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Iterative receivers for SFBC-OFDM systems with adaptive training scheme

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Abstract: This paper considers the design of iterative receivers for space-frequency block-coded orthogonal frequency division multiplexing (SFBC-OFDM) systems in unknown wireless dispersive fading channels. An iterative joint channel estimation and symbol detection algorithm is derived. In the algorithm, the channel estimator performs alternately in two modes. During the training mode, the channel state information (CSI) is obtained by a discrete-Fourier-transform-based channel estimator and the noise variance and covariance matrix of the channel response is estimated by the proposed method. In the data transmission mode, the CSI and transmitted data is obtained iteratively. In order to suppress the error propagation caused by a random error in identifying symbols, a simple error propagation detection criterion is proposed and an adaptive training scheme is applied to suppress the error propagation. Both theoretical analysis and simulation results show that this algorithm gives better bit-error-rate performance and saves the overhead of OFDM systems.

Key words: space-frequency block-coded orthogonal frequency division multiplexing (SFBC-OFDM); iterative receiver; channel estimation; adaptive training scheme

The combined application of multi-input and multi-output (MIMO) antenna technology and orthogonal frequency division multiplexing (OFDM) modulation yields improvements to communication capacity and quality to meet the requirements of the next generation of wireless systems. Recently, several studies addressing the design and application of MIMO-OFDM have been conducted[1-3]. In Ref. [4], it is shown that applying the simple orthogonal space frequency coding technique first proposed in Ref. [5] for OFDM systems can have better performance in fast fading environments and other implementation advantages.

In this paper, we focus on the design of iterative receivers for space-frequency block-coded orthogonal frequency division multiplexing (SFBC-OFDM) systems in unknown wireless dispersive fading channels with outer channel coding. The iterative joint channel estimation and symbol detection algorithm is derived. The initial channel estimation is achieved by using the training sequence[6], which not only simplifies the initial channel estimation, but also attains the best estimation performance. By this training sequence we also present a method to estimate the noise variance and covariance matrix of the channel response, which are needed by the subsequent iterative algorithm. In order to suppress the error propagation caused by a random error in identifying symbols, an adaptive training scheme is applied in the system. For this improved scheme, when a decision-error is estimated by a simple criterion, one more training sequence is required to be sent by the transmitter to update the initial channel estimation. The simulation results show that the improved algorithm can prevent error propagation effectively and consequently improve the bit-error-rate (BER) performance and keep a low system overhead.

In this paper, $(\cdot)^{\mathrm{T}}$, $(\cdot)^{\mathrm{H}}$ and $(\cdot)^{*}$ denote the transposition, Hermitian transposition and conjugate, respectively; if $\boldsymbol{x}$ is a given $M\times 1$ vector, $\mathrm{diag}(\boldsymbol{x})$ is the $M\times M$ diagonal matrix with the entries of $\boldsymbol{x}$ on the main diagonal; and $\boldsymbol{I}_M$ is the $M\times M$ identity matrix.

1 System Model

Consider an SFBC-OFDM system with two transmission antennas and one reception antenna. As shown in Fig. 1, the information bits are encoded by an outer-channel-code encoder and then interleaved. The interleaved code bits are modulated by an M-phase shift keying (MPSK) modulator. Finally, the modulated MPSK symbols are encoded by an SFBC encoder and transmitted from two transmission antennas. For simplification, we assume that a rate 1/2 convolutional code is applied as the error correcting code. To minimize the latency of the system, coding is assumed to be performed over a single OFDM symbol. The interleaver is assumed to be a random interleaver.

