Time Domain Equalization Method for DFTS-OFDM Signal without GI under Highly Mobile Environments

Pongsathorn Reangsuntea[1], **Pisit Boonsrimuang**[2],
Kazuo Mori[3], and **Hideo Kobayashi**[4], Non-members

ABSTRACT

In highly time-varying fading channel, the Discrete Fourier Transform Spreading Orthogonal Frequency Division Multiplexing (DFTS-OFDM) signal would be damaged significantly by the inter-channel interference (ICI) due to the loss of orthogonality among subcarriers which leads a fatal degradation of bit error rate (BER) performance. To solve this problem, this paper proposes a time domain equalization (TDE) technique in conjunction with a time domain channel impulse response (CIR) estimation method for the DFTS-OFDM signal without using a guard interval (GI). The features of proposed method is to employ a time domain training sequence (TS) both for the estimation of time domain CIR at every sampling time and for removing the inter-symbol interference (ISI) incurred in the multipath fading channel. The proposed method also employs the TDE with a maximum likelihood (ML) estimation method in the demodulation of received time domain signal at every symbol instead of using the conventional Minimum Mean Square Error-Frequency Domain Equalization (MMSE-FDE) method. This paper presents various simulation results to demonstrate the effectiveness of proposed demodulation method for the DFTS-OFDM signal without GI as comparing with the conventional MMSE-FDE and TDE methods both of using GI under highly mobile environments.

Keywords: DFTS-OFDM, Time Domain Equalization (TDE), Channel Impulse Response (CIR), Time Domain Training Sequence (TS)

1. INTRODUCTION

Discrete Fourier Transform Spreading Orthogonal Frequency Division Multiplexing (DFTS-OFDM) technique has been received a lot of attentions as an attractive alternative to OFDM technique for its lower Peak to Averaged Power Ratio (PAPR) and robustness to the multipath fading. The DFTS-OFDM can achieve better bit error rate (BER) performance than that for OFDM by using the Minimum Mean Square Error Frequency Domain Equalization (MMSE-FDE) method [1-5]. From the above advantages, the DFTS-OFDM has been adopted as the standard transmission technique for the uplink from the user terminal to the base station in the 4th generation mobile communication system (LTE: Long Term Evolution) [6-7].

The DFTS-OFDM signal experiences severe inter-channel interference (ICI) due to the Doppler frequency spreads in highly time-varying fading channel where the time domain channel impulse response (CIR) would be no more constant even during one DFTS-OFDM symbol period [8-9]. Accordingly the channel frequency response (CFR) which be employed in the MMSE-FDE at the receiver is also changed relatively during one DFTS-OFDM symbol period. From this fact, it is hard to compensate the ICI due to the Doppler frequency spreads by using the conventional MMSE-FDE method which leads the fatal degradation of BER performance in highly time-varying fading channel.

To solve the above problem, a time domain equalization (TDE) method of using GI have been proposed for the DFTS-OFDM signal [10-11]. The proposed TDE method for the DFTS-OFDM signal with GI can achieve better BER performance than the conventional MMSE-FDE method in highly time varying fading channel. Because the TDE method can track the variation of CIR at every sampling time basis. However, in the quasi-static condition, the proposed TDE method with GI shows the worse BER performance than the conventional MMSE-FDE method which can achieve the frequency diversity gain. This is the reason that the demodulation of conventional TDE method with GI is operated in the time domain which is the different from the MMSE-FDE method which is operated in the frequency domain.

To improve the BER performance for the DFTS-OFDM signal furthermore than the conventional TDE method of using GI in highly time-varying fading channel, this paper proposes a time domain equalization (TDE) technique without using GI. In the proposed method, a time domain training sequence

Manuscript revised on October 18,2015.
[1,3,4] The authors are with Electrical and Electronic Engineering, Graduate School of Engineering, Mie University, Tsushi, 514-8507 Japan., Email:pongsathorn@com.elec.mieu.ac.jp, kmori@elec.mie-u.ac.jp and koba@elec.mie-u.ac.jp
[2] The author is with Telecommunication Engineering, Faculty of Engineering, King Mongkuts Institute of Technology Ladkrabang, 10520 Thailand., Email:kbpisit@kmitl.ac.th

(TS) is added at the both ends of data symbol instead of adding the GI at the transmitter to compensate the inter-symbol interference (ISI) occurred due to the unemployment of GI [12]. In the proposed TDE method, the CIR is estimated at every sampling time by using the TS in highly time-varying fading channel. This paper also proposes the TDE method with a maximum likelihood (ML) estimation method for the received time domain signal after removing the ISI which can achieve better BER performance than the conventional MMSE-FDE and TDE methods both employing the GI in highly time-varying fading channel.

The rest of this paper is organized as follows. Section 2 presents a problem of conventional MMSE-FDE method of using GI for the DFTS-OFDM signal in highly time-varying fading channel. Section 3 proposes a time domain CIR estimation method at every sampling time by using the time domain TS which is also used in the mitigation of both ISI and ICI in the proposed ML based TDE method. Section 4 proposes the ML based TDE method in conjunction with the proposed time domain CIR estimation method. Section 5 presents various computer simulation results to verify the effectiveness of proposed ML based TDE demodulation method as comparing with the conventional MMSE-FDE and TDE methods of using the GI, and Section 6 draws some conclusions.

2. PROBLEM OF MMSE-FDE METHOD FOR DFTS-OFDM SIGNAL WITH GI

When assuming the quasi-static or slow time-varying fading channel, the time domain CIR can be considered as a constant during one DFTS-OFDM symbol period. Fig. 1 shows a schematic diagram for the relationships between the CIR in the time domain and the CFR in the frequency domain where the CFR can be obtained by performing a DFT to the CIR at a certain sampling time during one symbol period. Fig. 1 (a) shows the CIR and its corresponding CFR in the quasi-static channel. From the figure, it can be seen that the CIRs at the three sampling times are almost constant. Accordingly the CFRs converted from the CIRs at these three sampling times during one symbol period are also almost the same responses as shown in Fig. 1 (a). From this fact, the received DFTS-OFDM signal affected by the slow time-varying fading distortion can be equalized precisely by using the MMSE-FDE method with the fixed CFR converted from the CIR at any sampling time during one symbol period.

