國立交通大學

電子工程學系　電子研究所

碩士論文

在時變多路徑通道之低複雜度載波間互擾效應消除於正交分頻多工系統

A Low-Complexity ICI Cancellation Scheme For OFDM In Time-Varying Multipath Channel

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中華民國九十六年一月

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在時變多路徑通道之低複雜度載波間互擾效應消除於正交分頻多工系統

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摘要

本篇論文應用在正交分頻多工（OFDM）系統於時變頻率選擇衰變通道. 隨著符號長度的增加，OFDM 可抵抗對於多重路徑干擾的效應。然而在行動通訊的應用上，時變通道破壞了子載波間的正交性，造成載波間的互相干擾（ICI）因而降低系統效能。在本篇論文中，利用線性模型去近似時變通道。基於這個假設，我們解出線性的通道參數進而重建通道矩陣，目的去消除載波間的互相干擾。模擬結果顯示所提出的方法可有效抑制因時變通道所造成的 ICI 效應以及降低錯誤率在高雜訊比仍高居不下的現象。

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ABSTRACT

This thesis considers an orthogonal frequency division multiplexing (OFDM) system over frequency selective time-varying fading channels. As the symbol duration increases, OFDM is robust to channel multipath dispersion. However, for mobile applications, channel variations within an OFDM block period destroy the orthogonality between subcarriers; the effect, known as Intercarrier Interference (ICI), will degrade the system performance. In this thesis, a linear model is assumed for the channel variation because of its simplicity. Based on the assumption, the proposed method solves the linear channel parameters of each path and creates a channel matrix which can be used for ICI cancellation. The simulation results show that the proposed method can effectively suppress the ICI effects induced by the time variant channels and reduce the error floor of bit error rate (BER).

誌謝

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Contents

LIST OF TABLES

LIST OF FIGURES

Chapter1

Introduction

1.1 Motivation

OFDM is a widely used and considered a promising technique for high speed data transmission in digital audio broadcasting (DAB) systems, digital video broadcasting (DVB)systems, and for wireless broadband access standard such as IEEE Std. 802.16 (WiMAX). OFDM is robust to channel multipath dispersion and results in a decrease in the complexity of equalizers for high delay spread. However, using conventional channel equalization without ICI compensation results in an error floor in mobile OFDM systems [1]. Channel estimation schemes for mobile OFDM systems, in contrast to stationary channel where solely the frequency selectivity has to be estimated, also have to estimate the time variation of the mobile channel. In the thesis, a new ICI reduction method is proposed, which reduces the error floor caused by the channel variation.

1.2 Literature survey

To mitigate the ICI caused by channel variation, many approaches have been proposed, e.g., polynomial cancellation coding [2], self-cancellation scheme [3], DFT-based channel estimator [4], minimum mean-squared error (MMSE) with successive detection [5]. The scheme in [5] have good performance but requires $\geq O\left(N^3\right)$ computational complexity, where N denoted the FFT size. When N becomes very large, the complexity will be impractical. In [6], the complexity of MMSE is reduced by deriving an optimal low-rank estimator with singular-value decomposition (SVD). The channel estimation techniques for

OFDM systems based on pilot arrangement are investigated in [7]. Several Doppler spectra are analyzed and compared in [8]. Some methods need the information of the Doppler frequency. A novel pilot-based estimation scheme is proposed in [9], which develops a simple Doppler frequency estimation scheme. When the OFDM symbol duration is less than 10% of the channel coherence time, [10] also argues that variation of the channel can be assumed in a linear fashion. In [11] it is concluded that if the symbol duration is less than 1% of the channel coherence time, the channel can be assumed constant during one symbol interval. In [12] it is proposed that utilize the combination of Viterbi-type Maximum Likelihood (ML) equalizer and Bessel model in pilot-aided channel estimation to mitigate ICI effect. In [13] and [14], the channel variation is respectively modeled as a basis expansion using discrete prolate spheroidal sequences (DPS) and first order Taylor approximation. In the thesis, we apply the linear model for channel variation and one-tap equalizer with ICI reduction. The proposed method has low complexity and outperforms [4] which has comparable performance to that of MMSE channel estimation method wile costs much less computation. Moreover, for practical mobile applications, the linear model is good enough for channel variation and higher order approximations are not necessary [12] [13] [14].

1.3 Thesis organization

This thesis is organized as follows: In Chapter 2, the system model over time-varying channel is described, including the ICI effect induced by channel variations. In Chapter 3, after the conventional channel estimations introduced, the new ICI reduction method is proposed. Then, the analysis of the computational complexity is shown. In Chapter 4, the simulation results are described and some conclusions are drawn at the end.

Chapter 2

OFDM Communication Systems in Wireless Channels

2.1 System model

Consider an OFDM system with N subcarriers signaling through a time-varying frequency-selective Rayleigh fading channel. The whole system in base band model is illustrated in Fig.1.

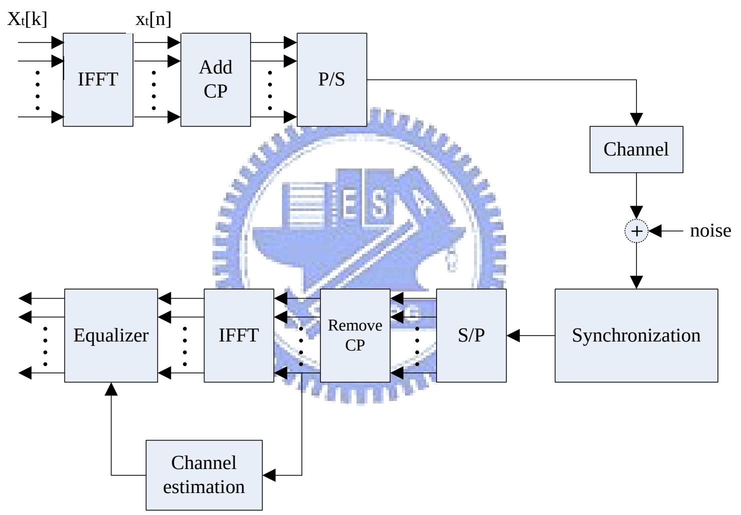

Fig. 1 System model

The data are modulated in blocks by means of a discrete Fourier transform (DFT),given by

