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Enhanced MMSE-FDE method for TS-OFDM under higher mobile environments

Pongsathorn Reangsuntea[1a)], Pisit Boonsrimuang[2], Kousuke Sanada[1], Kazuo Mori[1], and Hideo Kobayashi[1]

[1] *Department of Electrical and Electronic Engineering, Graduate School of Engineering, Mie University, Mie 514–8507, Japan*

[2] *Department of Telecommunication Engineering, Faculty of Engineering, King Mongkut's Institute of Technology Ladkrabang (KMITL), Bangkok, Thailand*

a) *pongsathorn@com.elec.mie-u.ac.jp*

Abstract: This paper proposes a minimum mean square error-frequency domain equalization (MMSE-FDE) method for training sequence inserted orthogonal frequency division multiplexing (TS-OFDM) signal under higher mobile environments. The salient features of proposed method are to enable the acquisition of frequency diversity gain by using an enhanced channel impulse response (CIR) matrix and to enable the reduction of complexity by using a fast algorithm for inverse matrix calculation. From the simulation results, this paper confirms that the proposed MMSE-FDE method can achieve better bit error rate (BER) performance than the conventional MMSE-FDE methods with keeping lower complexity under higher mobile environments.

Keywords: MMSE-FDE, TS-OFDM, CIR matrix, time-varying channels

Classification: Wireless Communication Technologies

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1 Problems of conventional FDE methods

Orthogonal frequency division multiplexing (OFDM) has been considered as one of efficient transmission techniques in the wireless mobile communications because of its robustness to multipath fading. However, the signal quality would be degraded under higher mobile environments accompanying with the Doppler frequency shift in which the property of orthogonality among OFDM subcarriers is no more satisfied due to the occurrence of inter-carrier interference (ICI) [1].

To solve the above ICI problem, various frequency domain equalization (FDE) methods have been proposed both for a cyclic prefix OFDM (CP-OFDM) and time domain training sequence OFDM (TS-OFDM) [2, 3, 4]. In [2], a low-complexity minimum mean square error-frequency domain equalization (MMSE-FDE) method was proposed for CP-OFDM. In this method, the full elements of channel frequency response (CFR) matrix which is converted from the time domain channel impulse response (CIR) is approximated by a banded matrix so as to employ a fast algorithm for inverse matrix calculation [5]. However, the CIR matrix is constructed by using only data period which leads the degradation of bit error rate (BER) performance in deep fading channels. To solve this problem, [3] proposed an enhanced FDE method for CP-OFDM in which the CIR matrix is constructed by using both data and CP periods to acquire the frequency diversity gain. However, this method requires the Moore-Penrose inverse matrix calculation for the non-square CFR matrix at every data symbol which leads significantly higher complexity. Furthermore, above two methods [2, 3] of using CP-OFDM are usually required to estimate the CIR by using pilot subcarriers in the frequency domain which leads the degradation of CIR estimation accuracy due to the occurrence of ICI in the received pilot subcarriers. To solve the above problems, [4] proposed a low-complexity FAST-MMSE-FDE method for TS-OFDM in which the TS is used in the estimation of CIR at every symbol and the enhanced CIR matrix at every sample time is constructed by using a linear interpolation between the CIRs estimated at two consecutives symbols. By assuming the liner changing of CIR within one symbol period, the low complexity Maclaurin's expansion approximation can be employed in the calculation of inverse matrix. Although this method can achieve the frequency diversity gain with lower complexity, the BER performance would be degraded in higher time-varying fading channels due to the assumption of linear changing of CIR and the approximation of first order Maclaurin's expansion.

To solve the above problems, this paper proposes a new MMSE-FDE method for TS-OFDM which can achieve better BER performance than the conventional MMSE-FDE methods with keeping lower complexity under higher mobile environments.

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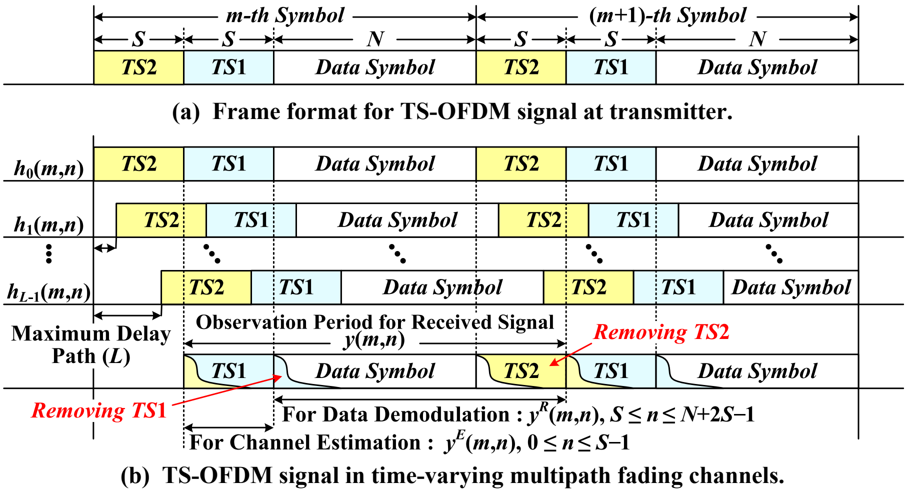

