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Adaptive channel estimation based on pilot signals and transform-domain processing in SISO/MIMO OFDM systems

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Abstract: Based on the transform-domain characteristics of pilot signals, a band suppression filter is used as a transform-domain filter to restrain the interference of noise in channel estimation. The performance effect on channel estimation for an orthogonal frequency division multiplex (OFDM) system by different energy coefficients in the transform domain and the energy coefficient under the different signal-to-noise ratios (SNR) are also analyzed. A new energy coefficient expression is deduced. It is theoretically proven that dynamically selecting an energy coefficient can significantly improve the performance of channel estimation. Simulation results show that the proposed algorithm can achieve better performance close to the theoretic bounds of perfect channel estimation. The algorithm is adapted to single-input single-output (SISO) OFDM and multi-input multi-output (MIMO) OFDM systems.

Key words: adaptive channel estimation; orthogonal frequency division multiplex (OFDM); transmit diversity

Orthogonal frequency division multiplex (OFDM) and multi-input multi-output (MIMO) were proven effective in resisting multipath fading and improving system capacity. The combination of the two techniques, MIMO OFDM, shows even higher performance[1]. OFDM has attracted much attention and gained many research productions. For single-input single-output (SISO) systems, several channel estimation methods can be found in Refs. [2 – 4], whereas for MIMO systems, some channel estimation algorithms can be found in Refs. [5 – 7]. In allusion to the approach in Ref. [8], adaptive channel estimation methods for SISO/MIMO OFDM systems are presented in this paper. The two algorithms give a band suppression filter to restrain the interference of noise in channel estimation and present new methods to design the filter in transform domain by detecting signal-to-noise ratio (SNR). Simulation results show that the proposed algorithm can achieve better performance close to the relevant theoretic bounds of perfect channel estimation.

1 System Description

1.1 SISO OFDM system model

The SISO OFDM system model[8] with N subcarriers is shown in Fig. 1. The receiving signal at time t,

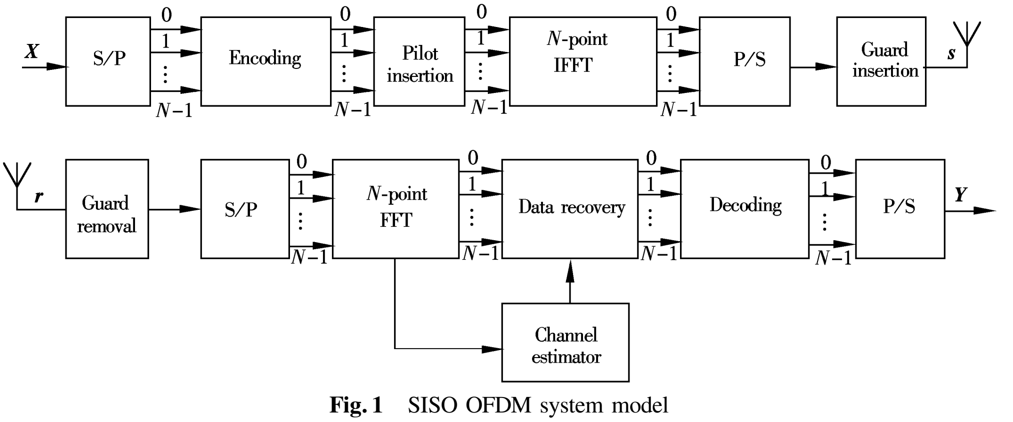

Fig. 1 SISO OFDM system model

$r(t)$, can be expressed as

$$r(t) = h(t) \otimes s(t) + n(t) \tag{1}$$

where $s(t)$ is the transmitted signal, $n(t)$ is the Gaussian white noise with variance σ_N^2, and $\otimes$ denotes convolution.

After removing the cyclic prefix (CP) at the receive antenna and coherent demodulating, the sampled

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output signal $Y(k)$ at tone k can be expressed as

$$Y(k) = X(k)H(k) + N(k) \tag{2}$$

where $X(k)$ is the transmitted data at tone k of the l-th OFDM symbol, $H(k)$ is the channel frequency response corresponding to tone k, and $N(k)$ is the matching filter's output of $n(t)$. At the receiver we can estimate the channel transfer function and correct the information data $X(k)$ according to $Y(K)$.

