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Research Article

Block Iterative/Adaptive Frequency-Domain Channel Estimation for Cyclic-Prefixed Single-Carrier Broadband Wireless Systems

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This paper presents a new block iterative/adaptive frequency-domain channel estimation scheme, in which a channel frequency response (CFR) is estimated iteratively by the proposed weighted element-wise block adaptive frequency-domain channel estimation (WEB-CE) scheme using the soft information obtained by a soft-input soft-output (SISO) decoder. In the WEB-CE, an equalizer coefficient is calculated by minimizing a weighted conditional squared-norm of the a posteriori error vector with respect to its correction term. Simulation results verify the superiority of the WEB-CE in a time-varying typical urban (TU) channel.

1. INTRODUCTION

Cyclic-prefixed single-carrier frequency-domain equalization (SC-FDE) has received enormous attention in recent years because of its efficient implementation and low peak-to-average power ratio (PAPR) characteristics over broadband wireless channels. On the other hand, it is noticed that most works have been performed with the assumption that a channel frequency response (CFR) is perfectly known to the receiver [1, 2]. In practice, this assumption is not valid since a real channel is unknown and time-varying. This, it highlights the need for the precise estimation of CFR. In [3, 4], an adaptive channel estimation scheme using the hard information from the decision was presented. In [5], an iterative (nonadaptive) channel estimation scheme using the soft information obtained from a soft-input soft-output (SISO) decoder was presented. Recently, an iterative/adaptive scheme which performs a channel estimation iteratively by employing an adaptive algorithm using the soft information was proposed in [6]. This work concludes that the iterative/adaptive approach is adequate to support a good channel tracking performance over time-varying channels. However, the scheme has not been applied to the SC-FDE. In this correspondence, an iterative/adaptive channel estimation scheme for the SC-FDE is studied. Moreover, a new block-type channel estimation scheme is studied in order to provide a better channel tracking performance.

In this paper, we propose a block iterative/adaptive CFR estimation scheme, in which a weighted element-wise block adaptive frequency-domain channel estimation (WEB-CE) using the soft information obtained from the SISO decoder is presented. It is found that the WEB-CE has a flexibility over the element-wise block length as compared to an approximated recursive least square (RLS)-CE of [6] by applying a weighted conditional least-square (LS) criterion formulated with the *a posteriori error* vector [7]. Moreover, mean square error (MSE) of the WEB-CE with respect to the element-wise block length is analyzed. Simulation results show that the WEB-CE yields good performance in a typical urban (TU) channel as the iteration number and element-wise block length increase.

The paper is organized as follows. The next section describes the system model. In Section 3, the derivation procedure and property of the proposed CFR estimation scheme are discussed. Simulation results are discussed in Section 4, and conclusions are drawn in Section 5.

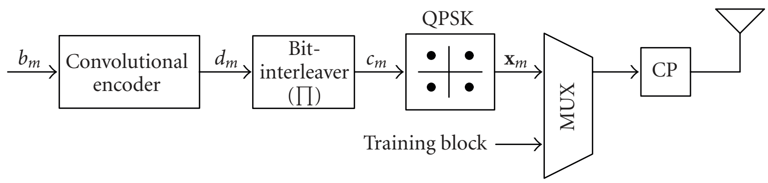

Figure 1: Block diagram of SC-FDE transmitter.

Notation

$E\{\cdot\}$ and $(\cdot)^T$ denote an expectation and a transpose operator, respectively. $(\cdot)^*$ denotes complex conjugates of a symbol, vector, and matrix. $\mathbf{I}_N$, $\mathbf{0}_N$, and $\text{diag}(\cdot)$ denote the identity matrix of size N, zero matrix of size N, and diagonal matrix, respectively.

2. SC-FDE SYSTEM MODEL

2.1. SC-FDE transmitter

The transmitter of SC-FDE is shown in Figure 1. The blocks of data bits $b_m \in \{0,1\}$ are encoded by the convolutional encoder. The code bits d_m are interleaved by an interleaver $\Pi(\cdot)$, $c_m = \Pi(d_m)$. QPSK mapper maps a pair of input bits to a symbol x_m from the symbol alphabet $\mathcal{S}$. (In this paper, a constant-amplitude PSK constellation is assumed.) Then, a block of signals $\mathbf{x}_n = [x_{n,0} \ \cdots \ x_{n,N-2} \ \ x_{n,N-1}]^T$ is structured with block length of N and a cyclic-prefix (CP) is inserted between blocks to prevent an interblock interference (IBI) at block time n. A training block (TB) that has block length of N is inserted every data frame that consists of data blocks in order to initialize the channel coefficients and track the channel variations.