$$x_t[n] = \sum_{k=0}^{N-1} X_t[k] \cdot \exp\left(-j\frac{2\pi kn}{N}\right) \qquad 0 \le n, k \le N-1 \tag{2.1}$$

where the subscript *t* represents the t^{th} OFDM frame. A cyclic prefix (CP) is inserted into the transmitted signal to prevent the intersymbol interference (ISI) between successive OFDM frames. After parallel to serial conversion, the signals are transmitted trough a frequency selective time-varying fading channel. At the receiver side, the received time-domain signal can be expressed as [15]

$$r_t[n]=\sum_{l=0}^{L-1} h_{n,l}^t \cdot x_t\left[|n-l|_N\right]+n_t[n] \quad 0 \le n \le N-1 \tag{2.2}$$

where $h_{n,l}^t$ represents the channel impulse response (CIR) of the l^{th} path during t^{th} OFDM frame, L represents the length of the frequency-selective fading channel, $n^t[n]$ represents the additive white Gaussian noise (AWGN) with zero mean and variance $E\left[|n_t[n]|^2\right]=\sigma^2$, and $|\bullet|_N$ represents the modulo-N operation. The fading channel coefficients $h_{n,l}^t$ are modeled as zero mean complex Gaussian random variables. According to Wide Sense Stationary Uncorrelated Scattering (WSSUS) assumption, the fading coefficients in different delay paths are statistically independent. In the time domain, the fading coefficients $h_{n,l}^t$ are correlated and have a Doppler power spectrum density modeled as in Jakes [16], given by

$$D(f)=\begin{cases} \dfrac{1}{\pi F_d\sqrt{1-\left(\dfrac{f}{F_d}\right)^2}} & |f| \le F_d \\ 0 & \text{otherwise} \end{cases} \tag{2.3}$$

where F_d is the Doppler bandwidth. Hence $h_{n,l}^t$ has an autocorrelation function given by

$$E\left[h_{n,l}^t \cdot h_{m,l}^{t\,*}\right]=J_0\left(2\pi(n-m)F_dT_s\right) \tag{2.4}$$

where $J_0(\bullet)$ is the first kind Bessel function of zero order and T_s is OFDM sample duration.. Then substitute equation (2.1) into (2.2):

$$r_t[n]=\frac{1}{N}\sum_{l=0}^{L-1}\sum_{k=0}^{N-1} X_t[k]\cdot h_{n,l}^t \exp\left(j\frac{2\pi k(n-l)}{N}\right)+n_t[n] \qquad 0 \le n \le N-1 \tag{2.5}$$

At the output of the DFT demodulator, the sequence can be expressed as

$$Y_t[m] = DFT_n\{r_t[n]\}$$
$$= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{l=0}^{L-1}\sum_{k=0}^{N-1} X_t[k]\cdot h_{n,l}^t \exp\left(j\frac{2\pi[k(n-l)+mn]}{N}\right) + w_t[n] \quad 0 \le n \le N-1 \qquad (2.6)$$

Where $Y_t[m]$ represent the received signal at the k^{th} subcarrier of the t^{th} OFDM frame and $w_t[n]$ is i.i.d complex AWGN noise due to the orthonormal transformation of the original noise $n_t[n]$. It also can be written in a concise matrix form as

$$\mathbf{Y_t} = \mathbf{H_t}\cdot\mathbf{X_t} + \mathbf{W_t} \qquad (2.7)$$

where $\mathbf{Y_t}$ and $\mathbf{W_t}$ are vectors of size $N\times 1$, and the channel matrix $\mathbf{H_t}$ is given by

$$\mathbf{H_t} = \left[H_{p,q}^t\right]_{N\times N}, \qquad H_{p,q}^t = \frac{1}{N}\sum_{n=0}^{N-1}\sum_{l=0}^{L-1} h_{n,l}^t \exp\left(j\frac{2\pi[(q-p)n-ql]}{N}\right) \qquad (2.8)$$

2.2 Inter-carrier interference in OFDM systems induced by the Doppler effect

OFDM is robust against frequency selective fading due to the increase of the symbol duration. With the introduction of the CP, the problem of the ISI in single-carrier systems can be greatly reduced. However, in mobile radio environment, multipath channels are usually time-varying. Channel variations within an OFDM block destroy the orthogonality among the subcarriers, resulting in ICI and performance degradation. Since the transmitted symbol duration is N times longer than that in a single-carrier system, the increase in the symbol duration makes the system more sensitive to the time variation of the channel. If ICI is modeled as an additive white Gaussian process and not adequately compensated, the ICI will lead to an error floor. Doppler frequency F_d is used to indicate the rate of the channel variation, which is proportional to vehicle velocity v and carrier frequency f_c, given by

$$F_d = \frac{v \cdot f_c}{c} \tag{2.9}$$

Therefore, the ICI induced by time-varying channel is determined by Doppler frequency F_d and the OFDM symbol duration T. In order to analyze the symbol energy distribution and ICI on one subcarrier, equation (2.6) can be modified as following form

$$Y_t[m] = H^t_{m,m} X_t[m] + \sum_{n=0, n \neq m}^{N-1} H^t_{m,n} X_t[m] + w_t[n] \qquad 0 \le m \le N-1 \tag{2.10}$$

The first term in (2.10) is the desired signal, the second term represents the ICI from the other subcarriers, and finally the third term is the additive noise. Hence, the energy of $X_t[n]$ leaked to the m^{th} subcarrier can be expressed as

$$\begin{aligned} P_{m,n} &= E\left[\left|H^t_{m,n} X_t[n]\right|^2\right] = E_s E\left[H^t_{m,n} H^{t\,*}_{m,n}\right] \\ &= \frac{E_s}{N^2} \sum_{k=0}^{N-1} \sum_{l=0}^{L-1} \sum_{k'=0}^{N-1} \sum_{l'=0}^{L-1} E\left[h^t_{k,l} h^{t\,*}_{k',l'}\right] \exp\left(-j\frac{2\pi}{N}\left[(l-k-l'+k')n + (k-k')m\right]\right) \\ &= \frac{E_s}{N^2} \sum_{k=0}^{N-1} \sum_{k'=0}^{N-1} J_0\left(2\pi F_d (k'-k)\frac{T_s}{N}\right) \exp\left(-j\frac{2\pi}{N}(k'-k)(n-m)\right) = P_l \end{aligned} \tag{2.11}$$

