Channel Estimation and ICI Cancellation for OFDM Systems in Doubly-selective Channels

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***Abstract*—In orthogonal frequency division multiplexing (OFDM) systems, time- and frequency-selective (or doubly-selective) fading leads to the loss of subcarrier orthogonality and the occurrence of inter-carrier interference (ICI), which increases an irreducible error floor in proportional to the normalized Doppler frequency offset. Channel estimation (CE) in rapidly time-varying multi-path scenarios is critical for ICI cancellation and coherent demodulation. In this paper, we introduce an iterative CE scheme to estimate time-varying channel parameters and a low-complexity equalization method to cancel ICI and detect data. The proposed CE technique performs an initial CE based on a piece-wise linear model. Then ICI is reconstructed and cancelled from the received signals. Estimating channel again by using signals of less interference, refined CE can be obtained. Meanwhile, a low-complexity equalizer is proposed to further improve BER performance. Finally, simulation results show good performance of the proposed CE method in doubly-selective fading channels.**

***Index Terms*—Channel estimation, doubly-selective channel, equalization, inter-carrier interference (ICI) cancellation, OFDM, time-varying frequency-selective channel.**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) is widely recognized as an efficient modulation method for wireless communication [1]. Thanks to IFFT and FFT operations, demodulation and modulation are efficiently implemented. Equalization can be easily achieved by a single tap filter on each subcarrier in OFDM systems. This has justified the candidates of OFDM for many recent and emerging broadband wireless standards such as IEEE 802.16e, IEEE 802.20 and 3GPP-LTE.

However, future wireless application is expected to operate at high transmit frequencies and mobility, resulting in the channel fading to both be time- and frequency-selective. In such cases, channel variations in an OFDM symbol destroy the orthogonality between subcarries, resulting in inter-carrier interference (ICI) and performance degradation. Therefore, in order to compensate ICI, a high quality of channel impulse information is required at the receiver.

Several approaches have been proposed for channel estimation (CE) of OFDM systems. A minimum mean-square error (MMSE) channel estimator is proposed in [2], which utilizes the time- and frequency-domain correlation functions of the fading channels. In [3], a low-rank approximation to the frequency-domain linear MMSE estimator is proposed. But OFDM systems that make use of both techniques still suffer from ICI, since no attempt is made to cancel it. To reduce the impact of ICI, time-domain channel estimators are proposed in [4] [5], which assume that the channel impulse response (CIR) varies in a linear fashion within the symbol duration. Based on a linear channel model, a two-stage CE method is proposed [6]. First, CE is performed at the pilot positions using conventional methods [7]. Then both adjacent symbols are utilized to acquire the slopes of channel variations. In [8], a potential candidate for the estimation is proposed by exploiting the time-variant nature of the channel as a provider of time diversity and reduces the complexity using the singular value decomposition method. As in the case of the schemes proposed in [2] [3], this technique requires knowledge of the channel statistic.

The techniques above basically perform CE based on the ICI-corrupted pilot symbols, upon which the channel matrix is reconstructed assuming a linear model for the channel variations. As a result, channel parameters are achieved from channel measurements affected by data-independent ICI. So the weak spot of the solutions is that it admits ICI in the first place, but allows for ICI to corrupt the pilots leading to degraded CE. In this paper, we present an iterative CE method to eliminate the impact of ICI on CE based on piece-wise linear model. Meanwhile, a low-complexity equalizer is proposed to further improve BER performance. The rest of the paper is organized as follow: Section II provides the OFDM system model in a doubly-selective environment. Section III describes the CE method in detail. In the section IV, several equalizers for combating ICI are described. Performance results are given in Section V and the conclusions are drawn in section VI.

Notation: Matrices and vectors are denoted by symbols in boldface and the superscript T and H represent transpose and Hermitian respectively. $\mathbf{I}_m$ denotes the $m \times m$ identity matrix. $E(\cdot)$, $\mathrm{diag}\{\cdot\}$ and $\frac{\partial f}{\partial x}$ mean the statistic expectation, diagonal matrix and partial derivative respectively. $[a:b](a<b)$ represents the vector $[a, a+1, \ldots, b-1, b]$.