When assuming the highly time-varying fading channel occurred in the data communications such as high speed vehicle to vehicle (V2V) and high speed trains communications [9], the time domain CIR would be no more constant even during one DFTS-OFDM symbol period. Fig. 2 shows an example of computer simulation results of time domain CIR for 1st to 3rd delay paths at every sampling time in the Rician multipath fading channel. In the simulation, the normalized Doppler frequency $f_{dmax}/\Delta f$ is taken by 5% where f_{dmax} is the maximum Doppler frequency spread and Δf is the subcarrier spacing of DFTS-OFDM signal. From the figure, it can be seen that the amplitude of time domain CIR is changing relatively even during one DFTS-OFDM symbol period. Fig. 1 (b) shows the relationships between the time domain CIR and its corresponding frequency domain CFR in higher time-varying fading channel. The time domain CIRs are changing during one symbol period and the frequency domain CFRs converted from the time domain CIRs at the different sampling times are much different as shown in Fig. 1 (b).

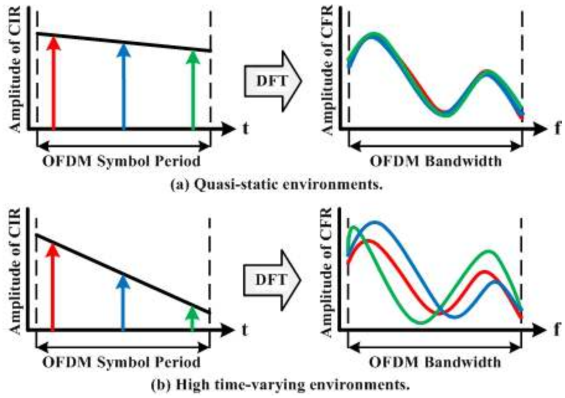

Fig.1: *Relationships between CIR and CFR.*

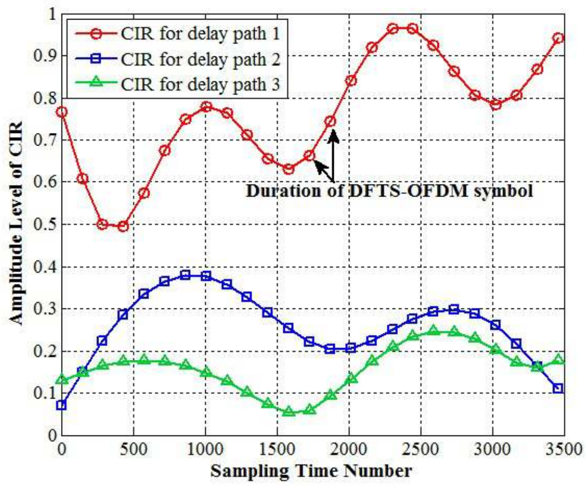

Fig.2: *Amplitude of CIR at every sampling time in high time-varying channel.*

From this fact, it is difficult to mitigate the time-varying fading distortion by the MMSE-FDE method of using one fixed CFR over one symbol period which leads the fatal degradation of BER performance.

To solve the above problem of MMSE-FDE method in highly time-varying fading channel, this paper proposes the ML based TDE method for the DFTS-OFDM signal without GI in conjunction with the time domain CIR estimation method. The salient feature of the proposed ML based TDE method is to achieve much better BER performance than the conventional MMSE-FDE method.

3. PROPOSAL OF TIME DOMAIN CIR ESTIMATION METHOD

Fig. 3 shows the frame format employed in the proposed TDE method. In the proposed frame format, the time domain training sequence (TS) is added at the both ends of every data symbol which be used in the estimation of CIR at every sampling time. The time domain TS can be also used as the role of GI which is usually employed in the conventional DFTS-OFDM signal to avoid the ISI. Here it should be noted that the ISI occurred in the received DFTS-OFDM signal without GI can be removed by using the TS1 and TS2 of which data patterns are known at the receiver.

Fig. 4 shows the structure of transceiver for the proposed TDE method of using the proposed CIR estimation method. At the transmission side, the input information data is encoded by a forward error correction (FEC) code [13]. The encoded M data are modulated by the quadrature-amplitude modulation (QAM) method. Then the modulated signal $x_D(m,n)$ at the $n-th$ sampling time of $m-th$ symbol is converted to the frequency domain signal $X_D(m,k)$ at the $k-th$ subcarrier by the $M-point$ DFT which is given by the following equation.

$$X_D(m,k) = \sum_{n=0}^{M-1} x_D(m,n) \cdot e^{-j\frac{2\pi kn}{M}},\ 0 \le k \le M-1 \quad (1)$$

The M subcarriers of $X_D(m,k)$ given in (1) are mapped into the certain frequency band within the frequency bandwidth with N subcarriers continuously from the subcarrier number N_{Z1} to N_{Z2} ($N_{Z2}-N_{Z1}+1 = M$) and the null subcarriers (zero padding) are added at the both ends of M data subcarriers. After the subcarrier mapping, the frequency domain signal $X(m,k_1)$ over N subcarriers can be given by,

$$X(m,k_1) = \begin{cases} 0(zero\ padding), & 0 \le k_1 \le N_{Z1}-1 \\ X_0(m,k_1)-N_{Z1}, & N_{z1} \le k_1 \le N_{Z2} \\ 0(zero\ padding), & N_{Z2}+1 \le k_1 \le N-1 \end{cases} \quad (2)$$

where $(N-M)$ is the number of zero padding (NZ) added at the both ends of M data subcarriers in the frequency domain. The frequency domain signal $X(m,k_1)$ including the zero padding given in (2) is converted to the time domain signal as similar to the conventional OFDM signal by the $N-point$ IFFT which can be given by,

$$x(m,n_1) = \frac{1}{N}\sum_{k_1=N_{Z1}}^{N_{z2}} x(m,k_1) \cdot e^{-j\frac{2\pi n_1 k_1}{N}},\ 0 \le n_1 \le N-1 \quad (3)$$

where $x(m,n_1)$ is the transmitted DFTS-OFDM signal in the time domain.

