A Novel Comb-Pilot Transform Domain Frequency Diversity Channel Estimation for OFDM System

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Abstract. *Due to implementation complexity, the transform domain channel estimation based on training symbols or comb-type pilots has been paid more attention because of its efficient algorithm FFT/IFFT. However, in a comb-type OFDM system, the length of the channel impulse response is much smaller than the pilot number. In this case, the comb-pilot transform domain channel estimation only works as interpolation like the Least Squares (LS) algorithm, but loses the noise suppression function. In this paper, we propose a novel frequency diversity channel estimation method via grouped pilots combining. With this estimator, not only the channel frequency response on non-pilot subcarriers can be interpolated, but also the noise can be better suppressed. Moreover, it does not need prior statistical characteristics of the wireless channel.*

Keywords

Channel estimation, transform domain, orthogonal frequency division multiplexing (OFDM), noise suppression.

1. Introduction

Orthogonal frequency division multiplexing (OFDM) has received a lot of interests recently since it can transmit data at a very high speed. It has been the key technique in wireless LAN (WLAN), 3rd Generation Mobile Group Long Term Evolution (3GPP LTE) and WiMAX for its advantages in transmitting over frequency selective fading channels, anti-interference of narrow band and combining MIMO easily [1], [2]. However, perfect channel state information (CSI) is required for signal detection and its accuracy determines the performance of OFDM communication system directly, especially for those with high order modulation or MIMO techniques.

The comb-pilot OFDM is widely used because it can estimate the time-varying wireless channel without frequency efficiency reduction [3], [4]. Then channel frequency response (CFR) of other non-pilot subcarriers used for transmitting data information can be obtained with linear interpolation [5] or spline interpolation [6] with the comb-pilots. These methods are performed in frequency domain. In recent years, more and more attention has been paid to transform domain channel estimation because of its efficient and fast algorithm via FFT/IFFT [7]-[10]. With this method, after the initial transformation of the training symbols or pilots, if the length of multipath wireless channel can be obtained, we can zero the terms out in transform domain (time domain) corresponding to noise while retaining the significant taps corresponding to the true delay taps. However, being lack of the knowledge on the order of wireless channel, when transform domain channel estimator is employed, it is difficult to zero-force the noise in transform domain in a comb-type pilot or training symbol OFDM systems. The traditional method is zero-padding at first, and then directly transform back to frequency domain, thus the noise remains on each subcarrier [8]. In this way, the comb-pilot transform domain channel estimation only works as interpolation on non-pilot subcarriers but causes large loss in noise suppression. If the distance of the pilots is less than the coherence bandwidth of the multipath wireless channel, the performance of this traditional method is the same as LS algorithm [11] for all pilots when the channel CIR taps are sample spaced. If the CIR taps are not sample spaced, error floor will occur. This will be analyzed further in the following sections of this paper.

In this paper, we propose a novel comb-pilot transform domain based frequency diversity channel estimation for OFDM systems. First of all, all the pilots in one OFDM symbol are divided into several groups, in which the distance between two adjacent pilots in a sub-group should be less than the coherence bandwidth of the multipath wireless channel. Then the pilots in the sub-group estimate the CRF of the wireless channel respectively so that several CFRs of the same wireless channel can be obtained. Finally, we can combine these CFRs with some rules, such as equiprobability rule. As a result, the combined CFR shows better performance than LS algorithm due to stronger noise suppression. Furthermore, the prior statistical characteristics of the multipath wireless channel is unnecessary while employing this method.

The rest of this paper is organized as follows. Section 2 describes the comb-pilot OFDM system and multipath channel model. In Section 3, the relationship between the comb-pilot transform domain channel estimation and the LS algorithm for all pilots is analyzed. And also the equivalence of these two methods is shown if the channel CIR taps are sample spaced. Then a novel comb-pilot transform domain frequency diversity channel estimation for OFDM system is presented. In Section 4 we compare the proposed algorithm results with the traditional methods using computer simulation, and the conclusions are drawn in Section 5.