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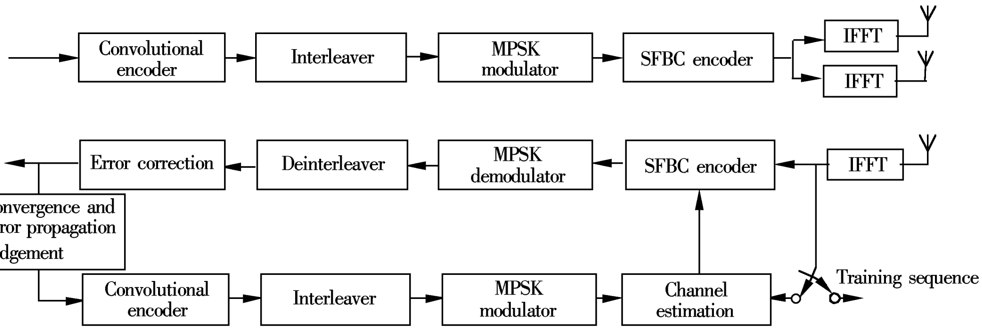

Fig. 1 Transmitter and receiver structure with two transmission antennas and one reception antenna

We denote N as the total number of subcarriers, or the FFT size. $\boldsymbol{X}_i$ is the SFBC symbol vector transmitted from the i-th antenna. The time domain channel impulse response between the i-th transmission antenna and the reception antenna can be modeled as an L-order tapped-delay-line given by $\boldsymbol{h}_i$, $\boldsymbol{h}_i = \{h_i(0), \ldots, h_i(l), \ldots, h_i(L-1)\}^{\mathrm{T}}$ which is a complex Gaussian vector, representing different path gains. The elements of $\boldsymbol{h}_i$ are uncorrelated with zero-mean and variation $\boldsymbol{\sigma}_i^2 = \{\sigma_0^2, \ldots, \sigma_l^2, \ldots, \sigma_{L-1}^2\}^{\mathrm{T}}$. Without loss of generality, we assume the total average power of the channel impulse response to be unity, i. e. $\sum_{l=0}^{L-1} \sigma_l^2 = 1$, so the correlation matrix of $\boldsymbol{h}_i$ is a diagonal matrix with the elements of $\boldsymbol{\sigma}_i^2$ on the main diagonal. $\boldsymbol{h}_i$ are assumed to be independent and identically distributed among different antenna couples.

At the receiver side, with the assumptions that the guard interval duration is longer than the channel maximum excess delay and that the channel is quasi-stationary (i. e. the channel does not change within one OFDM symbol duration), then the demodulated signal at the receiving antenna can be represented by

$$\boldsymbol{Y} = \sum_{i=1}^{2} \operatorname{diag}(\boldsymbol{X}_i)\boldsymbol{H}_i + \boldsymbol{z} \tag{1}$$

where $\boldsymbol{z}$ is a white complex Gaussian noise vector with covariance matrix $\sigma_z^2 \boldsymbol{I}_N$, $\boldsymbol{H}_i$ is the channel frequency response between the i-th transmission antenna and the reception antenna, which can be written in matrix notation as follows:

$$\boldsymbol{H}_i = \boldsymbol{W}_N \boldsymbol{h}_i \tag{2}$$

where $\boldsymbol{W}_N$ is the $N \times L$ Fourier transform matrix.

Assume the channel frequency responses between adjacent subcarriers are nearly constant, i. e. $H_i(n) \approx H_i(n+1)$, $n = 0, 2, \ldots, N-2$. Then the channel frequency response vector $\boldsymbol{H}_i$ can be divided into the even and odd component vectors. The corresponding elements of two vectors are equal. Each component vector can be achieved by performing an $N/2$-point discrete Fourier transform (DFT) on vector $\boldsymbol{h}_i$, i. e.,

$$\boldsymbol{G}_i = \boldsymbol{W}_{\frac{N}{2}} \boldsymbol{h}_i \tag{3}$$

where $\boldsymbol{W}_{\frac{N}{2}}$ is the $\frac{N}{2} \times L$ Fourier transform matrix made of the odd rows of $\boldsymbol{W}_N$.

