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Co-channel interference rejection for MIMO-OFDM systems

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Abstract: A method based on the maximum *a posteriori* probability (MAP) criterion is proposed to estimate the channel frequency response (CFR) matrix and interference-plus-noise spatial covariance matrix (SCM) for multiple input and multiple output orthogonal frequency division multiplexing (MIMO-OFDM) systems. An iterative solution is proposed to solve the MAP-based problem and an interference rejection combining (IRC) receiver is derived to suppress co-channel interference (CCI) based on the estimated CFR and SCM. Furthermore, considering the property of SCM, i. e., Hermitian and semi-definite, two schemes are proposed to improve the accuracy of SCM estimation. The first scheme is proposed to parameterize the SCM via a sum of a series of matrices in the time domain. The second scheme measures the SCM on each subcarrier as a low-rank model while the model order can be chosen through the penalized-likelihood approach. Simulation results are provided to demonstrate the effectiveness of the proposed method.

Key words: channel estimation; spatial covariance matrix (SCM) estimation; interference rejection; interference rejection combining (IRC) receiver

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In order to satisfy the increasing demand for supporting more users with high data rates, modern cellular systems such as WiMax and LTE-advanced reuse the same frequency spectrum at each cell[1]. Such systems will encounter serious impairment due to co-channel interference (CCI) from neighboring cells. Hence, interference rejection is valuable and essential to interference-limited systems.

Some CCI suppression techniques, i. e., IRC[2] and fast-ML[3], model the CCI as a zero-mean, time uncorrelated and spatially colored stationary Gaussian random process, which needs to estimate the spatial covariance matrix (SCM) of interference-plus-noise and the channel frequency response (CFR) matrix of desired users. The conventional solution for the SCM estimation problem is the sample SCM estimate which is the maximum likelihood (ML) estimate in the case of Gaussian signals. However, in practical OFDM systems, the number of available residual samples is limited, thus only a rough estimation of SCM can be obtained and other measures should be taken into account to improve the accuracy of SCM estimation. Larsson[4] proposed a method by using a low-order time-domain model to improve the SCM estimation based on the fact that the SCM is the FFT of power spectrum of the interference-plus-noise. A structured model for SCM estimation was proposed in Ref. [5], which was obtained by eigenvalue decomposition (EVD) of the residual sample covariance matrix. Raghavendra et al. [6] proposed a method by parameterizing SCM on each subcarrier as a combination of several low-rank models, and each model has different probabilities. An improved SCM estimation exploiting the frequency correlation of the interferer channels was discussed in Ref. [7].

As mentioned above, the CFR matrix of desired users is also required for interference rejection. Conventional channel estimations in MIMO-OFDM systems have been investigated intensively[8]. However, in these studies only thermal noise was considered; i. e., the SCM was treated as an identity matrix, which clearly degraded the accuracy of the channel estimation when CCI was present. Also, on the other hand, channel estimation errors will result in error residual samples for SCM estimation. In this paper, a method based on the maximum *a posteriori* probability (MAP) is proposed to effectively reduce the estimation errors of both CFR and SCM. Furthermore, we propose two schemes exploiting the number of interferers and the correlation between SCMs on different subcarriers. The first scheme transforms SCMs of all subcarriers back to the time domain via IFFT to obtain the correlation matrices followed by a temporal low-pass smoothing. Then, we represent each correlation matrix as a sum of a series of matrices described in Ref. [9] to maintain the semi-definite structure of SCM. The second scheme structures the SCM as a low-rank model with a few parameters through eigenvalue decomposition, and then smooths the low-rank matrix in the time domain. The low-rank model is somewhat similar to Ref. [5]. Ref. [5] assumed that the channels were frequency non-selective while there is no such assumption in this article. Both the two schemes not only keep the semi-definite of the obtained SCM, but also decrease the parameters to be

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estimated, which in turn improves the accuracy of SCM estimation. Finally, interference rejection combining (IRC) is employed on the receiver side for CCI suppression by using the estimated CFR and SCM.