1.2 MIMO OFDM system model

The MIMO OFDM system model[9] is shown in Fig. 2, with N_t transmit antennas, N_r receive antennas and N subcarriers.

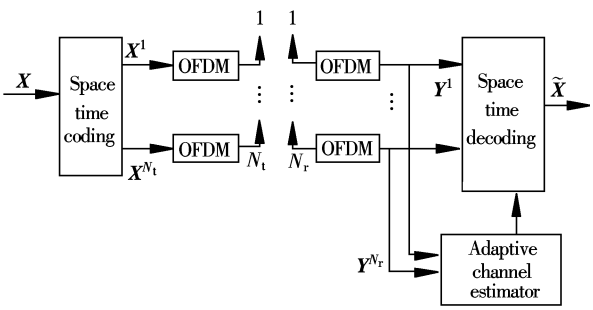

Fig. 2 MIMO OFDM system model

Suppose the OFDM symbol which is transmitted from the r-th antenna at time index n is denoted by the $N \times 1$ vector $\boldsymbol{X}^r(n)$. Before transmission, this vector is processed by an IFFT, and a cyclic prefix of length N_{CP} is added. We assume that $N_{\mathrm{CP}} \geqslant L$, where L is the maximum length of all channels. After removing the cyclic prefix at the q-th receive antenna, we obtain the $N \times 1$ vector $\boldsymbol{y}^q(n)$, which can be expressed as

$$\boldsymbol{y}^q(n) = \sum_{r=1}^{N_t} \boldsymbol{H}_{\mathrm{cir}}^{q,r} \boldsymbol{F}^{\mathrm{H}} \boldsymbol{X}^r(n) + \boldsymbol{\eta}^q(n) \tag{3}$$

where $\boldsymbol{F}$ denotes the $N \times N$ unitary FFT matrix. We define $\boldsymbol{H}_{\mathrm{cir}}^{q,r}$ as

$$\boldsymbol{H}_{\mathrm{cir}}^{q,r} = \boldsymbol{R}_{\mathrm{CP}} \boldsymbol{H}_{\mathrm{ISI}}^{q,r} \boldsymbol{T}_{\mathrm{CP}} \tag{4}$$

where $\boldsymbol{H}_{\mathrm{ISI}}^{q,r}$ is a $(N + N_{\mathrm{CP}}) \times (N + N_{\mathrm{CP}})$ matrix representing the inter-symbol interference (ISI) between OFDM symbols from the r-th transmit antenna to the q-th receive antenna, which can be written as (to simplify notation, we will omit the antenna index r and q)

$$\boldsymbol{H}_{\mathrm{ISI}} = \begin{bmatrix} h(0) & & & & \\ \vdots & \ddots & & & \\ h(L-1) & & \ddots & & \\ & \ddots & & \ddots & \\ & & \ddots & & \ddots \\ & & & h(L-1) \quad \dots & h(0) \end{bmatrix} \tag{5}$$

$\boldsymbol{T}_{\mathrm{CP}}$ denotes the cyclic prefix add matrix,

$$\boldsymbol{T}_{\mathrm{CP}} = [\boldsymbol{O}_{N_{\mathrm{CP}} \times (N - N_{\mathrm{CP}})} \quad \boldsymbol{I}_{N_{\mathrm{CP}} \times N_{\mathrm{CP}}} \quad \boldsymbol{I}_{N \times N}] \tag{6}$$

$\boldsymbol{R}_{\mathrm{CP}}$ denotes the cyclic prefix remove matrix,

$$\boldsymbol{R}_{\mathrm{CP}} = [\boldsymbol{O}_{N \times N_{\mathrm{CP}}} \quad \boldsymbol{I}_N] \tag{7}$$