According to the operation of the space-frequency encoder described in Ref. [4], the symbol vector $\boldsymbol{X}$ is divided into the even component vector $\boldsymbol{X}_{\mathrm{e}}$ and the odd component vector $\boldsymbol{X}_{\mathrm{o}}$. Similarly, $\boldsymbol{X}_{i,\mathrm{e}}$, $\boldsymbol{X}_{i,\mathrm{o}}$ denote the even and odd component vectors of $\boldsymbol{X}_i$ transmitted from the i-th transmission antenna. Then the output of the space-frequency encoder can be expressed in terms of the even and odd component vectors as

$$\boldsymbol{X}_{1,\mathrm{e}} = \boldsymbol{X}_{\mathrm{e}},\ \boldsymbol{X}_{1,\mathrm{o}} = -\boldsymbol{X}_{\mathrm{o}}^*,\ \boldsymbol{X}_{2,\mathrm{e}} = \boldsymbol{X}_{\mathrm{o}},\ \boldsymbol{X}_{2,\mathrm{o}} = \boldsymbol{X}_{\mathrm{e}}^* \tag{4}$$

The demodulated signal at the receiving antenna and the noise vector $\boldsymbol{Z}$ can also be rewritten in terms of the even and odd component vectors as $\boldsymbol{Y} = [\boldsymbol{Y}_{\mathrm{e}}^{\mathrm{T}} \quad \boldsymbol{Y}_{\mathrm{o}}^{\mathrm{T}}]^{\mathrm{T}}$ and $\boldsymbol{Z} = [\boldsymbol{Z}_{\mathrm{e}}^{\mathrm{T}} \quad \boldsymbol{Z}_{\mathrm{o}}^{\mathrm{T}}]^{\mathrm{T}}$. Considering the assumption about channel frequency response between adjacent subcarriers and the SFBC coding constraints of Eq. (4), Eq. (1) can be further expressed as

$$\boldsymbol{Y} = \boldsymbol{S}\boldsymbol{W}\boldsymbol{h} + \boldsymbol{z} \tag{5}$$

with

$$\boldsymbol{S} = \begin{bmatrix} \operatorname{diag}(\boldsymbol{X}_{\mathrm{e}}) & \operatorname{diag}(\boldsymbol{X}_{\mathrm{o}}) \\ \operatorname{diag}(-\boldsymbol{X}_{\mathrm{o}}^*) & \operatorname{diag}(\boldsymbol{X}_{\mathrm{e}}^*) \end{bmatrix}$$
$$\boldsymbol{W} = \begin{bmatrix} \boldsymbol{W}_{\frac{N}{2}} & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{W}_{\frac{N}{2}} \end{bmatrix},\quad \boldsymbol{h} = \begin{bmatrix} \boldsymbol{h}_1 \\ \boldsymbol{h}_2 \end{bmatrix} \tag{6}$$

2 Iterative Joint Decoding and Channel Estimation

In this section, we consider the iterative receiver design for SFBC-OFDM systems in fast fading channels as depicted in Fig. 1. In our algorithm, the channel estimator performs alternately in two modes. During the training period, the CSI is obtained by a DFT-based channel estimator assisted by the optimum training sequence[6]. When the systems are in the data transmission mode, the CSI and transmitted data is obtained iteratively. It consists of an ML SFBC decoder and a hard quantization ML outer-channel-code decoder.

2.1 Iterative SFBC decoder

Eq. (5) represents the overall input/output relation of the system as a function of the channel impulse response. According to the maximum likelihood criterion, the joint optimal solution of $\boldsymbol{X}$ and $\boldsymbol{h}$ can be obtained by

$$(\hat{\boldsymbol{X}}, \hat{\boldsymbol{h}}) = \arg\max_{\boldsymbol{X},\boldsymbol{h}} \{p(\boldsymbol{Y} \mid \boldsymbol{X}, \boldsymbol{h})\} \tag{7}$$

The solution of (7) can be achieved by an iteration algorithm. This algorithm ensures attainment of at least a local maximum. Theoretically, it maintains non-decreasing likelihood at each iteration step[7].