where $E_s = E[|X_t[n]|^2]$ and $l = n - m$

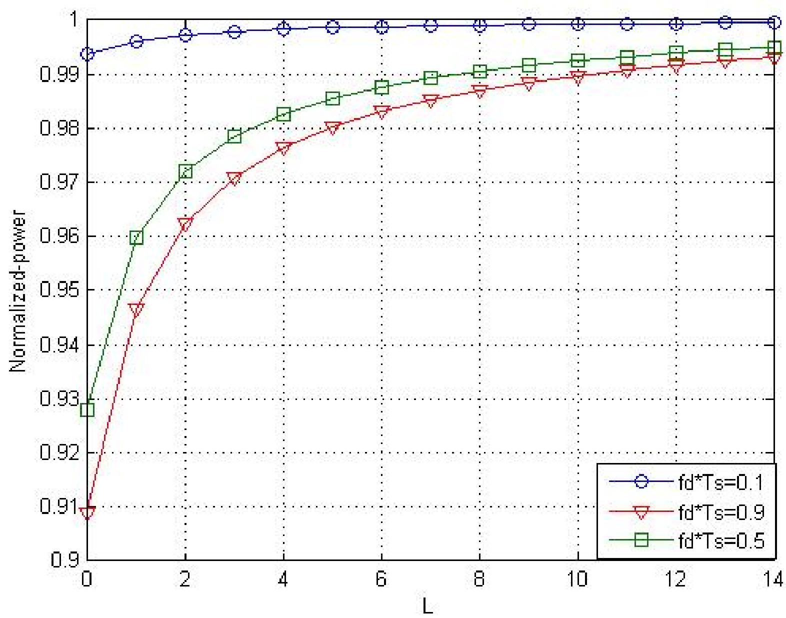

Fig. 2 Normalized symbol energy distribution

The energy of $X_t[n]$ distributed to subcarrier $n-v$ to $n+v$ can be expressed as

$$\Phi_v = \sum_{l=-v}^{v} P_l = \frac{E_s}{N^2}\sum_{k=0}^{N-1}\sum_{k'=0}^{N-1} J_0\left(2\pi F_d\left(k'-k\right)\frac{T_s}{N}\right)\sum_{l=-v}^{v}\exp\left(-j\frac{2\pi\left(k'-k\right)l}{N}\right) \tag{2.12}$$

The normalized symbol energy distribution Φ_v/E_s is depicted in Fig. 2. It indicates that most of the subcarrier's energy is spread over itself and its nearby subcarriers when $f_d T_s < 1$, and more than 99% of the subcarrier's energy is distributed on itself and its two nearby subcarriers when $f_d T_s = 0.1$. Therefore, the ICI on one subcarrier mostly comes from only few neighborhood subcarriers.

Chapter 3

ICI Mitigation in Doppler Spread Channel

3.1 Pilot-based channel estimation

Pilot-based approaches are widely used to estimate channel properties and correct the received signal [14]. Usually, a comb-type pilot subcarriers arrangement is adopted, as depicted in Fig. 3. For each transmitted symbol, pilot signals are uniformly distributed within an OFDM symbol, while null and data tones are assigned to other subcarriers.

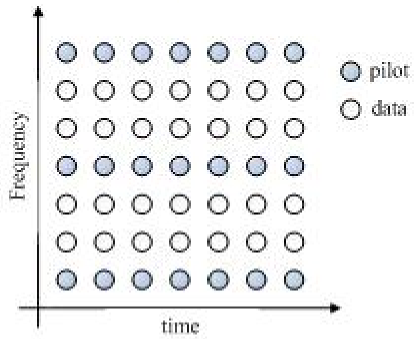

Fig. 3 Comb-type pilot arrangement

Without using any knowledge of statistics of the channels, the estimate of the channel at pilot subcarriers based on least square (LS) estimation is given by

$$\hat{H}_p[m] = \frac{Y_p[m]}{P[m]}$$
$$Y_p[m] = Y\left[\frac{N}{N_p}m\right] \qquad m = 0,1,2\cdots,N_p-1 \tag{3.1}$$

where N is the number of subcarriers, N_p is the number of pilot subcarriers and $P[m]$ is the pilot data. Thus, channel estimation with comb-type pilot-aided symbol requires an interpolation technique in order to obtain the channel information at null and data subcarriers. Before further discussion, assume that the number of pilots is large enough such that aliasing of channel impulse response will not occur. The channel estimation at the $\frac{N}{N_p}m+l$ -th subcarrier using linear interpolation is given by [17]

$$\hat{H}\left[\frac{N}{N_p}m+l\right]=\hat{H}_p\left[\frac{N}{N_p}m\right]+\left(\hat{H}_p\left[\frac{N}{N_p}(m+1)\right]-\hat{H}_p\left[\frac{N}{N_p}m\right]\right)\left(l/\left(N/N_p\right)\right) \quad 1\le l\le \frac{N}{N_p}-1 \qquad (3.2)$$

The second-order interpolation performs better than the linear interpolation method, where channel estimates at the data subcarriers are obtained by weighted linear combination of the three adjacent pilot estimates, given by [7]

$$\hat{H}\left[\frac{N}{N_p}m+l\right]=c_1\hat{H}_p\left[\frac{N}{N_p}(m-1)\right]+c_0\hat{H}_p\left[\frac{N}{N_p}m\right]+c_{-1}\hat{H}_p\left[\frac{N}{N_p}(m+1)\right]$$

$$where \begin{cases} c_1=\dfrac{\alpha(\alpha-1)}{2} \\ c_0=-(\alpha-1)(\alpha+1) \\ c_{-1}=\dfrac{\alpha(\alpha+1)}{2} \end{cases} , \alpha=\frac{l}{N} \qquad (3.3)$$

The low-pass interpolation method is performed by inserting zeros into the LS estimates at pilot subcarriers and then applying a low-pass finite-length impulse response (FIR) filter, which allows the original data to pass through unchanged and interpolates such that the mean-square error (MSE) between the interpolated points and their ideal values is minimized. The time domain interpolation [4] is a high-resolution interpolation based on zero-padding and DFT/IDFT. First, it converts $\hat{H}_p[m]$ to time domain by IDFT :

$$G_p[n]=\sum_{k=0}^{N_p-1}H_p[m]\exp(j\frac{2\pi kn}{N_p}) \qquad n=0,1,\cdots N_p-1 \qquad (3.4)$$

Then, based on the basic multi-rate signal processing properties, the N_p -sample time domain sequence $G_p[n]$ is extended to an N-sample sequence $G_N[q]$ by padding with

$N - N_p$ zeros samples at the "high frequency" region around $N_p/2$, given by

$$G_N[q] = \begin{cases} G_p[q] & 0 \le q \le N_p/2 \\ 0 & N_p/2 < q \le N - N_p/2 \\ G_p\left[q - N + N_p/2\right] & N - N_p/2 < q \le N-1 \end{cases} \tag{3.5}$$

Finally, the estimate of the channel at all frequencies is obtained by:

$$H[k] = \sum_{n=0}^{N-1} G_N[n] \exp(-j\frac{2\pi nk}{N}) \qquad 0 \le k \le N-1 \tag{3.6}$$

The performance among the comb-type estimation techniques usually ranks from the best to the worst as follows: low-pass, time-domain, second-order, and linear.