1 System Model

Consider a MIMO-OFDM system equipped with transmit antennas N_t and receiving antennas N_r. Assume that there are N_c subcarriers for each OFDM symbol and the cyclic prefix is long enough to avoid inter-symbol interference (ISI). The channel is assumed to be a constant during one block of OFDM symbols, but changes independently from block to block. Here, one block of OFDM symbols consists of pilot symbols N_p followed by data symbols N_d. The received signal on pilot symbols is

$$\boldsymbol{y}_{k,n} = \boldsymbol{H}_k \boldsymbol{x}_{k,n} + \boldsymbol{z}_{k,n} \qquad 0 \leqslant k \leqslant N_c - 1,\ 0 \leqslant n \leqslant N_p - 1 \quad (1)$$

where $\boldsymbol{y}_{k,n} \in \mathbf{C}^{N_r \times 1}$ and $\boldsymbol{x}_{k,n} \in \mathbf{C}^{N_t \times 1}$ denote the received signal and transmitted pilot symbol on the k-th subcarrier of the n-th OFDM symbol in the frequency domain; $\boldsymbol{H}_k \in \mathbf{C}^{N_r \times N_t}$ is the MIMO channel matrix; $\boldsymbol{z}_{k,n} \sim \mathrm{CN}(\boldsymbol{0}, \boldsymbol{\Sigma}_k) \in \mathbf{C}^{N_r \times 1}$ denotes the received interference-plus-noise, which is modeled as a colored Gaussian noise. The relationship between CFR and CIR is given by

$$\boldsymbol{H}_k = \sum_{l=0}^{L-1} \boldsymbol{\Phi}_l \mathrm{e}^{-\mathrm{j}2\pi kl/N_c} = [\boldsymbol{\Phi}_0, \ldots, \boldsymbol{\Phi}_{L-1}] \cdot (\boldsymbol{f}_k \otimes \boldsymbol{I}_{N_t}) \quad (2)$$

where $\boldsymbol{\Phi}_l$ is the $N_r \times N_t$ time domain channel matrix; $\boldsymbol{f}_k = [\mathrm{e}^0, \mathrm{e}^{-\mathrm{j}2\pi kl/N_c}, \ldots, \mathrm{e}^{-\mathrm{j}2\pi k(L-1)/N_c}]^{\mathrm{T}}$; and L is the number of channel taps for the desired user. Substituting Eq. (2) into Eq. (1), we can obtain

$$\boldsymbol{y}_{k,n} = \boldsymbol{X}_{k,n} \boldsymbol{h} + \boldsymbol{z}_{k,n} \quad (3)$$

where $\boldsymbol{h} = \mathrm{vec}\{[\boldsymbol{\Phi}_0, \ldots, \boldsymbol{\Phi}_{L-1}]\}$, $\boldsymbol{X}_{k,n} = (\boldsymbol{x}_{k,n}^{\mathrm{T}} \otimes \boldsymbol{I}_{N_r})(\boldsymbol{f}_k^{\mathrm{T}} \otimes \boldsymbol{I}_{N_r N_t})$.

2 Estimation of CIR and SCM based on MAP Criterion

Collecting all the $N_c N_p$ received pilots together, the least square (LS) estimation of $\boldsymbol{h}$ is given as

$$\tilde{\boldsymbol{h}} = (\boldsymbol{X}^{\mathrm{H}} \boldsymbol{X})^{-1} \boldsymbol{X}^{\mathrm{H}} \boldsymbol{y} \quad (4)$$

where $\boldsymbol{y} = [\boldsymbol{y}_{0,0}^{\mathrm{T}}, \ldots, \boldsymbol{y}_{N_c-1,N_p-1}^{\mathrm{T}}]^{\mathrm{T}}$ and $\boldsymbol{X} = [\boldsymbol{X}_{0,0}^{\mathrm{T}}, \ldots \boldsymbol{X}_{N_c-1,N_p-1}^{\mathrm{T}}]^{\mathrm{T}}$. Next, the SCM of interference-plus-noise on the k-th subcarrier is estimated from the residuals,

$$\tilde{\boldsymbol{\Sigma}}_k = \frac{1}{N_p} \sum_{n=0}^{N_p-1} \tilde{\boldsymbol{z}}_{k,n} \tilde{\boldsymbol{z}}_{k,n}^{\mathrm{H}} = \frac{1}{N_p} \sum_{n=0}^{N_p-1} (\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n} \tilde{\boldsymbol{h}})(\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n} \tilde{\boldsymbol{h}})^{\mathrm{H}} \quad (5)$$

