbol replica block. The data symbol replica generation will be presented in the next subsection. $\hat{\mathbf{D}}^{(i)}$ is transformed by an N_c-point IFFT to the time-domain decision variable block $\hat{\mathbf{d}}^{(i)}$ associated with the data symbol block $\mathbf{d}$ as

$$\begin{aligned}\hat{\mathbf{d}}^{(i)}&=[\hat{d}^{(i)}(0), \ldots, \hat{d}^{(i)}(t), \ldots, \hat{d}^{(i)}(N_c-1)]^T=\mathbf{F}^H\hat{\mathbf{D}}^{(i)}\\&=\sqrt{\frac{2E_s}{T_s}}\mathbf{d}+\sqrt{\frac{2E_s}{T_s}}\{\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{H}\mathbf{W}_t\mathbf{F}-\mathbf{I}\}\{\mathbf{d}-\tilde{\mathbf{d}}^{(i-1)}\}\\&\quad+\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{F}\mathbf{n},\end{aligned} \tag{12}$$

where $\tilde{\mathbf{d}}^{(i-1)}=[\tilde{d}^{(i-1)}(0), …, \tilde{d}^{(i-1)}(t), …, \tilde{d}^{(i-1)}(N_c-1)]^T$ $=\mathbf{F}^H\tilde{\mathbf{D}}^{(i-1)}$ and the first, second, and third terms denote the desired signal, residual ISI after the receive FDE & FDIC at the ith iteration, and noise, respectively.

2.4 Turbo Decoding

Turbo decoding is carried out using the bit LLRs obtained from $\hat{\mathbf{d}}^{(i)}$. The bit LLR, $\Lambda_n(x)$, associated with the xth bit of the nth data symbol in a block is computed as

$$\Lambda_n^{(i)}(x)\approx\frac{\left|\hat{d}^{(i)}(n)-d_{b_{n,x}=0}^{\min}\right|^2}{2\{\sigma^{(i)}\}^2}-\frac{\left|\hat{d}^{(i)}(n)-d_{b_{n,x}=1}^{\min}\right|^2}{2\{\sigma^{(i)}\}^2}, \tag{13}$$

where $x=0\sim\log_2 M-1$ and $n=0\sim N_c-1$ (M is the modulation level), $b_{n,x}$ denotes the xth bit of the nth data symbol and $d_{b_{n,x}=0(\text{or }b_{n,x}=1)}^{\min}$ denotes the most probable symbol that gives the minimum Euclidean distance from $\hat{d}^{(i)}(n)$ among all the candidate symbols with $b_{n,x}=0$ (or $b_{n,x}=1$). $2\{\sigma^{(i)}\}^2$ represents the variance of the noise plus the residual ISI at the ith iteration stage. The bit LLR sequence is then de-interleaved and de-punctured.

The turbo decoder considered in this paper comprises two maximum a *posteriori* probability (MAP) decoders that are connected via interleaver/de-interleaver each other [11]. The bit LLRs given by Eq. (13) is used by the decoders. In addition, the first MAP decoder uses the output LLRs from the second MAP decoder in the previous iteration as a *priori* information to compute the a *posteriori* LLRs. Then, the second MAP decoder computes the improved a *posteriori* LLRs using the output LLRs from the first MAP decoder as a *priori* information. The resultant bit LLR sequences are interleaved to generate the data symbol replicas. The bit LLR, associated with the xth bit of the nth data symbol of the data symbol block after the decoding, can be expressed as

$$\lambda_n^{(i)}(x)=\ln\frac{p^{(i)}(b_{n,x}=1)}{p^{(i)}(b_{n,x}=0)}, \tag{14}$$

where $p^{(i)}(b_{n,x}=1\,(or\,b_{n,x}=0))$ represents the a *posteriori* probability of $b_{n,x}=1$ (or $b_{n,x}=0$) after the decoding.