2. Comb-Pilot OFDM System Model

In an OFDM transmitter, comb-pilots are inserted in frequency domain after binary bit stream is mapped. Here, we consider N_p pilots locate in each OFDM symbol and equidistance between two pilots is $\Delta P = N/N_p$ without considering the virtual subcarriers and DC tone. The data subcarrier and the inserted pilot signal in frequency domain can be expressed as:

$$X(k) = \begin{cases} X_p(m\cdot\Delta P) & m \text{ is an integer,} \quad 0 \le m \le N_p - 1 \\ X_D & \text{otherwise} \end{cases} \tag{1}$$

where $X(k)$ is the signal carried by a subcarrier in frequency domain, X_D denotes the data and X_p represents the pilot. Then OFDM transmitter uses an inverse IDFT (IFFT) with size of N to modulate each subcarrier. Thus the time domain signal $x(n)$ can be expressed as:

$$x(n) = \frac{1}{N}\sum_{k=0}^{N-1} X(k)exp(j2\pi\frac{kn}{N}) \quad 0 \le n \le N-1 \tag{2}$$

where n is the signal index of an OFDM symbol in time domain. Then cyclic prefix (CP) whose length is much longer than the maximum delay tap of multipath wireless channel is added to each OFDM symbol to avoid inter-symbol interference. After that the OFDM symbol passes through the multipath channel. For simplicity, the channel tap is sample spaced. Otherwise the interpolation can be applied to obtain the spaced tap gain. Here, we assume the channel is a time-varying multipth channel. The multipath channel can be modeled as a tapped delay line structure as [8]:

$$h(n) = \sum_{i=0}^{r-1} h_i \exp\left(j\frac{2\pi}{N}f_{Di}Tn\right)\delta(\lambda - \tau_i) \tag{3}$$

where r is the total number of resolvable delay paths in channel. h_i is the complex gain of i-th propagation path. f_{Di} is the i-th path Doppler frequency shift caused by the relative motion of the transceivers. T is the duration of each OFDM symbol. λ is the delay spread index, and τ_i is the i-th path delay time normalized by sampling interval. Then the received signal is as follows:

$$y(n) = x(n) * h(n) + w(n) \tag{4}$$

where $'*'$ denotes the convolution operation, $w(n)$ is additive white Gaussion noise (AWGN).

Because of CP, linear convolution of the transmitted signal $x(n)$ and the channel impulse response $h(n)$ changes into circular convolution. At the receiver side, after perfect synchronization and CP removal, the received signal is demodulated by DFT (FFT) with size of N. And then the signal on each subcarrier in frequency domain can be written as:

$$Y(k) = X(k)H(k) + W(k) \tag{5}$$

where $H(k)$ is the CFR of the subcarrier, $W(k)$ is the AWGN with zero mean and variance σ^2 for the k-th subcarrier.

3. Comb-Pilot Transform Domain Frequency Diversity Channel Estimation

3.1 Unification of Comb-Pilot Traditional Transform Domain Channel Estimation with Frequency Domain LS Approach

According to (5), the CFR of the pilot subcarrier based LS algorithm is given as [11]:

$$\begin{aligned} \tilde{H}_p(k) &= X_p^*(k)Y_p(k) \\ &= |X_p(k)|^2 H_p(k) + X_p^*(k)W_p(k) \\ &= H_p(k) + W_p'(k) \end{aligned} \tag{6}$$

where the subscript p indicates that this subcarrier is used for pilot and $W_p'(k)$ also follows AWGN distribution. Here, BPSK is employed for pilot, then $|X_p(k)|^2 = 1$.

Actually, if the channel CIR taps are sample spaced, and the distance between two adjacent pilots is less than the coherence bandwidth, comb-pilot traditional transform domain channel estimation and frequency domain LS approach perform equivalently. With the assumption that channel taps are sample spaced, the mean square error (MSE) of LS approach is written as [7]:

$$\text{MSE} = \frac{\beta}{\text{SNR}} \tag{7}$$

where $\text{SNR} = E\{|x_k|^2\}/\sigma^2$ is the average signal to noise ratio. Here, x_k is transmitted signal on each subcarrier. We assume that all the tones are distributed independently and use the same constellation. β is a constellation factor and expressed as $\beta = E\{|x_k|^2\}E\{|x_k|^{-2}\}$. For BPSK and QPSK, $\beta = 1$, for 16QAM, $\beta = 17/9$.