The time domain training sequences TS1 and TS2 with the length of N_{TS} samples are added at the both ends of every data symbol as shown in Fig. 3 which be used in the estimation of CIR. The time domain TS added at the both ends of data symbol can be also used as the role of GI in the conventional DFTS-OFDM signal to avoid the ISI. The transmitted time domain signal $x_T(m,n_2)$ at the n_2-th sampling time of $m-th$ symbol including both TS1 and TS2 can be expressed by,

$$x_T(m,n_2) = \begin{cases} d(m,n_2), & 0 \le n_2 \le N_{TS}-1 \\ x(m,n_2-N_{TS}), & N_{TS} \le n_2 \le N+N_{TS}-1 \\ d_2(m,n_2-N-N_{TS}), & N+N_{TS} \le n_2 \le N+2N_{TS}-1 \end{cases} \quad (4)$$

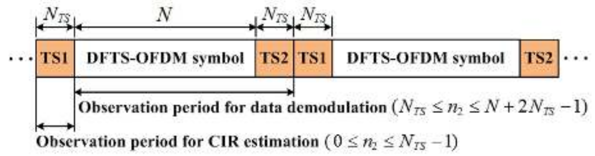

Fig.3: *Proposed frame format in the time domain.*

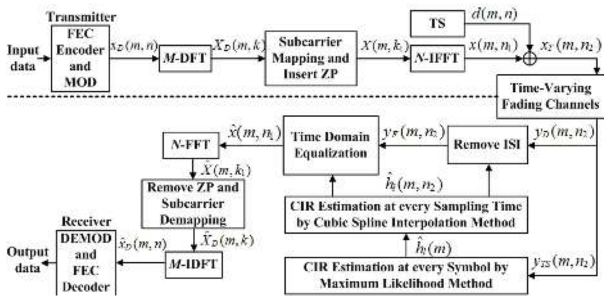

Fig.4: *Structure of transceiver for proposed DFTS-OFDM system.*

where $d_1(m,n)$ and $d_2(m,n)$ are the time domain TS1 and TS2 with the length of N_{TS} samples of which data patterns are known at the receiver. For simplicity, this paper assumes the data patterns both for TS1 and TS2 are the same as $d(m,n)$. The length of N_{TS} should be taken longer than the length of delay paths (L) which is the same condition of the conventional GI. Assuming that the timing synchronization for the received signal is perfect at the receiver, the received time domain TS signal after passing through the multipath fading channel can be expressed by,

$$y_{TS}(m,n_2) = \sum_{l=0}^{L-1} h_l(m)\cdot d(m, m_2-1) + w(m, m2), \quad (5)$$
$$= (0 \le n_2 \le N_{TS}-1)$$

where L is the number of delay paths, $h_l(m)$ is the time domain CIR for the $l-th$ delay path at the $m-th$ symbol in the time-varying fading channel and $w(m,n_2)$ is an additive white Gaussian noise (AWGN) in the time domain. Here it should be noted that the CIR is assumed to be constant during the short time period of TS1 and TS2 even in highly time-varying fading channel. Assuming the time domain CIR $\hat{h}_l(m)$ as the unknown parameters, the expected received time domain TS signal $\hat{y}_{TS}(m,n_2)$ can be expressed by,

$$\hat{y}_{TS}(m,n_2) = \sum_{i=0}^{N_{TS}-1} \hat{h}_l(m)\cdot d(m,n_2-1), 0 \le n_2 \le N-TS-1 \quad (6)$$

The unknown parameters for the time domain CIR $\hat{h}_l(m)$ at the $m-th$ symbol can be estimated by solving the following maximum likelihood (ML) equation [14] under the constraint condition with minimizing the difference between the actual received TS signal $y_{TS}(m,n_2)$ in (5) and the expected received TS signal $\hat{y}TS(m,n_2)$ in (6).

$$\Upsilon = \arg\min_{\hat{h}_l(m)} \left[\sum_{n_2=0}^{N_{TS}-1} |y_{TS}(m,n_2) - \hat{y}_{TS}(m,n_2)|^2 \right] \quad (7)$$

In this paper, we employ a simple solution for solving ML equation in (7) by using the following simultaneous equations.

$$y_{TS}(m,n_2) = \hat{y}_{TS}(m,n_2) \quad for \quad 0 \le n_2 \le N_{TS}-1 \quad (8)$$

By substituting $\hat{y}_{TS}(m,n_2)$ given in (6) into (8), (8) can be expressed by the following matrix operations.

$$\underbrace{[\hat{h}_l(m)]}_{N_{TS}\times 1} = \underbrace{[d(m,n_2-l)]^{-1}}_{N_{TS}\times N_{TS}} \cdot \underbrace{[y_{TS}(m,n_2)]}_{N_{TS}\times 1} \quad (9)$$

By using (9), the time domain CIR $\hat{h}_l(m)$ can be estimated at every symbol. The number of delay paths to be estimated in (9) is assumed by N_{TS} because the actual number of delay paths occurred in the real channel is unknown at the receiver. Here it should be noted that since the data pattern of time domain TS $d(m,n_2)$ is known at the receiver, the inverse matrix of $[d(m,n_2-l)]-1$ in (9) can be calculated in advance. The calculated inverse matrix can be used for every received data symbol which enables the considerable reduction of computation complexity in the estimation of CIR at every symbol by using (9).