2.5 Symbol Replica Generation

The frequency-domain data symbol replica block $\tilde{\mathbf{D}}^{(i-1)}$ of Eq. (11) is obtained similar to [8]. The nth data symbol replica, $\tilde{d}^{(i-1)}(n)$, is given as

$$\tilde{d}^{(i-1)}(n)=\sum_{d\in Y}d\prod_{b_{n,x}\in d}p^{(i-1)}(b_{n,x}\,|\,\lambda_n^{(i-1)}(x)), \tag{15}$$

where d represents the candidate symbol having $b_{n,x}=0$ or 1 in the symbol set Y and $p^{(i-1)}(b_{n,x}=0$ and $1)$ are given by

$$\begin{cases}p^{(i-1)}(b_{n,x}=0\,|\,\lambda_n^{(i-1)}(x))=-\dfrac{1}{2}\tanh\left(\dfrac{\lambda_n^{(i-1)}(x)}{2}\right)+\dfrac{1}{2}\\[2ex] p^{(i-1)}(b_{n,x}=1\,|\,\lambda_n^{(i-1)}(x))=\dfrac{1}{2}\tanh\left(\dfrac{\lambda_n^{(i-1)}(x)}{2}\right)+\dfrac{1}{2},\end{cases} \tag{16}$$

since

$$p^{(i-1)}(b_{n,x}=0\,|\,\lambda_n^{(i-1)}(x))+p^{(i-1)}(b_{n,x}=1\,|\,\lambda_n^{(i-1)}(x))=1. \tag{17}$$

According to Eqs. (16) and (17), $\tilde{d}^{(i-1)}(n)$ is given as [8]

$$\tilde{d}^{(i-1)}(n)=\begin{cases}\dfrac{1}{\sqrt{2}}\left\{\tanh\left(\dfrac{\lambda_n^{(i-1)}(0)}{2}\right)+j\tanh\left(\dfrac{\lambda_n^{(i-1)}(1)}{2}\right)\right\}\text{ for QPSK},\\[2ex] \dfrac{1}{\sqrt{10}}\left\{\tanh\left(\dfrac{\lambda_n^{(i-1)}(0)}{2}\right)\left(2+\tanh\left(\dfrac{\lambda_n^{(i-1)}(1)}{2}\right)\right)\right\}\\[2ex] +\dfrac{j}{\sqrt{10}}\left\{\tanh\left(\dfrac{\lambda_n^{(i-1)}(2)}{2}\right)\left(2+\tanh\left(\dfrac{\lambda_n^{(i-1)}(3)}{2}\right)\right)\right\}\\ \qquad\qquad\qquad\text{for 16QAM}.\end{cases} \tag{18}$$

Using Eq. (18), the frequency-domain symbol replica is generated as $\tilde{\mathbf{D}}^{(i-1)}=\mathbf{F}\tilde{\mathbf{d}}^{(i-1)}$.

3. FDE WEIGHTS

3.1 Receive FDE Weight

The receive FDE weight $\mathbf{W}_r^{(i)}$ at the ith iteration stage is derived. A concatenation of the transmit FDE and the propagation channel is viewed as an equivalent channel. The error vector between $\hat{\mathbf{d}}^{(i)}$ and $\mathbf{d}$ is given as

$$\begin{aligned}\mathbf{e}^{(i)} &= \hat{\mathbf{d}}^{(i)} - \mathbf{d}/\sqrt{2E_s/T_s} \\ &= \{\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{H}\mathbf{W}_t\mathbf{F}-\mathbf{I}\}\{\mathbf{d}-\tilde{\mathbf{d}}^{(i-1)}\}+\left(\frac{2E_s}{T_s}\right)^{-\frac{1}{2}}\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{F}\mathbf{n}.\end{aligned} \tag{19}$$

The mean square error (MSE), $e^{(i)}$, is given as

$$e^{(i)} = tr[E(\mathbf{e}^{(i)}\{\mathbf{e}^{(i)}\}^H)]. \tag{20}$$

Using Eq. (20), MMSE receive FDE weight can be solved similar to [8] as

$$\mathbf{W}_r^{(i)} = \rho^{(i-1)}\mathbf{W}_t^H\mathbf{H}^H\{\rho^{(i-1)}\mathbf{H}\mathbf{W}_t\mathbf{W}_t^H\mathbf{H}^H+\gamma^{-1}\mathbf{I}\}^{-1}, \tag{21}$$

where $\gamma=(E_s/N_0)$ and

$$E(\{\mathbf{D}-\tilde{\mathbf{D}}^{(i-1)}\}\{\mathbf{D}-\tilde{\mathbf{D}}^{(i-1)}\}^H) = \rho^{(i-1)}\cdot\mathbf{I} \tag{22}$$