2.2. SC-FDE receiver

Suppose that an overall channel impulse response (CIR) containing the combined effects of transmit and receive filters, multipath fading and sampling is $\mathbf{h}_n = [h_{n,0} \ \cdots \ h_{n,L-2} \ \ h_{n,L-1}]^T$, where L is the total length of the CIR, at a block time n, and it is constant within one block. It is also supposed that the CP length is equal or larger than $L-1$. At the receiver, discarding the CP, the time-domain received block $\mathbf{y}_n$ can be expressed in a matrix form as

$$\mathbf{y}_n = \mathbf{H}_n^c \mathbf{x}_n + \mathbf{w}_n, \tag{1}$$

where $\mathbf{w}_n = [w_{n,0} \ \cdots \ w_{n,N-2} \ \ w_{n,N-1}]^T$ is an additive white noise vector, and $\mathbf{H}_n^c$ is a circular matrix. It is assumed that $E\{\mathbf{x}_n\mathbf{x}_n^*\} = \sigma_x^2\mathbf{I}_N$, $E\{\mathbf{x}_n\mathbf{w}_n^*\} = \mathbf{0}_N$, and $E\{\mathbf{w}_n\mathbf{w}_n^*\} = \sigma_w^2\mathbf{I}_N$, where σ_x^2 and σ_w^2 denote the signal power and noise variance, respectively. Then, the discrete-time Fourier transform (DFT) matrix $\mathbf{Q} = (1/\sqrt{N})e^{-j2\pi il/N}$ (for $0 \le i$, $l \le N-1$) yields the frequency-domain block as follows:

$$\mathbf{Y}_n = \mathbf{H}_n \mathbf{X}_n + \mathbf{W}_n, \tag{2}$$

where $\mathbf{Y}_n = \mathbf{Q}\mathbf{y}_n = [Y_n[0] \ \cdots \ Y_n[N-2] \ \ Y_n[N-1]]^T$, $\mathbf{X}_n = \mathbf{Q}\mathbf{x}_n = [X_n[0] \ \cdots \ X_n[N-2] \ \ X_n[N-1]]^T$, $\mathbf{W}_n = \mathbf{Q}\mathbf{w}_n = [W_n[0] \ \cdots \ W_n[N-2] \ \ W_n[N-1]]^T$, and $\mathbf{H}_n = \mathbf{Q}\mathbf{H}_n^c\mathbf{Q}^* = \text{diag}(H_n[0],\ldots,H_n[N-2],H_n[N-1])$, where $H_n[i]$ represents the channel frequency response at the ith frequency bin.

From (2), the frequency-domain estimate of the transmitted data block, $\hat{\mathbf{X}}_n = [\hat{X}_n[0] \ \cdots \ \hat{X}_n[N-2] \ \ \hat{X}_n[N-1]]^T$, is performed by employing FDE at a block time n. In this paper, a zero-forcing FDE (ZF-FDE) is considered. (Note that the minimum-mean-square error (MMSE) equalization can be used in place of the ZF equalization.) The received data at the lth frequency bin is described as

$$\hat{X}_n[l] = \frac{Y_n[l]}{H_n[l]}, \quad 0 \le l \le N-1. \tag{3}$$

After the inverse DFT (IDFT) operation, an extrinsic log-likelihood ratio (LLR) of a coded bit is computed [8]. The extrinsic LLR value is de-interleaved and fed to an SISO decoder. Then, the SISO decoder outputs the a priori LLR, which is interleaved to compute the soft information, that is, a priori mean vector $E\{\mathbf{x}_n\} = [\overline{x}_{n,0} \ \cdots \ \overline{x}_{n,N-2} \ \ \overline{x}_{n,N-1}]^T$, in which $\overline{x}_{n,i}$ is defined as [8]

$$\overline{x}_{n,i} = \sum_{s\in\mathcal{S}} s \cdot P(x_{n,i} = s), \tag{4}$$

where s denotes a symbol $\mathcal{S}$. Correspondingly, the DFT of the soft information such as the a priori mean vector is given by

$$\begin{aligned} \overline{\mathbf{X}}_n &= \mathbf{Q}E\{\mathbf{x}_n\} \\ &= \left[\overline{X}_n[0] \ \cdots \ \overline{X}_n[N-2] \ \ \overline{X}_n[N-1]\right]^T. \end{aligned} \tag{5}$$

The soft information is used to estimate $H_n[l]$ in (3).

3. ITERATIVE/ADAPTIVE CFR ESTIMATION SCHEME

3.1. WEB-CE using the soft information

The coefficient update algorithm of the WEB-CE at lth frequency bin is expressed as [7]

$$\hat{H}_n[l] = \hat{H}_{n-1}[l] + \Delta\hat{H}_n[l], \tag{6}$$

where $\Delta\hat{H}_n[l]$ is the correction term of $\hat{H}_{n-1}[l]$. From (2), we define $1 \times B$ input vectors $\hat{H}_n[l]$ at a block time n as follows:

$$\begin{aligned}\mathbf{Y}_n[l] &= \left[Y_n[l] \quad Y_{n-1}[l] \quad \cdots \quad Y_{n-B+1}[l]\right],\\ \mathbf{X}_n[l] &= \left[X_n[l] \quad X_{n-1}[l] \quad \cdots \quad X_{n-B+1}[l]\right],\\ \mathbf{W}_n[l] &= \left[W_n[l] \quad W_{n-1}[l] \quad \cdots \quad W_{n-B+1}[l]\right],\end{aligned} \tag{7}$$

