

Low Complexity Equalization Method for ZP-OFDM under Highly Mobile Environments

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Abstract—Many Interferences from multipath fading channels in the mobile channels, there are many types of research trying to compensate for such interferences. The Zero-padded OFDM (ZP-OFDM) is a popular method that has much attention as an efficient technique to improve the performances. In the high-speed mobile channel, the Doppler frequency destroyed orthogonality among subcarriers and inter-carrier interference (ICI), It would degrade the bit error rate (BER) performance. We proposed both equalization (TDE) method and CIR estimation with less computation complexity for the ZP-OFDM signal that can solve this problem. The salient feature of our proposed method is using a training sequence (TS) instead of using Zero-padded which enables for estimation of CIR. And we show that the proposed method shows the less computational complexity when compared with the conventional frequency-domain equalization, the proposed TDE method employs an iterative solution instead of using the direct inverse matrix method. We validate the proposed methods presented by computer simulations under highly mobile environments. The simulation results show a comparison of the effectiveness of the proposed TDE method and conventional TDE methods.

Keywords— ZP-OFDM, ICI, Training sequence (TS), MMSE-TDE, MMSE-FDE

I. INTRODUCTION

Recently, the Zero-padded OFDM (ZP-OFDM) systems have an excellent performance to compensate for the multipath fading channels and more interest that commonly used cyclic-prefix OFDM (CP-OFDM) [1]. The OLA, LS or MMSE equalizers require the CIRs at every sampling time for compensation the ICI in the ZP-OFDM signal. The that the CIRs at every sampling time is known perfect [2-5] for each OFDM symbol at the receiver. The time-domain CIR estimation technique is considering an efficient method for usage of frequency bandwidth, especially under highly mobile environments.

In [6], we proposed the ICI mitigation method both for channel estimation and TDE methods under highly mobile environments. The proposed TDE method can compensate both of inter-symbol interference (ISI) and ICI. However, in demodulation processing, this method requires calculating the direct inverse at every OFDM symbol, which leads to high computation complexity and not suitable in a practical system.

The proposed method enables us to use the CIR estimation in compensation of both inter-symbol interference (ISI) and ICI. In contrast to the existing TDE method with direct inverse matrix calculation, our proposed method operates an iterative method calculation [7] in the time domain, which allows a low-complexity implementation. The proposed low complexity TDE method in conjunction with the CIR estimation method of using a time-domain TS, which can achieve BER performance similarly as compared with conventional demodulation with direct inverse matrix calculation for ZP-OFDM even under highly mobile environments.

This paper is organized as follows. Section II shows the proposed system model. Section III proposes the CIR estimation and low complexity TDE method in highly mobile environments which suitable for our transmission systems. Section IV, presents various computer simulation results to verify the effectiveness of the proposed method as compared with conventional methods, and Section V draws some conclusions.

II. SYSTEM MODEL

The proposed OFDM method can be modeled as the baseband equivalent system that shows in figure 1. The input data input to the forward error correction (FEC) code [8] and modulation. The quadrature-amplitude modulation (QAM) modulates the encoded K data. $X_D(m,k)$ is the modulated signal at the k -th subcarrier of m -th symbol, which mapped into the N subcarriers in the frequency bandwidth from the subcarrier number N_{Z1} to N_{Z2} ($N_{Z2}-N_{Z1}+1=K$). The null subcarriers (N_Z : zero paddings) are added at both ends of K data subcarriers, and It can reject the aliasing occurring at the output of digital to analog (D/A) converter. After that, the signal $X(m,k_1)$ over N subcarriers is transformed to time-domain signal by N -point IDFT which can be expressed by,

$$x_T(m, n_1) = \frac{1}{\sqrt{N}} \sum_{k_1=N_{Z1}}^{N_{Z2}} X(m, k_1) \cdot e^{j \frac{2\pi n_1 k_1}{N}}, 0 \leq n_1 \leq N-1 \quad (1)$$

where $x_T(m, n_1)$ is an OFDM symbol in the time domain passed the IDFT processing at the n_1 -th sampling time of m -th symbol. In the frame structure, the time domain TS [6] is added at both ends of every data symbol. The added TS will be used for both the estimation of CIR and use as the role of CP or ZP, which is

