Sparse channel recovery with inter-carrier interference self-cancellation in OFDM

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Abstract: A new sparse channel estimation method of orthogonal frequency division multiplexing (OFDM) system based on inter-carrier interference (ICI) self-cancellation is investigated. Firstly, based on the characteristic that the ICI generated by a subcarrier to the two adjacent subcarriers is approximately equal, a data pair with opposite sign and equal magnitude is modulated onto two adjacent subcarriers as pilot pair to eliminate the effect of ICI on pilots. Secondly, a new OFDM channel estimation model based on linear time-varying (LTV) model and compressed sensing (CS) is constructed, which obtains the mean of the gains of the multipath. Finally, a pilot pair optimization algorithm based on two layers loop is used to realize the minimization of the mutual coherence of the measurement matrix. For time-varying channel scenes with different numbers or delay of multipath and maximum Doppler frequency shift, the performances of several channel estimation methods are verified by simulation. The result shows that the new method has obvious advantage in both the performance of the channel estimation and the spectral efficiency.

Keywords: orthogonal frequency division multiplexing (OFDM), channel estimation, compressed sensing (CS), inter-carrier interference (ICI), pilot design.

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1. Introduction

There are two main problems for orthogonal frequency division multiplexing (OFDM) system channel estimation in high speed environment [1,2]. Firstly, the value of channel impulse response (CIR) is not constant during one OFDM symbol period, and the number of parameters to be estimated is very huge. Secondly, the orthogonality between the subcarriers is destroyed due to the Doppler frequency shift, which causes inter-carrier interference (ICI).

For the first issue, the channel matching model is usually used to reduce the number of parameters to be estimated in time-varying channels, such as linear time-varying (LTV)

channel model and basis expansion model (BEM). The LTV model is relatively simple, and when the normalized maximum Doppler frequency shift is less than 0.2, the performance of LTV is perfect and suitable for most mobile communication occasions [3–6]. The LTV model was first proposed by Mostofi in [3]. The main idea of the method is that the variation of each CIR is assumed to be a linear approximation with a constant slope in one OFDM symbol, then the averages and slopes of each multi-path is estimated by equal space pilots during three successive OFDM symbols. Because the pilots are affected by the ICI from the unknown data, the performance of channel estimation is not ideal. As for the second issue, there are several main ICI cancellation methods based on the LTV model, including iterative method [4], window-LTV [5] and ICI self-cancellation [6,7]. However, the iteration increases the amount of computation and reduces the timeliness. The window function is added into the OFDM system, which reduces the ICI of out band, but needs to change the structure of OFDM system. The ICI self-cancellation method in [6] and [7], uses '1' and '–1' as two adjacent pilots to realize the ICI self-cancellation, but introduces notable performance degrade in long-delay channel, as shown in Section 5. In addition, the common drawbacks of these methods include: the pilot number cannot be smaller than the channel length, and the system frequency utilization is lower. And the threshold is used to determine the effective taps, and the error is large under low signal to noise ratio (SNR).

In recent years, a large number of experimental results show that the wireless multipath channel with scattering environment distribution has obvious sparse characteristics and strong temporal correlation [8,9]. The number of multipath is far smaller than the length of the channel. The change of multipath delay can be considered as constant during one or even a few OFDM symbols [10,11]. The compressed sensing (CS) theory can be applied in the channel estimation to gain better performance, and the ICI

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can also be reduced through reasonable pilot design, which is called sparse recovery of time-varying channel [12–14]. However, the computation of sparse recovery based on the BEM is still very large. And the research into the sparse channel estimation based on the LTV model is very little. Finally, the pilot pattern design is important for the reconstruction of sparse channel, however, at present, the researches on this problem are all aimed at the untime-varying channel, and only consider the mutual coherence of the measurement matrix, ignoring the influence of ICI, which is not suitable for sparse reconstruction of time-varying channels [14–16].

In view of the above problems, this paper analyzes the characteristics of ICI, uses the special pilot pattern to eliminate the ICI from unknown data. And the estimation of the mean values of multi-path gains is converted to a sparse reconstruction problem.

The rest of this paper is organized as follows: The signal model is described in Section 2. In Section 3, the ICI characteristic of OFDM system is analyzed and an ICI mitigation based on pilot pair is discussed. Section 4 presents the new sparse recovery model and the pilot optimization algorithm. In Section 5, the simulation results are given. Finally, Section 6 concludes this paper.