where B is the element-wise block length of the WEB-CE. In (7), it is noticed that $\mathbf{X}_n[l]$ may actually not be known to the receiver, thus $\mathbf{X}_n[l]$ is alternatively defined as [6]

$$\mathbf{X}_n[l] = \overline{\mathbf{X}}_n[l] + \mathbf{V}_n[l], \tag{8}$$

where $\overline{\mathbf{X}}_n[l] = [\overline{X}_n[l] \quad \overline{X}_{n-1}[l] \quad \cdots \quad \overline{X}_{n-B+1}[l]]$ and $\mathbf{V}_n[l] = [V_n[l] \quad V_{n-1}[l] \quad \cdots \quad V_{n-B+1}[l]]$, whose element $V_{n-k}[l]$ has zero mean and variance given by

$$E\{V_{n-k}[l]V^*_{n-k}[l]\} = E\{X_{n-k}[l]X^*_{n-k}[l]\} - \overline{X}_{n-k}[l]\overline{X}^*_{n-k}[l]. \tag{9}$$

In addition, it is assumed that $V_{n-k}[l]$ is independent of $\overline{X}_{n-k}[l]$. Furthermore, we define $1 \times B$ *a priori error* vector $\mathbf{e}_n[l]$ and *a posteriori error* vector $\overline{\mathbf{e}}_n[l]$ as follows [7]:

$$\begin{aligned}\mathbf{e}_n[l] &= \mathbf{Y}_n[l] - \hat{H}_{n-1}[l]\mathbf{X}_n[l],\\ \overline{\mathbf{e}}_n[l] &= \mathbf{e}_n[l] - \Delta\hat{H}_n[l]\mathbf{X}_n[l].\end{aligned} \tag{10}$$

From (7)–(10), *a posteriori error* vector-based weighted conditional LS criterion is defined as [6, 7]

$$J_n[l] = E\{\overline{\mathbf{e}}_n[l]\mathbf{F}\overline{\mathbf{e}}^*_n[l] \mid \mathbf{Y}_n[l], \hat{H}_{n-1}[l], \mathbf{X}_n[l]\}, \tag{11}$$

where a squared norm of $\overline{\mathbf{e}}_n[l]$ is attenuated geometrically by $B \times B$ diagonal matrix $\mathbf{F} = \mathrm{diag}(1, \ldots, \beta^{B-1})$ with the forgetting factor β that satisfies $0 < \beta \le 1$. Using (8) and (10), (11) can be rewritten as

$$\begin{aligned}J_n[l] &= E\left\{\overline{\mathbf{e}}_n[l]\mathbf{F}\overline{\mathbf{e}}^*_n[l] \mid \underbrace{\mathbf{e}_n[l], \Delta\hat{H}_n[l], \mathbf{V}_n[l], \overline{\mathbf{X}}_n[l]}_{G_n}\right\}\\ &= E\{\mathbf{e}_n[l]\mathbf{F}\mathbf{e}^*_n[l] \mid G_n\}\\ &\quad + E\{\Delta\hat{H}_n[l]\mathbf{X}_n[l]\mathbf{F}\mathbf{X}^*_n[l]\Delta\hat{H}^*_n[l] \mid G_n\}\\ &\quad - E\{\Delta\hat{H}_n[l]\mathbf{X}_n[l]\mathbf{F}\mathbf{e}^*_n[l] \mid G_n\}\\ &\quad - E\{\mathbf{e}_n[l]\mathbf{F}\mathbf{X}^*_n[l]\Delta\hat{H}^*_n[l] \mid G_n\}.\end{aligned} \tag{12}$$

In (12), using (8), the second term is computed as

$$\begin{aligned}&E\{\Delta\hat{H}_n[l]\mathbf{X}_n[l]\mathbf{F}\mathbf{X}^*_n[l]\Delta\hat{H}^*_n[l] \mid G_n\}\\ &= \Delta\hat{H}_n[l]\underbrace{E\{(\overline{\mathbf{X}}_n[l] + \mathbf{V}_n[l])\mathbf{F}(\overline{\mathbf{X}}^*_n[l] + \mathbf{V}^*_n[l])\}}_{P^s_n}\Delta\hat{H}^*_n[l]\\ &= \Delta\hat{H}_n[l]P^s_n\Delta\hat{H}^*_n[l].\end{aligned} \tag{13}$$

Using (9), P^s_n can be expressed as

$$\begin{aligned}P^s_n &= \overline{\mathbf{X}}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l] + E\{\mathbf{V}_n[l]\mathbf{F}\mathbf{V}^*_n[l]\}\\ &= \sum_{k=0}^{B-1}\beta^k(\overline{X}_{n-k}[l]\overline{X}^*_{n-k}[l] + E\{V_{n-k}[l]V^*_{n-k}[l]\})\\ &= \sum_{k=0}^{B-1}\beta^k\mathbf{q}_l E\{\mathbf{x}_{n-k}\mathbf{x}^*_{n-k}\}\mathbf{q}^*_l\\ &= \sigma^2_x\left(\frac{1-\beta^B}{1-\beta}\right),\end{aligned} \tag{14}$$

where $\mathbf{q}_l$ denotes the lth row vector of $\mathbf{Q}$. In (12), using (8) and (10), the third term is given by

$$\begin{aligned}&E\{\Delta\hat{H}_n[l]\mathbf{X}_n[l]\mathbf{F}\mathbf{e}^*_n[l] \mid G_n\}\\ &\quad = \Delta\hat{H}_n[l](\overline{\mathbf{X}}_n[l]\mathbf{F}\mathbf{Y}^*_n[l] + E\{\mathbf{V}_n[l]\mathbf{F}\mathbf{Y}^*_n[l]\} - P^s_n\hat{H}^*_{n-1}[l]).\end{aligned} \tag{15}$$