3.2 The proposed method

First we will recapitulate why the LS estimator presented in the previous section is not adequate at combating ICI. The LS estimator with 1-D interpolation compensates for the frequency-selectivity fading channel, assuming that the channel is stationary during one symbol interval. Usually, if $f_d T$ is les than 0.01, the channel can be assumed constant during one symbol interval. But the equalizer considers the ICI as an additive Gaussian random process, the performance of equalizer degrades significantly due to ICI for larger channel variation, as for $f_d T \ge 0.01$.

To fully count the ICI effect, (2.7) should be used to solve $\mathbf{X_t}$, and it is necessary to estimate the channel matrix $\mathbf{H_t}$ then calculate its matrix inverse. Nevertheless, accurate estimation of the transfer function requires complete knowledge of the time-variation of the CIR for each OFDM symbol, which is not usually available. When the OFDM symbol duration is smaller than 10% of the channel coherence time, the variation of the channel during a block period can be assumed in a linear model [10]. By utilizing the above assumption, the estimation problem of the channel matrix $\mathbf{H_t}$ can be greatly simplified, since the value of the slope of the linear model uniquely determines the ICI. The component of the channel matrix $\mathbf{H_t}$ can be expressed as

$$
\begin{aligned}
H^{t}_{p,q} &= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{l=0}^{L-1} h^{t}_{n,l}\exp\left(-j\frac{2\pi}{N}\left[(p-q)n+ql\right]\right) \\
&= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{l=0}^{L-1}(a_l + n\cdot b_l)\exp\left(-j\frac{2\pi}{N}\left[(p-q)n+ql\right]\right) \\
&= \frac{1}{N}\sum_{l=0}^{L-1}\left(a_l\cdot\sum_{k=0}^{N-1}\exp\left(-j\frac{2\pi(p-q)n}{N}\right)+b_l\cdot\sum_{k=0}^{N-1}k\exp\left(-j\frac{2\pi(p-q)n}{N}\right)\right)\exp(-j\frac{2\pi ql}{N}) \qquad (3.7)\\
&= \begin{cases} B_q\Phi_{p-q} & \text{p} \neq \text{q} \\ A_q + \dfrac{N-1}{2}B_q & \text{p=q} \end{cases}
\end{aligned}
$$

where

$$
\begin{aligned}
A_q &= \sum_{n=0}^{L-1} a_l\exp(-j\frac{2\pi ql}{N}) \\
B_q &= \sum_{n=0}^{L-1} b_l\exp(-j\frac{2\pi ql}{N}) \qquad (3.8)\\
\Phi_q &= \frac{1}{\exp(-j\frac{2\pi q}{N})-1}
\end{aligned}
$$

It is shown in (3.7) that the main diagonal of the estimated channel matrix depends on the average of the channel frequency response during one block period and the ICI term is only determined by the channel variation and OFDM parameters. Then substitute (3.7) and (3.8) into (2.7),we get

$$
\mathbf{Y_t} = \mathbf{H_t}\mathbf{X_t} + \mathbf{N}
$$

$$
= \left\{\begin{pmatrix} A_0 + \frac{N-1}{2}\cdot B_0 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & A_{N-1} + \frac{N-1}{2}\cdot B_{N-1} \end{pmatrix} + \left(\begin{pmatrix} 0 & \Phi_{-1} & \cdots & \Phi_{-N+1} \\ \Phi_1 & 0 & \cdots & \Phi_{-N+2} \\ \vdots & \vdots & \ddots & \vdots \\ \Phi_{N-1} & \Phi_{N-2} & \cdots & 0 \end{pmatrix}\right)\bullet\begin{pmatrix} B_0 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & B_{N-1} \end{pmatrix}\right\}\bullet\begin{pmatrix} X[0] \\ X[1] \\ \vdots \\ X[N-1] \end{pmatrix} + N \qquad (3.9)
$$

Since Φ_q only depends on OFDM parameters, it is possible to precalculate it once at initialization. There are several schemes proposed in the following. The procedure for the estimation of the channel matrix is shown in Fig. 4 and described as follows.

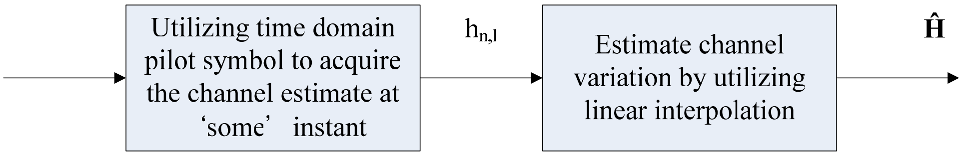

Fig. 4 Block diagram of the estimation of the channel matrix

First, a time domain pilot signal is inserted at the end of every symbol, as shown in Fig. 5. A similar method can be found in [10].

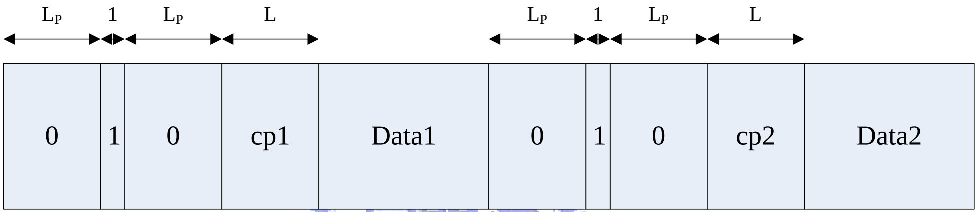

Fig. 5 Transmitted data format 1

The pilot symbol is composed of $2L_P+1$ samples. The first L_P samples are used to avoid ISI, while remaining L_P+1 samples are inserted for CIR estimation. Then, by comparing the CIR changes between the received signals corresponding to the $(i-1)$th pilot symbol and ith pilot symbol for each path, the CIR variation during the block period is estimated using linear interpolation. Although using pseudo-delta function to do the channel estimation is straightforward, it may have some drawbacks. In order to coincide with the signal power spectrum density, the pseudo-noise can be used as time domain pilot in place of the delta function, as shown in Fig.6.