After the eigenvalue decomposition of $\boldsymbol{H}_{\mathrm{cir}}^{q,r}$, we can obtain

$$\boldsymbol{H}_{\mathrm{cir}}^{q,r} = \boldsymbol{F}^{\mathrm{H}} \mathrm{diag}\{\sqrt{N}[\boldsymbol{h}_{q,r}^{\mathrm{T}} \quad \boldsymbol{0}_{1 \times (N-L)}]^{\mathrm{T}}\} \boldsymbol{F} \tag{8}$$

where $\boldsymbol{h}_{q,r} = \{h_{q,r}(0), h_{q,r}(1), \dots, h_{q,r}(L-1)\}^{\mathrm{T}}$ is the $L \times 1$ vector representing the length L channel impulse response from the r-th transmit antenna to the q-th receive antenna.

Using Eq. (8), and taking the FFT of $\boldsymbol{y}^q(n)$, then Eq. (3) can be written as

$$\begin{aligned} \boldsymbol{Y}^q(n) = & \sum_{r=1}^{N_t} \mathrm{diag}\{\sqrt{N}[\boldsymbol{h}_{q,r}^{\mathrm{T}} \quad \boldsymbol{0}_{1 \times (N-L)}]^{\mathrm{T}}\} \times \\ & \boldsymbol{X}^r(n) + \boldsymbol{N}^q(n) \end{aligned} \tag{9}$$

where $\boldsymbol{N}^q(n)$ is the Fourier transform of $\boldsymbol{\eta}^q(n)$.

We take the two-branch transmit diversity with one receiver[10] OFDM system as an example. Assume that $\boldsymbol{X}_1$ and $\boldsymbol{X}_2$ are adjacent $N \times 1$ vectors. After the space-time-block-code process, we obtain the coded matrix:

$$\boldsymbol{G} = \begin{bmatrix} \boldsymbol{X}_1 & \boldsymbol{X}_2 \\ -\boldsymbol{X}_2^* & \boldsymbol{X}_1^* \end{bmatrix} \tag{10}$$

Each line of the matrix is transmitted at different time index. At the receiver we obtain

$$\left.\begin{aligned} \boldsymbol{Y}_1 &= \boldsymbol{H}_1 \boldsymbol{X}_1 + \boldsymbol{H}_2 \boldsymbol{X}_2 + \boldsymbol{N}_1 \\ \boldsymbol{Y}_2 &= -\boldsymbol{H}_1 \boldsymbol{X}_2^* + \boldsymbol{H}_2 \boldsymbol{X}_1^* + \boldsymbol{N}_2 \end{aligned}\right\} \tag{11}$$

where $\boldsymbol{H}_1$ and $\boldsymbol{H}_2$ represent the channel frequency response from the transmit antennas to the receive antenna, respectively; $\boldsymbol{N}_1$ and $\boldsymbol{N}_2$ denote the noise vectors at different time indices.

Based on the decoding method of the two-branch transmit diversity with one receiver system[11], we can obtain the estimate of $\boldsymbol{X}_1$ and $\boldsymbol{X}_2$:

$$\left.\begin{aligned} \tilde{\boldsymbol{X}}_1 &= \boldsymbol{H}_1^* \boldsymbol{Y}_1 + \boldsymbol{H}_2 \boldsymbol{Y}_2^* \\ \tilde{\boldsymbol{X}}_2 &= \boldsymbol{H}_2^* \boldsymbol{Y}_1 - \boldsymbol{H}_1 \boldsymbol{Y}_2^* \end{aligned}\right\} \tag{12}$$

Using Eq. (11), Eq. (12) can be written as

$$\left.\begin{aligned} \tilde{\boldsymbol{X}}_1 &= (\boldsymbol{H}_1^* \boldsymbol{H}_1 + \boldsymbol{H}_2^* \boldsymbol{H}_2) \boldsymbol{X}_1 + \boldsymbol{H}_1^* \boldsymbol{N}_1 + \boldsymbol{H}_2 \boldsymbol{N}_2^* \\ \tilde{\boldsymbol{X}}_2 &= (\boldsymbol{H}_1^* \boldsymbol{H}_1 + \boldsymbol{H}_2^* \boldsymbol{H}_2) \boldsymbol{X}_2 - \boldsymbol{H}_1 \boldsymbol{N}_2^* + \boldsymbol{H}_2^* \boldsymbol{N}_1 \end{aligned}\right\} \tag{13}$$

Then, we finally obtain the estimate of transmit vectors according to the received signals.