The LS estimation in Eq. (4) is not accurate due to the interference and because it also degrades the performance of SCM estimation for the inaccuracy of estimated residuals $\tilde{\boldsymbol{z}}_{k,n}$. We propose a method based on MAP criterion to estimate $\boldsymbol{h}$ and $\boldsymbol{\Sigma}_k$. With the statistical description $\boldsymbol{z}_{k,n} \sim \mathrm{CN}(\boldsymbol{0}, \boldsymbol{\Sigma}_k)$ and Eq. (3), the likelihood function of $\boldsymbol{h}$ and $\boldsymbol{\Sigma}_k$ is

$$f(\boldsymbol{y} \mid \boldsymbol{h}, \boldsymbol{\Sigma}_k) = \prod_{k=0}^{N_c-1} \prod_{n=0}^{N_p-1} |\pi \boldsymbol{\Sigma}_k|^{-1} \cdot \exp[-(\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n}\boldsymbol{h})^{\mathrm{H}} \boldsymbol{\Sigma}_k^{-1} (\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n}\boldsymbol{h})] \quad (6)$$

where $\boldsymbol{y} = [\boldsymbol{y}_{0,0}^{\mathrm{T}}, \ldots, \boldsymbol{y}_{N_c-1,N_p-1}^{\mathrm{T}}]^{\mathrm{T}}$. The channel vector $\boldsymbol{h}$ is modeled as $\boldsymbol{h} \sim \mathrm{CN}(\bar{\boldsymbol{h}}, \boldsymbol{R}_h)$, where $\bar{\boldsymbol{h}}$ is the mean vector and $\boldsymbol{R}_h$ is the correlation matrix. Since the statistics of the channel remain the same for a long time, we assume that $\bar{\boldsymbol{h}}$ and $\boldsymbol{R}_h$ are all known at the receiver. We treat $\boldsymbol{\Sigma}_k$ as a random matrix with the uniform distribution property which is independent of channel vector $\boldsymbol{h}$. Then, the *a posteriori* probability of $\boldsymbol{h}$ and $\boldsymbol{\Sigma}_k$ is given by

$$f(\boldsymbol{h}, \boldsymbol{\Sigma}_k \mid \boldsymbol{y}) = \frac{f(\boldsymbol{y} \mid \boldsymbol{h}, \boldsymbol{\Sigma}_k) f(\boldsymbol{h}) f(\boldsymbol{\Sigma}_k)}{\mathrm{f}(\boldsymbol{y})} \propto f(\boldsymbol{y} \mid \boldsymbol{h}, \boldsymbol{\Sigma}_k) f(\boldsymbol{h}) = \prod_{k=0}^{N_c-1} \prod_{n=0}^{N_p-1} |\pi \boldsymbol{\Sigma}_k|^{-1} \exp[-(\boldsymbol{h} - \bar{\boldsymbol{h}})^{\mathrm{H}} \boldsymbol{R}_h^{-1} (\boldsymbol{h} - \bar{\boldsymbol{h}})] \cdot \exp[-(\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n}\boldsymbol{h})^{\mathrm{H}} \boldsymbol{\Sigma}_k^{-1} (\boldsymbol{y}_{k,n} - \boldsymbol{X}_{k,n}\boldsymbol{h})] \, |\pi \boldsymbol{R}_h|^{-1} \quad (7)$$

Since $\boldsymbol{\Sigma}_k$ is positive semi-definite, the joint MAP estimation of $\boldsymbol{h}$ and $\boldsymbol{\Sigma}_k$ can be expressed as

$$\begin{aligned} &\{\hat{\boldsymbol{h}}, \hat{\boldsymbol{\Sigma}}_k\} = \max_{\boldsymbol{h}, \boldsymbol{\Sigma}_k} \ln f(\boldsymbol{h}, \boldsymbol{\Sigma}_k \mid \boldsymbol{y}) \\ &\text{s.t.}\ \ \boldsymbol{\Sigma}_k \geqslant 0 \qquad k = 0, 1, \ldots, N_c - 1 \end{aligned} \quad (8)$$

In the next paragraph, we will propose two schemes to deal with the SCM, which not only move away the constraint in Eq. (8), but also decreases the number of estimated parameters of $\boldsymbol{\Sigma}_k$ ($k = 0, 1, \ldots, N_c - 1$) to offer a better estimation performance. Then we design an iterative algorithm to solve Eq. (8).