Given the initial CSI $\hat{\boldsymbol{h}}^{[0]}$, the symbol vector $\hat{\boldsymbol{X}}$ can be obtained conveniently according to Alamouti's combining scheme[5]. The estimate of $\boldsymbol{X}$ during the $(i+1)$-th iteration can be expressed as

$$\begin{bmatrix} \hat{\boldsymbol{X}}_{\mathrm{e}}^{[i+1]} \\ \hat{\boldsymbol{X}}_{\mathrm{o}}^{[i+1]} \end{bmatrix} = \begin{bmatrix} \mathrm{diag}((\boldsymbol{W}_{\frac{N}{2}} \hat{\boldsymbol{h}}_1^{[i]})^*) & \mathrm{diag}(\boldsymbol{W}_{\frac{N}{2}} \hat{\boldsymbol{h}}_2^{[i]}) \\ \mathrm{diag}((\boldsymbol{W}_{\frac{N}{2}} \hat{\boldsymbol{h}}_2^{[i]})^*) & \mathrm{diag}(-\boldsymbol{W}_{\frac{N}{2}} \hat{\boldsymbol{h}}_1^{[i]}) \end{bmatrix} \begin{bmatrix} \boldsymbol{Y}_{\mathrm{e}} \\ \boldsymbol{Y}_{\mathrm{o}}^* \end{bmatrix} \tag{8}$$

After that, the estimated symbol vector $\hat{\boldsymbol{X}}^{[i+1]}$ is deinterleaved and decoded by a hard quantization ML outer-channel-code decoder. Instead of being regarded as the detection results, the detected data is fed back to reconstruct the estimate of $\boldsymbol{S}$ in Eq. (6) and then is used to estimate the channel attenuation again. According to the Bayesian philosophy[7], the conditional mean $\hat{\boldsymbol{h}}^{[i]}$ obtained during the i-th iteration is

$$\hat{\boldsymbol{h}}^{[i]} = \frac{1}{\sigma_z^2} \boldsymbol{\Sigma} \boldsymbol{W}^{\mathrm{H}} (\hat{\boldsymbol{S}}^{[i]})^{\mathrm{H}} \boldsymbol{Y} \tag{9}$$

with

$$\boldsymbol{\Sigma} = \left(\boldsymbol{R}_{\boldsymbol{h}}^{-1} + \frac{1}{\sigma_z^2} \boldsymbol{W}^{\mathrm{H}} (\hat{\boldsymbol{S}}^{[i]})^{\mathrm{H}} (\hat{\boldsymbol{S}}^{[i]}) \boldsymbol{W} \right)^{-1} \tag{10}$$

where $\boldsymbol{R}_{\boldsymbol{h}}$ and σ_z^2 denote the covariance matrix of $\boldsymbol{h}$ and noise variance, respectively, which can be measured with the aid of training sequence.

Due to the orthogonality property of SFBC as well as the constant amplitude modulation, $\boldsymbol{W}^{\mathrm{H}} (\hat{\boldsymbol{S}}^{[i]})^{\mathrm{H}} (\hat{\boldsymbol{S}}^{[i]}) \boldsymbol{W}$ results to $N\boldsymbol{I}_{2L}$ regardless of the transmitted symbol. (Noting that the average energy of the symbols transmitted from each antenna is normalized to 1/2, so that the total energy of each transmitted symbol is to be one). Therefore no matrix inversion is involved in the calculation of $\hat{\boldsymbol{h}}^{[i]}$ and the computation is also numerically more stable. With the updated channel estimation information, we can get a new estimation of $\boldsymbol{X}$. The iteration will be repeated until a stopping criterion is reached.