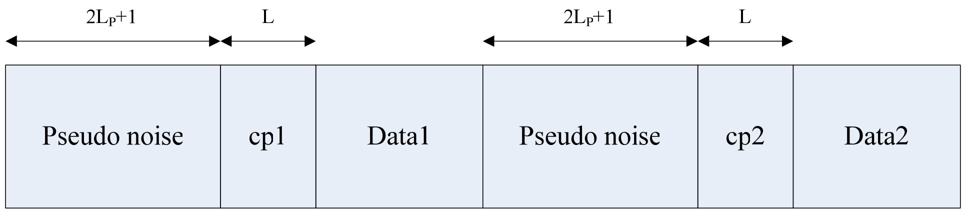

Fig. 6 Transmitted data format 2

Before further discussion, we assume that the length of time domain pilot is M- sample. Assuming the channel variation during the M samples can be negligible. The time domain convolution can be expressed as a matrix vector multiplication. The linear convolution matrix is formed from the time domain pilot.The least-square channel estimate, assuming $\mathbf{P}^{\mathbf{H}}\mathbf{P}$ has full rank, is given by

$$\hat{\mathbf{h}} = \left(\mathbf{P}^{\mathbf{H}}\mathbf{P}\right)^{-1}\mathbf{P}^{\mathbf{H}}\mathbf{r} \tag{3.10}$$

and the corresponding MSE is given by $\sigma_n^2 tr\left\{\left(\mathbf{P}^{\mathbf{H}}\mathbf{P}\right)^{-1}\right\}$. Based on minimizing the channel estimation MSE, it can be achieved if and only if $\mathbf{P}^{\mathbf{H}}\mathbf{P}$ has equal eigenvalues. This is achieved when

$$\mathbf{P}^{\mathbf{H}}\mathbf{P} = E_P\mathbf{I} \tag{3.11}$$

where $\mathbf{P}$ denotes the linear convolution matrix. It can be observed that the time domain pilot should be a shift-orthogonal sequence. The corresponding minimum MSE is $L\sigma_n^2 / E_P$. Then, the CIR variation for each path during the OFDM symbol can be estimated with linear interpolation, given the LS estimates at the inserted pilot symbols. Finally, the proposed method reconstructs the channel matrix and calculates its matrix inverse. Since the channel matrix can have a large size, it is difficult to process in real time. Based on the structure of the channel matrix $\mathbf{H}_t$, whose energy is concentrated on the neighborhood of the main diagonal, the computation complexity can be reduced by considering only the elements nearby the main diagonal and ignoring remaining elements, as shown in Fig. 7.

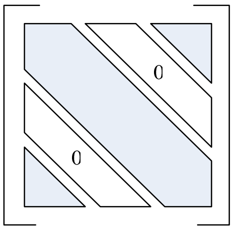

Fig. 7 Banded channel matrix with color region corresponding to ICI concentration

Another way to avoid straight matrix inversion is to divide the task into ICI reduction and a simple one-tap equalizer. In the following, the ICI reduction procedure will be described in detail. According to equation (2.7), it can be found that ICI component is determined by not only the variation of the channel but also the transmitted data. The decision-feedback techniques can be utilized to acquire the estimates of the transmitted data, then the estimated channel matrix in (3.9) is used to subtract the ICI components from the received signal. The resulting ICI reduction method is as follows

$$\tilde{Y}_t[m] = Y_t[m] - \sum_{n=0, n\neq m}^{N-1} \hat{H}_{m,n}^t \hat{X}_t[m] \qquad 0 \le m \le N-1 \tag{3.12}$$

$$\hat{X}_t[m] = \mathrm{Dec}\left(\frac{1}{\hat{\mathrm{H}}_{\mathrm{m,m}}} Y_t[m]\right) \tag{3.13}$$

where Dec denotes a slicer in the demapper. Assuming that the number of data carriers is large, the effect of incorrect hard decisions will not have a significant impact. Finally, the ICI-cancelled $\tilde{Y}_t[m]$ can be used in a conventional one-tap equalizer. For severe Doppler effects, it also can be combined with an iterative method to enhance the ICI estimation accuracy, as shown in Fig. 8. In addition, the hard decision in (4.1) can be replaced by a MMSE equalization, given by

$$\hat{X}_t[m] = \frac{\hat{\mathrm{H}}_{\mathrm{m,m}}^* Y_t[m]}{\left|\hat{\mathrm{H}}_{\mathrm{m,m}}\right|^2 + 1/SNR} \tag{3.14}$$

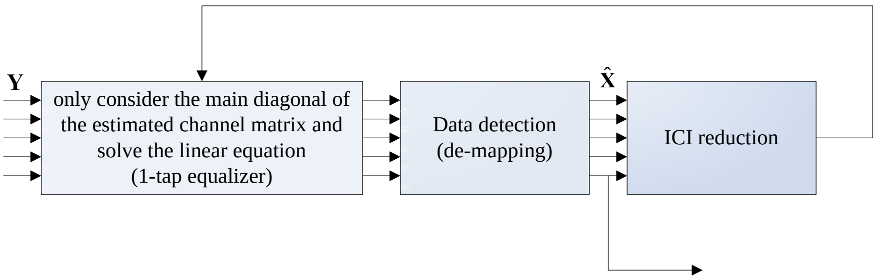

Fig. 8 Block diagram of the proposed ICI-reduction method

In the same manner, the computation complexity of the ICI reduction can be reduced by considering only the ICI component due to the nearby q subcarriers without degrading much system performance, given by

$$\tilde{Y}_t[m] = Y_t[m] - \sum_{n=0,|n-m|\le q/2}^{N-1} \hat{H}_{m,n}^t \hat{X}_t[m] \qquad 0 \le m \le N-1 \qquad (3.15)$$

Even, it can be ignored the off-diagonal elements of the estimated channel matrix and use one-tap equalizer without ICI reduction. It can be observed in (3.7) that the main diagonal elements are the average of the channel frequency response during one block period. Based the linear property, it can be obtained by

$$diag(\mathbf{H_t}) = DFT_N\left\{\frac{\vec{h}_t + \vec{h}_{t+1}}{2}\right\} \qquad (3.16)$$

where the subscript t denotes the t^{th} OFDM frame and $diag()$ denotes the main diagonal of the matrix. Then, utilize one-tap equalizer without ICI reduction.