Fig. 1. Frame format of TS-OFDM signal at transmitter (a) and receiver (b).

2 Proposed MMSE-FDE method for TS-OFDM

Fig. 1(a) shows the frame format for TS-OFDM. At the transmitter, the modulated data $X(m,k)$ at the k-th subcarrier of m-th symbol is converted to the time domain by N-point inverse fast Fourier transform (IFFT) which is given by,

$$x(m,n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(m,k) \cdot e^{j\frac{2\pi kn}{N}}, \quad 0 \le n \le N-1 \tag{1}$$

where $x(m,n)$ is the transmitted time domain signal at the n-th time sample. Then, the same data pattern of TS1 and TS2 with the length of S samples are added at both ends of data symbol which is given by,

$$x^T(m,n) = \begin{cases} TS1 : b(n), & 0 \le n \le S-1 \\ \text{Data} : x(m,n-S), & S \le n \le N+S-1 \\ TS2 : b(n-N-S), & N+S \le n \le N+2S-1 \end{cases} \tag{2}$$

where $b(n)$ is the TS signal and $x^T(m,n)$ is the transmitted signal including data, TS1 and TS2. Here the length of S is taken by longer than the length of delay path (L) to remove the interferences of TS1 and TS2 from the received data symbol.

Fig. 1(b) shows the received TS-OFDM signal in the time-varying multipath fading channels. The received signal $y(m,n)$ from $n = 0$ to $N+2S-1$ is divided into two observation periods $y^E(m,n)$ with length of S samples and $y^R(m,n)$ with the length of $P = N+S$ samples as shown in Fig. 1(b) which are used in the CIR estimation and the data equalization, respectively.

2.1 Low complexity enhanced MMSE-FDE method

The received signal $y^R(m,n)$ with the observation period of P $(= N+S)$ samples can be given by,

$$y^R(m,n) = \sum_{l=0}^{L-1} h_l(m,n) \cdot x^T(m,n-l) + w(m,n), \quad S \le n \le N+2S-1 \tag{3}$$

where $h_l(m,n)$ is the CIR at every sample time for the l-th delay path and $w(m,n)$ is the additive white Gaussian noise (AWGN) with variance of σ^2. Here it should be

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noted that the received data signal in (3) includes the interferences of TS1 and TS2 incurred at the start and last of data symbol as shown in Fig. 1(b). To remove these interferences from the received signal, it is required to estimate the CIR $h_l(m,n)$ precisely at every sample time. The authors have proposed the CIR estimation method at every sample time by using the cubic interpolation method for the CIR estimated at every symbol by using the TS1 signal $y^E(m,n)$ [6]. By using the estimated CIR $\hat{h}_l(m,n)$ at every sample time and the known data pattern of TS1 and TS2 given in (2), the interferences can be removed by,

$$y^D(m,n) = \begin{cases} y^R(m,n) - \sum_{l=n-S+1}^{S-1} \hat{h}_l(m,n) \cdot b(n-l), & S \leq n \leq 2S-2 \\ y^R(m,n), & 2S-1 \leq n \leq N+S-1 \\ y^R(m,n) - \sum_{l=0}^{n-N-S} \hat{h}_l(m,n) \cdot b(n-N-S-l), & \\ & N+S \leq n \leq N+2S-1 \end{cases} \tag{4}$$

where $y^D(m,n)$ is the received signal after removing the interferences of TS1 and TS2. From (4), $y^D(m,n)$ can be expressed by the following matrix operation.