Traditional comb-pilot transform domain channel estimation algorithm is expressed as [12]:

$$\mathbf{H}_{DFT} = \sqrt{N_P/N}\mathbf{F}\mathbf{D}\mathbf{F}_p^H\widetilde{\mathbf{H}}_{LS} \tag{8}$$

where $\mathbf{F}$ is the unitary FFT matrix with elements $e^{-j2\pi ik/N}$. $(\cdot)^H$denotes conjugate-transpose operation. $\mathbf{F}_p^H$ is the unitary IFFT matrix with elements $e^{j2\pi ik/N_p}/N_p$ and $\widetilde{\mathbf{H}}_{LS}$ is the channel estimation on pilot subcarriers with LS algorithm. D is an upper triangular diagonal matrix used for zero-padding from N_p to N as interpolation function, which is:

$$\mathbf{D} = \begin{bmatrix} \mathbf{I}_{N_p \times N_p} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} \end{bmatrix}_{N \times N} \tag{9}$$

where $\mathbf{I}_{N_p \times N_p}$ is an identity matrix with size of $N_p \times N_p$. If some information such as the order of wireless channel is known at the receiver side, N_p can decrease to some point for noise suppression.

Here, we define the auto-covariance matrix of CFR as

$$\mathbf{R}_{HH} = E\{\mathbf{H}\mathbf{H}^H\} = E\{\mathbf{F}\mathbf{h}(\mathbf{F}\mathbf{h})^H\} = \mathbf{F}\mathbf{R}_{hh}\mathbf{F}^H \tag{10}$$

where $\mathbf{H}$ is a vector of CFR in frequency domain. $\mathbf{h}$ is a vector of CIR in time domain. $\mathbf{R}_{hh}$ is the auto-covariance matrix of CIR and a Hermitian matrix. Then the singular value decomposition (SVD) is performed over Hermitian $\mathbf{R}_{hh}$ as:

$$\mathbf{R}_{hh} = \mathbf{U}\sum\mathbf{U}^H \tag{11}$$

where $\mathbf{U}$ is a unitary matrix, and $\sum$ is a diagonal matrix containing corresponding singular values of $\mathbf{R}_{hh}$ in descending order. Here, if the channel taps are sample spaced, SVD of Hermitian $\mathbf{R}_{hh}$ results in as many non-zero significant singular values as the significant CIR taps. However, this assumption is not always satisfied. Therefore, the concentrative power of CIR tap leaks to other taps so that the channel taps are correlated in some degree. Because more non-zero significant singular values occur in SVD, it brings an error floor in *MSE* that will be analyzed as below.

Now we directly give the explicit expression of *MSE* for comb-pilot traditional transform domain channel estimation (8) as:

$$\mathrm{MSE} = \frac{\beta}{\mathrm{SNR}} + \sum_{j=N_p}^{N-1} \rho_j. \tag{12}$$

It can be found that (12) can be considered as two parts. The first part β/SNR is determined when system is designed. The second part $\sum_{j=N_p}^{N-1} \rho_j$ corresponds to the channel singular value related to channel taps in time domain. As we mentioned above, N_p is larger than the length of channel taps. If channel taps are sample spaced, the power will concentrate only on practical channel taps. Because the length of CP is larger than the maximum delay tap of the wireless channel, it can be obtained that the singular value $\rho_j = 0$, $j = N_p$, $\cdots$, $N-1$. Hence, (12) becomes $\mathrm{MSE} = \beta/\mathrm{SNR}$. According to (7), it is obvious that in this case comb-pilot traditional transform domain channel estimation and frequency domain LS approach perform equivalently. However, if the channel taps are not sample spaced, in other words, sample time and channel are not matched, some of singular values are not equal to zero. So $\rho_j \neq 0, j > r$, where r indicates the total amount of resolvable delay paths

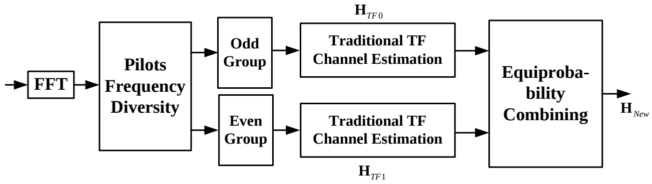

Fig. 1. Block diagram of the proposed estimator with diversity.

of the channel described in (3). This may lead to a result that $\sum_{j=N_p}^{N-1} \rho_j \neq 0$. In this case, when SNR is relatively high, the small term $\sum_{j=N_p}^{N-1} \rho_j$ is considerably more than β/SNR. This part is difficult to handle. Here, we define this component as:

$$\textbf{\textit{error floor}} = \sum_{j=N_p}^{N-1} \rho_j. \tag{13}$$

Here, ***error floor*** is invariant in a special channel, then (12) can be written as:

$$\mathrm{MSE} = \frac{\beta}{\mathrm{SNR}} + \textbf{\textit{error floor}}. \tag{14}$$

Therefore on the point of view of performance, we can estimate CFR by this comb-pilot transform domain method as follows:

$$\mathbf{H}_{TF} = \mathbf{H}_{LS} + \underline{\mathbf{e}} = \mathbf{H} + \mathbf{W} + \underline{\mathbf{e}} \tag{15}$$

where $\underline{\mathbf{e}}$ is the error floor caused by leakage when estimating CFR mentioned above.