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II. System Model

In this section, we introduce a discrete-time baseband equivalent OFDM system under consideration. N input symbols are transformed into an N-point symbol block $\mathbf{X}=\left[X_0,X_1,\ldots,X_{N-1}\right]^T$ by serial-to-parallel converter, assuming $E\left(\mathbf{XX}^H\right)=E_s\mathbf{I}_N$, where E_s is the symbol energy. Then it is transformed into an N-point time-domain sample block $\mathbf{x}=\left[x_0,x_1,\ldots,x_{N-1}\right]^T$ by IFFT and the nth sample can be expressed as

$$\begin{aligned} x_n &= \mathrm{IDFT}\{\mathbf{X}\} \\ &= \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}X_k e^{j2\pi nk/N},\ n=0,1,\ldots,N-1 \end{aligned} \tag{1}$$

In order to eliminate intersymbol interference (ISI), a cyclic prefix (CP) is inserted. The length G of CP is chosen to be longer than the multi-path spread of the channel.

The doubly-selective fading channel is given by $h(n,l)$, where $h(n,l)$ represents the channel impulse response (CIR) at lag l and instant n. The nth received sample can be written as

$$y_n=\sum_{l=0}^{L-1}h(n,l)x_{n-l}+w_n,\ 0\le n\le N-1 \tag{2}$$

where L denotes the maximum delay spread and w_n is the additive white Gaussian noise with variance σ^2. After removal of the CP, the received samples are demodulated by taking the DFT, i.e.,

$$\begin{aligned} Y_m &= \mathrm{DFT}\{y_k\}=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}y_k e^{-j2\pi mk/N} \\ &= \frac{1}{N}\sum_{n=0}^{N-1}\left(\sum_{k=0}^{N-1}\sum_{l=0}^{L-1}h(n,l)e^{-j2\pi(l-k)n/N}e^{-j2\pi mk/N}\right)X_n+W_m \\ &= H_{m,m}X_m+\sum_{n=0,n\ne m}^{N-1}H_{m,n}X_n+W_m,\ 0\le m\le N-1 \end{aligned} \tag{3}$$

where

$$H_{m,n}=\frac{1}{N}\sum_{k=0}^{N-1}\sum_{l=0}^{L-1}h(n,l)e^{-j2\pi(l-k)n/N}e^{-j2\pi mk/N} \tag{4}$$

$$W_m=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}w_k e^{-j2\pi mk/N} \tag{5}$$

Note that $H_{m,m}X_m$ is the expected ICI-free term and $\sum_{n=0,n\ne m}^{N-1}H_{m,n}X_n$ represents the ICI term. Furthermore

$$H_{m,m}=\sum_{l=0}^{L-1}h_{ave}(l)e^{-j2\pi ml/N} \tag{6}$$

where $h_{ave}(l)=\frac{1}{N}\sum_{n=0}^{N-1}h(n,l)$ is the average of the lth channel tap over the time duration of $0\le n<N\times T_s$ and T_s is the sampling period. Hence the OFDM symbol duration is $T=(N+G)T_s$. We define $h(n,l)$ as

$$h(n,l)=h_{ave}(l)+\Delta h(n,l),\ 0\le n\le N-1, 0\le l\le L-1 \tag{7}$$

As was demonstrated by previous work [5], the expected output signals are decided by the average of the channel impulse response over an OFDM symbol $h_{ave}(l)$, the ICI are decided by the offset of the channel impulse response with respect to the average $\Delta h(n,l)$.

III. Channel Estimation

In this section, we address the challenging problem of doubly-selective fading channel estimation for OFDM systems and we propose a promising technique based on the assumption that channel varies linearly for most practical Doppler spread.

Regarding the ICI as additional Gauss noise and substituting (6) into (3), (3) could be rewritten as

$$\begin{aligned} Y_m &= \left(\sum_{l=0}^{L-1}h_{ave}(l)e^{-j2\pi ml/N}\right)X_m+e_m \\ &= \sqrt{N}\mathbf{F}(m,1:L)\mathbf{h}_{ave}X_m+e_m \end{aligned} \tag{8}$$

where

$$e_m=\sum_{n=0,n\ne m}^{N-1}H_{m,n}X_n+W_m \tag{9}$$

and $\mathbf{F}$ is the DFT matrix, $\mathbf{F}_L$ is the first L columns of $\mathbf{F}$, $\mathbf{F}_L(m,:)$ represents the mth row of $\mathbf{F}_L$, $\mathbf{h}_{\mathbf{ave}}=\left[h_{ave}(0),h_{ave}(1),\ldots,h_{ave}(L-1)\right]^T$.