2 Adaptive Channel Estimation Methods

2.1 Adaptive channel estimation method of SISO OFDM system

For simplicity, we start with the adaptive channel estimation of the SISO OFDM system and then extend it to the MIMO OFDM system.

From the received pilot signals we can estimate $H(k)$. Here comb-pattern[11] is adopted. There are in total N_{P} pilot subcarriers in each OFDM signal. Assume N/N_{P} is an integer, and all the pilot signals have an equal complex value c, then the modulated OFDM signal on the k-th subcarrier can be expressed as

$$X(k)=X\left(m\frac{N}{N_{\mathrm{P}}}+i\right)=\begin{cases}c & i=0\\ \text{data} & i=1,2,\ldots,N/N_{\mathrm{P}}-1\end{cases} \tag{14}$$

where $0 \leqslant m \leqslant N_{\mathrm{P}}-1$.

Assume that the channel transfer function at the pilot subcarriers can be expressed as an $N_{\mathrm{P}}\times 1$ vector,

$$\boldsymbol{H}_{\mathrm{P}}=\{H_{\mathrm{P}}(0),\ H_{\mathrm{P}}(1),\ldots,H_{\mathrm{P}}(N_{\mathrm{P}}-1)\}^{\mathrm{T}}=\left\{H(0),H\left(\frac{N}{N_{\mathrm{P}}}\right),\ldots,H_{\mathrm{P}}\left((N_{\mathrm{P}}-1)\frac{N}{N_{\mathrm{P}}}\right)\right\}^{\mathrm{T}} \tag{15}$$

Then the received frequency-domain signal vector $\boldsymbol{Y}_{\mathrm{P}}=\{Y_{\mathrm{P}}(0),Y_{\mathrm{P}}(1),\ldots,\ Y_{\mathrm{P}}(N_{\mathrm{P}}-1)\}^{\mathrm{T}}$ can be obtained as

$$\boldsymbol{Y}_{\mathrm{P}}=\boldsymbol{X}_{\mathrm{P}}\boldsymbol{H}_{\mathrm{P}}+\boldsymbol{N}_{\mathrm{P}} \tag{16}$$

where

$$\boldsymbol{X}_{\mathrm{P}}=\mathrm{diag}\{X_{\mathrm{P}}(0),X_{\mathrm{P}}(1),\ldots,X_{\mathrm{P}}(N_{\mathrm{P}}-1)\}=\mathrm{diag}\left\{X(0),X\left(\frac{N}{N_{\mathrm{P}}}\right),\ldots,X\left((N_{\mathrm{P}}-1)\frac{N}{N_{\mathrm{P}}}\right)\right\}$$

$\boldsymbol{X}_{\mathrm{P}}$ denotes the signals at the pilot subcarriers and $\boldsymbol{N}_{\mathrm{P}}$ denotes the Gaussian white noise vector at the pilot subcarriers.

According to Ref. [8], we study the transform-domain processing of channel frequency response. Fig. 3 gives an example of $|G_{N_{\mathrm{P}}}(p)|$.

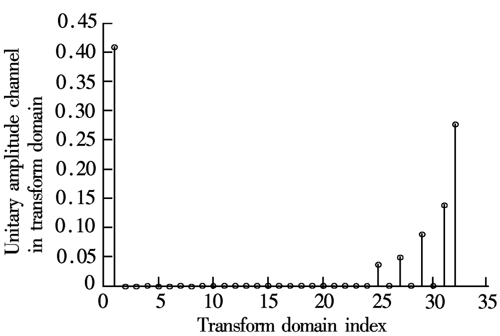

Fig. 3 An example of $|G_{N_{\mathrm{P}}}(p)|$

A new efficient band suppression filtering is adopted here instead of the traditional lowpass filtering. According to Fig. 3, the cut-off frequency p_{c1} can be selected as 0, and p_{c2} can be selected with respect to different values of the energy parameter R'.

$$\left[|G(0)|^2+\sum_{p=N_{\mathrm{P}}-p_{\mathrm{c2}}}^{N_{\mathrm{P}}-1}|G(p)|^2\right]\Big/\sum_{p=0}^{N_{\mathrm{P}}-1}|G(p)|^2=R' \tag{17}$$

where the numerator is the energy in passband, the denominator the total energy.