2.2 DFT-based initialization of CSI

The initial CSI $\hat{\boldsymbol{h}}^{[0]}$ is obtained by the DFT-based channel estimator with the aid of the optimum training sequence[6]. If we denote $\boldsymbol{P}_1$ as the $N \times 1$ training sequence for the first transmission antenna in the frequency domain, the training sequence for the second transmission antenna $\boldsymbol{P}_2$ is formulated as the following rule:

$$\boldsymbol{P}_2 = \boldsymbol{F}\boldsymbol{P}_1 \tag{11}$$

where $\boldsymbol{F}$ denotes an $N \times N$ diagonal matrix with each diagonal entry equal to $\exp\left(-\frac{\mathrm{j}2\pi n\lambda}{N}\right)$ for the frequency domain index $0 \leqslant n < N$ and λ is equal to $N/2$. Without loss of generality, we assume that the number of subcarriers N is even and $\lambda > L$.

During the training period Eq. (1) can be expressed as

$$\boldsymbol{Y} = \sum_{i=1}^{2} \mathrm{diag}(\boldsymbol{P}_i) \boldsymbol{H}_i + \boldsymbol{Z} \tag{12}$$

Then the sum of the two antennas' frequency response can be estimated by

$$\begin{aligned} \hat{H}_{\mathrm{total}}(n) &= \frac{Y(n)}{P_1(n)} = H_1(n) + \\ & H_2(n) \exp\left(-\frac{\mathrm{j}2\pi n\lambda}{N}\right) + \frac{Z(n)}{P_1(n)} \qquad 0 \leqslant n < N \end{aligned} \tag{13}$$

Since the training sequence is designed specifically for each antenna, the individual channel impulse response is easily distinguished in the time domain. Then the frequency domain estimated in Eq. (13) will be transformed into the time domain by N-point IDFT,

$$\hat{h}_{\mathrm{total}}(k) = h_1(k) + h_2(k-\lambda) + \tilde{Z}(k) \qquad 0 \leqslant k < N \tag{14}$$

If the duration of channel impulse response does not exceed L and $\lambda > L$, the initial CSI $\hat{\boldsymbol{h}}^{[0]}$ can be expressed by

$$\hat{\boldsymbol{h}}^{[0]} = \begin{bmatrix} \{\hat{h}_{\mathrm{total}}(0), \ldots, \hat{h}_{\mathrm{total}}(l), \ldots, \hat{h}_{\mathrm{total}}(L-1)\}^{\mathrm{T}} \\ \{\hat{h}_{\mathrm{total}}(\lambda), \ldots, \hat{h}_{\mathrm{total}}(\lambda+l), \ldots, \hat{h}_{\mathrm{total}}(\lambda+L-1)\}^{\mathrm{T}} \end{bmatrix} \tag{15}$$

Then the noise variance and $\boldsymbol{R}_{\boldsymbol{h}}$ can be obtained by

$$\hat{\sigma}_z^2 = \left(\sum_{k=L}^{\lambda-1} |\hat{h}_{\mathrm{total}}(k)|^2 + \sum_{k=\lambda+L}^{N-1} |\hat{h}_{\mathrm{total}}(k)|^2 \right) \frac{N}{N-2L} \tag{16}$$

$$\hat{\boldsymbol{R}}_h = \hat{\boldsymbol{h}}^{[0]} (\hat{\boldsymbol{h}}^{[0]})^{\mathrm{H}} \tag{17}$$

2.3 Adaptive training scheme

The iterative receiver uses the current decisions to update the CSI and then it is applied in symbol detection for the next OFDM symbol. Although the error correcting codes and iterative algorithm can alleviate error propagation caused by erroneous decisions, the more effective methods should be considered.

In Ref. [8], the suppression of error propagation is achieved by inserting the training sequence periodically (with one training block sent per 10 OFDM blocks), which leads to more system overhead. In our improved algorithm, except for the first training sequence, the other training sequences are transmitted according to the request of the decoder. The maximum iteration number is $i_{\max}$. Once the algorithm does not converge during two successive OFDM symbols after $i_{\max}$ iterations, we consider that error propagation occurs. Then one more training sequence is required to be sent by the transmitter to update the initial channel estimation. The simulation results show that this simple improvement can prevent error propagation effectively, while keeping low system overhead.