Assuming that N_P pilot tones are uniformly inserted into X_k, which are placed at subcarriers $\mathbf{\Theta}=\{m(1),\ldots,m(N_P)\}$, and we obtain from (8)

$$\begin{aligned} \mathbf{Y}_P &= \begin{bmatrix} Y_{m(1)} \\ \vdots \\ Y_{m(N_P)} \end{bmatrix} = \mathrm{diag}\left\{\sqrt{N}\mathbf{F}_L(\mathbf{\Theta},:)\mathbf{h}_{ave}\right\}\begin{bmatrix} X_{m(1)} \\ \vdots \\ X_{m(N_P)} \end{bmatrix}+\mathbf{e} \\ &= \sqrt{N}\mathrm{diag}\left\{X_{m(1)},\ldots,X_{m(N_P)}\right\}\mathbf{F}_L(\mathbf{\Theta},:)\mathbf{h}_{ave}+\mathbf{e} \\ &= \mathbf{\Xi}\mathbf{h}_{ave}+\mathbf{e} \end{aligned} \tag{10}$$

where $\mathbf{\Xi}=\sqrt{N}\mathrm{diag}\left\{X_{m(1)},\ldots,X_{m(N_P)}\right\}\mathbf{F}_L(\mathbf{\Theta},:)$. We can obtain the LS estimate without the inversion [9] if $\left|X_{m(k)}\right|=1$

$$\mathbf{h}_{ave}=\mathbf{\Xi}^H\mathbf{Y}_P \tag{11}$$

To perform the linearization, knowledge of the channel at one time instant is necessary. Approximation error would be minimized if we approximate $h(N/2-1,l)$ with the estimation of $h_{ave}(l)$. $\hat{h}(N/2-1,l)\approx\hat{h}_{ave}(l)$, so we get an estimate of the channel at the midpoint. As is shown in the Fig. 1, the slopes can be acquired by utilizing adjacent symbols.

$$\hat{\alpha}_1(l)=\frac{\hat{h}(N/2-1,l)-\hat{h}^{prev.}(N/2-1,l)}{T} \tag{12}$$

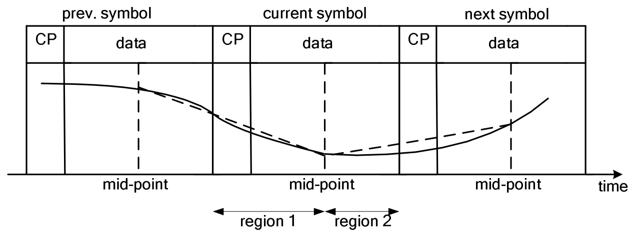

Fig. 1. Piece-wise linear model.

$$\hat{\alpha}_2(l) = \frac{\hat{h}^{next}(N/2-1,l) - \hat{h}(N/2-1,l)}{T} \quad (13)$$

Then $h(n,l)$ can be approximated as follows:

$$\hat{h}(n,l) = \begin{cases} \hat{h}(N/2-1,l) + (n+1-N/2)\times\hat{\alpha}_1(l)\times T_s, 0 \le n \le N/2-1 \\ \hat{h}(N/2-1,l) + (n+1-N/2)\times\hat{\alpha}_2(l)\times T_s, N/2 \le n \le N \end{cases} \quad (14)$$

But CE is performed based on the ICI-corrupt pilot symbols. Channel parameters are still achieved from channel measurements affected by data-independent ICI. So we propose an iterative way to cancel ICI from the received signals and estimate CIR again using signals of less interference to make the CE more precise.