2.1 Scheme 1

The received interference-plus-noise term $\boldsymbol{z}_k$ (ignoring subscript n) can be written as

$$\boldsymbol{z}_k = \sum_{l=0}^{L_I-1} \boldsymbol{v}_l \mathrm{e}^{-\mathrm{j}2\pi k/N_c} \quad (9)$$

where $\boldsymbol{v}_l$ is the time domain interference-plus-noise and L_I is the maximum length of the interference channels. Hence, the SCM on the k-th subcarrier can be represented as

$$\boldsymbol{\Sigma}_k = E\{\boldsymbol{z}_k \boldsymbol{z}_k^{\mathrm{H}}\} = E\left\{ \sum_{l=0}^{L_I-1} \boldsymbol{v}_l \mathrm{e}^{-\mathrm{j}2\pi kl/N_c} \sum_{l'=0}^{L_I-1} \boldsymbol{v}_{l'} \mathrm{e}^{-\mathrm{j}2\pi kl/N_c} \right\} = \sum_{l=0}^{L_I-1} \sum_{l'=0}^{L_I-1} E\{\boldsymbol{v}_l \boldsymbol{v}_{l'}\} \mathrm{e}^{-\mathrm{j}2\pi kl/N_c} \mathrm{e}^{-\mathrm{j}2\pi kl'/N_c} = \sum_{l=0}^{L_I-1} \sum_{l'=0}^{L_I-1} \boldsymbol{V}(1, 1') \mathrm{e}^{-\mathrm{j}2\pi k(l-l')/N_c} \quad (10)$$

It is assumed that the transmitted time domain interference on different sampling times is uncorrelated and the received interference is the linear convolution of the transmitted interference and the CIR, so $\boldsymbol{V}(l, l') = \sigma^2 \boldsymbol{I}_{N_r}$ if $l \neq l'$, where σ^2 is the noise variance. Thus, by replacing

$l-l'$ with n, Eq. (10) can be simplified as

$$\boldsymbol{\Sigma}_k = \sum_{n=-L_1+1}^{L_1-1} \boldsymbol{V}_n \mathrm{e}^{-\mathrm{j}2\pi kn/N_c} \tag{11}$$

Taking Eq. (11) into consideration, a general method to refine the estimated SCM emerges. The estimated SCMs can be first transformed to the time domain via IFFT to obtain the correlation matrices $\{\tilde{\boldsymbol{V}}_n, -N_c/2 \leqslant n \leqslant N_c/2-1\}$, after which the matrices $\{\tilde{\boldsymbol{V}}_n, |n| > L_1 - 1\}$ are assumed to be zeros. Then from Eq. (11), the accuracy-improved estimation of SCM can be obtained. However, one problem remains. The estimated matrices $\{\tilde{\boldsymbol{V}}_n, |n| \leqslant L_1 - 1\}$ cannot be guaranteed to always be an auto-correlation sequence, thus transforming $\{\tilde{\boldsymbol{V}}_n, |n| \leqslant L_1 - 1\}$ back into the frequency domain via FFT cannot guarantee the obtained SCM on each subcarrier to be positive semi-definite, which is an essential feature of a covariance matrix. Here we provide a solution method. According to Ref. [9], matrix $\boldsymbol{V}_n$ can be represented by matrix $\boldsymbol{G}_m$ as

$$\boldsymbol{V}_n = \sum_{m=n}^{L_1-1} \boldsymbol{G}_m \boldsymbol{G}_{m-n}^H \tag{12}$$