typically used in the conventional CP-OFDM or ZP-OFDM signal. It can help to avoid the ISI occurred in the multipath fading channels. The frame structure of the proposed method, the observation period for data demodulation is more extended than N samples, in which the system performance is increased when the length of TS1 and TS2 is increasing. The transmitted time-domain signal $x(m, n_2)$ at the n_2 -th sampling time of m -th symbol inserted both TS1 and TS2 can be shown by equation,

$$x(m, n_2) = \begin{cases} d_1(m, n_2), & 0 \leq n_2 \leq N_{TS} - 1 \\ x_T(m, n_2 - N_{TS}), & N_{TS} \leq n_2 \leq N + N_{TS} - 1 \\ d_2(m, n_2 - N - N_{TS}), & N + N_{TS} \leq n_2 \leq N + 2N_{TS} - 1 \end{cases} \quad (2)$$

where $d_1(m, n_2)$ and $d_2(m, n_2 - N - N_{TS})$ are the identical time-domain TS1 and TS2. The TS1 and TS2 are the known data patterns at the receiver. For simplicity, we assume the known data patterns both for TS1 and TS2 are the same patterns as $d(m, n)$. The length of N_{TS} is the same condition both CP-OFDM and ZP-OFDM signal, which should be taken longer than the sample of delay paths (L). we assume the timing synchronization for the received signal is perfect at the receiver, the received time-domain OFDM signal $y(m, n_2)$ at n_2 -th sampling time of m -th symbol after passing through the multipath fading channels can be formulated by the equation,

$$y(m, n_2) = \sum_{l=0}^{L-1} h_l(m, n_2) \cdot x(m, n_2 - l) + z(m, n_2), \quad (3)$$

$$0 \leq n_2 \leq N + 2N_{TS} - 1$$

where $h_l(m, n_2)$ is the CIR with time-varying fading channels, $x(m, n_2 - l)$ is the data symbol in the time domain, and $z(m, n_2)$ is additive white Gaussian noise (AWGN) of variance σ^2 . From (3) and Fig. 2, the received time-domain signal $y(m, n_2)$ can be separated into two parts. In the first part, $y_{TS}(m, n_2)$ is considered from 0 to $N_{TS} - 1$ period for the CIR estimation. Other parts, the data demodulation $y_D(m, n_2)$ is considered from 0 to $N + N_{TS} - 1$ period, which shows the details in Fig.2.

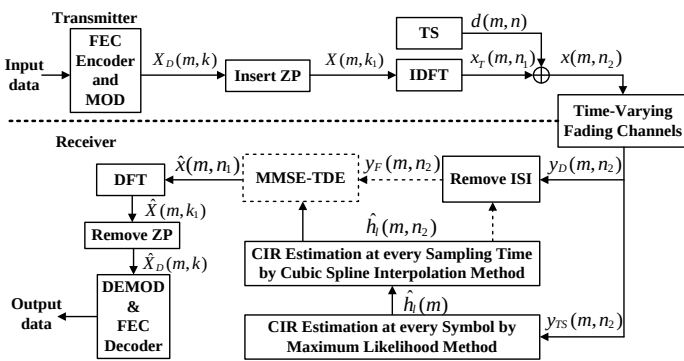


Figure 1. System model for proposed MMSE-TDE method.

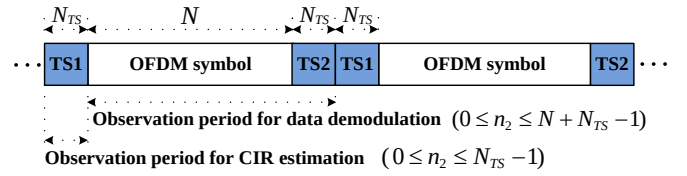


Figure 2. The proposed OFDM signal of frame structure with TS1 and TS2.

III. PROPOSED METHOD

A. Proposed CIR Estimation Method

From Eq. (3), it assumes that the short period of TS1 and TS2 are the constant under the highly time-varying fading channels. The received time-domain TS signal $y_{TS}(m, n_2)$ passed through the unknown CIR $\hat{h}_l(m)$, it can be given by the following equation,

$$\hat{y}_{TS}(m, n_2) = \sum_{l=0}^{N_{TS}-1} \hat{h}_l(m) \cdot d(m, n_2 - l), \quad 0 \leq n_2 \leq N_{TS} - 1 \quad (4)$$