After getting an estimate of the CIR according to (11), (12), (13) and (14), substitute (14) into (3) and we can obtain the frequency domain received signals as follow

$$\mathbf{Y} = \mathbf{H}_{ave}\mathbf{X} + (\mathbf{C}_1 \times \mathbf{H}_{r1} + \mathbf{C}_2 \times \mathbf{H}_{r2}) \times \mathbf{X} + \mathbf{W} \quad (15)$$

where $\mathbf{H}_{ave}$, $\mathbf{H}_{r1}$ and $\mathbf{H}_{r2}$ are all diagonal matrixes. They can be easily show as

$$\mathbf{H}_{ave} = \text{diag}\Big\{\text{FFT}\Big[\hat{h}(N/2-1,0), \hat{h}(N/2-1,1), \ldots, \hat{h}(N/2-1,L-1), 0, \ldots, 0\Big]\Big\} \quad (16)$$

$$\mathbf{H}_{r1} = \text{diag}\Big\{\text{FFT}\Big[\hat{\alpha}_1(0), \ldots, \hat{\alpha}_1(L-1), 0, \ldots, 0\Big]\Big\} \quad (17)$$

$$\mathbf{H}_{r2} = \text{diag}\Big\{\text{FFT}\Big[\hat{\alpha}_2(0), \ldots, \hat{\alpha}_2(L-1), 0, \ldots, 0\Big]\Big\} \quad (18)$$

$\mathbf{C}_1$ and $\mathbf{C}_2$ are fixed matrixes which do not depend on the channel or data. They can be easily shown as follow

$$\mathbf{C}_1 = \mathbf{F} * \text{diag}\{[-N/2+1:0], 0, \ldots, 0\} * \mathbf{F}^H \quad (19)$$

$$\mathbf{C}_2 = \mathbf{F} * \text{diag}\{0, \ldots, 0, [1:N/2]\} * \mathbf{F}^H \quad (20)$$

Then an iterative CE scheme is presented for reducing the effect of ICI based on (15). The method proceeds as follows

1) An estimate of the CIR , $\hat{h}(N/2-1,l)$ and $\hat{\alpha}(l)$, is obtained from pilots using the method discussed above;
2) Neglecting the ICI contribution, we obtain a one-tap equalization matrix $\mathbf{H}_{ave}^{-1}$, where $\mathbf{H}_{ave}$ is diagonal matrix given by (16). An estimate of the transmitted block $\hat{\mathbf{X}}$ can be obtained from $\mathbf{Y}$ through one-tap equalization and hard decision $\hat{\mathbf{X}} = \text{Dec}\{\mathbf{H}_{ave}^{-1}\mathbf{Y}\}$.
3) Reconstruct the ICI by $\hat{\alpha}(l)$ and detected data $\hat{\mathbf{X}}$, then subtract it from the received signal $\mathbf{Y}$ as
$$\mathbf{Y}' = \mathbf{Y} - (\mathbf{C}_1 \times \mathbf{H}_{r1} + \mathbf{C}_2 \times \mathbf{H}_{r2}) \times \hat{\mathbf{X}} \quad (21)$$
4) Use $\mathbf{Y}'$ to perform an ICI free estimation of $\hat{h}(N/2-1,l)$ and $\hat{\alpha}(l)$ as outlined above.
5) Repeat the steps above for two or three times.
6) Output $\hat{h}(N/2-1,l)$ and $\hat{\alpha}(l)$, end iteration process.

At last the computational complexity of the proposed CE method is evaluated. There is no inversion in the whole CE process. $\mathbf{C}_1$ and $\mathbf{C}_2$ are fixed matrixes that can be precalculated and stored in the receiver. The computation complexity cost of FFT or IFFT is $o(N)$. $\mathbf{H}_{ave}$, $\mathbf{H}_{r1}$ and $\mathbf{H}_{r2}$ are all diagonal matrixes and the multiplication of any of them by $\hat{\mathbf{X}}$ is still a diagonal matrix, so the complexity of $\mathbf{C}_1 \times \mathbf{H}_{r1} \times \hat{\mathbf{X}}$ or $\mathbf{C}_2 \times \mathbf{H}_{r2} \times \hat{\mathbf{X}}$ is $o(N)$. Hence the total complexity of the proposed CE method is $o(N)$.

IV. EQUALIZATION

A. MMSE Equalizer

We define $\mathbf{H} = \mathbf{H}_{ave} + \mathbf{C}_1 \times \mathbf{H}_{r1} + \mathbf{C}_2 \times \mathbf{H}_{r2}$ and obtain from (15)

$$\mathbf{Y} = \mathbf{H}\mathbf{X} + \mathbf{W} \quad (22)$$

In the case of a time-invariant channel, $\mathbf{H}$ is diagonal and there is no ICI, so single-tap equalization is sufficient. However, in the time-varying case, $\mathbf{H}$ is no longer diagonal and significant ICI occurs. Motivated by this fact, several approaches have been proposed to combat ICI over rapidly time-varying channels.