with [8]

$$\begin{aligned}\rho^{(i-1)} &= E[|D(k)-\tilde{D}^{(i-1)}(k)|^2] = E[|d(t)-\tilde{d}^{(i-1)}(t)|^2] \\ &\approx \begin{cases} 1-|\tilde{d}^{(i-1)}(t)|^2 & \text{for QPSK}, \\ \frac{4}{10}\tanh\left(\frac{\lambda_n^{(i-1)}(1)}{2}\right)+\frac{4}{10}\tanh\left(\frac{\lambda_n^{(i-1)}(3)}{2}\right) & \\ +1-|\tilde{d}^{(i-1)}(t)|^2 & \text{for 16QAM}. \end{cases}\end{aligned} \tag{23}$$

3.2 Transmit FDE Weight

From the previous subsection, it can be understood that the receive FDE weight is updated for each iteration stage, including $\rho^{(i-1)}$ which indicates the accuracy of the symbol replicas generated in the previous iteration. In contrast to the receiver side, transmit FDE weight cannot be updated. In [9], we derived the transmit FDE weight for the given receive FDE weight. However, the weight updating of receive FDE & FDIC makes it difficult to find the optimal transmit FDE weight. In this paper, to derive the transmit FDE weight, we introduce a virtual receiver having a receive FDE & FDIC into the transmitter. This virtual receiver is assumed to reduce the residual ISI by a factor of $1-\sqrt{\rho^{tx}}$, i.e., the received signal after the receive FDE & FDIC in the virtual receiver is assumed to be given as

$$\hat{\mathbf{D}}^{tx} = \sqrt{\frac{2E_s}{T_s}}\mathbf{W}_r^{tx}\mathbf{H}\mathbf{W}_t\mathbf{D}-\sqrt{\frac{2E_s}{T_s}}\left[1-\sqrt{\rho^{tx}}\right]\{\mathbf{W}_r^{tx}\mathbf{H}\mathbf{W}_t-\mathbf{I}\}\mathbf{D}+\mathbf{W}_r^{tx}\mathbf{N}, \tag{24}$$

where the second term represents subtraction of the residual ISI replica as in the receiver given by Eq. (10). Similar to Eq. (21), the receive FDE weight in the virtual receiver is derived as

$$\mathbf{W}_r^{tx} = \rho^{tx}\mathbf{W}_t^H\mathbf{H}^H\{\rho^{tx}\mathbf{H}\mathbf{W}_t\mathbf{W}_t^H\mathbf{H}^H+\gamma^{-1}\mathbf{I}\}^{-1}. \tag{25}$$

Replacing $\hat{\mathbf{d}}^{(i)}$ of Eq. (19) by $\hat{\mathbf{d}}^{tx}=\mathbf{F}^H\hat{\mathbf{D}}^{tx}$, the MSE for the given the receive FDE weight of the virtual receiver is given as

$$\begin{aligned}e^{tx} &= tr[\{\mathbf{W}_r^{tx}\mathbf{H}\mathbf{W}_t-\mathbf{I}\}\rho^{tx}\{\mathbf{W}_r^{tx}\mathbf{H}\mathbf{W}_t-\mathbf{I}\}^H]+\gamma^{-1}tr[\mathbf{W}_r^{tx}\{\mathbf{W}_r^{tx}\}^H] \\ &= \gamma^{-1}\rho^{tx}\cdot tr[(\rho^{tx}\mathbf{H}\mathbf{W}_t\mathbf{W}_t^H\mathbf{H}^H+\gamma^{-1}\mathbf{I})^{-1}].\end{aligned} \tag{26}$$

The optimal transmit FDE weight that minimizes e^{tx} in Eq. (26) can be obtained, similar to [9], by solving the convex optimization problem [12] under the power constraint given by Eq. (5) as (the derivation is omitted for the sake of brevity)

$$\mathbf{W}_t = diag\{\sqrt{|W_t(0)|^2},\ \ldots,\ \sqrt{|W_t(k)|^2},\ \ldots,\ \sqrt{|W_t(N_c-1)|^2}\} \tag{27}$$

with

$$|W_t(k)|^2 = \max\left\{\frac{1}{\mu}\frac{1}{\sqrt{\gamma}|H(k)|}-\frac{1}{\rho^{tx}}\frac{1}{\gamma|H(k)|^2},\ 0\right\}, \tag{28}$$

where μ is a constant determined to satisfy Eq. (5).