3.3 Computation complexity analysis

Utilizing the time domain pilot inserted at the end of every OFDM symbol to acquire the time domain channel estimates in (3.10) requires L^2 multiplications. The matrix $\mathbf{\Phi}$ can be pre-calculated at the initialization. Based on the linear property of the channel variation, the computation of the channel matrix in (3.8) and (3.9) approximately requires

$2N\log_2(N)$ multiplications. The method that directly uses the inversion of the estimated channel matrix requires $O(N^3)$ computational complexity. The method using one-tap equalizer with ICI reduction approximately requires $2N\log_2(N)+2N+N^2$ multiplications. The simplified method in (3.14) and (3.15) using one-tap equalizer with ICI reduction which only consider the elements of the estimated channel matrix in the adjacent of the main diagonal requires $2N\log_2(N)+2N+qN$ multiplication. As the size of the channel matrix increases, the simplified method conserves more computation complexity than that of which utilizes the whole estimated channel matrix to do ICI-reduction. The method using one-tap equalizer without ICI reduction approximately requires only $N\log_2(N)+N$ multiplication. The complexity of the proposed methods are summarized in Table 1

Table 1 Computational complexity analysis

Estimation Scheme	Computational complexity	Comments
One-tap equalizer without ICI reduction	$L^2+N\log_2(N)+N$	Low complexity
One-tap equalizer with ICI reduction	$2L^2+2N\log_2(N)+2N+N^2$	High complexity
One-tap equalizer with partial ICI reduction	$2L^2+2N\log_2(N)+2N+qN$	Moderate complexity
One-tap equalizer with DFT-based channel estimation	$L\log_2(L)+N\log_2(N)+N$	Low complexity

Chapter4

Simulations and Discussions

4.1 Simulation result

1) System parameters:

OFDM system parameters used in the simulations are illustrated in Table 2. Since the aim is to observe channel estimation performance, it assumed to be perfect synchronization in the simulations. Moreover, the guard interval is assumed to be longer than the maximum delay spread of the channel. Simulations are carried out for various signal-to-noise ratio (SNR) and Doppler spreads.

Table 2 Simulation parameters

Parameters	**Specifications**
FFT Size	128
Pilot Ratio	1/16
Guard Interval	16
Carrier frequency	5M
Bandwidth	500K
Signal Constellation	QPSK,16-QAM
Channel Model	Jakes

2) Channel model

Assume a symbol spaced, tap-delay-line channel model of 4 paths, where each channel tap is generated with the Doppler spectrum based on the Jakes' model. The generation of the tap gains is illustrated in Fig. 9.

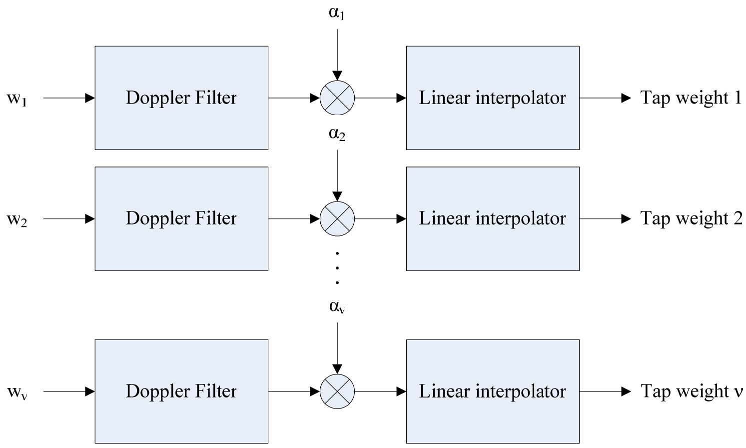

Fig. 9 Generation of the tap weight processes

It start with a set of independent, zero-mean complex Gaussian white noise processes, which are filtered to produce the appropriate Doppler spectrum, as depicted by the Jakes model. These are then scaled to produce the desired power profile. In the simulations, the path gains follow the exponentially-decayed power profile as depicted by

$$E\left[\left|\alpha_l(t)\right|^2\right] = e^{-\frac{\tau_l}{\sigma}} \tag{4.1}$$

where $\alpha_l(t)$ is the path gain, τ_l is the path time delay and σ is the power delay constant. We choose the power delay time constant such that the last path power is 20dB below the first path. Before further discussion, we assume the overhead of the proposed method and DFT-based method are the same. Moreover, the suboptimal time domain pilot we used is the product of the shift-orthogonal sequence and a finite-duration window function, such as Hanning window.

Fig. 10 shows the BER performance for the case $f_d T_s = 0.04$. The modulation scheme used for this simulation is 16-QAM. Assume the pilot subcarriers are equispaced along the frequency domain such that DFT-based channel estimator can be used. The DFT-based channel estimation is also shown as reference, which has comparable performance to that of MMSE channel estimation method. In addition, the overhead of the proposed method are the same as that of the DFT-based method in the simulation. From Fig. 13, it can be observed that the ICI-reduction algorithm effectively reduces the error floor, as is more evident especially in high SNR than low SNR. The method using the one-tap equalizer with ICI reduction has comparable performance to that of directly using the inversion of the estimated channel matrix, while costs much less computation.