$$\mathbf{y}^D_{P\times 1} = \mathbf{h}_{P\times N}\mathbf{x}_{N\times 1} + \mathbf{w}_{P\times 1} \tag{5}$$

where $\mathbf{x}$ with the size of $N \times 1$ corresponds to the transmitted time domain signal $x(m,n)$ in (1) and $\mathbf{h}$ is the enhanced CIR matrix with the size of $P \times N$ including the CIR at every sample time during the data and TS2 periods. The time domain signal in (5) is converted to the frequency domain by $P \times P$-DFT matrix which can be expressed by,

$$\begin{aligned} \mathbf{Y}^D_{P\times 1} &= \mathbf{F}_{P\times P}\mathbf{y}^D_{P\times 1} + \mathbf{F}_{P\times P}\mathbf{w}_{P\times 1} = \mathbf{F}_{P\times P}\mathbf{h}_{P\times N}\mathbf{F}^H_{N\times P}\mathbf{F}_{P\times N}\mathbf{x}_{N\times 1} + \mathbf{W}_{P\times 1} \\ &= \mathbf{H}_{P\times P}\mathbf{X}^T_{P\times 1} + \mathbf{W}_{P\times 1} \end{aligned} \tag{6}$$

where $(\cdot)^H$ is the Hermitian transpose, $\mathbf{F}$ and $\mathbf{F^H}$ are the DFT and IDFT matrixes, $\mathbf{H}$ is the square matrix of CFR with the size of $P \times P$ and $\mathbf{W}$ is the noise component. $\mathbf{X^T}$ is the frequency domain signal with the size of $P \times 1$ which includes the data symbol and its delay component after removing the interferences of TS1 and TS2 as shown in Fig. 1(b). The frequency domain signal $\hat{\mathbf{X}}^{\mathbf{T}}$ in (6) can be equalized with the MMSE-FDE method which is given by,

$$\hat{\mathbf{X}}^T_{P\times 1} = (\mathbf{H}^H_{P\times P}\mathbf{H}_{P\times P} + \sigma^2\mathbf{I}_P)^{-1}\mathbf{H}^H_{P\times P}\mathbf{Y}^D_{P\times 1} \tag{7}$$

where $\mathbf{I}$ is the Identity matrix and σ^2 is the variance of AWGN. The calculation of inverse matrix in (7) requires the order of complexity $O[P^3]$ which leads significantly higher complexity. To reduce the complexity, the full elements of CFR matrix $\mathbf{H}$ is approximated by a banded matrix $\mathbf{B}$ with the lower and upper bandwidth of Q_1 so as to employ the fast algorithm for inverse matrix calculation [5]. By suing the banded matrix $\mathbf{B}$, (7) can be approximated by,

$$\hat{\mathbf{X}}^T_{P\times 1} \approx (\mathbf{B}^H_{P\times P}\mathbf{B}_{P\times P} + \sigma^2\mathbf{I}_P)^{-1}\mathbf{B}^H_{P\times P}\mathbf{Y}^D_{P\times 1} \approx \mathbf{G}^{-1}_{P\times P}\mathbf{B}^H_{P\times P}\mathbf{Y}^D_{P\times 1} \tag{8}$$

where $\mathbf{G} = (\mathbf{B}^H\mathbf{B} + \sigma^2\mathbf{I})$ is the Hermitian banded matrix. The order of complexity required in the fast algorithm for calculation of inverse matrix $\mathbf{G}$ becomes $O[3(Q_1+1)^2P]$ which is relatively lower than the direct inverse matrix calculation

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[5]. In the demodulation of data information, the estimated $\hat{\mathbf{X}}^{T}$ in (8) is converted to the time domain signal $\hat{\mathbf{x}}^{T}$ by P-points IDFT. Here the time domain signal $\hat{x}^{T}(m,n)$ from $n = 0$ to $N-1$ corresponds to the data information $\hat{x}(m,n)$ in (1). From this fact, the equalized data information $\hat{X}(m,k)$ can be obtained from the time domain signal $\hat{x}^{T}(m,n)$ by N-points FFT.

2.2 Computation complexity for proposed MMSE-FDE method

Table I shows the order of complexities for the proposed, conventional MMSE-FDE [2] and the FAST-MMSE-FDE [4] methods. The order of complexities are evaluated for the processing loads required in the removing of ISI, construction of CFR matrix, calculation of inverse matrix and equalizations. In the table, the orders of complexities for N-point FFT and P-points DFT are evaluated by $O[N\log_2 N]$ and $O[P^2]$, respectively. In the next section, the order of complexities for all methods shown in Table I are evaluated by assuming the actual parameters.

Table I. Order of complexities for proposed and conventional methods.