Furthermore, assuming $H_n \simeq \hat{H}_{n-1}$ [6], $E\{\mathbf{V}_n[l]\mathbf{F}\mathbf{Y}^*_n[l]\}$ in (15) can be expressed as

$$\begin{aligned}E\{\mathbf{V}_n[l]\mathbf{F}\mathbf{Y}^*_n[l]\} &= E\{\mathbf{V}_n[l]\mathbf{F}(\mathbf{X}^*_n[l]H^*_n[l] + \mathbf{W}^*_n[l])\}\\ &= E\{\mathbf{V}_n[l]\mathbf{F}\mathbf{V}^*_n[l]\}\hat{H}^*_{n-1}[l].\end{aligned} \tag{16}$$

Then, (15) can be rewritten as

$$\begin{aligned}&E\{\Delta\hat{H}_n[l]\mathbf{X}_n[l]\mathbf{F}\mathbf{e}^*_n[l] \mid G_n\}\\ &\quad = \Delta\hat{H}_n[l](\overline{\mathbf{X}}_n[l]\mathbf{F}\mathbf{Y}^*_n[l] - \overline{\mathbf{X}}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l]\hat{H}^*_{n-1}[l]).\end{aligned} \tag{17}$$

The computation of the fourth term in (12) can be done similarly as that of the third term as follows:

$$\begin{aligned}&E\{\mathbf{e}_n[l]\mathbf{F}\mathbf{X}^*_n[l]\Delta\hat{H}^*_n[l] \mid G_n\}\\ &\quad = (\mathbf{Y}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l] - \hat{H}_{n-1}[l]\overline{\mathbf{X}}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l])\Delta\hat{H}^*_n[l].\end{aligned} \tag{18}$$

Finally, substituting the results of (13), (14), (17), and (18) into (12), and then minimizing $J_n[l]$ with respect to $\Delta\hat{H}^*_n[l]$, $\Delta\hat{H}_n$ in (6) is computed:

$$\begin{aligned}\frac{\partial J_n[l]}{\partial\Delta\hat{H}^*_n[l]} &= \Delta\hat{H}_n[l]P^s_n - \mathbf{Y}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l] - \hat{H}_{n-1}[l]\overline{\mathbf{X}}_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l]\\ &= \Delta\hat{H}_n[l]P^s_n - (\mathbf{Y}_n[l] - \hat{H}_{n-1}[l]\overline{\mathbf{X}}_n[l])\mathbf{F}\overline{\mathbf{X}}^*_n[l]\\ &= \Delta\hat{H}_n[l]P^s_n - \mathbf{e}^s_n[l]\mathbf{F}\overline{\mathbf{X}}^*_n[l]\\ &= 0,\end{aligned} \tag{19}$$

where $1 \times B$ *a priori soft error* vector $\mathbf{e}^s_n[l]$ is defined as

$$\mathbf{e}^s_n[l] = \mathbf{Y}_n[l] - \hat{H}_{n-1}[l]\overline{\mathbf{X}}_n[l]. \tag{20}$$

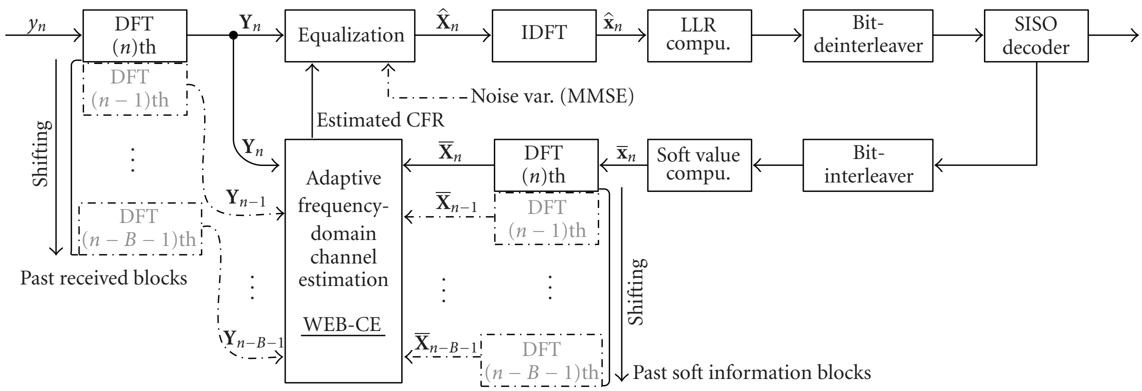

Figure 2: Iterative receive structure employing the WEB-CE when the iteration number is greater than one. Dotted blocks correspond to registers (or buffers) preserving the past information.