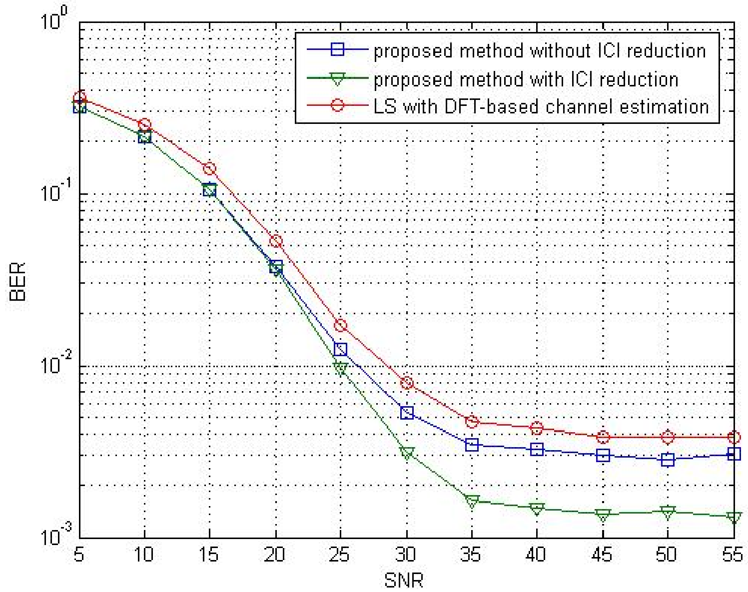

Fig. 10 BER performance of ICI reduction algorithm with $f_d T_s = 0.04$

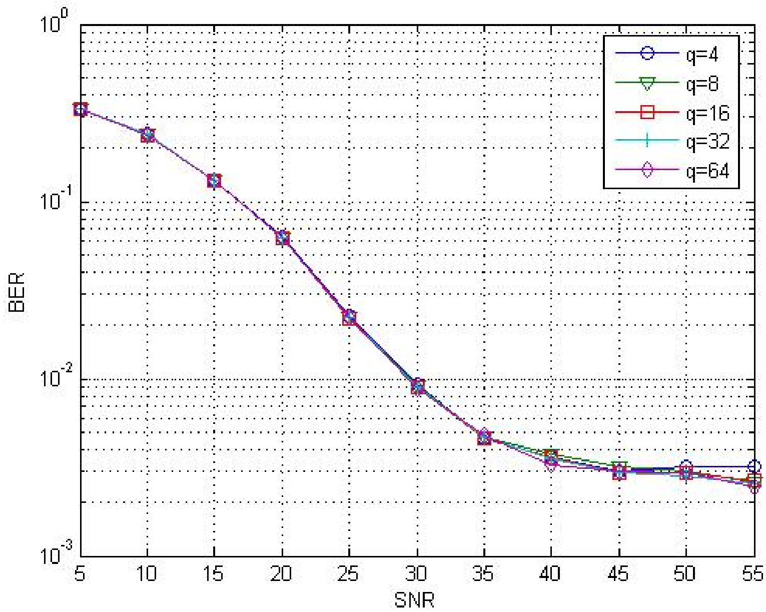

Fig. 11 The comparison of partial ICI reduction method

Fig. 11 also shows the BER performance for the case $f_d T_s = 0.04$ with choosing different number q of the nearby subcarriers for ICI reductions. From Fig. 11, it can be observed that the simplified scheme gives comparable results to the method using whole estimated channel matrix but has low computation complexity.

Fig. 12 and Fig.13 shows the BER performance respectively for the case $f_d T_s = 0.1$. and $f_d T_s = 0.2$. From Fig.12, it can be observed that the proposed method From Fig.13, it can be observed that the proposed method gives rise to an error floor as the assumption of the linear property of the channel variation no longer holds.

Fig. 14 and Fig. 15 shows the BER performance for the case $f_d T_s = 0.04$ respectively under various conditions of carrier frequency offset and timing offset . From Fig. 14, it can be observed that the proposed method is robust to the effect of carrier frequency offset. From Fig. 15, it can be shown that the proposed method is sensitive to the timing offset.