After the filtering, the noise component is reduced to $p_{\mathrm{c2}}/N_{\mathrm{P}}$ of its original value. The band suppression filtering can be realized as

$$\widetilde{G}_N(p)=\begin{cases}G_{N_{\mathrm{P}}}(p) & p=0;N-p_{\mathrm{c2}}\leqslant p\leqslant N-1\\ 0 & \text{otherwise}\end{cases} \tag{18}$$

Then Eq. (17) can be written as

$$R'=\frac{1+N_{\mathrm{CP}}\sigma_{\mathrm{N}}^2/N_{\mathrm{P}}}{1+\sigma_{\mathrm{N}}^2} \tag{19}$$

where N_{CP}, N_{P}, σ_{N}^2 are the length of CP, the number of pilot subcarriers and the variance of Gaussian white noise, respectively.

With respect to any different SNR of the receiver, Eq. (19) can be written in another form:

$$R'=1-\frac{(N_{\mathrm{P}}-N_{\mathrm{CP}})/N_{\mathrm{P}}}{1+r_{\mathrm{SNR}}} \tag{20}$$

where $r_{\mathrm{SNR}}=1/\sigma_{\mathrm{N}}^2$.

It is obvious that the proposed method is adaptive and the cut-off frequency p_{c2} can be dynamically designed according to the SNR of receiver.

2.2 Adaptive channel estimation method of MIMO OFDM system

We now consider the adaptive channel estimation of the MIMO OFDM system and the optimal pilot design. Based on the study of the MIMO OFDM system, we take the two-branch transmit diversity with two-receiver OFDM system as an example. The coded matrix is

$$\boldsymbol{G}=\begin{bmatrix}\boldsymbol{X}_1 & \boldsymbol{X}_2\\ -\boldsymbol{X}_2^* & \boldsymbol{X}_1^*\end{bmatrix} \tag{21}$$

$Y_{j,m}$ denotes the signal of the j-th receive antenna at time m, and $H_{i,j}$ is the channel frequency response of the k-th subcarrier from the i-th transmit antenna to the j-th receive antenna (for simplicity, we omit the frequency index k). At the receiver we can obtain

$$\begin{Bmatrix}Y_{1,1}\\ Y_{1,2}\end{Bmatrix}=\begin{bmatrix}X_1 & X_2\\ -X_2^* & X_1^*\end{bmatrix}\begin{Bmatrix}H_{1,1}\\ H_{2,1}\end{Bmatrix}+\begin{Bmatrix}N_1\\ N_2\end{Bmatrix} \tag{22}$$

$$\begin{Bmatrix}Y_{2,1}\\ Y_{2,2}\end{Bmatrix}=\begin{bmatrix}X_1 & X_2\\ -X_2^* & X_1^*\end{bmatrix}\begin{Bmatrix}H_{1,2}\\ H_{2,2}\end{Bmatrix}+\begin{Bmatrix}N_3\\ N_4\end{Bmatrix} \tag{23}$$

We take receive antenna 1 as an example. Using Eq. (22), we can obtain the least-square (LS) estimation of $\{H_{1,1},\ H_{2,1}\}^{\mathrm{T}}$:

$$\begin{Bmatrix}\widetilde{H}_{1,1}\\ \widetilde{H}_{2,1}\end{Bmatrix}=\begin{bmatrix}X_1 & X_2\\ -X_2^* & X_1^*\end{bmatrix}^{-1}\begin{Bmatrix}Y_{1,1}\\ Y_{1,2}\end{Bmatrix}=\begin{Bmatrix}H_{1,1}\\ H_{2,1}\end{Bmatrix}+\begin{Bmatrix}X_1 & X_2\\ -X_2^* & X_1^*\end{Bmatrix}^{-1}\begin{Bmatrix}N_1\\ N_2\end{Bmatrix} \tag{24}$$