The $\hat{h}_l(m)$ is unknown parameters of CIR at the m -th symbol, which can be estimated by solving the Maximum Likelihood (ML) equation. The differentiation from the actual received TS signal $y_{TS}(m, n_2)$ and the expected received TS signal $\hat{y}_{TS}(m, n_2)$ is the CIR estimation, which can express as the following equation,

$$Y = \arg \min_{\hat{h}_l(m)} \left[\sum_{n_2=0}^{N_{TS}-1} |y_{TS}(m, n_2) - \hat{y}_{TS}(m, n_2)|^2 \right] \quad (5)$$

Then, we employ a simple solution for solving ML equation in Eq.(5) by the following equations.

$$y_{TS}(m, n_2) = \hat{y}_{TS}(m, n_2) \quad \text{for} \quad 0 \leq n_2 \leq N_{TS} - 1 \quad (6)$$

By substituting Eq.(4) into Eq.(6). And the result can be shown by the following equation,

$$[\hat{h}_l(m)] = [d(m, n_2 - l)]^{-1} \cdot [y_{TS}(m, n_2)] \quad (7)$$

where $[\cdot]^{-1}$ is inverse matrix operation. $\hat{h}_l(m)$ is the CIR estimation every symbol and N_{TS} is the actual number of delay paths that have not known at the receiver in the real channel. Here it should be noted that the inverse matrix of $[d(m, n_2 - l)]^{-1}$ in (7) can be calculated in advance from the known data pattern of TS $d(m, n_2)$ at the receiver and it can be used for every symbol of the received data. By using the $\hat{h}_l(m)$ of every symbol, we can estimate the CIR at every sampling time $\hat{h}_l(m, n_2)$ from the cubic spline interpolation function of mathematics over several estimated CIRs of $\hat{h}_l(m)$. The estimation accuracy of time-domain CIR at every sampling time will be evaluated in section IV.

B. Proposed TDE Method with Direct Inverse Matrix

The received demodulated time-domain OFDM signal in high-speed mobile environments can be given by,

$$y_D(m, n_2) = \sum_{l=0}^{L-1} h_l(m, n_2) \cdot x(m, n_2 - l) + z(m, n_2), \quad (8)$$

$$0 \leq n_2 \leq N + N_{TS} - 1$$

From (8), the received time-domain signal $y_D(m, n_2)$ includes the ISI of TS, it consists of the start and end of data symbol from the TS1 and TS2, respectively. By using the estimated CIR $\hat{h}_l(m, n_2)$ at every sampling time and the $d(m, n_2)$ is the known data pattern both for TS1 and TS2, the ISI of TS added at both ends of the received signal can be removed by the following equation.

$$y_F(m, n_2) = \begin{cases} y_D(m, n_2) - \sum_{l=n_2+1}^{N_{TS}-1} \hat{h}_l(m, n_2) \cdot d(m, n_2 + N_{TS} - l), & (0 \leq n_2 \leq N_{TS} - 2) \\ y_D(m, n_2), & (N_{TS} - 1 \leq n_2 \leq N - 1) \\ y_D(m, n_2) - \sum_{l=0}^{n_2-N} \hat{h}_l(m, n_2) \cdot d(m, n_2 - N - l), & (N \leq n_2 \leq N + N_{TS} - 2) \end{cases} \quad (9)$$

When every data symbol is demodulation, we omitted the parameter m -th symbol. The received time domain can be simply in vector equation as shown below,

$$y_F = h x + z \quad (10)$$

Here $y_F = [y_F(0), \dots, y_F(n_2), \dots, y_F(N+N_{TS}-2)]^T$ is the received signal vector in the time-domain, where $[\cdot]^T$ represents the transpose of matrix operation. $x = [x(0), \dots, x(n_2), \dots, x(N-1)]^T$ is the transmitted signal vector in the time-domain, which equate the $x_T(m, n_1)$ parameter in (1). Furthermore, h is the CIR matrix in the time-domain after removing ISI with the size of $(N+N_{TS}-1) \times (N)$ which equivalent to the CIR matrix of the ZP-OFDM signal [3-5]. From (10), the conventional OLA, LS and MMSE equalizers operate both in the frequency-domain or in the time-domain that can be applied for demodulation of the transmitted data information. This paper considers a time-domain MMSE ICI equalization, which calculates coefficient equalization using the linear equalizer matrix W , then $W y_F$ is closest to x in the MMSE sense can be expressed by,

$$W_{MMSE} = \arg \min_W \{E |W y_F - x|^2\} \quad (11)$$

The solution for minimization is obtained equation as

$$W_{MMSE} = (h^H h + \sigma^2 I)^{-1} h^H \quad (12)$$

where $(\cdot)^H$ is the Hermitian transpose matrix operation, σ^2 is a variance of AWGN, and I is an identity matrix. By using the estimated CIR $\hat{h}_l(m, n_2)$ at every sampling time, the MMSE equalizer output equals

$$\hat{x}_{MMSE} = W_{MMSE} y_F = (h^H h + \sigma^2 I)^{-1} h^H y_F \quad (13)$$

where $\hat{x}_{MMSE}$ is an expected transmitted time-domain signal.