From Eqs. (11) and (12), we obtain $\boldsymbol{\Sigma}_k = \boldsymbol{\Sigma}_k^{1/2}\boldsymbol{\Sigma}_k^{\mathrm{H}/2}$, and

$$\boldsymbol{\Sigma}_k^{1/2} = \sum_{n=0}^{L_1-1} \boldsymbol{G}_n \mathrm{e}^{-\mathrm{j}2\pi kn/N_c} = \boldsymbol{G}(\boldsymbol{f}_{L_1,k} \otimes \boldsymbol{I}_{N_r}) \tag{13}$$

where $\boldsymbol{G} = [\boldsymbol{G}_0, \ldots, \boldsymbol{G}_{L_1-1}]$, $\boldsymbol{f}_{L_1,k} = [1, \mathrm{e}^{-\mathrm{j}2\pi k 1/N_c}, \ldots, \mathrm{e}^{-\mathrm{j}2\pi k(L_1-1)/N_c}]^{\mathrm{T}}$. Finally, the SCM on the subcarrier k can be expressed as

$$\boldsymbol{\Sigma}_k = \boldsymbol{G}\boldsymbol{Q}_k\boldsymbol{G}^{\mathrm{H}} \tag{14}$$

where $\boldsymbol{Q}_k = \boldsymbol{f}_{L_1,k}\boldsymbol{f}_{L_1,k}^{\mathrm{H}} \otimes \boldsymbol{I}_{N_r}$. Clearly, in Eq. (14) $\boldsymbol{\Sigma}_k \geqslant 0$ for all k and then the constrained optimization problem in Eq. (8) can be converted to an unconstrained optimization problem,

$$\{\hat{\boldsymbol{G}}, \hat{\boldsymbol{h}}\} = \max_{\boldsymbol{G},\boldsymbol{h}} \ln f(\boldsymbol{h}, \boldsymbol{G} \mid \boldsymbol{y}) \tag{15}$$

where $f(\boldsymbol{h}, \boldsymbol{G} \mid \boldsymbol{y})$ is similar to Eq. (7).

It is difficult to solve $\boldsymbol{h}$ and $\boldsymbol{G}$ at the same time. We take a more general method which solves one while treats the other as constant, and the iterative solution is considered. It is simple to find that, when $\boldsymbol{G}$ is assumed to be fixed, and let the derivation of Eq. (15) with respect to $\boldsymbol{h}^*$ to be zeros, the estimation of CIR is

$$\tilde{\boldsymbol{h}} = \left(\sum_{k=0}^{N_c-1}\sum_{n=0}^{N_p-1} \boldsymbol{X}_{k,n}^{\mathrm{H}}(\boldsymbol{G}\boldsymbol{Q}_k\boldsymbol{G}^{\mathrm{H}})^{-1}\boldsymbol{X}_{k,n} + \boldsymbol{R}_h^{-1}\right)^{-1} \cdot \left(\sum_{k=0}^{N_c-1}\sum_{n=0}^{N_p-1} \boldsymbol{X}_{k,n}^{\mathrm{H}}(\boldsymbol{G}\boldsymbol{Q}_k\boldsymbol{G}^{\mathrm{H}})^{-1}\boldsymbol{y}_{k,n} + \boldsymbol{R}_h^{-1}\bar{\boldsymbol{h}}\right) \tag{16}$$

In the same way, we can obtain the estimation of matrix $\boldsymbol{G}$. If we estimate $\boldsymbol{\Sigma}_k$ (for $k = 0, 1, \ldots, N_c - 1$), the number of unknown parameters is $N_r N_r N_c$. Now, we estimate matrix $\boldsymbol{G}$ instead, and the number of unknown parameters is reduced to $N_r N_r L_1$. In a real scenario, $L_1 \ll N_c$, so the proposed method can improve the accuracy of the estimation effectively as it can significantly reduce the unknown parameters.

2.2 Scheme 2

Scheme 1 exploits the correlation of SCMs on different subcarriers to reduce the unknown parameters while the Scheme 2 in this subsection further exploits the number of interferers to improve the estimation accuracy. Assume that the interferer transmission is synchronized with that of the desired user; i. e., the cyclic prefix of the interference signal lines up with that of the desired signal. We assume that each interferer has only one antenna without loss of generality. Then, according to Eq. (1), the interference-plus-noise can be written as

$$\boldsymbol{z}_{k,n} = \sum_{i=0}^{d} \boldsymbol{g}_{k,i} \boldsymbol{s}_{k,n,i} + \boldsymbol{e}_{k,n}$$
$$0 \leqslant k \leqslant N_c - 1,\ 0 \leqslant n \leqslant N_p - 1 \tag{20}$$