MMSE ICI equalization amounts for finding the linear matrix $\mathbf{G}$ which minimizes $E\{\|\mathbf{G}\mathbf{Y} - \mathbf{X}\|^2\}$. The resulting MMSE equalizer equals

$$\hat{\mathbf{X}} = \mathbf{G}\mathbf{Y} = \left(\mathbf{H}^H\mathbf{H} + \sigma^2\mathbf{I}_N\right)^{-1}\mathbf{H}^H\mathbf{Y} \quad (23)$$

B. PMMSEOSC Equalizer

In [8], an MMSE with successive cancellation (MMSESC) method to mitigate ICI was proposed. The MMSE and MMSESC have good performance but require $\ge o(N^3)$ computational complexity. In this paper, exploiting the facts the ICI on one subcarrier mainly from only several neighborhood subcarriers [10], a low complexity partial minimum mean square with ordered successive cancellation (PMMSEOSC) method in frequency domain is proposed. Every time it detects the symbol which has the highest signal to interference and noise ratio (SINR)

$$\arg\max_k \text{SINR}_k = \arg\max_k \frac{|\mathbf{H}_{k,k}|^2 E_s}{E_s \sum_{n,n\ne k}^{N-1} |\mathbf{H}_{k,n}|^2 + \sigma^2} \quad (24)$$

Define SINR_k as the function $f(|\mathbf{H}_{k,k}|)$ and find the partial derivative of $f(|\mathbf{H}_{k,k}|)$ with respect to $|\mathbf{H}_{k,k}|$,

$$\frac{\partial f}{\partial\left|\mathbf{H}_{k,k}\right|}=\frac{2\left|\mathbf{H}_{k,k}\right|E_s}{E_s\sum_{n,n\neq k}^{N-1}\left|\mathbf{H}_{k,n}\right|^2+\sigma^2}>0 \tag{25}$$

Obviously $f\left(\left|\mathbf{H}_{k,k}\right|\right)$ is a monotonic increasing function with respect to $\left|\mathbf{H}_{k,k}\right|$. Hence, we only need one permuting operation to detect data in succession. The procedure proceeds as follows.

First, calculate SINRs of all symbols in an OFDM block. Sort the vector $\mathbf{SINR}=\left[\mathrm{SINR}_0,\ldots,\mathrm{SINR}_{N-1}\right]$ and obtain the indeces by sort operation $\mathbf{index}=\mathrm{sort}\left(\mathbf{SINR}\right)$.

Then, for $i=0:N-1$, $k=\mathbf{index}_i$

1) Detecting the symbol X_k based on partial MMSE based on D neighborhood subcarriers. Define $\mathbf{X}_k=\mathbf{X}(\boldsymbol{\rho})$, $\mathbf{Y}_k=\mathbf{Y}(\boldsymbol{\rho})$, $\mathbf{W}_k=\mathbf{W}(\boldsymbol{\rho})$ and $\mathbf{H}_k=\mathbf{H}(\boldsymbol{\rho},\boldsymbol{\rho})$, where $\boldsymbol{\rho}$ is a vector with mth element $\boldsymbol{\rho}_m=(k-D-1+m)\bmod N, 0\le k\le N-1, 0\le m\le 2D+1$, $\mathbf{X}_k$, $\mathbf{Y}_k$ and $\mathbf{W}_k$ denote the subvetors of $\mathbf{X}$, $\mathbf{Y}$ and $\mathbf{W}$ respectively, $\mathbf{H}_k$ denotes the submatrix of $\mathbf{H}$. Then we have
$$\mathbf{Y}_k=\mathbf{H}_k\mathbf{X}_k+\mathbf{W}_k \tag{26}$$
$\mathbf{X}_k$ is detected by partial MMSE detector
$$\hat{\mathbf{X}}_k=\left(\mathbf{H}_k^H\mathbf{H}_k+\sigma^2\mathbf{I}_K\right)^{-1}\mathbf{H}_k^H\mathbf{Y}_k \tag{27}$$
where $\hat{\mathbf{X}}_k=\left[\hat{X}_{k_0},\ldots,\hat{X}_{k_{2D}}\right]^T$. Then we get the estimation of X_k through $\hat{X}_k=\hat{X}_{k_D}$.
2) Hard decision $\hat{X}_k=\mathrm{Dec}\left\{\hat{X}_k\right\}$.
3) Updating the received signal $\mathbf{Y}$ and the channel matrix $\mathbf{H}$ by $\mathbf{Y}=\mathbf{Y}-\mathbf{H}(:,k)\hat{X}_k$ and $\mathbf{H}(:,k)=0$, where $\mathbf{H}(:,k)$ represents the kth column vector of $\mathbf{H}$.