Equation (28) indicates that the transmit FDE weight depends on the value of ρ^{tx}. The value of ρ^{tx} indicates the reliability of the interference cancellation at the receiver, which is predicted by the transmitter. If the transmitter believes that the receiver can perfectly cancel the residual ISI, ρ^{tx} is set to be small. On the other hand, if the transmitter believes that the receiver cannot cancel the residual ISI at all, ρ^{tx} is set to be 1. The optimal ρ^{tx} may depend on the instantaneous channel condition, signal-to-noise power ratio, data modulation, the number of iterations of the receive FDE & FDIC and so on. Therefore, it is quite difficult to analytically find the optimal ρ^{tx} and hence, in this paper, by computer simulation, we find ρ^{tx} such that the average BER is minimized for the given average transmit E_s/N_0.

4. SIMULATION RESULTS

The BER performance of SC using the joint iterative Tx/Rx FDE & FDIC is evaluated by computer simulation. The channel is assumed to be an L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile. K=2048 with R=1/2 (coded) ~ 1 (uncoded) and log-MAP decoding is assumed. The number of iteration of the receiver is set to I=6. N_c=256, N_g=32 are considered. QPSK and 16QAM are assumed for data modulation.

Figures 2 and 3 compare the achievable average BER performances of the joint iterative Tx/Rx FDE & FDIC and the conventional iterative receive FDE & FDIC. The optimal values of ρ^{tx} are used for all the average E_b/N_0 of the joint iterative Tx/Rx FDE & FDIC. It is seen that the coded BER performance can be significantly improved by using the proposed scheme, with the help of channel coding gain, compared to that obtained by the conventional iterative receive FDE & FDIC. When QPSK data modulation is used, the joint iterative Tx/Rx FDE & FDIC can reduce the required E_b/N_0 for achieving BER=10^{-3} by about 1.0dB and 0.8dB compared to the conventional iterative receive FDE & FDIC for R=1/2 and R=3/4, respectively. When 16QAM data modulation is used, the joint iterative Tx/Rx FDE & FDIC can reduce the required E_b/N_0 by about 0.5dB and 1.0dB for R=1/2 and R=3/4, respectively. However, in the uncoded case, the BER performance improvement is smaller since the LLRs for computing the residual ISI replicas cannot be improved by the decoders. This is more evident when 16QAM data modulation is used and the joint iterative Tx/Rx FDE & FDIC provides almost the same BER performance as the conventional iterative receive FDE & FDIC.

5. CONCLUSION

In this paper, we proposed joint iterative Tx/Rx FDE & FDIC for SC block transmissions. The receive FDE weight and the residual ISI replica are updated in each iteration. To optimize the transmit FDE weight, we introduced a virtual receiver having a receive FDE & FDIC into the transmitter. In the virtual receiver at the transmitter, the same degree of the residual ISI cancellation is assumed to compute the transmit FDE weight, based on the MMSE criterion, matched to the receive FDE & FDIC. We evaluated the BER performance improvement by computer simulation. It was shown that the joint iterative Tx/Rx FDE & FDIC provides much better BER performance than the iterative receive FDE & FDIC.

In this paper, we assumed the perfect knowledge of CSI for computing the set of FDE weights. In [13], we investigated how the channel estimation (CE) error affects the BER performance of SC using joint Tx/Rx FDE. However, the performance of joint iterative Tx/Rx FDE & FDIC with imperfect CE has not been studied yet. This is left as an important future work.

6. REFERENCES

[1] S. Sesia, I. Toufik, and M. Baker, *LTE, The UMTS Long Term Evolution: From Theory to Practice*, Wiley, 2009.

[2] J. G. Proakis, *Digital Communications*, 4th edition, McGraw-Hill, 2001.

[3] D. Falconer, S. L. Ariyavistakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," IEEE Commun., Mag., Vol.40, No.4, pp.58-66, April 2002.

[4] F. Adachi, D. Garg, S. Takaoka, and K. Takeda, "Broadband CDMA techniques," IEEE Wireless Commun., Mag., Vol.12, No.2, pp.8-18, April 2005.