3.2 Proposed Comb-Pilot Transform Domain Frequency Diversity Channel Estimation

According to (14), it can be found that MSE is decreasing with SNR increasing if SNR is not at high level. In this case, the transmitted signal power is determined when the system is designed. We improve the performance originating from frequency diversity for noise power reduction. Here, we divide all the pilots into L groups. The new distance ΔP_L between two adjacent pilots and the number of pilots N_{p_L} in one sub-group can be defined as:

$$\Delta P_L = L \cdot \Delta P \tag{16}$$

$$N_{p\_L} = N_p / L \tag{17}$$

It is assumed that the new distance between two adjacent pilots is less than the coherence bandwidth of the wireless channel. L-group pilots can be expressed as X_{p_l}, $l = 0, 1, \cdots, L-1$. For simplicity, we take $L = 2$ and the diagram of our method is shown in Fig. 1.

The pilots from the two groups can be obtained as $X_{p\_0} \cup X_{p\_1} = X_p$, where

$$\begin{cases} X_{p\_0}(i) = X_p(2i-1) \\ X_{p\_1}(i) = X_p(2i) \end{cases} \tag{18}$$

where $1 \leq i \leq N_p/2$. It can be found that X_{p_0} and X_{p_1} correspond to the pilots in the odd location and even location respectively. According to these two grouped pilots, we can apply them for two channel estimators respectively with comb-pilot traditional transform domain channel estimation. When both the CFRs are obtained, we can combine them with some combining rules, such as equiprobability rule. This result is as follows:

$$\mathbf{H}_{New} = \alpha \mathbf{H}_{TF0} + (1-\alpha)\mathbf{H}_{TF1} \tag{19}$$

where $\mathbf{H}_{New}$ is the combined channel estimation proposed by us, $\mathbf{H}_{TF0}$ and $\mathbf{H}_{TF1}$ are the odd and even group channel estimations, α is the weight of the odd group channel estimation. Because of unification of comb-pilot traditional transform domain channel estimation with frequency domain LS approach, formula (19) becomes:

$$\mathbf{H}_{New} = \mathbf{H} + \underline{\mathbf{e}} + \alpha \mathbf{W}_0 + (1-\alpha)\mathbf{W}_1 \tag{20}$$

where $\mathbf{W}_0$ and $\mathbf{W}_1$ are independent AWGN with zero mean and variance σ^2 respectively. When $\alpha = 1/2$ which means equiprobability, (20) can be expressed as:

$$\mathbf{H}_{New} = \mathbf{H} + \underline{\mathbf{e}} + \frac{1}{2}(\mathbf{W}_0 + \mathbf{W}_1). \tag{21}$$

It can be found that $(\mathbf{W}_0 + \mathbf{W}_1)/2$ is also AWGN, but with zero mean and variance $\sigma^2/2$. Because of unification which has been analyzed above, MSE of our proposed frequency diversity channel estimation can be obtained as:

$$\mathrm{MSE} = \frac{1}{2}\frac{\beta}{\mathrm{SNR}} + \boldsymbol{error\ floor}. \tag{22}$$

This is an exciting result. It shows that when all the pilots are divided into two groups, we can say that the pilots can use two frequency diversity to estimate the channel. If the distance between two adjacent pilots in sub-group is less than the coherence bandwidth of the channel, the diversity gain of MSE will be 3 dB. When four frequency diversity is applied, the diversity gain of MSE will be 6 dB. When eight frequency diversity is used, the diversity gain of MSE will be 9 dB.

Theoretically, regardless of complexity, when the distance between adjacent two pilots in sub-group is less than the coherence bandwidth of channel, L frequency diversity will bring $10log(1/L)$ dB diversity gain of :

$$\mathrm{MSE} = \frac{1}{L}\frac{\beta}{\mathrm{SNR}} + \boldsymbol{error\ floor}. \tag{23}$$

This diversity gain of MSE will improve bit error rate (BER) performance greatly. Furthermore, another merit of (23) is that it employs no prior statistical characteristics of the multipath wireless channel.