2. System model

2.1 OFDM model

We assume an OFDM system with N subcarriers. $x[n]$ and $y[n]$ denote the transmitted symbol and the received symbol in time domain respectively, and $n \in [0, N-1]$. Then the received OFDM symbol $y[n]$ after cyclic prefix (CP) removal is given by

$$y[n] = \sum_{l=0}^{L-1} h[n, l] x[n-l] + w[n] \quad (1)$$

where $h[n, l]$ represents the complex zero-mean unit-variance CIR tap at lag l ($l = 0, 1, \dots, L-1$) and time instant n , and L is the maximum multipath delay. $w[n]$ is the additive noise. Taking the discrete Fourier transform (DFT) of (1), we can get

$$Y_k = G_{k,k} X_k + \sum_{m=0, m \neq k}^{N-1} G_{k,m} X_m + W_k \quad (2)$$

where X_k and Y_k denote the modulated signal and demodulated signal of the k th carrier respectively, $G_{k,k} X_k$ is the ideal received signal in the case of ICI-free channel, $\sum_{m=0, m \neq k}^{N-1} G_{k,m} X_m$ is the one in the case of ICI, and W_k is additive noise. $G_{k,m}$ ($m \neq k$) and $G_{k,k}$ are represented

respectively as

$$G_{k,m} = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} h[n, l] e^{-j2\pi n(k-m)/N} e^{-j2\pi ml/N} \quad (3)$$

$$G_{k,k} = \sum_{l=0}^{L-1} h^{\text{ave}}(l) e^{-j2\pi kl/N} \quad (4)$$

where $h^{\text{ave}}(l) = \frac{1}{N} \sum_{n=0}^{N-1} h[n, l]$ denotes the time average of $h[n, l]$ over one OFDM symbol. The $h^{\text{ave}}(l)$ for all paths can be denoted as $\mathbf{h}^{\text{ave}} = [h^{\text{ave}}(0), h^{\text{ave}}(1), \dots, h^{\text{ave}}(L-1)]$. Because the channel is sparse, most of components of $\mathbf{h}^{\text{ave}}$ are zero, or approximately zero, which means $\mathbf{h}^{\text{ave}}$ is a sparse vector.

2.2 LTV channel model

For the LTV channel model, CIR can be equivalent to a linear change during one OFDM symbol period. The LTV model can be divided into two kinds: single line and double line, also called piece-wise. The former is to estimate the slope by using the CP of the current OFDM symbol, and the latter is using three consecutive symbols. In contrast, the model error of the latter is much smaller, and the performance is much higher, which is more widely used, shown as Fig. 1.

Fig. 1 Piece-wise LTV model

In the piece-wise LTV model, $h[n, l]$ of current symbol can be denoted as

$$\begin{cases} h[n, l] = h\left[\frac{N}{2} - 1, l\right] + s_l^1 \times \left(n - \frac{N}{2} + 1\right), \\ n \in \left[0, \frac{N}{2} - 1\right] \\ h[n, l] = h\left[\frac{N}{2} - 1, l\right] + s_l^2 \times \left(n - \frac{N}{2} + 1\right), \\ n \in \left[\frac{N}{2}, N - 1\right] \end{cases} \quad (5)$$

where s_l^1 and s_l^2 are the slopes of region 1 and region 2.

They can be obtained by using adjacent symbols.

$$\begin{cases} s_l^1 = \frac{h\left[\frac{N}{2}-1, l\right] - h^{\text{previous}}\left[\frac{N}{2}-1, l\right]}{N} \\ s_l^2 = \frac{h^{\text{next}}\left[\frac{N}{2}-1, l\right] - h\left[\frac{N}{2}-1, l\right]}{N} \end{cases} \quad (6)$$