By using the estimated $[\hat{h}_l(m)]$ at every symbol, the CIR at every sampling time $[\hat{h}_l(m,n_2)]$ can be estimated by using the cubic spline interpolation method over several estimated CIRs of $[\hat{h}_l(m)]$. Fig. 5 shows a schematic diagram for the cubic spline interpolation method to estimate the time domain CIR at every sampling time. The estimation accuracy of time domain CIR at every sampling time will be evaluated in Section 5.

4. PROPOSAL OF TDE METHOD FOR DFTS-OFDM SIGNAL WITHOUT GI

Assuming that the length of time domain TS (N_{TS}) is longer than the maximum length of delay paths (L) in the time-varying fading channel, the received time domain data signal $y_D(m,n_2)$ at every sampling time during the observation period for the data demodulation from N_{TS} to $N+2N_{TS}-1$ as shown in Fig. 3 can be expressed by,

$$y_D(m,n_2) = \sum_{l=0}^{L-1} h_l(m,n_2)\cdot x_T(m,n_2-l) + w(m,n_2),$$
$$(N_{TS} \le n_2 \le N+2N_{TS}-1) \quad (10)$$

where $x_T(m,n_2)$ is the transmitted time domain signal given in (4) and $h_l(m,n_2)$ is the time domain CIR at every sampling time occurred in the actual time-varying fading channel. Fig. 6 shows the received time domain signal in time-varying fading channel. From the figure, it can be seen that the received time domain data signal $y_D(m,n_2)$ in (10) includes the ISI of TS which are added at the start and end of data symbol from the TS1 and TS2, respectively. As shown in Fig. 6, by using the estimated CIR $\hat{h}_l(m,n_2)$ at every sampling time and the data pattern of $d(m,n_2)$ both for TS1 and TS2 which are known at the receiver, the ISI added at the both ends of time domain data signal can be removed by the following equation.

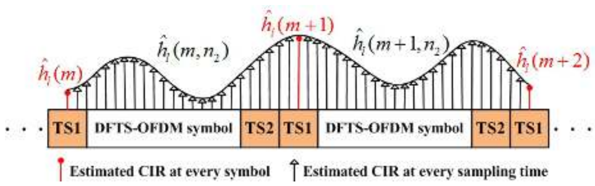

Fig.5: *Estimation of CIR at every sampling time by using cubic spline interpolation method.*

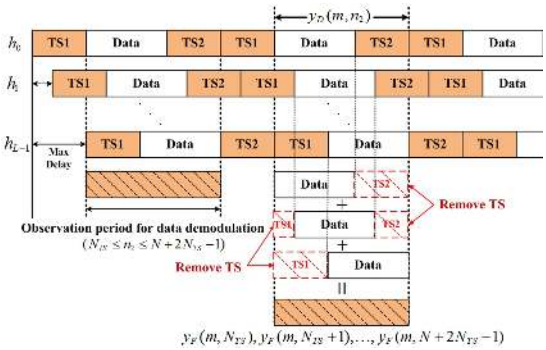

Fig.6: *Time domain received signal in multipath fading channel.*

$$
y_F(m,n_2) = \begin{cases} y_D(m,n_2) - \displaystyle\sum_{l=n_2-N_{TS}+1}^{N_{TS}-1} \hat{h}_l(m,n_2)\cdot d(m,n_2-1), \\ \qquad (N_{TS} \le n_2 \le 2N_{TS}-2) \\ y_D(m,n_2), \qquad (2N_{TS}-1 \le n_2 \le N+N_{TS}-1) \\ y_D(m,n_2) - \displaystyle\sum_{l=0}^{n_2-N-N_{TS}-1} \hat{h}_l(m,n_2)\cdot d(m,n_2-N-N_{TS}-1), \\ \qquad (N+N_{TS} \le n_2 \le N+2N_{TS}-2) \end{cases} \tag{11}
$$

In the derivation of (11), the following relationships are used.

$$
n_2-l \le N_{TS}-1, \quad x_T(m,n_2-l) = d(m,n_2-l) \tag{12}
$$

$$
n_2-l \le N+N_{TS}-1, \quad x_T(m,n_2-1) = d(m,n_2-N-N_{TS}-l) \tag{13}
$$

where $y_F(m,n_2)$ in (11) is the received time domain signal after removing the ISI of TS from the actual received time domain signal $y_D(m,n_2)$ in (10). From (11) and Fig. 6, it can be observed that the superimposed received time domain signal consists of transmitted data symbols with having the different delay times which corresponds to the original transmitted signal given in (3). When employing the conventional TDE method of using GI, the received time domain signal in the time-varying fading channel includes the partial signal within GI at the start of every delay path signal. From this fact, it can be expected that the proposed TDE method without GI shows the better BER performance than the conventional TDE method with GI in the ML based demodulation method.

When the transmitted time domain data $x(m,n_1)$ given in (3) is assumed as the unknown parameters, the expected received time domain data signal $\hat{y}_E(m,n_2)$ without including the ISI which corresponds to the actual received time domain signal after removing the ISI in (11) can be given by,

$$
\hat{y}_E(m,n_2) = \begin{cases} \displaystyle\sum_{l=0}^{n_2-N_{TS}} \hat{h}_l(m,n_2)\cdot \hat{x}(m,n_2-N_{TS}-l), \\ \qquad (N_{TS} \le n_2 \le 2N_{TS}-2) \\ \displaystyle\sum_{l=0}^{N_{TS}-1} \hat{h}_l(m,n_2)\cdot \hat{x}(m,n_2-N_{TS}-l) \\ \qquad (2N_{TS}-1 \le n_2 \le N+N_{TS}-1) \\ \displaystyle\sum_{l=n_2-N-N_{TS}+1}^{N_{TS}-1} \hat{h}_l(m,n_2)\cdot \hat{x}(m,n_2-N_{TS}-l), \\ \qquad (N+N_{TS} \le n_2 \le N+2N_{TS}-2) \end{cases} \tag{14}
$$