Equalization methods	Processing loads		
	Removing of ISI	Construction of CFR matrix	Inverse matrix calculation and equalization
Conv. MMSE-FDE for CP-OFDM [2]	N/A	$O[(2Q_1+3)N\log_2 N + N(Q_1+1)^2 + N(Q_1+1)]$	$O[N^2 + 3N(Q_1+1)^2]$
Conv. FAST-MMSE-FDE for TS-OFDM [4]	$O[S^2 - S]$	$O[P^2 + 2PS + P]$	$O[(2Q_2+1)P^2 + (2Q_2+3)P + PN]$
Prop. MMSE-FDE for TS-OFDM	$O[S^2 - S]$	$O[P^2 + (2Q_1+2)PS + P(Q_1+1)^2 + P(Q_1+1)]$	$O[2P^2 + N\log_2 N + 3P(Q_1+1)^2]$

N is the number of FFT/IFFT points, S is the length of TS and CP, $P = N + S$, Q_1 is the lower and upper bandwidths of banded matrix, and Q_2 is the order of Maclaurin's expansion.

3 Performance evaluations for proposed MMSE-FDE method

This section presents simulation results for the proposed method as comparing with the conventional methods. In the simulation, the modulation method is QPSK, OFDM bandwidth is 5 MHz, radio frequency is 5 GHz, number of data subcarriers is $N = 128$, and length of TS and CP are taken by $S = CP = 16$. The lower and upper bandwidths of banded matrix for the conventional [2] and proposed MMSE-FDE are taken by $Q_1 = 3$, 5 and 7. The Maclaurin's expansion order for the conventional FAST-MMSE-FDE is $Q_2 = 1$ [4]. The time-varying fading channel is modeled by the exponential power delay profile having the power decay constant -1 dB with the length of delay path $L = 14$. The normalized Doppler frequency $R_D = f_{\mathrm{dmax}}/\Delta f$ is employed as the measure of mobile environments in which f_{dmax} is the maximum Doppler frequency and Δf is the OFDM subcarrier spacing.

Fig. 2 shows the BER performances for the proposed and conventional methods when changing C/N at $R_D = 5\%$. From the figure, it can be observed that the proposed method with $Q_1 = 5$ can achieve better BER performance than

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the conventional methods. Here it should be noted that the degradation of BER performances as compared with the proposed method are caused by the lower CIR estimation accuracy of using pilot subacarriers and no acqisition of frequency diversity gain in the conventional MMSE-FDE for CP-OFDM, and the assumption of linear changing of CIR and the approximation of first order Maclaurin's expansion in the FAST-MMSE-FDE for TS-OFDM, respectively.

From Table I and assuming the above actual parameters, the ratios of complexity for the proposed method over the conventional MMSE-FDE and FAST-MMSE-FDE methods are 2.4 and 1.1 times, respectively. From these complexities and BER performances in Fig. 2, it can be concluded that the proposed method can achieve better BER performance with small increment of complexity under higher mobile environments.

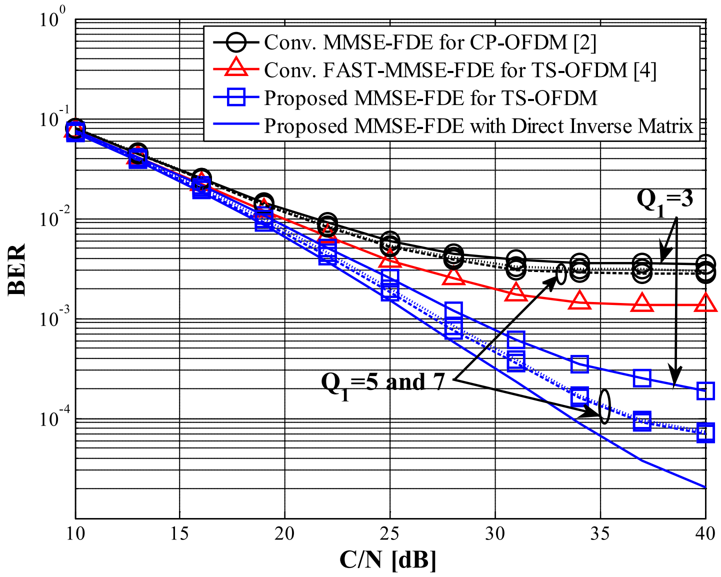

Fig. 2. BER performance of proposed method when changing C/N at $R_D = 5\%$.

4 Conclusion

This paper proposed the low complexity enhanced MMSE-FDE method for TS-OFDM. From the computer simulation results, this paper concluded that the proposed method can achieve better BER performance than the conventional methods with keeping lower complexity under higher mobile environments.

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