From (13), the expected transmitted frequency domain signal $\hat{X}_{MMSE}$ can be obtained by converting the time domain data signal to frequency domain signal with N -point DFT. From (13), it also can be seen that the equalization in the time domain (TDE) shows the simplicity structure, especially under a highly mobile environment because of it easy to construct the CIR matrix. However, the total computational complexity for the proposed MMSE-TDE method is still high with order $N^3 + N^2$ at every OFDM symbol which may not be suitable for a practical system.

C. Proposed TDE with Iterative Method

To solve the high complexity problem, we propose an iterative method for solving the simultaneous equation in (13). From (13), the unknown parameters $\hat{x}_{MMSE}$ can be solved by using the inverse matrix of $(h^H h + \sigma^2 I)$ which be expressed by,

$$\underbrace{[\hat{x}(m, n_1)]}_{N \times 1} = \underbrace{[A_m(s, n_1)]}_{N \times N}^{-1} \cdot \underbrace{[b(m, s)]}_{N \times 1} \quad (14)$$

where $\hat{x}(m, n_1)$ is expected transmitted time domain signal with $N \times N$ matrix, $[A_m(s, n_1)]$ is the term of $(h^H h + \sigma^2 I)$, and $[b(m, s)]$ is the term of $(h^H y_F)$ as in (13), respectively. The coefficients matrix $[A_m(s, n_1)]$ in (14), which is the square matrix with a size of $N \times N$. This means that matrix $[A_m(s, n_1)]$ can be solved by using an iterative calculation. The conjugate gradient squared (CGS) algorithm [7] is well known as one of the iterative methods which can solve a system of N linear equations with N unknown parameters. Let us consider the system $A \hat{x} = b$, where A corresponds to $A_m(s, n_1)$ in (14) and it is the Hermitian and positive definite matrix with the size of $N \times N$. The exact CGS solution can be obtained after at most N steps. Hence, stopping the iteration after $N_{iter} (< N)$ steps would yield an approximate solution. In the demodulation of every data symbol, the CGS algorithm minimizes the cost function in a reduced-rank Krylov subspace iteratively. When the spectral condition number of the matrix A is too high, a preconditioned matrix D is employed which is called the precondition CGS (PCGS) algorithm. The PCGS algorithm solves the system by,

$$D^{-1} A \hat{x} = D^{-1} b \quad (15)$$

where the inversion of matrix D should be a computationally efficient operation. For the rest of our analysis, we assume for simplicity that matrix $D(m)$ is the diagonal matrix of $A_m(s, n_1)$ at m -th symbol. The initial solution of $\hat{x}(m, n_1)$ is given by,

$$\underbrace{[\hat{x}(m, n_1)]^{(0)}}_{N \times 1} = \underbrace{[D(m)]}_{N \times N}^{-1} \cdot \underbrace{[b(m, s)]}_{N \times 1} \quad (16)$$

The reduced-rank PCGS method is described analytically in Table 1. The repetition of the PCGS algorithm is stopped when the following normalized mean square error (NMSE) between the i -th and $(i+1)$ -th solutions of $[\hat{x}(m, n_1)]$ is smaller than the predetermined threshold level (TOL).

$$NMSE = \frac{\sum_{n_1=0}^{N-1} \left| [\hat{x}(m, n_1)]^{(i)} - [\hat{x}(m, n_1)]^{(i+1)} \right|^2}{\sum_{n_1=0}^{N-1} \left| [\hat{x}(m, n_1)]^{(i)} \right|^2} \quad (17)$$

In the proposed iterative method with the PCGS algorithm, the following procedures are repeated up to either the value of $NMSE$ achieve better than the predetermined value of threshold level (TOL) or the number of iterations reaches to the predetermined maximum number (N_{max}).