2.4 Algorithm summary

The proposed algorithm is summarized as follows:

Step 1 Set $i=0$, transmit the optimum training sequence and then perform the DFT-based channel estimation. The initial CSI $\hat{\boldsymbol{h}}^{[0]}$, noise variance and $\boldsymbol{R}_h$ are obtained by Eqs. (15) to (17).

Step 2 During the data transmission mode, given $\hat{\boldsymbol{h}}^{[i]}$, obtain the estimation of $\hat{\boldsymbol{X}}^{[i+1]}$ according to Eq. (8). After deinterleaving and error correcting, the detected data is re-encoded to reconstruct $\hat{\boldsymbol{S}}^{[i+1]}$.

Step 3 Feed $\hat{\boldsymbol{S}}^{[i+1]}$ to the channel estimator to update the CSI according to Eq. (9) and set $i=i+1$.

Step 4 If $i<i_{\max}$ and the iteration does not converge, i. e. $\hat{\boldsymbol{X}}^{[i]} \neq \hat{\boldsymbol{X}}^{[i-1]}$, go to step 2. If $i \leqslant i_{\max}$ and the iteration converges, i. e. $\hat{\boldsymbol{X}}^{[i]} = \hat{\boldsymbol{X}}^{[i-1]}$, stop the iteration, set $i=0$, $\hat{\boldsymbol{h}}^{[0]} = \hat{\boldsymbol{h}}^{[i]}$, go to step 2 and begin a new iteration process for the next OFDM symbol. If $i=i_{\max}$ and iteration does not converge, stop the iteration and check the decoding state of the previous OFDM symbol. If the algorithm does not converge during the previous OFDM symbol after the $i_{\max}$ iteration, the error propagation is predicted so go to step 1; otherwise set $i=0$, $\hat{\boldsymbol{h}}^{[0]} = \hat{\boldsymbol{h}}^{[i]}$, go to step 2 and begin a new iteration process for the next OFDM symbol.

3 Simulation Results

The proposed iterative algorithm is applied in an SFBC-OFDM system with 2×1 antenna configuration. The carrier frequency band is at 2.4 GHz and the bandwidth is 800 kHz. The total number of the OFDM subcarrier is $N=128$; the guard interval is $L=16$. QPSK modulation is applied. A rate 1/2 convolutional code is applied as the error correcting code and the space-frequency block code proposed in Ref. [5] is used in our simulation. A four-tap Rayleigh fading channel has been investigated and vehicle speed is set to be 100 km/h. We assume channel parameters corresponding to different transmission and reception antenna pairs are independent but with the same statistical properties.

In each independent simulation, 1 000 OFDM symbols are transmitted. The performance averaged over independent simulations is evaluated. To illustrate the performance improvement of the proposed algorithm, two training schemes have been investigated: the scheme with the fixed training interval and the improved adaptive training scheme.

In the first experiment, we consider the SFBC-OFDM system with the fixed training interval, the training interval M_t is set to be 10, i. e. one training block is sent every 10 OFDM blocks. Fig. 2 shows the average bit-error-rate (BER) of the proposed iterative algorithm. It is shown that our proposed iterative receiver with $i_{\max}=2$ has SNR gains of about 2 dB performance improvement with respect to the case with no iterations and the increase of $i_{\max}$ shows some moderate improvement. So the maximum iteration time is set to be 2 in the following experiments, which makes the algorithm have low complexity and short latency.