Table 1: Computational complexity of approximated RLS-CE and WEB-CE algorithms.

Algorithm	Overall number of complex Mul./Add.
Approximated RLS-CE	$N\log_2 N + 3N$ (Mul.) $2N\log_2 N + 2N$ (Add.)
WEB-CE	$N\log_2 N + 3BN$ (Mul.) $2N\log_2 N + 2BN$ (Add.)

From (19), $\Delta\widehat{H}_n[l]$ is given by

$$\Delta\widehat{H}_n[l] = \mathbf{e}_n^s[l]\mathbf{F}\overline{\mathbf{X}}_n^*[l](P_n^s)^{-1}. \quad (21)$$

Consequently, substituting (21) into (6), we can formulate WEB-CE algorithm as follows:

$$\widehat{H}_n[l] = \widehat{H}_{n-1}[l] + \mathbf{e}_n^s[l]\mathbf{F}\overline{\mathbf{X}}_n^*[l](P_n^s)^{-1}. \quad (22)$$

The overall iterative receive structure employing the WEB-CE is depicted in Figure 2.

3.2. Properties of WEB-CE

(1) *Flexibility*: In (14), (21), and (22), assuming $B = 1$, it is found that the WEB-CE is similar to the approximated RLS-CE (or least mean square (LMS)-CE) of [6] defined as

$$\widehat{H}_n^{\mathrm{RLS}}[l] = \widehat{H}_{n-1}^{\mathrm{RLS}}[l] + e_n^{\mathrm{RLS},s}[l]\overline{X}_n^*[l](P_n^{\mathrm{RLS},s})^{-1}, \quad (23)$$

where $(P_n^{\mathrm{RLS},s})^{-1} \approx (1-\beta)/\sigma_x^2$, and $e_n^{\mathrm{RLS},s}[l] = Y_n[l] - \widehat{H}_{n-1}^{\mathrm{RLS}}[l]\overline{X}_n[l]$. As a result, we can see that the WEB-CE has flexibility over the block length B as compared to the approximated RLS-CE.

(2) *Computational complexity*: The computational complexity of the WEB-CE and the approximated RLS-CE are compared with respect to a complex operator. We only focus on the required computation number of the channel estimators, that is, (22) and (23). It is assumed that the DFT operator requires $(N/2)\log_2 N$ complex multiplications (Mul.) and $N\log_2 N$ complex additions (Add.) per input symbols, that is, $\overline{\mathbf{X}}_n$ and $\mathbf{Y}_n$. It is noticed in Figure 2 that the dotted DFT blocks would not be considered in computing the computational complexity because these correspond to the register (or buffer). In the WEB-CE, the following steps are necessary for each block length of B.

(i) $\mathbf{e}_n^s[l]$ in (22) requires B complex multiplications and B complex additions, respectively, when $\mathbf{Y}_n[l]$ and $\overline{\mathbf{X}}_n[l]$ are given. Considering the block length of N, this step requires $(N\log_2 N + BN)$ complex multiplications and $(2N\log_2 N + BN)$ complex additions.

(ii) (22) requires $2B$ complex multiplications and B complex additions, respectively, where the term $(P_n^s)^{-1}$ is a constant from (14). Therefore, this step requires $2BN$ complex multiplications and BN complex additions.

The computational complexity of the approximated RLS-CE can be computed identically. As a result, the overall computational complexities of the WEB-CE and the approximated RLS-CE are shown in Table 1, which indicates that the computational load involved in the WEB-CE is linearly proportional to B.

(3) *MSE analysis*: The MSE $\varepsilon_n[l]$ of the WEB-CE with respect to B is analyzed in the Appendix. From (A.7), the MSE is given by

$$\varepsilon_n[l] = \sigma_w^2\frac{(1-\beta)(1+\beta^B)}{(1+\beta)(1-\beta^B)}. \quad (24)$$

When $\sigma_w^2 = 0.1, 0.0316$ and 0.01, the MSE (dB) curves versus β and B is depicted in Figure 3, where it is found that the MSE decreases exponentially as B increases for a given β.

4. SIMULATION RESULTS

We evaluate the performance of SC-FDE employing WEB-CE in the slow time-varying channels [3]. In the simulation, a recursive systematic convolutional encoder with a generator $G = [7\ 5]$ and BCJR SISO decoding algorithm were employed. S-random interleaver with a construction $S = 0.5\sqrt{0.5\cdot I}$ was used, where $I = 1024$ is a block size of the