To simplify Eq. (24), the optimal pilot sequences can be designed by choosing disjoint sequences, i. e., zeros in one pilot sequence, with equal complex value X_{C} in another one, and vice versa. Then Eq. (24) can be written as (assume $X_1=X_{\mathrm{C}}, X_2=0$)

$$\begin{Bmatrix}\widetilde{H}_{1,1}\\ \widetilde{H}_{2,1}\end{Bmatrix}=\begin{bmatrix}X_C & 0\\ 0 & X_C^*\end{bmatrix}^{-1}\begin{Bmatrix}Y_{1,1}\\ Y_{1,2}\end{Bmatrix}=\begin{Bmatrix}H_{1,1}\\ H_{2,1}\end{Bmatrix}+\begin{bmatrix}\dfrac{1}{X_C} & 0\\ 0 & \dfrac{1}{X_C^*}\end{bmatrix}\begin{Bmatrix}N_1\\ N_2\end{Bmatrix} \tag{25}$$

Based on the above consideration, pilot sequences can be designed as

$$\left.\begin{aligned} \boldsymbol{X}_1 &= \{X_C, \text{data}, \ X_C, \ \text{data}, \ldots, X_C, \text{data}\} \\ \boldsymbol{X}_2 &= \{0, \text{data}, 0, \text{data}, \ldots, \ 0, \text{data}\} \end{aligned}\right\} \quad (26)$$

where $\boldsymbol{X}_1, \boldsymbol{X}_2$ are $N \times 1$ vectors (N_P pilot subcarriers). The variance of pilot signals does not affect the channel estimation complexity.

Using Eq. (25), we can obtain the rough estimate of $\boldsymbol{H}_{1,1}$, $\boldsymbol{H}_{2,1}$, thus, the LS estimate from the transmit antennas to the receive antennas can be expressed as two $P \times 1$ vectors $\hat{\boldsymbol{H}}_{1,1}$, $\hat{\boldsymbol{H}}_{2,1}$. The proposed method based on pilot signals and adaptive transform-domain processing is depicted in Fig. 4. We take the LS estimate $\hat{\boldsymbol{H}}_{q,r}$ as an example (from the q-th transmit antenna to the r-th receive antenna), and give three key points as follows:

1) The transform-domain representation of the $N_P \times 1$ vector $\hat{\boldsymbol{H}}_{q,r}$ is

$$\tilde{\boldsymbol{G}}_{q,r} = \boldsymbol{F}_{N_P} \hat{\boldsymbol{H}}_{q,r} \quad (27)$$

where $\boldsymbol{F}_{N_P}$ denotes the $N_P \times N_P$ unitary FFT matrix.

2) After transform-domain filtering, we can obtain $\hat{\boldsymbol{G}}_{q,r}$. The filter design is realized in a similar way of SISO OFDM system. The cut-off frequency $p_2^{q,r}$ is adjusted by energy parameter $R'_{q,r}$ with respect to the following formula:

$$R'_{q,r} = \frac{1 + \dfrac{N_{CP}}{N_P \rho}\sigma_N^2}{1 + \dfrac{1}{\rho}\sigma_N^2} = \frac{\rho + \dfrac{N_{CP}}{N_P}\sigma_N^2}{\rho + \sigma_N^2} \quad (28)$$

where ρ denotes the fixed power dedicated for pilot sequence, N_{CP} is the length of CP, N_P is the number of pilot subcarriers, and σ_N^2 is the variance of Gaussian white noise. According to the SNR of the receiver, Eq. (28) can also be written in another form:

$$R'_{q,r} = 1 - \frac{(N_P - N_{CP})/N_P}{1 + r_{SNR}} \quad (29)$$

where $r_{SNR} = 1/\sigma_N^2$.