[5] F. Adachi, H. Tomeba, and K. Takeda, "Introduction of frequency-domain signal processing to broadband single-carrier transmissions in a wireless channel," IEICE Trans. Commun., Vol. E92-B. No. 9, pp. 2789-2808, Sept. 2009.

[6] S. Tomasin and N. Benvenuto,"Iterative design and detection of a DFE in the frequency domain," IEEE Trans. Commun., Vol. 53, No. 11, pp. 1867-1875, Nov. 2005.

[7] R. Dinis, P. Silva, and T. Araujo, "Joint turbo equalization and cancelation of nonliear distortion effects in MC-CDMA signals," Proc. International Conference on Signal and Image Processing, Honolulu, Hawaii, USA, 2006.

[8] K. Takeda and F. Adachi, "Iterative frequency-domain inter-chip interference cancellation for DS-CDMA," Proc. IEEE 8th Wireless Personal Multimedia Communications (WPMC), Aalborg, Denmark, pp. 759-763, Sept. 2005.

[9] K. Takeda, H. Tomeba, and F. Adachi, "Multicode DS-CDMA with joint transmit/receive frequency-domain equalization," Proc. IEEE 20th Personal Indoor and Mobile Radio Communications (PIMRC), Tokyo, Japan, Sept. 2009.

[10] C. Berrou, A. Glavieux, and P. Thitimajshima, "Near Shannon limit error-correcting coding and decoding: Turbo-codes. 1", Proc. IEEE International Conference on Communications, Vol. 2, pp. 1064-1070, Geneva, Switzerland, May 1993.

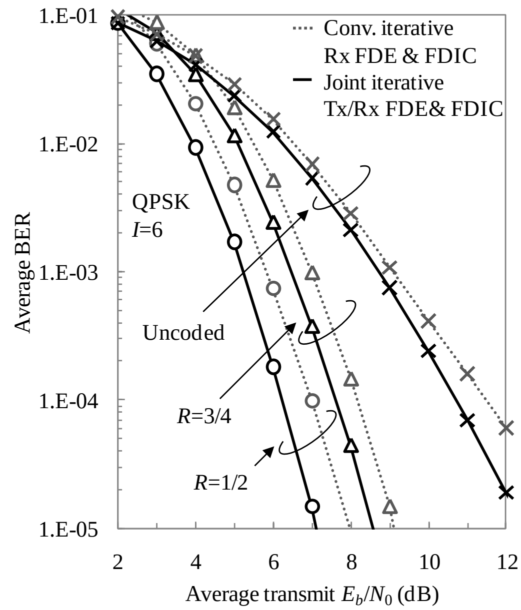

Figure 2. BER performance with QPSK.

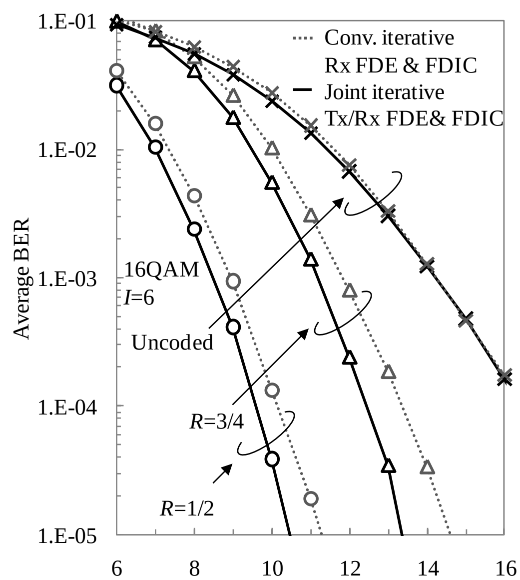

Figure 3. BER performance with 16QAM.

[11] D. Raphaeli and Y. Zarai, "Combined turbo equalization and turbo decoding," Proc. IEEE Global Telecommunications Conference (GLOBECOM), Vol. 2, No. 3-8, pp. 639-643, Nov. 1997.

[12] S. Boyd and L. Vandenverghe, *Convex optimization*, Cambridge University Press, 2004.

[13] K. Takeda and F. Adachi, "Impact Of Channel Estimation Error On The Single-carrier Joint Tx/Rx MMSE-FDE," to be presented at IEEE VTS APWCS2010, Kaohsiung, Taiwan, 20-21 May 2010.