where d is the number of interferers; $\boldsymbol{g}_{k,i} \in \mathbf{C}^{N_r \times 1}$ denotes the channel vector of the i-th interferer on the k-th subcarrier; $s_{k,n,i}$ is the transmitted data for interferer i with a zero mean and unit transmission energy. Signals from different interferers are assumed to be uncorrelated, i. e., $E\{s_{k,n,i}s_{k,n,j}^*\} = \delta(i-j)$, where δ denotes the Kronecker delta function. $\boldsymbol{e}_{k,n} \in \mathbf{C}^{N_r \times 1}$ is additive white Gaussian noise (AWGN) vector with a zero mean and covariance matrix $\sigma^2 \boldsymbol{I}_{N_r}$. The SCM of interference-plus-noise on the k-th subcarrier can be expressed as

$$\boldsymbol{\Sigma}_k = E\{\boldsymbol{z}_{k,n}\boldsymbol{z}_{k,n}^{\mathrm{H}}\} = \boldsymbol{\Psi}_k + \sigma^2\boldsymbol{I}_{N_r} \qquad k = 0, 1, \ldots, N_c - 1 \tag{21}$$

where $\boldsymbol{\Psi}_k = \sum_{i=0}^{d} \boldsymbol{g}_{k,i}\boldsymbol{g}_{k,i}^{\mathrm{H}}$ is the instantaneous SCM of interference. It is simple to conclude that the rank of $\boldsymbol{\Psi}_k$ is no greater than the number of interferers d. So a reasonable decomposition of the SCM is

$$\boldsymbol{\Sigma}_k = \boldsymbol{G}_k\boldsymbol{G}_k^{\mathrm{H}} + \sigma^2\boldsymbol{I}_{N_r} \qquad k = 0, 1, \ldots, N_c - 1 \tag{22}$$

where $\boldsymbol{G}_k \in \mathbf{C}^{N_r \times d}$ is made up of d dominant eigenvalues and corresponding eigenvectors of matrix $\boldsymbol{\Sigma}_k$. Here we suppose that $d \leqslant N_r$, and if $d > N_r$, the dimension of $\boldsymbol{G}_k$ is $N_r \times N_r$. σ^2 is the noise variance, which can be obtained by a long statistics estimation of the thermal noise. The same as $\boldsymbol{\Sigma}_k$ is described in Scheme 1, the frequency domain matrix $\boldsymbol{G}_k$ can be represented as a truncated Fourier sum:

$$\boldsymbol{G}_k = \sum_{n=-L_1+1}^{L_1-1} \boldsymbol{A}_n \mathrm{e}^{-\mathrm{j}2\pi kn/N_c} = \boldsymbol{A}(\boldsymbol{f}_{L_1,k} \otimes \boldsymbol{I}_{\mathrm{d}}) \tag{23}$$

where $\boldsymbol{A} = [\boldsymbol{A}_{-L_1+1}, \ldots, \boldsymbol{A}_0, \ldots, \boldsymbol{A}_{L_1-1}]$, $\boldsymbol{f}_{L_1,k} = [\mathrm{e}^{\mathrm{j}2\pi k(L_1-1)/N_c}, \ldots, 1, \ldots, \mathrm{e}^{-\mathrm{j}2\pi k(L_1-1)/N_c}]^{\mathrm{T}}$. Substituting Eq. (23) into Eq. (11), we obtain

$$\boldsymbol{\Sigma}_k = \boldsymbol{A}\boldsymbol{Q}_k\boldsymbol{A}^{\mathrm{H}} + \sigma^2\boldsymbol{I}_{N_r} \tag{24}$$

where $\boldsymbol{Q}_k = (\boldsymbol{f}_{L_I,k} \otimes \boldsymbol{I}_d)(\boldsymbol{f}_{L_I,k} \otimes \boldsymbol{I}_d)^{\mathrm{H}}$. Clearly, the structure of SCM in Eq. (24) is a Hermitian positive semi-definite matrix. Then the constrained optimization problem in Eq. (8) can be equivalent to an unconstrained optimization problem,

$$\{\hat{\boldsymbol{A}}, \hat{\boldsymbol{h}}\} = \min_{\boldsymbol{A},\boldsymbol{h}} \ln f(\boldsymbol{h}, \boldsymbol{A} \mid \boldsymbol{y}) \tag{25}$$

where $f(\boldsymbol{h}, \mathrm{A} \mid \boldsymbol{y})$ is similar to Eq. (7).