The unknown parameter of time domain data signal $\hat{x}(m,n_1)$ which corresponds to (3) can be estimated by solving the following maximum likelihood (ML) equation under the constraint with minimizing the difference between the actual received time domain signal $y_F(m,n_2)$ in (13) and the expected received time domain signal $\hat{y}_E(m,n_2)$ in (14).

$$
\Upsilon = \arg\min_{\hat{x}(m,s)} \left[\sum_{n_2=N_{TS}}^{N+2N_{TS}-2} |y_F(m,n_2) - \hat{y}_E(m,n_2)|^2 \right] \tag{15}
$$

The maximum likelihood (ML) equation in (15) can be converted to the simultaneous equations by taking the partial differentiation for the unknown parameters $\hat{x}^*(m,s)$ which can be expressed by the following equation.

$$
\frac{\partial \Upsilon}{\partial \hat{x}^*(m,s)} = \frac{\partial \left(\displaystyle\sum_{n_2=N_{TS}}^{N+2N_{TS}-2} |y_F(m,n_2) - \hat{y}_E(m,n_2)|^2 \right)}{\partial \hat{x}^*(m,s)} = 0, \quad (0 \le s \le N-1) \tag{16}
$$

where $*$ represents the conjugate complex number. By using (16), the ML equation (15) can be expressed by the following simultaneous equations with N unknown time domain data information of $\hat{x}(m,n_1)$.

$$
\underbrace{[b(m,s)]}_{N\times 1} = \underbrace{[A_m(s,n_1)]}_{N\times N} \cdot \underbrace{[\hat{x}(m,n_1)]}_{N\times 1} \tag{17}
$$

where $b(m,s)$ and $A_m(s,n_1)$ can be given by,

$$\begin{aligned} b(m,s) &= \sum_{n_2=N_{TS}}^{N+2N_{TS}-2} y_F(m,n2)\frac{\partial \hat{y}_E^*(m,n_2)}{\partial \hat{x}^*(m,s)} \\ &= \underbrace{[h_l^H(m,s)]}_{N\times(N+N_{TS}-1)} \cdot \underbrace{[y_F(m,n_2)]}_{(N\times N_{TS}-1\times N)}, 0\leq s\leq N-1 \end{aligned} \tag{18}$$

$$\begin{aligned} A_m(s,n_1) &= \sum_{n_2=N_{TS}}^{N+2N_{TS}-2} \hat{y}_E(m,n2)\frac{\partial \hat{y}_E^*(m,n_2)}{\partial \hat{x}^*(m,s)} \\ &= \underbrace{[h_l^H(m,s)]}_{N\times(N+N_{TS}-1)} \cdot \underbrace{[h_l(m,n_1)]}_{(N\times N_{TS}-1\times N)}, 0\leq n_1\leq N-1 \end{aligned} \tag{19}$$

where the parameter of s and n_1 can be given by,

$$s = n_2 - n_{TS} - l \quad \text{in} \quad \frac{\partial \hat{y}_E^*(m,n_2)}{\partial \hat{x}^*(m,s)} \tag{20}$$

$$n_1 = n_2 - N_{TS} - l \quad \text{in} \quad \hat{y}_E(m,n_2) \tag{21}$$

From (17), the unknown parameters $[\hat{x}(m,n_1)]$ can be solved by using the inverse matrix of $[A_m(s,n_1)]$ which can be given by,

$$\underbrace{[\hat{x}(m,n_1)]}_{N\times 1} = \underbrace{[A_m(s,n_1)]^{-1}}_{N\times N} \cdot \underbrace{[b(m,s)]}_{N\times 1} \tag{22}$$

where $[\cdot]^{-1}$ represents the inverse matrix operation. Here it should be noted that the inverse matrix calculation in (22) would be the dominant factor in the computational complexity required for the receiver. The computation complexity for the inverse matrix calculation is estimated by the order of $O(N^3)$ when the matrix size is $N \times N$. As one of solutions to reduce the computation complexity for the inverse matrix calculation, we proposed the iterative based TDE method for OFDM signal in [14]. The feature of this method is to employ the iterative method for solving the simultaneous equations instead of using the direct calculation of inverse matrix. From the evaluation results in [14], we concluded that the iterative method can achieve almost the same accuracy with smaller computation complexity than the direct calculation of inverse matrix. From the reason that, we are focusing on the feasibility study for the potential capability of proposed TDE method in this paper. The investigation of iterative method for the DFTS-OFDM signal with the proposed TDE method is subject to the future work.

The time domain signal $\hat{x}(m,n_1)$ estimated in (22) is converted to the frequency domain signal $\hat{X}(m,k1)$ by N-point FFT. After subcarrier de-mapping in the frequency domain data, the data information $\hat{x}_D(m,n)$ can be obtained from the frequency domain signal $\hat{X}_D(m,k)$ by $M-point$ IDFT as shown in Fig. 4. The proposed TDE method with the estimated CIR at every sampling time can mitigate both for the ISI and ICI at the sampling time basis during one DFTS-OFDM symbol period. From this fact, it can be expected that the proposed TDE method without GI can achieve better BER performance than the conventional MMSE-FDE and conventional TDE method of using GI under highly mobile environments.

In this section we derive the mathematical form for the ML based CIR estimation method of using the TS and the ML based TDE demodulation method in high time-varying fading channel as given in (7) and (15), respectively. From these ML equations, we derive the simultaneous equations theoretically which can obtain the solutions by using the Moore Penrose inverse matrix calculation for the non-square matrix and the inverse matrix calculation for the square matrix as given in (9) and (22), respectively.

Here it should be noted that the proposed method is required to store the received time domain signal at every symbol basis at the receiver. The data demodulation is conducted at every symbol for the stored signal of which store and processing technique is also employed in the conventional DFTS-OFDM receiver. However the proposed method is required for the additional processing's in the estimation of CIR at every sampling time, the removing of ISI and the calculation of inverse matrix at every symbol. From these processing, the additional delay would be occurred in the proposed method as compared with the conventional method. The hardware size of transceiver for the proposed method may not be increased so much as compared with that for the conventional method by using the state-of-the-art digital signal processing (DSP) device.