Here we evaluate the computational complexity of PMMSEOSC equalizer. Calculating SINRs of all symbols needs N multiplicative operations, and sort operation needs $N\log_2 N$ compare operation (adopting "quick sort"). Detecting all symbols by partial MMSE needs $o\left(K^3N\right)$ operations ($K\ll N$). So the total complexity is $o\left(N\right)$, and it is much lower than MMSE or MMSESC, which are $o\left(N^3\right)$.

V. SIMULATION RESULTS

In this section, we demonstrate the relative performance of the above proposed CE technique for OFDM systems through extensive computer simulations. A symbol spaced, tap-delay channel model of 6 paths is considered for time-varying multi-path channel. A bandwidth of 5 MHz with a carrier frequency at 2 GHz is assumed. The total bandwidth was divided into 128 sub-bands, hence the size of FFT becomes 128 ($N=128$) and the length of cyclic prefix (CP) is 16. 16 pilot tones are inserted into OFDM symbol equidistantly for CE. The modulation scheme used here is QPSK. The Doppler frequency is taken into account up to $f_d=648$ Hz, resulting in the normalized frequency shift $f_D=f_dNT_s\approx 0.1$, it corresponds to a mobile speed $V=350$ km/h. Perfect carrier and symbol synchronization are assumed.

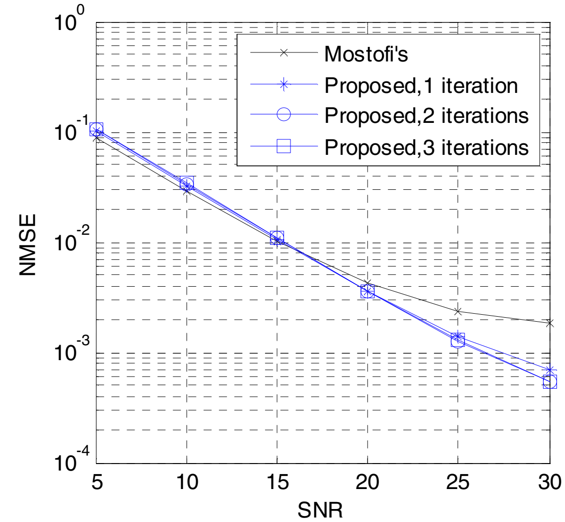

Fig. 2. Channel estimation performance, $f_D=0.05$.

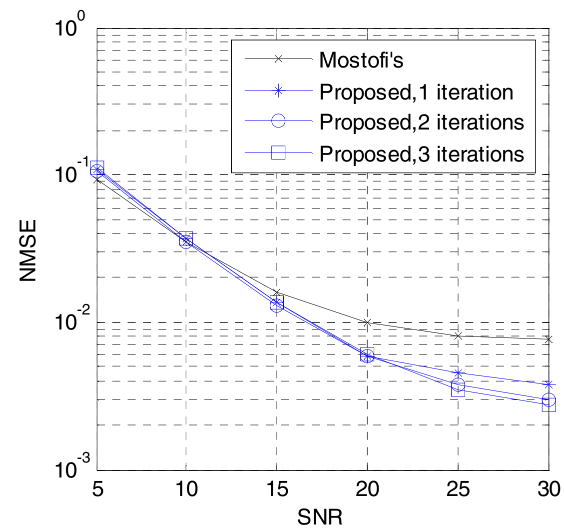

Fig. 3. Channel estimation performance, $f_D=0.1$.