- Step1: The maximum iteration number is set to N_{max} ($N_{max}=5$), the threshold level is set to TOL and the initial solution of $[\hat{x}(m, n_1)]^{(0)}$ is given by (43).
- Step2: Calculate the i -th solution of $[\hat{x}(m, n_1)]^{(i)}$ by the PCGS algorithm and calculate the $NMSE$ by (17).
- Step3: Compare the $NMSE$ obtained at the i -th iteration with the predetermined threshold level of TOL . If the $NMSE$ is less than TOL , the $[\hat{x}(m, n_1)]^{(i)}$ is output as the estimated data information. If not, repeat the same procedures. If the number of iterations reaches to predetermined N_{max} , $[\hat{x}(m, n_1)]^{(N_{max})}$ is output as the estimated data information.

From Table I, the computational complexity is determined by initialization, line 5 and line 12. Specifically, the total computational complexity order for the demodulation of each OFDM symbol can be expressed by $N_{iter}(2N^2+10N)+N^2+2N$. The following equation defines the ratio of computation complexity between TDE with the inverse matrix and TDE with the iterative method for the whole OFDM frame.

$$R = \frac{N_S \cdot \{N_{Aver} \cdot (2N^2 + 10N) + N^2 + 2N\}}{N_S \cdot \{N^3 + N^2\}} \quad (18)$$

$$= \frac{N_{Aver} \cdot (2N^2 + 10N) + N^2 + 2N}{N^3 + N^2}$$

TABLE I PRECONDITIONED CONJUGATE GRADIENT SQUARED (PCGS) ALGORITHM

PCGS Algorithm	Complexity
Initialization :	
$\hat{x}_0$ is an initial guess;	
$r_0 = b - A\hat{x}_0$;	$O(N^2)$
$\tilde{r}_0 = r_0$; $\rho_0 = (r_0, \tilde{r}_0)$;	$O(N)$
$\beta_{-1} = \rho_0$; $p_{-1} = q_0 = 0$;	$O(N)$
1. for $i = 0, 1, 2, \dots, N_{max}$ do	
2. $u_i = r_i + \beta_{i-1} q_i$;	$O(N)$
3. $p_i = u_i + \beta_{i-1} (q_i + \beta_{i-1} p_{i-1})$;	$O(N)$
4. $\hat{p} = D^{-1} p_i$;	$O(N)$
5. $\hat{v} = A\hat{p}$;	$O(N^2)$
6. $\alpha_i = \rho_i / (\tilde{r}_0, \hat{v})$;	$O(N)$
7. $q_0 = u_i - \alpha_i \hat{v}$;	$O(N)$
8. $\hat{u} = D^{-1} (u_i + q_{i+1})$;	$O(N)$
9. $\hat{x}_{i+1} = \hat{x}_i + \alpha_i \hat{u}$;	$O(N)$
10. if $NMSE$ in (17) > Threshold level (TOL)	$O(N)$
11. return; else	
12. $r_{i+1} = r_i - \alpha_i A\hat{u}$;	
13. $\rho_{i+1} = (r_0, \tilde{r}_{i+1})$;	$O(N^2)$
14. $\alpha_i = \rho_{i+1} / \rho_i$;	$O(N)$
15. end for	$O(N)$

where N_S is a number of data symbols in one OFDM frame and N_{Aver} is the average number of iterations in one OFDM frame.

IV. PERFORMANCE EVALUATIONS

This section explains the various computer simulation to verify the effectiveness of proposed TDE with a direct inverse matrix method [6] as compared with the proposed TDE with an iterative method for ZP-OFDM. For both cases, we employ the convolutional code is taken by 1/2. A generator encodes the data information in the octal form as (133, 171) while decoding is done by a hard decision Viterbi decoder.

The modulation in the ZP-OFDM employed the 16QAM, $M=64$ subcarriers, $K=96$ within frequency bandwidth 1 MHz, and the radio frequency operates at 5.9 GHz. The number of DFT/IDFT takes by $N=128$ points, the length of zero-padded samples and time-domain TS are the same as $N_{TS}=16$. The communication channel is modeled by the Rayleigh multipath fading channels having a uniform power delay profile with a length of delay paths $L=14$. In the following performance evaluations, the normalized Doppler frequency $R_D = f_{dmax}/\Delta f$ (%) is employed as the definition of high-speed mobile conditions where f_{dmax} is the maximum Doppler frequency and Δf is the subcarrier spacing. Here, it should be noted that the conventional 1-Tap FDE and TDE of the OLA methods are also shown so as to compare with the proposed TDE method.