3) After zero padding, we take the N-point IFFT of $\hat{\boldsymbol{G}}_{q,r}$ and finally obtain $\tilde{\boldsymbol{H}}_{q,r}$.

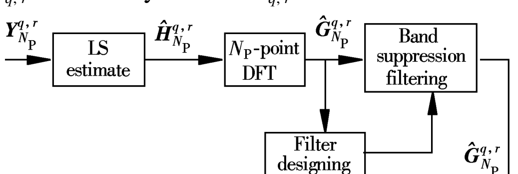

Fig. 4 Adaptive transform-domain processing of $\hat{\boldsymbol{H}}_{q,r}$

3 Simulation

In our simulation, the MIMO OFDM system is equipped with two transmit antennas and two receive antennas. Space-time-block-code with BPSK modulation is used in the system. The entire channel bandwidth is 20 MHz. The total number of all subcarriers in each OFDM symbol is 256, and the number of pilot subcarriers is 32. The length of cyclic prefix is 0.5 μs, and we assume the channel delay is 0.3 μs. The average power distribution of a time-variant multipath channel model[12] is given as follows:

$$\boldsymbol{h} = \{1, 0.7071, 0.3253, 0.1273, 0, 0, 0.0495\} \quad (30)$$

Fig. 5 and Fig. 6 show that the adaptive channel estimation methods have offered a better performance gain in the SISO/MIMO OFDM systems than the traditional transform-domain processing methods.

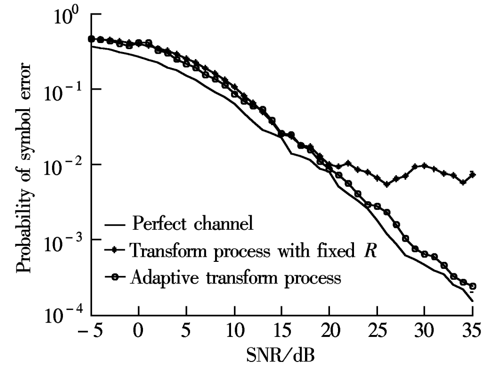

Fig. 5 Comparison of BER in SISO OFDM system

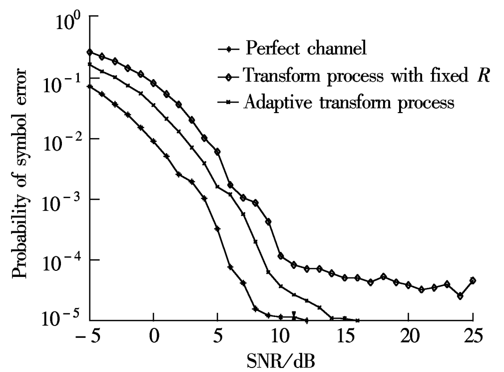

Fig. 6 Comparison of BER in 2I2O OFDM system

4 Conclusion

In this paper, we have presented adaptive channel estimation methods for SISO/MIMO OFDM systems. By adjusting the energy parameter according to the SNR of the receiver in order to improve the filtering effect, the proposed channel estimation methods based on optimum pilot-aided and transform-domain filtering have been seen to significantly outperform the traditional transform-domain processing in SISO/MIMO OFDM systems. Simulation results have demonstrated that the adaptive channel estimation methods can achieve better performance close to the theoretic bound of ideal channel estimation.

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基于导频和变换域的 SISO/MIMO OFDM 系统自适应信道估计

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摘要:根据导频信号在变换域中的频响特性,将系统的变换域滤波器设计成具有更好滤波效果的“带阻滤波器”;在小信噪比和大信噪比条件下,分析了变换域能量参数对系统信道估计性能的影响,推导了新的能量参数表达式,理论证明了根据接收端检测到的信噪比自适应选择能量参数可以显著改善信道估计的性能. 仿真结果表明,提出的信道估计算法具有较好的性能,在一定程度上可以接近于理想信道估计的性能,适用于 SISO OFDM 系统和 MIMO OFDM 系统.

关键词:自适应信道估计;正交频分复用(OFDM);发射分集

中图分类号:TN914. 3