5. PERFORMANCE EVALUATIONS

Although it should be desirable if the CIR estimation accuracy and BER performance of proposed method can be evaluated theoretically by using the mathematical closed form. However it is very hard to derive the mathematical model for the proposed TDE method of using the CIR estimation method especially in high time-varying fading channel. The main purpose of this section is to demonstrate the effectiveness of proposed method in high time-varying fading channel. This section evaluates the performance of proposed method by using the computer simulation results which is usually employed in the basic design of new transmission technique. In this section, various computer simulations are conducted to evaluate the performance of proposed TDE method as comparing with the conventional MMSE-FDE and TDE methods of using GI in highly time-varying fading channel. The simulation parameters to be used in the following evaluations are listed in Table 1. The proposed channel estimation method of using the time domain training sequence (TS) is employed in the estimation of time domain CIR for the conventional TDE of using GI and proposed TDE without using GI methods. The proposed channel estimation method is also used in the estimation of channel frequency response (CFR) for the conventional MMSE-FDE method in order to compare the BER performances with the proposed TDE method fairly. The communication channel is modelled by the Rician multipath fading which is usually experienced in the communications by the user on the highly moving vehicle or train [9]. In the following evaluations, the normalized Doppler frequency $R_D = f_{dmax}/\Delta f$ is employed as the measure of mobile conditions where fdmax is the maximum Doppler frequency spread and Δf is the DFTS-OFDM subcarrier spacing.

The estimation accuracy for the time domain CIR at every sampling time is evaluated by the normalized mean square error (NMSE) which is given by,

Table 1: *Accuracy Comparison of Two Service Areas.*

Parameters		Value
No. of FFT/IFFT points (N)		128
No. of DFT/IDFT points (M)		96
No. of zero padding (NZ)		32
Modulation for data subcarrier		16QAM
No. of data symbols per one frame (NS)		33
Allocated bandwidth		1 MHz
Radio frequency		5.9 GHz
Forward Error Correction (FEC) Code [13]		
Encoding		Convolution
FEC rate		1/2
Constraint length		7
Decoding		Viterbi with hard decision
Interleaver		Matrix with one frame
Conventional MMSE-FDE and TDE method	Modulation for TS	16QAM
	Length of TS (NTS)	16
	Length of GI	16
	Symbol duration (TD)	132μs
Proposed TDE	Modulation for TS	16QAM
	Length of TS (NTS))	16
	Symbol duration (TD)	120μs
Rician multipath fading channel model		
Rice factor (K)		6dB
Delay profile		Exponential
Decay constant		-1dB
No. of delay paths (L)		14
No. of scattered rays		20

$$NMSE = \frac{\sum_{l=0}^{L-1}\sum_{m=0}^{N_s-1}\sum_{n_2=0}^{N+2N_{TS}-1} |\hat{h}_l(m,n_2) - h_l(m,n_2)|^2}{\sum_{l=0}^{L-1}\sum_{m=0}^{N_s-1}\sum_{n_2=0}^{N+2N_{TS-1}} |h_l(m,n_2)|^2} \tag{23}$$

where N_S is the number of DFTS-OFDM symbols per one frame.

Fig. 7 shows the time domain CIR estimation accuracy at every sampling time which is evaluated by the normalized mean square error (NMSE) for the proposed CIR estimation method of using the time domain TS when changing the normalized Doppler frequency R_D and operation C/N. Here it should be noted that the CIR estimation accuracy at every sampling time is evaluated taking into account both the estimation accuracy for the estimated CIR at every symbol by using time domain TS and the estimated CIR at every sampling time by using the cubic interpolation method. From the figure, it can be observed that the proposed method can keep higher estimation accuracy when R_D is up to 20% which corresponds to the vehicle speed of 381km/hr even at lower operation C/N.

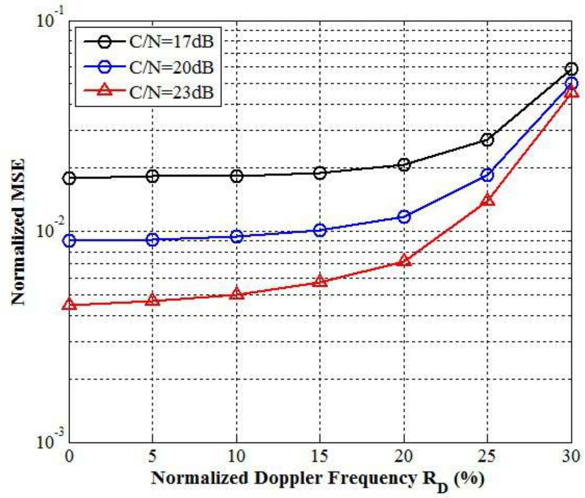

Fig. 7: *CIR estimation accuracy (NMSE) for proposed method when changing the normalized Doppler frequency RD and operation C/N.*

From these results, it can be confirmed that the proposed CIR estimation method of using the time domain TS can be used in the proposed TDE method to mitigate the inter-symbol interference (ISI) and inter-channel interference (ICI) precisely when R_D is up to 20%.