Approximation error would be minimized, if we approximate $h\left[\frac{N}{2}-1, l\right]$ with the estimation of $h^{\text{ave}}(l)$ [3]. That is $h\left[\frac{N}{2}-1, l\right] \approx h^{\text{ave}}(l)$, so the relationship between the received signal and the sent signal in frequency domain can also be expressed as

$$\mathbf{Y} = \mathbf{H}^{\text{ave}} \mathbf{X} + (\mathbf{C}^{s1} \mathbf{H}^{s1} + \mathbf{C}^{s2} \mathbf{H}^{s2}) \mathbf{X} + \mathbf{W} \quad (7)$$

where $\mathbf{Y} = [Y_0, Y_1, \dots, Y_{N-1}]^T$, $\mathbf{X} = [X_0, X_1, \dots, X_{N-1}]^T$, $\mathbf{W} = [W_0, W_1, \dots, W_{N-1}]^T$, $\mathbf{H}^{\text{ave}} = \text{diag}\{\text{FFT}([h^{\text{ave}}(0), \dots, h^{\text{ave}}(L-1), 0, \dots, 0])\}$, $\mathbf{H}^{s1} = \text{diag}\{\text{FFT}([s_0^1, s_1^1, \dots, s_{L-1}^1, 0, \dots, 0])\}$, $\mathbf{H}^{s2} = \text{diag}\{\text{FFT}([s_0^2, s_1^2, \dots, s_{L-1}^2, 0, \dots, 0])\}$. $\mathbf{C}^{s1}$ and $\mathbf{C}^{s2}$ are fixed matrixes, and the special values are shown in [3].

Obviously, it is a key issue to estimate $h^{\text{ave}}(l)$ of one OFDM symbol for the channel estimation based on the LTV model.

3. ICI characteristic and self-cancellation method

3.1 ICI characteristic of OFDM system

When the angular spread is small for non-resolvable components of a channel, the discrete-time CIR of a time-varying multipath channel [6,18] can be also expressed as

$$h[n, l] = \sum_{u=0}^{L-1} a_u e^{\frac{j2\pi\Delta f_u n}{N}} \delta(l-u) \quad (8)$$

where a_u is the complex number for the u th resolvable multipath (tap), Δf_u is the corresponding Doppler frequency normalized by one subcarrier spacing. By substituting (8) into (3), we can get

$$\begin{aligned} G_{k,m} &= \frac{1}{N} \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} h[n, l] e^{-j2\pi n(k-m)/N} e^{-j2\pi ml/N} = \\ &= \frac{1}{N} \sum_{l=0}^{L-1} \left\{ a_l \frac{\sin[\pi(m + \Delta f_l - k)]}{N \sin[\frac{\pi}{N}(m + \Delta f_l - k)]} \right. \\ &\quad \left. e^{-j2\pi ml/N} e^{j\frac{\pi(N-1)}{N}(m + \Delta f_l - k)} \right\}. \end{aligned} \quad (9)$$

If we only take the l th tap into consideration, and use $S_l(k, m)$ to denote the ICI coefficient between the k th subcarrier and the m th subcarrier, then the $G_{k,m}$ can be taken as a linear combination of $S_l(k, m)$ ($l = 0, 1, \dots, L-1$).

$$S_l(k, m) = \frac{\sin[\pi(m + \Delta f_l - k)]}{N \sin[\frac{\pi}{N}(m + \Delta f_l - k)]} e^{-j2\pi ml/N} e^{j\frac{\pi(N-1)}{N}(m + \Delta f_l - k)} \quad (10)$$

Obviously, the amplitude of $S_l(k, m)$ only depends on $|m - k|$. Fig. 2 shows $|S_l(k, m)|$ for different m and k in the case of $\Delta f_l = 0.1$.

(a) All values of k and m , for $N=64$ and $\Delta f_l=0.1$

(b) An example for a given $m=0, 16$ and 32

Fig. 2 Value of $|S_l(k, m)|$ for different m and k

According to Fig. 2, we can draw two conclusions: First, the ICI from a certain subcarrier mainly affects its adjacent tones. Second, for the majority of $m - k$, the value of $|S_l(k, m)|$ changes little. For a given m and most k , $|S_l(k, m)| \approx |S_l(k+1, m)|$. Because the $G_{k,m}$ can be seen as a linear combination of $|S_l(k, m)|$, and due to the sparsity of wireless channel, for $l = 0, 1, \dots, L-1$, most of a_l is zero, then the ICI generated by a subcarrier to the two adjacent subcarriers is approximately equal. That is, for a given m and most of k , $G_{k,m} \approx G_{k+1,m}$.