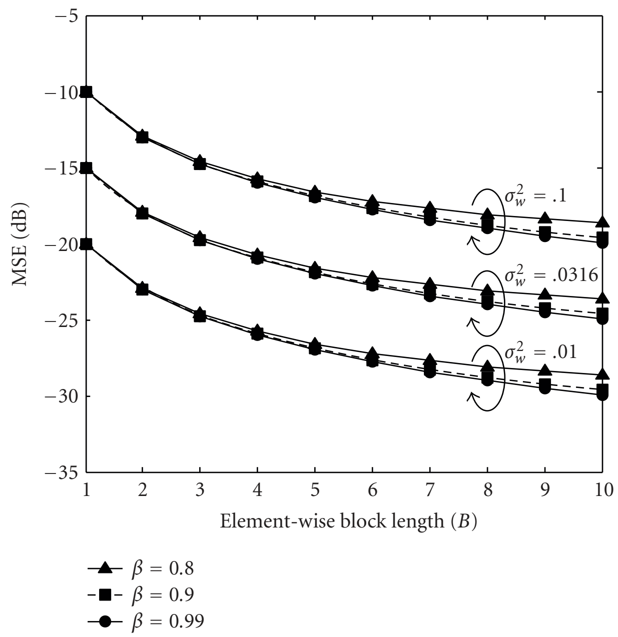

FIGURE 3: The MSE (dB) curves versus β and B when $\sigma_w^2 = 0.1, 0.0316$ and 0.01.

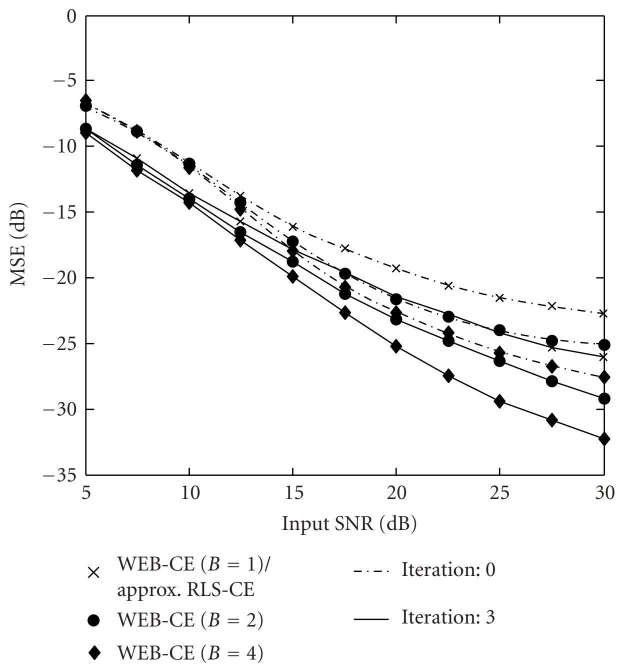

FIGURE 5: MSE performance of the WEB-CE ($B = 1$)/approx. RLS-CE, and WEB-CE ($B = 2, 4$) when $f_d = 5$ Hz.

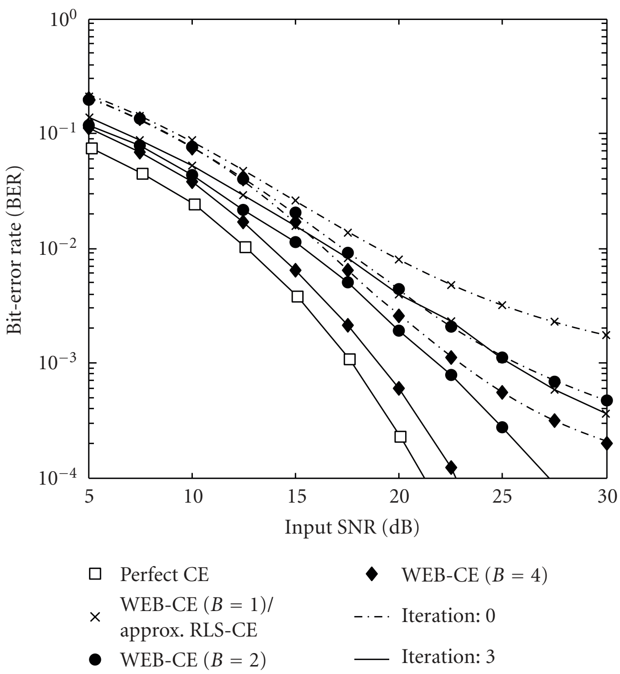

FIGURE 4: BER performance of ZF-FDE with the WEB-CE ($B = 1$)/approx. RLS-CE, and WEB-CE ($B = 2, 4$) when $f_d = 5$ Hz.

interleaver [8]. One block consists of 64 symbols ($N = 64$), where the symbol duration is $T_s = 3.69\,\mu$s. A normalized root mean square (rms) delay spread $\tau = 0.2886$ and radio channel memory of 3 (sample delay) are assumed for the TU channel as in Jake's model [7]. The CP length of 3 was set to eliminate the IBI. A data frame consists of 16 data blocks and one TS block. In the simulations, the bit error rate (BER) was measured to evaluate the performance of the ZF-FDE with respect to B of the WEB-CE. In addition, the MSE of channel estimate, that is, MSE between the perfect CFR and the estimated CFR by the WEB-CE, was also measured.