Fig. 8 shows the BER performances for the proposed TDE method as comparing with conventional MMSE-FDE and TDE methods of using GI when changing the normalized Doppler frequency R_D at the operation C/N=20dB. In the figure, the BER performance for the proposed TDE method of using the ideal CIR at every sampling time is also shown for the purpose of comparison with the proposed TDE method of using the estimated CIR. From the figure, it can be observed that the proposed TDE method with the estimated CIR shows better BER performance than the conventional TDE method with the estimated CIR. The proposed method can keep better performance until the normalized Doppler frequency R_D =20% with a little degradation of BER performance as compared with that in the static condition (R_D =0%). On the contrary the BER performance of conventional MMSE-FDE method is degraded drastically when the normalized Doppler frequency RD is larger than 5% (vehicle speed ≈ 95km/hr.). The conventional TDE method shows better BER performance than the conventional MMSE-FDE method when the normalized Doppler frequency RD is larger than 6%. However, the conventional TDE method shows worse BER performance than the MMSE-FDE method in the slow time-varying fading channel ($R_D < 5\%$). This is the reason that the conventional TDE method of using GI cannot obtain the frequency diversity gain like the MMSE-FDE method. On the contrary the proposed TDE method without GI which can compensate the ISI by using the estimated CIR and the known TS at the receiver shows the better BER performance than the MMSE-FDE method even in the slow time-varying fading channel. From these results, it can be concluded that the proposed TDE method without GI can achieve better BER performance than the conventional MMSE-FDE and TDE methods especially in highly time-varying fading channel. From the above results, it can be expected that the proposed method can be employed in the high speed vehicle to vehicle (V2V) communications even when both cars are running in the opposite direction. For instance, when both cars are running at the speed of 150km/hr in the opposite direction, the relative speed becomes 300km/hr. The proposed method can be also employed in the data communications for the users in the high speed train running at higher than the speed of 300km/hr.

Figs. 9 and 10 show the BER performances for the proposed TDE method as comparing with conventional MMSE-FDE and TDE methods when changing C/N at the normalized Doppler frequency R_D =10% (vehicle speed ≈ 190km/hr.) and R_D =15% (vehicle speed ≈ 286km/hr.), respectively. From the figures, it can be observed that the BER performance for the proposed TDE method shows much better BER performance than the conventional MMSE-FDE and TDE methods.

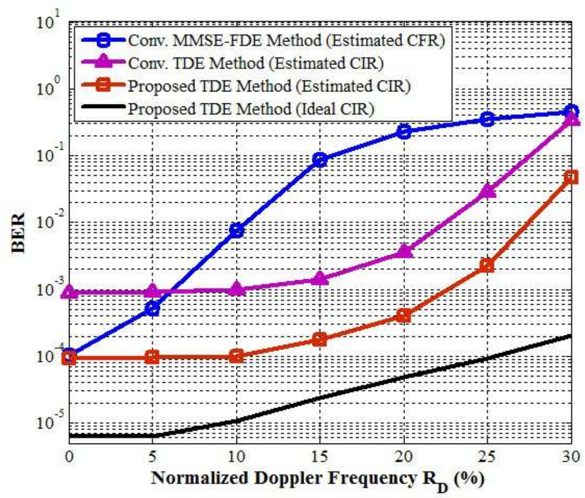

Fig.8: *BER performance for proposed TDE method when changing R_D at C/N=20dB.*

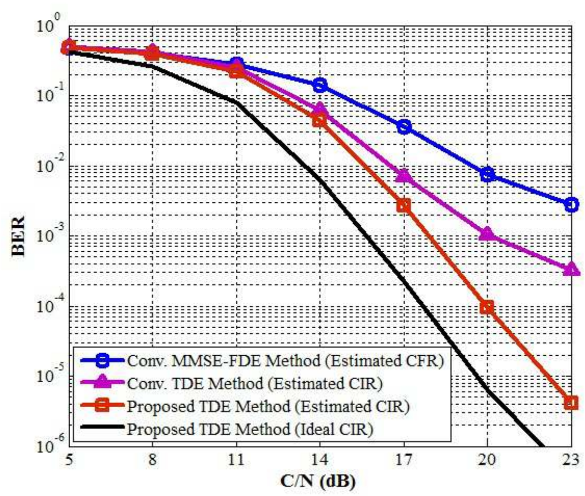

Fig.9: *BER performance for proposed TDE method when changing C/N at R_D =10% (vehicle speed ≈ 190km/hr.).*

6. CONCLUSION

This paper proposed the ML based TDE method in conjunction with the time domain CIR estimation method for the DFTS-OFDM signal without GI. The salient feature of proposed method is to employ a time domain training sequence (TS) both for the estimation of CIR at every sampling time and for removing the inter-symbol interference (ISI). The proposed ML based TDE method can track the time-varying CIR at every sampling time basis for the received time domain signal after removing the ISI. From the various computer simulation results, it can be concluded that the proposed ML based TDE method with the CIR estimation method by using the time domain training sequence (TS) can achieve much better BER performance than the conventional MMSE-FDE and TDE methods of using GI in highly time-varying fading channel.

ACKNOWLEDGMENT

The authors would like to thank to the Japanese Government (Monbukagakusho: MEXT) Scholarships who supported this research.

References

[1] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," *IEEE Commun. Mag.*, vol. 40, pp. 58-66, Apr. 2002.

[2] N. Benvenuto and S. Tomasin, "On the comparison between OFDM and single carrier modula-

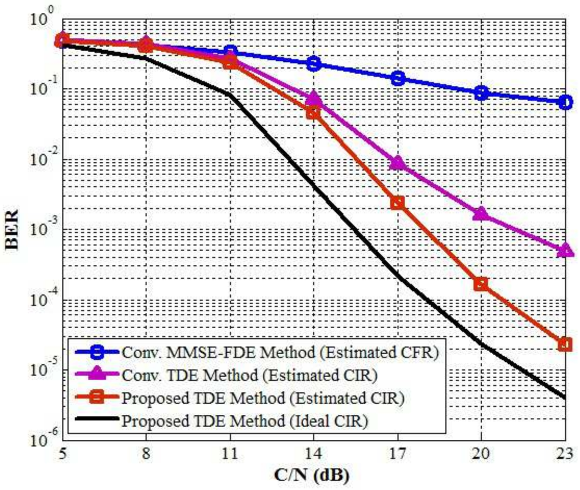

Fig.10: *BER performance for proposed TDE method when changing C/N at R_D =15% (vehicle speed ≈ 286km/hr.).*

tion with a DFE using a frequency-domain feed forward filter," *IEEE Trans. Commun. Mag.*, vol. 50, pp. 947-955, Jun. 2002.