3.2 ICI mitigation based on pilots pair

Based on the analysis of Section 3.1, when we design the pilot pattern, a data pair $(-1, 1)$ is modulated onto two adjacent subcarriers $(k_P, k_P + 1)$ as a pilot pair, that is to say, $X(k_P) = -X(k_P + 1)$, $k_P \in \mathbf{P}$, where $\mathbf{P} = \{P_1, P_2, \dots, P_{N_P}\}$ is the pilots location set, and N_P is the toll number of pilots, $0 \leq P_1, P_1 + 1 = P_2, P_3 + 1 = P_4, \dots, P_{N_P-1} + 1 = P_{N_P}, P_{N_P} \leq N - 1$. Then according to (2), the pilots received on the k_P th and $(k_P + 1)$ th sub-carrier can be expressed as follows:

$$Y_P(k_P) = X(k_P)G_{k_P, k_P} + \sum_{m=0, m \neq k_P}^{N-1} X(m)G_{k_P, m} + W(k_P) \quad (11)$$

$$Y_P(k_P + 1) = X(k_P + 1)G_{k_P+1, k_P+1} + \sum_{m=0, m \neq k_P+1}^{N-1} X(m)G_{k_P+1, k_P+1} + W(k_P + 1). \quad (12)$$

Through (11) and (12), we can get

$$Y_P(k_P) - Y_P(k_P + 1) = X(k_P)[G_{k_P, k_P} + G_{k_P+1, k_P+1}] + ICI_{k_P} - ICI_{k_P+1} + W(k_P) - W(k_P + 1) \quad (13)$$

where $ICI_{k_P} = \sum_{m=0, m \neq k_P}^{N-1} X(m)G_{k_P, m}$, $ICI_{k_P+1} = \sum_{m=0, m \neq k_P+1}^{N-1} X(m)G_{k_P+1, m}$. Based on the conclusion of Section 3.1, for a given m and most of k , $G_{k, m} \approx G_{k+1, m}$, $ICI_{k_P} - ICI_{k_P+1} \approx 0$. Obviously, compared with (11) or (12), the effect of ICI on pilots in (13) is much less, and the performance of channel estimation will be improved.

4. A new channel sparse recovery method

4.1 New sparse recovery model based on pilot pair

For (13), the classical channel estimation model based on CS theory, such as in [12–14], is not suitable. Therefore, we must build a new channel sparse recovery model.

Because the ICI is very little in (13), it can be treated as noise. Assuming $W'(k_P, k_P + 1) = ICI_{k_P} - ICI_{k_P+1} + W(k_P) - W(k_P + 1)$, substituting (3) and (4) into (13), we can get

$$Y_P(k_P) - Y_P(k_P + 1) = X(k_P) \left[\sum_{l=0}^{L-1} h^{\text{ave}}(l) e^{-j2\pi k_P l / N} + \sum_{l=0}^{L-1} h^{\text{ave}}(l) e^{-j2\pi (k_P+1) l / N} \right] + W'(k_P, k_P + 1). \quad (14)$$

Based on the pilot pair design principle of Section 3.2, we divide the pilot location set $\mathbf{P}$ into two subsets, $\mathbf{P}_{\text{odd}} = \{P_1, P_3, \dots, P_{N_P-1}\}$ and $\mathbf{P}_{\text{even}} = \{P_2, P_4, \dots, P_{N_P}\}$. Then (14) can be written in matrix notation as

$$\mathbf{Y}_{\text{odd}} - \mathbf{Y}_{\text{even}} = \mathbf{X}_{\text{odd}} \mathbf{F}_{\text{odd}} \mathbf{h}^{\text{ave}} - \mathbf{X}_{\text{even}} \mathbf{F}_{\text{even}} \mathbf{h}^{\text{ave}} + \mathbf{W}_{\text{odd}} - \mathbf{W}_{\text{even}} \quad (15)$$

where for $i \in \mathbf{P}_{\text{odd}}$ and $j \in \mathbf{P}_{\text{even}}$, $\mathbf{X}_{\text{odd}}$ and $\mathbf{X}_{\text{even}}$ are both the subset of $\mathbf{X}$, and written as diagonal matrix, whose diagonal elements are all X_i and X_j , respectively. $\mathbf{Y}_{\text{odd}}$ and $\mathbf{Y}_{\text{even}}$ are column vectors, which are composed of all Y_i and Y_j , respectively. $\mathbf{W}_{\text{odd}}$ and $\mathbf{W}_{\text{even}}$ are column vectors, which are composed of all W_i and W_j , respectively. $\mathbf{F}_{\text{odd}}$ is a sub-matrix selected by the row indices $\mathbf{P}_{\text{odd}}$ and column indices $[0, 1, \dots, L-1]$ from $\mathbf{F}_{N \times N}$, $\mathbf{F}_{\text{odd}}$ is a sub-matrix selected by the row indices $\mathbf{P}_{\text{even}}$ and column indices $[0, 1, \dots, L-1]$ from $\mathbf{F}_{N \times N}$, where $\mathbf{F}_{N \times N}$ is the standard DFT matrix.