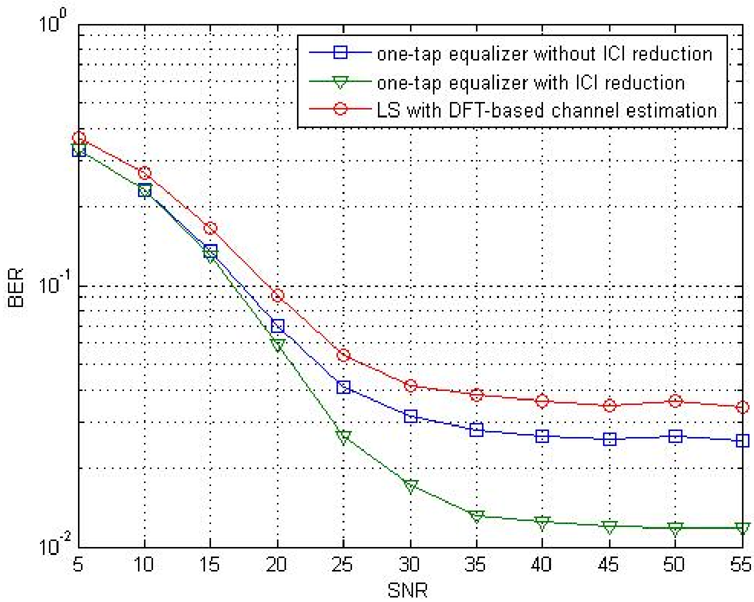

Fig. 12 BER performance of ICI reduction algorithm with $f_d T_s = 0.1$

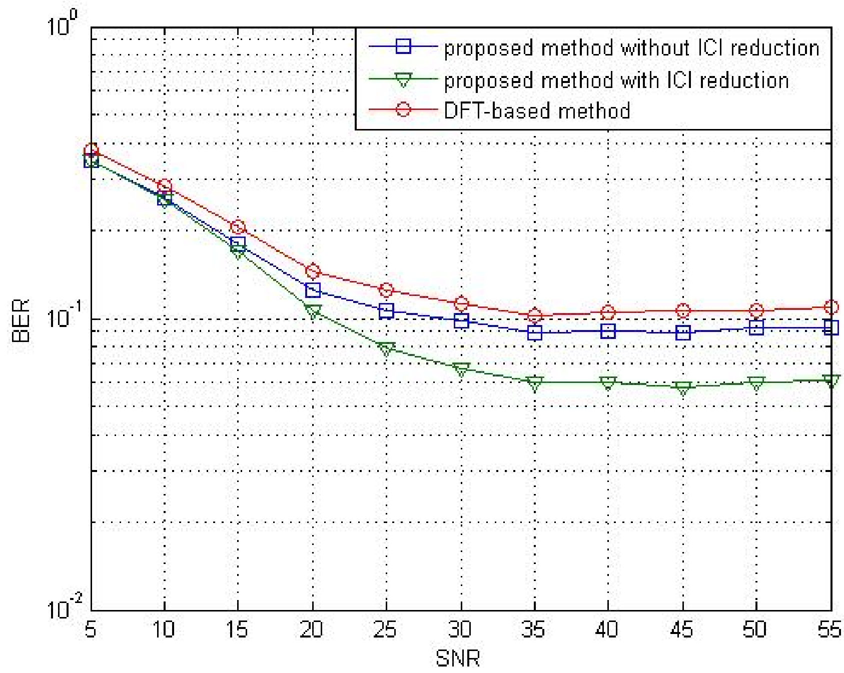

Fig. 13 BER performance of ICI reduction algorithm with $f_d T_s = 0.2$

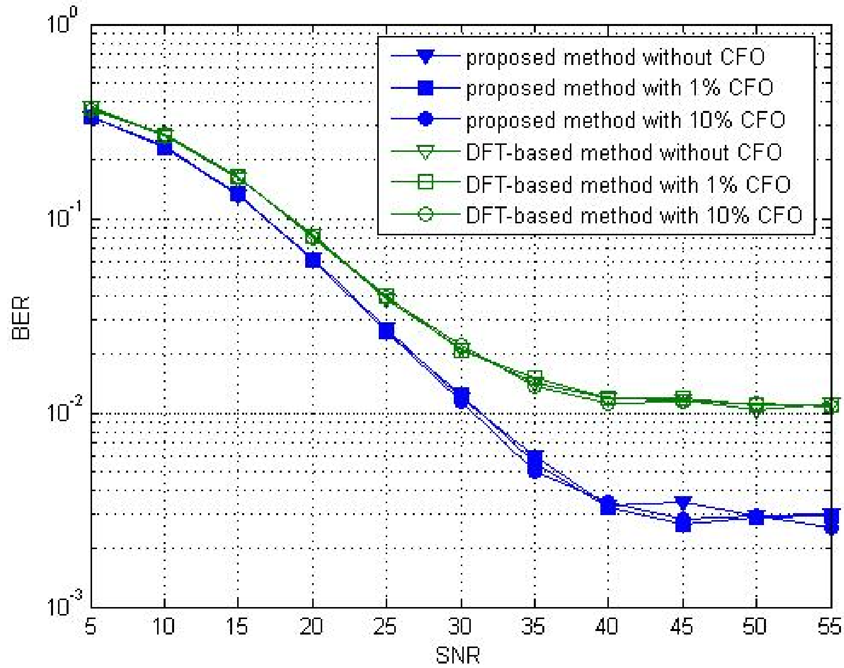

Fig. 14 BER performance of proposed method with $f_d T_s = 0.04$ **and various CFO**

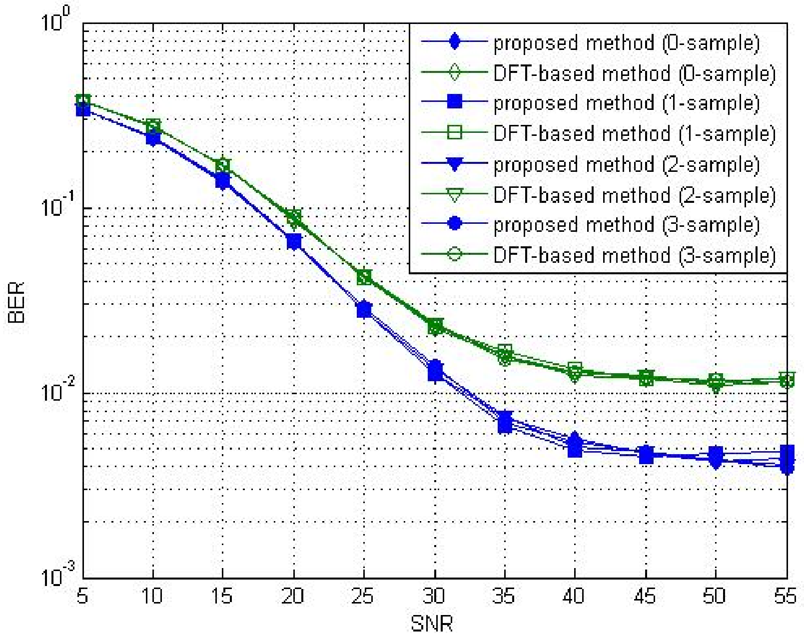

Fig. 15 BER performance of proposed method with $f_d T_s = 0.04$ and various timing offset

Consider that the tolerance interval to symbol timing error is k samples and the maximum channel length is L. The comparison among the proposed methods and DFT-based method are summarized as in Table 3, including overhead, computational complexity and BER performance.

From Table 3, it can be concluded that the overhead of the proposed method is slightly more than that of the DFT-based method but the performance of the proposed method is much better than that of the DFT-base method. Moreover, it is worthwhile to increase the computational complexity for the sake of performance, as shown in Table 3.

Table 3 Overhead, complexity and performance of the various schemes

Estimation Scheme	Overhead	Complexity	Performance
One-tap equalizer without ICI reduction	Moderate $(2L+3K-1)$	Low	Moderate
One-tap equalizer with ICI reduction	Moderate $(2L+3K-1)$	High	Very good
One-tap equalizer with partial ICI reduction	Moderate $(2L+3K-1)$	Moderate	Very good
One-tap equalizer with DFT-based method	Low $(2L+2K-1)$	Low	Moderate

4.2 Conclusion

Usually, DFT-based receivers considers ICI as an additive noise and it is not adequately compensated. In this thesis, we propose the time domain pilot-based estimation scheme combined an iterative method to suppress the ICI induced by time varying channels. It is shown through simulation that ICI can be compensated by the proposed method if the normalized Doppler frequency change is in the range of ($0.01 \le f_d T_s \le 0.1$). The reason is that for the channel with a very high Doppler frequency, the assumption that the channel parameters vary in a linear fashion within a block period is no longer a good approximation and gives rise to an error floor. On the other hand, the channel can be assumed invariant during a block period if the normalized Doppler frequency less than 0.01.Under these circumstances, the proposed method only slightly outperforms the DFT-based method. Furthermore, we also proposed a simplified method using a one-tap equalizer with partial ICI reduction; it has comparable performance to that using whole estimated channel matrix for ICI reduction.

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