The same as in Scheme 1, taking the derivative with respect to $\boldsymbol{h}$ and $\boldsymbol{A}$, we can obtain the estimation of CIR and matrix $\boldsymbol{A}$, and thus SCM. In this proposed method, we only need to estimate matrix $\boldsymbol{A}$ with $N_r d(2L_I - 1)$ unknown parameters instead of matrices $\boldsymbol{\Sigma}_k$ (for $k = 0, 1, \ldots, N_c - 1$) with $N_r N_r N_c$ unknown parameters. In most cases, Scheme 2 is more accurate compared to Scheme 1 since it estimates fewer unknown parameters and exploits more priori information such as the number of interferers.

We have assumed so far that the multi-tap L_I of interference channels and the number of interferers d are known. In practice, they can be estimated by the MDL criterion[10-11].

3 Receiver with Co-Channel Interference

Interference rejection combining (IRC) is an efficient CCI suppression technique employed at the receiver side. It is based on the minimum mean square error (MMSE) criterion. Following Ref. [12], assuming that both the true $\boldsymbol{H}$ and the true $\boldsymbol{\Sigma}$ are known, the IRC received data symbol is given as

$$\hat{\boldsymbol{x}} = \hat{\boldsymbol{H}}^{\mathrm{H}}(\hat{\boldsymbol{H}}\hat{\boldsymbol{H}}^{\mathrm{H}} + \hat{\boldsymbol{\Sigma}})^{-1}\boldsymbol{y} \tag{26}$$

4 Simulation Results

Simulations are presented in this section to describe the performance. We consider a block-fading MIMO-OFDM system with $N_c = 64$ subcarriers. The desired user goes through a channel with 8 taps in an exponential power delay profile (PDP). The interferers are equipped with one antenna. The signals are modulated to QPSK signal constellation. The channel coding is a convolutional code with a rate of 1/2.

We compare the FER performance under different scenarios and different estimation methods under the following conditions: 1) The ideal instantaneous SCM is known and the channel is estimated according to Eq. (16); 2) The ideal long-term SCM (defined as $E\{\boldsymbol{\Sigma}_k\}$) for each subcarrier k is known and the channel is estimated according to Eq. (16); 3) The channel and instantaneous SCM are estimated by the algorithm discussed in Ref. [7]; 4) The channel and instantaneous SCM are estimated by the algorithm discussed in Ref. [4]; 5) The first scheme we proposed; 6) The second scheme we proposed.

In Fig. 1, the mean square error (MSE) of the CFR is simulated under a different number of iterations. It is simple to conclude that the proposed algorithm has stable convergence after one iteration.

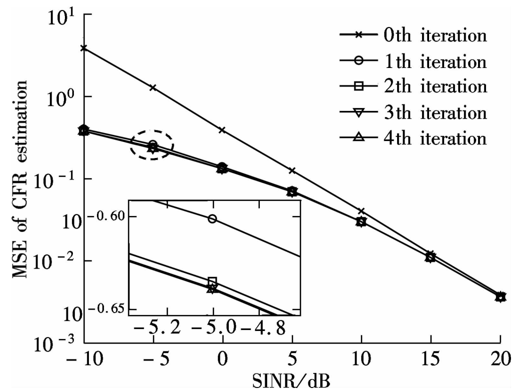

Fig. 1 MSE of CFR under different number of iterations

Fig. 2 compares the MSE of SCM under different schemes. The proposed schemes in this article show great performance and Scheme 2 outperforms Scheme 1. The MSE of SCM by using long-term SCM is the largest, for the long-term SCM is unable to characterize the instantaneous property of SCM in this case (INR = 5dB).