Assume $X_i = 1, i \in \mathbf{P}_{\text{odd}}$, and $X_j = -1, j \in \mathbf{P}_{\text{even}}$, then (15) can be simplified as

$$\mathbf{Y}_{\text{odd}} - \mathbf{Y}_{\text{even}} = \mathbf{A} \mathbf{h}^{\text{ave}} + \mathbf{W}' \quad (16)$$

where $\mathbf{Y}' = \mathbf{Y}_{\text{odd}} - \mathbf{Y}_{\text{even}}$, $\mathbf{A} = \mathbf{F}_{\text{odd}} - \mathbf{F}_{\text{even}}$ and $\mathbf{W}' = \mathbf{W}_{\text{odd}} - \mathbf{W}_{\text{even}}$.

Obviously, (16) satisfies the CS model, and the sparse recovery algorithm, such as orthogonal matching pursuit (OMP) can be employed to reconstruct $\mathbf{h}^{\text{ave}}$ with fewer pilots and higher accuracy. $\mathbf{A}$ is called measurement matrix.

4.2 Pilot pair optimization algorithm

In channel estimation based on CS theory, in order to make $\mathbf{A}$ satisfy restricted isometry property (RIP) [17], a universal approach is to minimize the mutual coherence of $\mathbf{A}$, which is denoted as

$$u(\mathbf{A}) = \max_{0 \leq c < r \leq L-1} |\mathbf{a}_c^H \mathbf{a}_r| = \max_{0 \leq c < r \leq L-1} \left| \sum_{s=1}^{N_P} e^{-j2\pi p_s(r-c)/N} \right| \quad (17)$$

where $\mathbf{a}_r$ and $\mathbf{a}_c$ are the r th and c th column vector of $\mathbf{A}$.

Because $\mathbf{A}$ only depends on the pilot location set $\mathbf{P}$, so the pilots optimization is to find a pilot location set $\mathbf{P}_{\text{opt}} = \arg \min_{\mathbf{P}} \mu(\mathbf{A})$.

Here, based on the special pilot pair in Section 3.2, a pilot pair optimization algorithm based on dual layer loops iteration is proposed, and is illustrated in Algorithm 1.

Algorithm 1 Pilot pair optimization based on dual layer loops iteration

Initialization:

(i) Variables: number of groups G_P , and $N_P = 2G_P$; length of each group L_g , and $L_g = N/G_P$; the sub-carries

set of all group Γ , the size is $G_P L_g$, and $\Gamma[m, n]$ is the index of the n th carrier of the m th group, the elements of the m th row of Γ are $(m-1)L_g + 1 : mL_g$; **pattern_best**, **pattern_last** and **pattern_tempt** are three vectors that save the current result, the result of the last loop and the tempt result of optimization.

(ii) Pilot pairs: for each group, the first and second carriers are used as pilots. The initial pilot pairs are $(1, 2)$, $(L_g + 1, L_g + 2), \dots, ((m-1)L_g + 1, (m-1)L_g + 2), \dots, ((G_P - 1)L_g + 1, (G_P - 1)L_g + 2)$, saved in the **pattern_best**.

Loop iteration:

while (1)

pattern_last= **pattern_best**;

for $i_1 = 1 : G_P$

for $i_2 = \Gamma[i_1, 1] : \Gamma[i_1, L_g]$

submit the $((i_1 - 1)L_g + 1)$ th, and the $((i_1 - 1)L_g + 2)$ th element of **pattern_last** with i_2 and $i_2 + 1$, then get a new pattern **pattern_tempt**.

According (17), compare the coherence of the measurement matrix of **pattern_tempt** and **pattern_last**.

If $\mu(\text{pattern_tempt}) < \mu(\text{pattern_last})$, **pattern_last** =

pattern_tempt.

endif

endfor

endfor

if $\mu(\text{pattern_best}) == \mu(\text{pattern_last})$, break;

else **pattern_best**= **pattern_last**;

endif

end while

5. Simulation results

5.1 Simulation setting

Two classical channel models are used: COST 207 RA (rural area) and COST 207 BU (bad urban). The parameters of channel and OFDM system are shown as Table 2 and Table 3, respectively.