[3] H. Kobayashi, T. Fukuhara, H. Yuan, and Y. Takeuchi, "Proposal of single carrier OFDM technique with adaptive modulation method," *in proc. of 57th IEEE on vehicular technology conference (VTC-Spring)*, vol. 3, pp. 1915-1919. Apr. 2003.

[4] F. Adachi, H. Tomeba, and K. Takeda, "Frequency-Domain Equalization for broadband Single-Carrier Multiple Access," *IEICE Trans. Commun.*, vol. E92-B, no. 5, pp. 1441-1456, May 2009.

[5] D. Y. Seol, U. K. Kwon, and G. H. Im, "Performance of single carrier transmission with cooperative diversity over fast fading channels," *IEEE Trans. Commun.*, vol. 57, no. 9, pp. 2799-2807, Sep. 2009.

[6] 3GPP TS 36.211, Evolved Universal Terrestrial Radio Access (EUTRA); Physical Channels and Modulation, 3GPP Standard, Rev. 10.0.0, 2011.

[7] A. Ghosh, R. Ratasuk, B. Mondal, N. Mangalvedhe, and T. Thomas, "LTE-advanced: next-generation wireless broadband technology," *IEEE Wireless Commun.*, vol. 17, no. 3, pp. 10-22, Jun. 2010.

[8] F. Hlawatsch, and G. Matz, "Wireless Communications Over Rapidly Time-Varying Channels," *Academic Press*, 2011.

[9] L. Yang, G. Ren, B. Yang, and Z. Qiu, "Fast Time-Varying Channel Estimation Technique for LTE Uplink in HST Environment," *IEEE Trans. Vehicular Technology*, vol. 61, no. 9, Nov. 2012.

[10] Y. Qi, P. Huang, and M. Rong, "Performance Comparison of TDE and FDE in Single-carrier System," *in proc. of 69th IEEE on vehicular technology conference (VTC-Spring)*, pp. 1-5. Apr. 2009.

[11] S. Yamasaki, and D. K. Asano, "Single-carrier transmission frequency-domain equalization based on a wiener filter," *in proc. of International Symposium on Communication and Information Technologies (ISCIT)*, pp.683-688, Oct. 2010.

[12] P. Reangsuntea, P. Boonsrimuang, K. Mori, and H. Kobayashi, "Time Domain Equalization Method for DFTS-OFDM Signal without GI under High Mobile Environments," *in proc. of 12th International Conference on Electrical Engineering/Electronics, Computer, Telecommunications and Information Technology (ECTI-CON)*, Jun. 2015.

[13] Andrew J. Viterbi, "Error Bounds for Convolutional Codes and an Asymptotically Optimum Decoding Algorithm," —em IEEE Transactions on Information Theory, Volume IT-13, pp. 260-269, Apr. 1967.

[14] P. Reangsuntea, P. Boonsrimuang, K. Mori, and H. Kobayashi, "Iterative Based Time Domain Equalization Method for OFDM Signal under High Mobile Environments," *in proc. of 8th International Conference on Signal Processing and Communication Systems (ICSPCS2014)*, pp. 1-6, Dec. 2014.

Pongsathorn Reangsuntea received the B. Eng and M. Eng degrees in Telecommunications Engineering from King Mongkuts Institute of Technology Ladkrabang (KMITL), Thailand in 2010, and 2012, respectively. He is currently working toward the Ph.D. degree at Mie University, Japan. His research interests include high-speed mobile communications and wireless LAN systems. He received IEEE VTS Japan 2015 Young Researcher's Encouragement Award in VTC2015-Spring, Glasgow, UK, May 2015.

Pisit Boonsrimuang received the B. Eng, M. Eng, and Doctor degrees in Telecommunications Engineering, in 1997, 2000, and 2007, respectively. He is currently an associate professor at the King Mongkuts Institute of Technology Ladkrabang (KMITL), Thailand. His research interests include transmission techniques for future multimedia wireless LAN systems and next generation of mobile communication systems. He received the Student Award of Young Research's Encouragement Award from IEICE Tokai branch and Doctor Thesis Award for information technology from Nation Research Council of Thailand (NRCT) in 2005 and 2008, respectively.

Kazuo Mori received the B.E. degree in computer engineering from Nagoya Institute of Technology, Japan, in 1986 and received the Ph.D. degree in information electronics engineering from Nagoya University, Japan in 2000. In 1986, He joined the Hyper-media Research Center, SANYO Electric Co., Ltd. and was engaged in research and development on telecommunication systems. From 1995 to 2000, he was a research engineer at YRP Mobile Telecommunications Key Technology Research Laboratories Co., Ltd., where he was engaged in research on mobile communication systems. Since August 2000 he has been with Mie University, Japan, where he is currently a Professor in the division of Electrical and Electronic Engineering, Graduate School of Engineering. During 2005–2006, he was a Visiting Research Fellow at King's College London, UK. His research interests include mobile communication systems, wireless sensor networks, radio resource management, and teletraffic evaluation. He received the Excellent Paper Award from IEICE, Japan in 2002. Dr. Mori is a member of IEEE.

Hideo Kobayashi received the B.E., M.E., and Ph.D in 1975, 1977 and 1989, respectively from Tohoku University in Japan. He joined KDD in 1977, and engaged in the research and development of digital fixed satellite and mobile satellite communications systems. From 1988 to 1990, he was with International Maritime Satellite Organization (INMARSAT) at its London Headquarters as the Technical Assignee and involved in the design and development of Inmarsat-M communications system. From 1997 to 1998, he was with the NEC ICO project office as the technical consultant at its London office and involved in the development of the ICO personal satellite communications system. Since 1998 he has been the Professor of Graduate School of Engineering, Mie University in Japan. His current research interests include mobile communications and wireless LAN systems. Dr. Kobayashi is a member of IEEE.