At first, the performance of the proposed CE is evaluated. We defined the average normalized mean square error ($\overline{NMSE}$) of CE as

$$\overline{NMSE}=\frac{1}{M}\sum_{i=1}^{M}\frac{\sum_{l=0}^{L-1}E\left\{\left|h(n,l)-\hat{h}(n,l)\right|^2\right\}}{\sum_{l=0}^{L-1}E\left\{\left|h(n,l)\right|^2\right\}} \tag{28}$$

where M denotes the number of OFDM symbols. Fig. 2 and 3 illustrate the $\overline{NMSE}$ with respect to SNR (dB) when f_D are 0.05 and 0.1 respectively. Compared with the method proposed

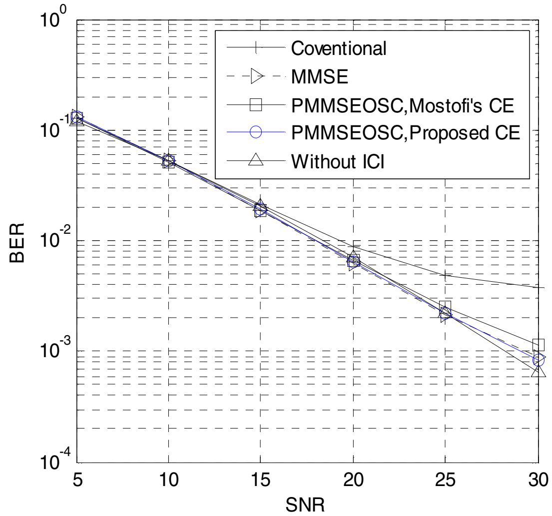

Fig. 4. BER for different equalizers and estimation schemes, $f_D = 0.05$.

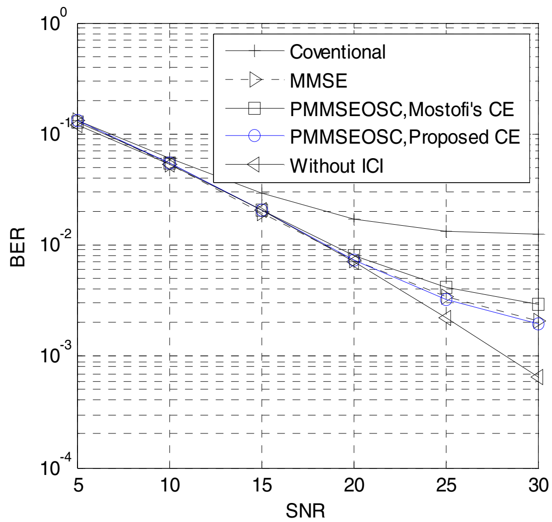

Fig. 5. BER for different equalizers and estimation schemes, $f_D = 0.1$.

in [6], our proposed method has a better performance in both cases. And two or three iterations are enough. The performance of Mosttofi's CE method is degraded due to pilots corrupted by ICI, but the proposed technique could cancel ICI and acquire a more accurate estimation to improve estimation performance by adopting an iterative way. Meanwhile, it is obvious that the higher the f_D , the poorer the performance of CE, which means the assumption of the linear model is invalid for the channel with a very high Doppler frequency shift.

Finally the equalization performance base on the estimated channel is tested. The BER performances are plotted in Fig. 4 and 5 for different values of Doppler frequency shift ($f_D = 0.05$, $f_D = 0.1$). As is seen from figures, conventional single-tap equalizer is not necessarily good for doubly-selective fading channels. The PMMSEOSC equalizer outperforms the conventional and nearly has the same performance as MMSE equalizer, but has much lower complexity than MMSE equalizer. As expected, the performance improves when using proposed CE method, because the estimation is less contaminated by ICI than using Mastofi's method.

VI. Conclusion

The assumption that the channel is stationary over an OFDM symbol results in an error floor owing to ICI in time-varying channel. In this case, the successful data detection poses some challenges in terms of ICI-free equalization and CE. In this paper, an iterative CE scheme was proposed based on a piece-wise linear model to approximate channel variations. Performance improvement was shown by simulation results in high Doppler spread environments. Meanwhile, a low-complexity equalizer is proposed to further improve BER performance. Anyway, the proposed techniques have good performance and low computational burdens.

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