4. Simulation Results

Our OFDM system parameters in simulation are listed in Tab. 1. Here, the virtual subcarriers and DC tone are ignored. The multipath channel is modeled as a tapped delay line structure. The delay time and average gain follow ITU-R vehicular A channel model listed in Tab. 2 [13]. The central carrier frequency is set to 2.3 GHz.

Fig. 2 and Fig. 3 present the MSE and BER performance of our proposed method when $L = 2$ and $L = 4$ and the traditional transform domain estimation vs SNR. The asterisk and the triangle line show the simulation results on condition corresponding to Case 1 and Case 2 in Tab. 1. In both cases, the amount of the pilots is not larger than the length of CP. The result of circle line shows the performance of our proposed algorithm corresponding to frequency diversity $L = 2 \quad (N_p = 64, \Delta P = 16, N_{p\_2} = 32, \Delta P_2 = 32)$. The result of square line shows the performance of our proposed algorithm corresponding to frequency diversity $L = 4 \quad (N_p = 128, \Delta P = 8, N_{p\_2} = 32, \Delta P_2 = 32)$. In Fig. 2 and Fig. 3, it can be found that according to the traditional transform domain estimation, when pilot amount is not larger than the length of CP, MSEs without frequency diversity perform the same as LS algorithm $\mathrm{MSE} = 1/\mathrm{SNR}$ in (7). We can see that no improvement is achieved even if pilot amount is increasing. When frequency diversity $L = 2$, it can be found that diversity gain of MSE is 3 dB compared to the traditional method. Diversity gain of MSE is 6 dB corresponding to frequency diversity $L = 4$. These gains improve the performance of BER, when $L = 2$ about 1.5 dB and when $L = 4$ more than 2 dB in Fig. 3.

Fig. 4 and Fig. 5 present the *MSE* and BER performance of our proposed method when $L = 8$ and the traditional transform domain estimation vs SNR. Here, the amount of pilots is larger than the length of CP. The square line corresponding to the result shows the performance of

Bandwidth	10 MHz
FFT/IFFT	1024
Length of CP	128
OFDM symbol duration	115.2 μs
Pilot/Data Modulation	BPSK($\beta = 1$)
N_p	Case 1: 64 Case 2: 128 Case 3: 256
ΔP	Case 1: 1024/64 = 16 Case 2: 1024/128 = 8 Case 3: 1024/256 = 4

Tab. 1. OFDM system parameters.

Tap	Relative Delay (ns)	Average Power (dB)
1	0	0.0
2	310	-1.0
3	710	-9.0
4	1090	-10.0
5	1730	-15.0
6	2510	-20.0

Tab. 2. ITU-R vehicular A channel model.

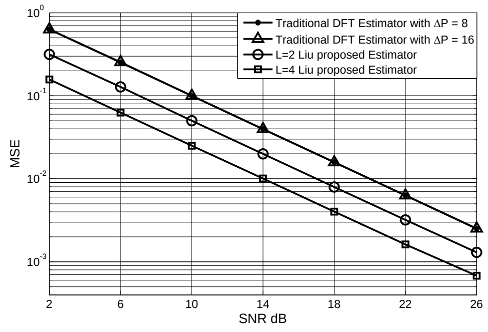

Fig. 2. *MSE* of the proposed channel estimation with different diversity and traditional transform domain estimation vs SNR.

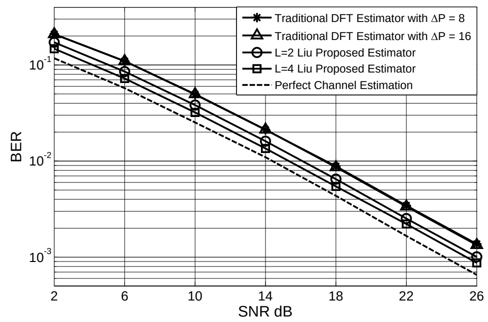

Fig. 3. BER of the proposed channel estimation with different diversity and traditional transform domain estimation vs SNR.