The normalized mean square error evaluates the estimation accuracy of the proposed CIR estimation method of using the time domain TS at every sampling time, which show in figure 3. It shows the normalized MSE when changing the normalized Doppler frequency R_D and C/N is taken by 16 dB, 18 dB, and 20 dB, respectively. The simulation results when assume R_D is up to 20% (vehicle speed ≈ 381 km/hr.) at lower operation C/N that the proposed method can show higher estimation accuracy. From these simulation results, we can be confirmed that our proposed CIR estimation can mitigate both ISI and ICI precisely even under high-speed mobile environments. From the figure, the proposed method has a higher accuracy for the estimated CIR when R_D is up to 20% (approximate speed ≈ 381 km/hr.) at lower operation C/N. From these results, we confirmed that the proposed method can be used to mitigate both the ISI and ICI under highly mobile environments.

Fig. 4 shows the BER performance for the proposed MMSE-TDE with iterative method when changing the threshold level (TOL) of $NMSE$ which to be used for stopping the iteration. In the simulation, the normalized Doppler frequency R_D is 15% which corresponds to the vehicle speed 286 km/h. In the figure, BER performance for the proposed TDE with the inverse matrix calculation method is also shown as the purpose of comparison with the proposed TDE with the iterative method. From Fig. 4, it can be observed that the BER performance of proposed TDE with the iterative method is very close to that for the inverse matrix calculation method when the threshold level TOL is less than 0.01 with regardless of operation C/N. From the results in Fig. 4, the threshold level TOL for the $NMSE$ is taken by $TOL=0.01$ in the following evaluations.

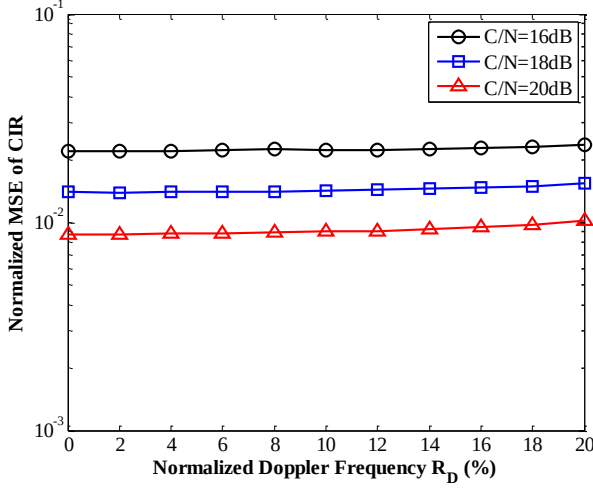


Figure 3. CIR estimation accuracy (NMSE) when changing the Doppler Frequency.

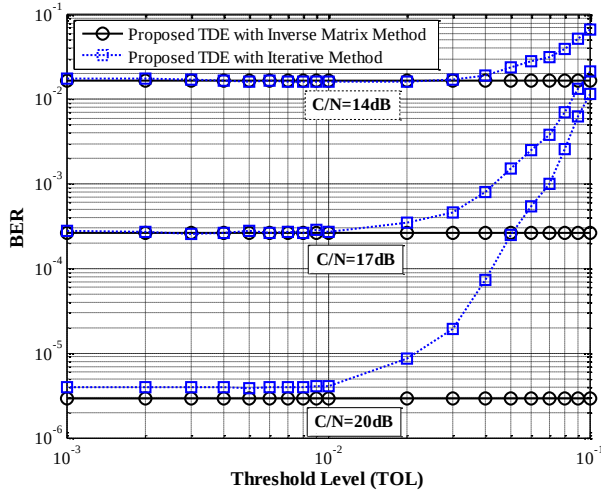


Figure 4. BER performance for the proposed MMSE-TDE method when changing TOL at $R_D = 15\%$.

Fig. 5 shows the average number of iterations for the proposed TDE with iterative method when changing the normalized Doppler frequency R_D . From the simulation results in Fig. 5, it can be observed that the average number of iterations N_{Aver} of one OFDM frame becomes larger when increasing normalized Doppler frequency R_D and decreasing the operation C/N. By using the average number of iterations obtained in Fig. 5 and equation (18), Table II shows the ratio of computation complexity (R) between the proposed TDE with the iterative method and the inverse matrix method. From simulation parameters in table 1, it can be concluded that the proposed TDE with the iterative method can reduce the computation complexity to 0.0782 (Approximately 7.8%) at C/N is 14dB and R_D is 20% which correspond to the vehicle speed 381 km/hrs as compared with inverse matrix calculation method.