Figure 4 depicts the BER curve of the ZF-FDE employing the WEB-CE ($B = 1, 2, 4$), where the iteration number is 0 and 3. In this figure, the maximum Doppler shift was set to $f_d = 5$ Hz and a parameter $\beta = 0.85$ was chosen through a simulation in order to allow a good tracking for the WEB-CE. In order to provide a reference BER, a perfect channel condition was also considered. Here, the approximated RLS-CE could be regarded equivalently as the WEB-CE ($B = 1$). It is readily shown that the performance of the ZF-FDE is improved as the iteration number increases. In particular, it is noticed that the WEB-CE ($B = 2, 4$) yields a significant improvement of performance as compared to the WEB-CE ($B = 1$), which is caused by the gain obtained from B (see Figure 3). It is also noticed that the error propagation due to the inaccurate past soft information [7] can be reduced as the iteration number increases.

Figure 5 depicts the corresponding MSE curve of Figure 4. It is shown that the WEB-CE ($B = 2, 4$) yields better MSE performance than the WEB-CE ($B = 1$).

Figures 6 and 7 depict the BER and MSE curves when $f_d = 10$ Hz, where the simulation parameters are the same as in Figures 4 and 5. From these figures, we can also see that the WEB-CE yields superior performance as the iteration number and B increase in the presence of a relatively fast fading. Furthermore, it is noticed from [9] that the TS blocks insertion, more often and using smaller N, can help improve the BER performance as the Doppler shift increases.

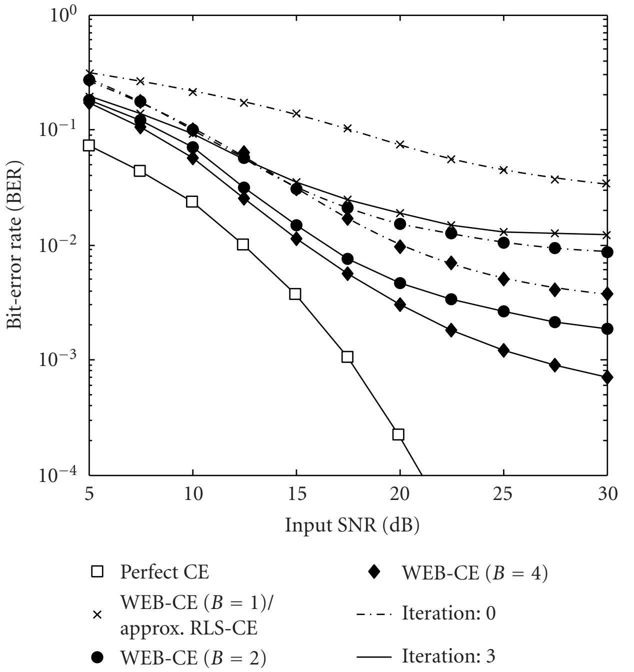

Figure 6: BER performance of ZF-FDE with the WEB-CE ($B = 1$)/approx. RLS-CE, and WEB-CE ($B = 2, 4$) when $f_d = 10$ Hz.

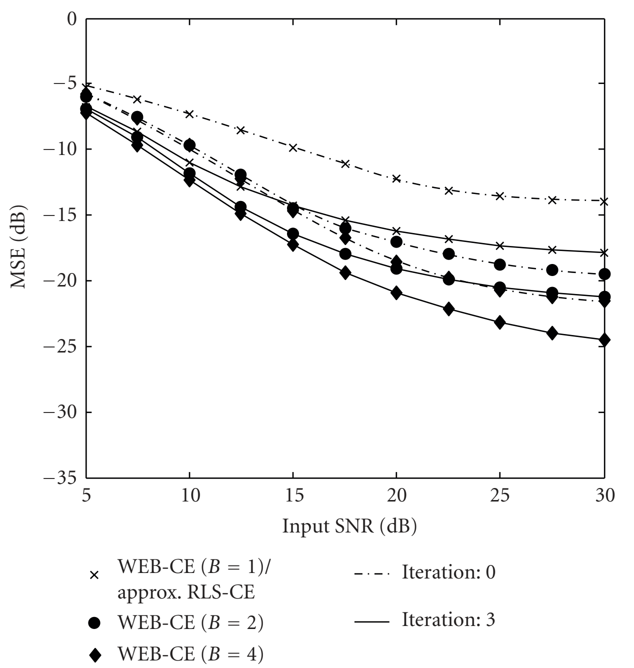

Figure 7: MSE performance of the WEB-CE ($B = 1$)/approx. RLS-CE, and WEB-CE ($B = 2, 4$) when $f_d = 10$ Hz.

Figure 8 depicts the performance comparison of the ZF-FDE with WEB-CE ($B = 4$) and the MMSE-FDE with WEB-CE ($B = 4$), where it is assumed that a noise variance is known to the receiver. Figure 8 shows that the MMSE-FDE provides the lower BER performance than the ZF-FDE.