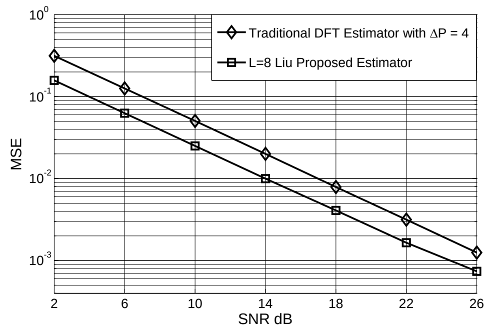

Fig. 4. *MSE* of the proposed channel estimation with diversity $L = 8$ and traditional transform domain estimation vs SNR.

our proposed algorithm corresponding to frequency diversity of $L = 8(N_p = 256, \Delta P = 4, N_{p\_8} = 32, \Delta P_8 = 32)$. The diamond line shows the traditional method. Here, when the amount of total pilots exceeds the length of CP, traditional transform domain estimation can employ noise suppression in time domain as zeroing the elements corresponding to those whose index exceeds CP length. It can be found that our proposed algorithm can also work better than this in noise suppression. However, when $L = 8$, the complexity of our proposed algorithm will be increased. In such case, a trade-off should be considered to obtain satisfying performance while maintaining the system complexity.

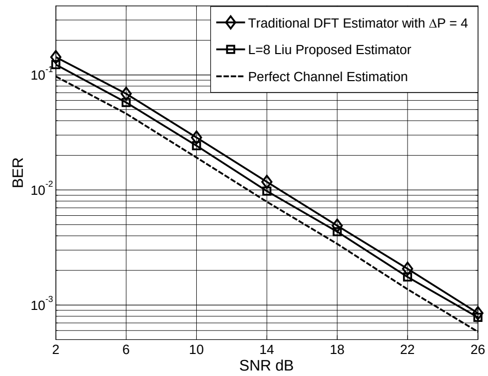

Fig. 5. BER of the proposed channel estimation with diversity $L = 8$ and traditional transform domain estimation vs SNR.

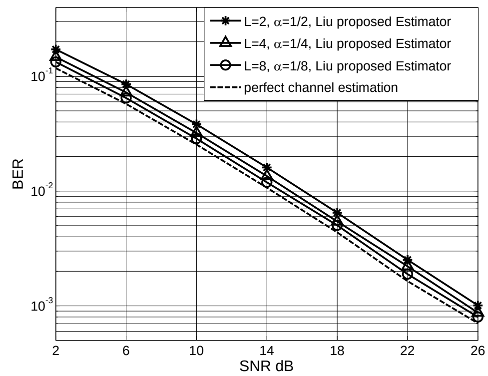

Fig. 6. *MSE* of the proposed channel estimation with diversity $L = 2, L = 4, L = 8$vs SNR.

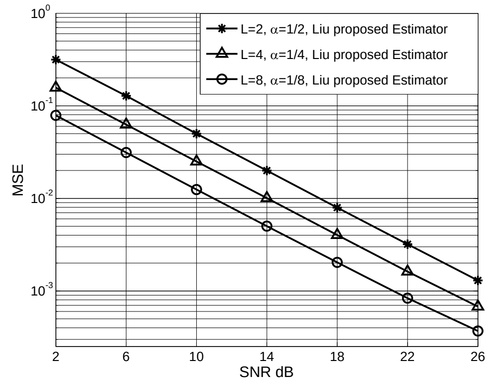

Fig. 7. BER of the proposed channel estimation with diversity $L = 2, L = 4, L = 8$ vs SNR.

Fig. 6 and Fig. 7 give the *MSE* and BER performance with diversity $L = 2, L = 4, L = 8$ vs SNR. We can conclude that our proposed estimator is better with increasing diversity because of better noise suppression. When $L = 8$, the BER gap between the perfect channel estimation is less than 0.8 dB.

Additionally, it should be pointed out that in Fig. 4 it can be found that error floor occurs in MSE and BER performance of our proposed algorithm when SNR exceeds 22 dB. This is caused by not sample spaced channel taps as mentioned above. When $L = 2$ and $L = 4$ this problem also exits, but does not stand out apparently when SNR is not high. This is difficult to handle because of a mismatch of the system parameters and the characteristics of the wireless channel.

5. Conclusion

In this paper, we propose a novel comb-pilot transform domain frequency diversity channel estimation for OFDM systems. If the distance between two adjacent pilots in subgroup is less than the coherence bandwidth of the channel, this algorithm can achieve better noise suppression. Theoretical analysis and simulation results show that this method directly reduces *MSE* compared with the traditional transform domain estimation, and significantly improves BER performance. Furthermore, this method does not need the prior statistical characteristics of the multipath wireless channel. This algorithm increases complexity in some degree but improves the performance. It can be applied on the base station side, on which some pre-equalization can be employed. This will also greatly reduce the complexity of channel estimation on the mobile phone side.

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