Next, we derive the IRC receiver and compare the FER of the above six schemes. In Fig. 3, the two proposed schemes providing 2 and 2.8 dB gain in SINR are compared to the method in Ref. [4] for the FER of 10^{-3},

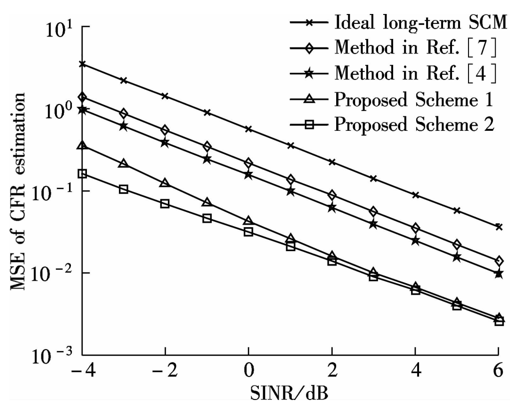

Fig. 2 MSE of SCM vs. SINR (INR = 5 dB, $N_p = 2$, $N_d = 10$, one interferer and $L_I = 2$)

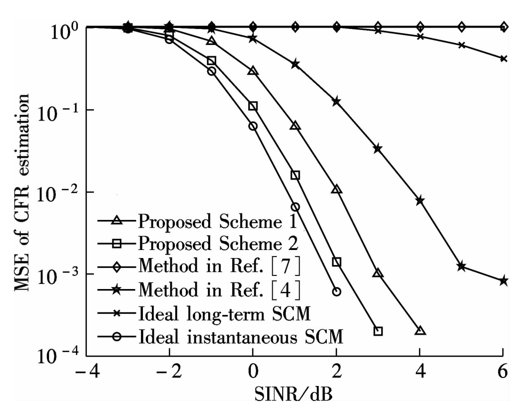

Fig. 3 FER vs. SINR (INR = 10 dB, $N_p = 1$, $N_d = 5$, one interferer and $L_I = 1$)

respectively, and there is less than a 0.5 dB gap between the second scheme we proposed and the method known as the ideal instantaneous SCM. The method in Ref. [7] with the worst-case performance for this method is invalid when the pilot symbol number is one. In Fig. 4, the method in Ref. [7] performs better when using more pilots.

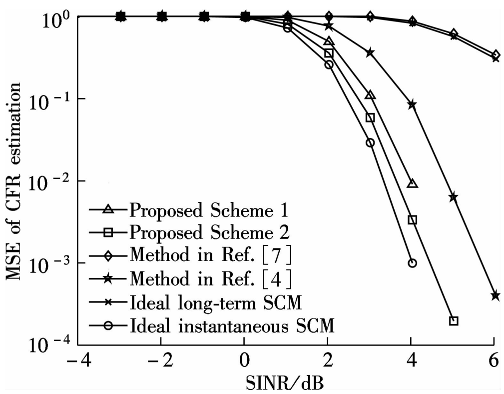

Fig. 4 FER vs. SINR (INR = 5 dB, $N_p = 2$, $N_d = 10$, one interferer and $L_I = 2$)

5 Conclusion

In this paper, the method based on the MAP probability function is proposed for channel and SCM estimation of MIMO-OFDM systems. The SCM and CFR are estimated iteratively. Meanwhile, considering the structure of the SCM, two approaches adopting different matrix handling strategies can effectively reduce the estimation errors of both CFR and SCM. This paper adopts the IRC receiver based on the estimated CFR and SCM. Simulation results show that the proposed method can significantly improve the system performance.

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MIMO-OFDM 系统中同信道干扰抑制

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摘要:针对 MIMO-OFDM 系统,提出了一种最大后验概率的信道矩阵和干扰协方差矩阵估计方法,并设计了迭代求解算法.利用所估计的信道矩阵和干扰协方差矩阵,采用 IRC 接收机完成同信道干扰的抑制.利用干扰协方差阵的共轭对称与半正定等特性,提出 2 种干扰协方差矩阵的处理方案以提高其估计精度.第 1 种方案将每个子载波上干扰协方差矩阵表征为一系列时域矩阵之和,第 2 种方案将每个子载波上的干扰协方差矩阵用低阶模型来建模,其中模型阶数通过最小描述长度算法估计.仿真结果表明了所提方案的有效性.

关键词:信道估计;干扰协方差估计;干扰抑制;IRC 接收机

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