南京航空航天大学

论 文 集

（二〇一〇年） 第22册

电子信息工程学院

（第3分册）

南京航空航天大学科技部编

二〇一一年五月

电子信息工程学院

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Wireless Pers Commun (2010) 53:603–612
DOI 10.1007/s11277-009-9706-3

Blind Carrier Frequency Offset Estimation for OFDM System with Multiple Antennas using Multiple-invariance Properties

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Abstract In this paper, we address the problem of carrier frequency offset (CFO) estimation for Orthogonal Frequency Division Multiplexing (OFDM) communications systems with multiple antennas. We reconstruct the received signal to form data model with multi-invariance property, and subsequently derive a multiple-invariance ESPRIT algorithm for CFO estimation. This algorithm has improved CFO estimation compared to ESPRIT method and maximum likelihood method. Simulation results illustrate performance of this algorithm.

Keywords Carrier frequency offset (CFO) · Multiple-invariance · OFDM · ESPRIT · Multiple antennas

1 Introduction

Orthogonal frequency division multiplexing (OFDM) is an efficient technique for high-speed digital transmission over multipath fading channels [1,2]. OFDM has been recently emerged as a promising technique for future mobile communications. As we all know that OFDM is highly sensitive to carrier frequency offset (CFO). In general, CFO estimation algorithm can be divided into two categories: the non-blind CFO estimation methods and blind CFO estimation methods. The non-blind CFO estimators are based on pilots[3–5] or on the cyclic prefix (CP) like Maximum Likelihood (ML) carrier-frequency offset estimator [6]. The blind CFO estimators contain MUSIC method [7], ESPRIT method [8], kurtosis-based CFO estimator [9], constant modulus-based CFO estimator [10], and cyclostationarity-based approach [11]. In contrast to the training-based and pilot-based methods, the blind CFO estimation methods improve bandwidth efficiency. Among the previous blind method, [9] assumes non-Gaussian sources, and uses kurtosis to measure non-Gaussianity, and

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[10,11] assume other special properties of the sources, like constant-modulus and cyclostationarity. MUSIC method [7] and ESPRIT method [8] are based on subspace technique and have the average CFO estimation performance.

Multiple-antenna receiver is useful in OFDM systems for providing receive diversity to overcome fading [12]. OFDM system with multiple antennas has been used in many communication systems such as Digital Video Broadcasting for Handheld (DVBH)[13], future mobile wireless communications [14], and so on. Similar to the single input single output (SISO) OFDM systems, the multiple-antenna diversity receivers of the OFDM system are sensitive to carrier synchronization errors. The CFO between the transmitter and the receiver must be estimated and compensated to ensure subcarrier orthogonality. In this paper, we address the problem of CFO estimation for OFDM systems with multiple antennas. We reconstruct the received signal to form data model with multiple-invariance property, and then multiple-invariance ESPRIT (MI-ESPRIT) algorithm for carrier frequency offset estimation is proposed. Comparing to ESPRIT method and ML method, we derive an algorithm with better CFO estimation performance.

This paper is structured as follows. Section 2 develops data model. Section 3 proposes algorithm. Section 4 presents simulation results, and Sect. 5 summarizes our conclusions.

Note We denote by $(.)^*$ the complex conjugation, by $(.)^T$ the matrix transpose, and by $(.)^H$ the matrix conjugate transpose. The notation $(.)^+$ refers to the Moore–Penrose inverse (pseudo inverse). $\|\ \|_F$ stands for Forbenius norm.

2 Data Model

We consider the uplink of an OFDM system, in which the transmitter has only a single transmitting antenna while the receiver is equipped with an array of I antennas. The number of subcarriers is N and a cyclic prefix of L sampling intervals is used. L is chosen to exceed the maximum delay spread. We denote the kth block to be transmitted by $\mathbf{s}(k) = [s_1(k), s_2(k), \ldots, s_P(k)]^T$. P parallel data are modulated onto P ($P < N$) subcarriers, the rest of which ($N - P$) subcarriers are virtual carriers. The multicarrier-modulated signal is padded with a cyclic prefix (CP). The output signal of insertion CP unit is $\mathbf{d}(k) = \mathbf{T}_{cp}\mathbf{F}_P\mathbf{s}(k)$, where $\mathbf{T}_{cp}$ is a matrix which is used to add CP, $\mathbf{F}_P \in \mathbb{C}^{N\times P}$ comprises the first P columns of the inverse discrete Fourier transform matrix. The signal $\mathbf{d}(k)$ is transmitted through the multipath fading channel. The received baseband signal of the ith antenna is $\mathbf{u}_i(k)$, and the output signal of the removing-CP unit is $\mathbf{x}_i(k) = \mathbf{T}_{rm}\mathbf{u}_i(k)$, where $\mathbf{T}_{rm}$ is a matrix which is used to remove CP . We assume that all the receive antenna are affected by the same CFO.