Fig. 6 shows the BER performances for the proposed TDE method both of the iterative and inverse matrix calculation

methods when changing C/N at $R_D = 15\%$ (Vehicle speed 286 km/hrs). From Fig. 6, it can be observed that the proposed TDE method can achieve much better BER performance than the conventional FDE method and the TDE with the OLA method [6]. From all the simulation results, we conclude that the proposed TDE with the iterative method can achieve less computation complexity and keeping the same BER performance as the inverse matrix calculation method.

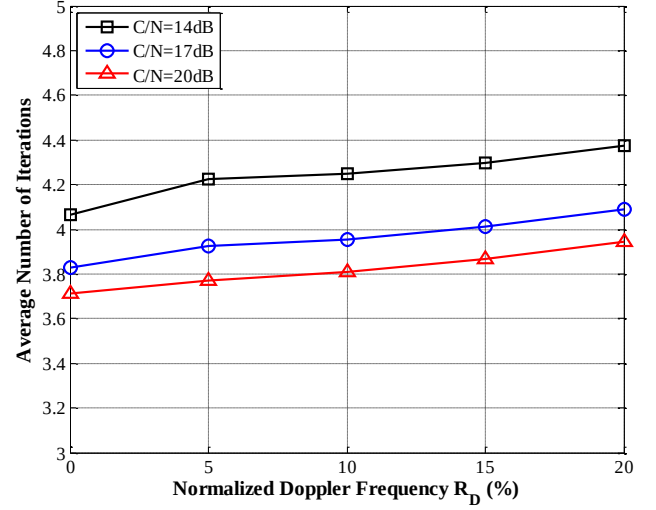


Figure 5. An average number of required iterations for the proposed method when changing R_D at $TOL = 0.01$.

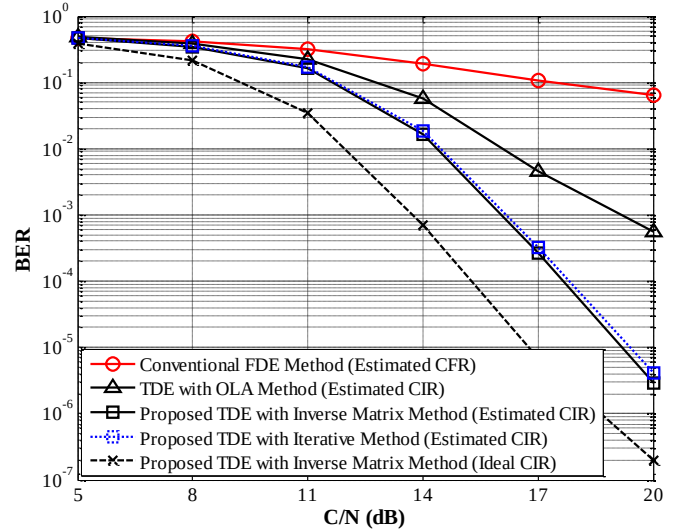


Figure 6. BER performance for the proposed method when changing R_D at C/N=20dB.

TABLE II COMPARISON OF COMPUTATION COMPLEXITY

C/N	Proposed iterative based TDE method ($N=128, TOL=0.01$)					
	$f_{dmax}/\Delta f = 10\%$		$f_{dmax}/\Delta f = 15\%$		$f_{dmax}/\Delta f = 20\%$	
	N_{Aver}	R	N_{Aver}	R	N_{Aver}	R
14dB	4.25	0.076	4.29	0.076	4.37	0.078
17dB	3.95	0.071	4.01	0.072	4.08	0.073
20dB	3.81	0.069	3.86	0.070	3.94	0.071

V. CONCLUSIONS

This paper proposes a low-complexity TDE method in conjunction with the CIR estimation method at every sampling time. By using the construction of the simultaneous equation, the coefficients matrix becomes the square matrix, which can be solved iteratively to enable much lower computation complexity. We use computer simulation to validate the performances of the proposed method compared with the conventional methods. From the computer simulation results, we conclude that the proposed method has the low-complexity TDE method and can achieve much better BER performance when compared with the conventional FDE and TDE with OLA methods even in highly mobile environments.

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LAN Systems.



mobile communications and wireless LAN systems.

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