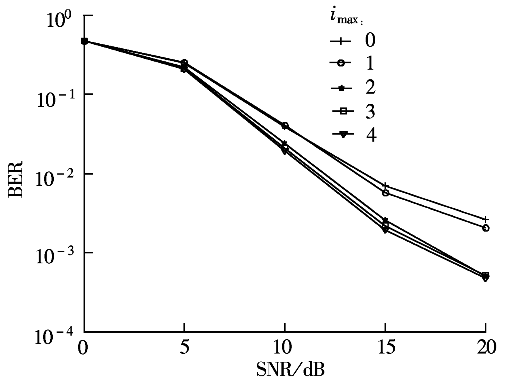

Fig. 2 BER performance with training interval $M_t=10$

Fig. 3 provides the comparisons of BER performance and the overhead ratio of the training sequence between the scheme with fixed training interval and the adaptive training scheme with error propagation detection, which is denoted by DEP $i_{\max}=2$. The overhead ratio of the training sequence is calculated as the ratio of the training block number to the number of OFDM blocks transmitted. The simulation results demonstrate that for systems with a fixed training interval, the BER performance decreases strictly as the training interval increases. Most of this degradation is due to more error propagation caused by the longer training interval. It is seen that the BER performance of the receivers with the improved training scheme outperforms the systems with a fixed training interval significantly. The simulation results also show that for the systems with error propagation detection, the overhead of the training sequence drops sharply as the SNR increases and keeps lower over the concerned range of SNR.

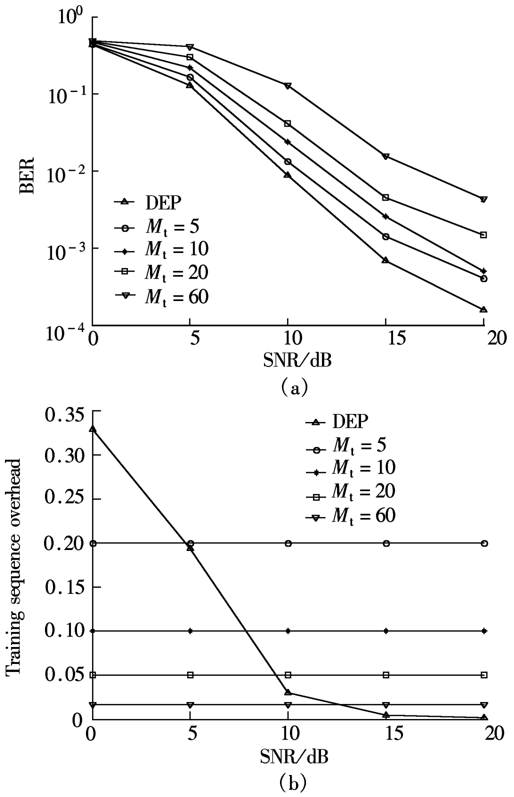

Fig. 3 Comparison of BER performance and overhead ratio with $i_{\max} = 2$

4 Conclusion

In this paper we have presented a low complexity iterative symbol detection and channel estimation algorithm for SFBC-OFDM systems. The methods to estimate the noise variance and covariance matrix of the channel response have been derived. An adaptive training scheme to suppress the error propagation is proposed and applied to improve the BER performance and keep low system overhead.

It has been shown by simulation results that the iterative algorithm brings better performance than that without iteration. Moreover the improved training scheme can prevent error propagation effectively, consequently improving the system performance significantly.

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SFBC-OFDM 系统中基于自适应训练机制的迭代接收机

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摘要: 研究了在空频块码正交频分复用(SFBC-OFDM)无线通信系统中适用于多径衰落信道下的迭代接收机的设计, 导出了一种迭代的联合信道估计与符号检测的算法. 在提出的算法中, 信道估计器交替地工作于2种模式. 在训练阶段, 采用基于DFT的估计器估计出信道状态信息, 并且采用所提出的算法估计出噪声方差和信道响应的互相关矩阵. 在数据传输模式下, 迭代地获得发送数据和信道状态信息. 为了抑制由于符号检测中误判引起的错误传播, 提出了一种简单的错误传播判定准则, 并使用了一种自适应的训练机制来抑制误差传播. 仿真结果显示, 与传统的迭代算法相比, 所提出的算法能够提供更好的误码性能, 且节约了系统开销.

关键词: SFBC-OFDM; 迭代接收机; 信道估计; 自适应训练机制

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