Table 2 COST 207 RA/BU channel model

Path	Path delay/ μ s	Path gain/dB	Doppler spectrum
1	0/0	0/-3	Rice/Jakes
2	0.4/0.4	-2/0	Jakes/Jakes
3	0.4/1.0	-10/-3	Jakes/Gauss1
4	0.6/1.6	-20/-5	Jakes/Gauss1
5	Null/5.0	Null/-2	Null/Gauss2
6	Null/6.6	Null/-4	Null/Gauss2

Table 3 OFDM system parameter setting

System bandwidth/MHz	Sample period/ns	Carriers space/kHz	Carriers number	Guard interval	Chanel code	System frequency/GHz	Modulation type
25.6	31.25	25	1 024	256	No	1-4	16QAM

According to the application requirements and channel parameters, the maximum moving speed of the mobile station is 50 m/s, the normalized Doppler frequency shift $v_{\max}$ is between 0.012 and 0.036, which satisfies the LTV channel model. Two extreme conditions are selected in simulation, that is, $v_{\max} = 0.012$ and $v_{\max} = 0.048$, which stand for the “small” and “large” normalized Doppler frequency shift, for RA and BU environment, four kind channels are denoted as RA-small, RA-large, BU-small, BU-large.

5.2 Performance comparasion of different methods

We compare the performance of three different LTV channel estimation methods, including the traditional method in [3], the improved method in [6] and [7], and the new method in this paper. The traditional method and the improved method all require the number of pilots not smaller than the channel length, for the worst case, the number of pilots used in these two methods is 256 ($N_P=256$), and both using comb-type pilot pattern with equal space, denoted as “Traditional-256” and “Improved-256”. For the new method presented in this paper, we use the pilot design optimization algorithm in Section 4.2 to obtain two

optimal pilot patterns with different lengths, $N_P = 192$ and $N_P = 256$, denoted as “New-192” and “New-256”. The mean square error (MSE) and bit error ratio (BER) of three methods are shown as from Fig. 3 to Fig. 6.

Through the comparison of MSEs and BERs in different channel environment, we can draw the following conclusions: First, with the increase of the number of multipath and Doppler shift, the performance of three methods reduces with different degrees.

Fig. 3 MSEs and BERs of three methods in COST 207RA-small ($v_{\max} = 0.012$)

Fig. 4 MSEs and BERs of three methods in COST 207RA ($v_{\max} = 0.048$)

Fig. 5 MSEs and BERs of three methods in COST 207BU ($v_{\max} = 0.012$)

Compared to the other two methods, the change of the traditional LTV method in [3] is not obvious, and it is much more robust to channel, however, whose performance is not the best in these scenarios. Second, for the improved method in [6] and [7], the effect of ICI is both taken into consideration during pilot design and equalization, so in the COST 207 RA channel, its performance is much better than the traditional method. However, the result in COST 207 BU channel is very poor, so it is only suitable for simple multipath channel with low delay. Third, the performance of the new method is much better than the traditional method, for all channel environments except the one with large delay and Doppler shift. Furthermore, compared with the latter, the former uses fewer pilots, and saves about 1/4. Compared with the improved method in [6] and [7], the result of the new method is improved with different degrees, especially for COST 207 BU channel. Finally, due to using the sparse reco-

very algorithm and pilot optimization algorithm, the new method does not have strict requirement on the number of pilots, and the performance difference between $N_P = 192$ and $N_P = 256$ is not significant, so it also improves the utilizing rate of frequency. Obviously, not only for the performance, but also the frequency utilization, the new method is all the best among the current channel estimation methods based on LTV.

Fig. 6 MSEs and BERs of three methods in COST 207BU ($v_{\max} = 0.048$)

6. Conclusions

Base on the principle of the LTV model and the ICI, we design a new pilot pattern with ICI self-elimination. Then, a new OFDM channel recovery model based on CS is reconstructed. Moreover, we also design a pilot pair optimization algorithm to realize the minimization of mutual coherence of the measurement matrix. In this paper, the

experiments are limited to four kinds of simulated communication environments, and results show the good performance of the new methods among all current methods based on the LTV model. We will investigate its effect on more realistic OFDM systems in the future.

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