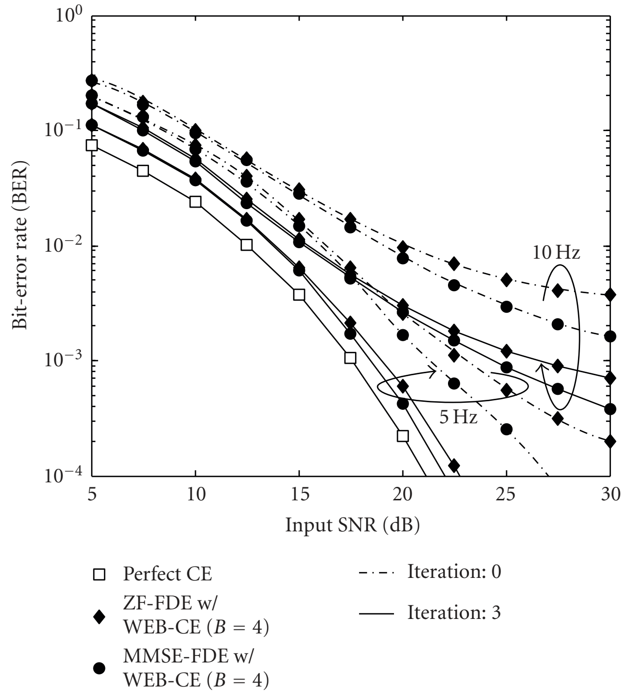

Figure 8: BER performance comparison of ZF-FDE w/WEB-CE ($B = 4$) and MMSE-FDE w/WEB-CE ($B = 4$) when $f_d = 5$ and 10 Hz.

5. CONCLUSIONS

In this paper, we proposed the novel block iterative/adaptive frequency-domain channel estimation named as the WEB-CE that utilizes the soft information in the time-varying channel. In the WEB-CE, the correction term of the coefficient was calculated by minimizing the weighted conditional *a posteriori error* vector-based LS criterion at each block iteration. It was found that the MSE of the WEB-CE decreases as B increases, which gives flexibility as compared to the approximated RLS-CE. From the simulation results, it was shown that the WEB-CE ($B > 1$) would be a good choice for the channel estimation scheme of the SC-FDE receiver in time-varying channels.

APPENDIX

MSE ANALYSIS OF THE WEB-CE

In (22), assuming that the channel is stationary, $\mathbf{X}_n$ is perfectly known to the receiver and the WEB-CE is on the steady state, we analyze the MSE $\varepsilon_n[l]$ defined as

$$\begin{aligned}\varepsilon_n[l] &= E\left\{\left(\widehat{H}_n^*[l] - H_n^*[l]\right)\underbrace{\left(\widehat{H}_n[l] - H_n[l]\right)}_{\Delta H_n}\right\} \\ &= E\{\Delta H_n^* \Delta H_n\},\end{aligned} \quad \text{(A.1)}$$

where $\widehat{H}_n[l]$ is defined as

$$\widehat{H}_n[l] = \widehat{H}_{n-1}[l] + \mathbf{e}_n[l]\mathbf{F}\mathbf{X}_n^*[l]P_n^{-1} \quad \text{(A.2)}$$

with $P_n = \mathbf{X}_n[l]\mathbf{F}\mathbf{X}_n^*[l] = \sum_k \beta^k \mathbf{q}_l \mathbf{x}_{n-k}\mathbf{x}_{n-k}^* \mathbf{q}_l^*$, and $\mathbf{e}_n[l] = \mathbf{Y}_n[l] - \widehat{H}_{n-1}[l]\mathbf{X}_n[l]$. Premultiplying both sides of (A.2) by P_n and using $\mathbf{e}_n[l]$, (A.2) can be rewritten as

$$\widehat{H}_n[l] = \mathbf{Y}_n[l]\mathbf{F}\mathbf{X}_n^*[l]P_n^{-1}. \tag{A.3}$$

Substituting $\mathbf{Y}_n[l] = H_n[l]\mathbf{X}_n[l] + \mathbf{W}_n[l]$ into (A.3), ΔH_n is expressed as

$$\Delta H_n = \mathbf{W}_n[l]\mathbf{F}\mathbf{X}_n^*[l]P_n^{-1}. \tag{A.4}$$

Correspondingly, the MSE $E\{\Delta H_n^* \Delta H_n\}$ is computed as

$$E\{\Delta H_n^* \Delta H_n\} = \sigma_w^2 P_n^{-1} \underbrace{E\{\mathbf{X}_n[l]\mathbf{F}^2\mathbf{X}_n^*[l]\}}_{P_n'} P_n^{-1}. \tag{A.5}$$

Here, assuming $E\{\mathbf{x}_{n-k}\mathbf{x}_{n-k}^*\} = \mathbf{I}_N$ for all k in (1), P_n' and P_n are given by

$$\begin{aligned} P_n' &= \frac{(1-\beta^{2B})}{(1-\beta^2)}, \\ P_n &= \frac{(1-\beta^{B})}{(1-\beta)}. \end{aligned} \tag{A.6}$$

Finally, substituting (A.6) into (A.5), the MSE of the WEB-CE is calculated as

$$\varepsilon_n[l] = \sigma_w^2 \frac{(1-\beta)(1+\beta^B)}{(1+\beta)(1-\beta^B)}. \tag{A.7}$$

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