Define $H_i(n) = \sum_{l=0}^{L_m-1} h_i(l)e^{-j2\pi nl/N}$ as the channel frequency response for the nth subcarrier, corresponding to the ith antenna, where $\{h_i(l)\}_{l=0}^{L_m-1}$ is discrete-time channel impulse response. The frequency domain channel vector for the ith receive antenna is $\mathbf{h}_i = [H_i(1), H_i(2), \ldots, H_i(P)]^T$. The frequency domain channel matrix for the multiple antenna receivers is

$$\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_I]^T \in \mathbb{C}^{I\times P} \tag{1}$$

The output signal of the ith antenna through removing CP is denoted as [8]

$$\mathbf{x}_i(k) = \mathbf{E}\mathbf{F}_P\mathrm{diag}(\mathbf{h}_i)\mathbf{s}(k)e^{j2\pi\Delta f(k-1)(N+L)} \tag{2}$$

where Δf is CFO, $\mathbf{E} = \mathrm{diag}\{1, e^{j2\pi\Delta f}, \ldots, e^{j(N-1)2\pi\Delta f}\} \in \mathbb{C}^{N\times N}$ is CFO matrix, diag($\mathbf{h}_i$) is

$$\mathrm{diag}(\mathbf{h}_i) = \begin{bmatrix} H_i(1) & \cdots & 0 \\ \vdots & H_i(n) & \vdots \\ 0 & \cdots & H_i(P) \end{bmatrix} \in \mathbb{C}^{P\times P} \tag{3}$$

We assume the channel parameter is constant for K blocks. Define the source matrix $\mathbf{S} = [\mathbf{s}(1), \mathbf{s}(2), \ldots, \mathbf{s}(K)]^T \in \mathbb{C}^{K\times P}, \mathbf{X}_i = [\mathbf{x}_i(1)\ \ \mathbf{x}_i(2)\ \ \cdots\ \ \mathbf{x}_i(K)]$, which is denoted as

$$\mathbf{X}_i = \mathbf{A}\mathrm{diag}(\mathbf{h}_i)\mathbf{B}^T, \quad i = 1, 2, \ldots, \mathrm{I} \tag{4}$$

where $\mathbf{B} = \mathrm{diag}\{1, e^{j2\pi\Delta f(N+L)}, \ldots, e^{j2\pi\Delta f(K-1)(N+L)}\}\mathbf{S} \in \mathbb{C}^{K\times P}, \quad \mathbf{A} = \mathbf{E}\mathbf{F}_P \in \mathbb{C}^{N\times P}$. Equation (4) can be denoted as

$$\mathbf{X}_i = \mathbf{A}D_i(\mathbf{H})\mathbf{B}^T, \quad i = 1, 2, \ldots, \mathrm{I} \tag{5}$$

where $D_i(.)$ is to extract the mth row of its matrix argument and construct a diagonal matrix out of it. The signal in (5) can also be expressed as trilinear model [15]

$$x_{n,k,i} = \sum_{p=1}^{P} a_{n,p}b_{k,p}h_{i,p}, \quad n = 1, \ldots, N, \ k = 1, \ldots, K, \ i = 1, \ldots, I \tag{6}$$

where $h_{i,p}$ stands for the (i, p) element of the matrix $\mathbf{H}$, and similarly for the others. N, K, I are the number of subcarriers, the blocks of source and the antennas, respectively. $\mathbf{X}_i = \mathbf{A}D_i(\mathbf{H})\mathbf{B}^T, \quad i = 1, 2, \ldots, \mathrm{I}$, can be regarded as slicing the trilinear model in a series of slices (matrix) along the antenna direction. The symmetry of the trilinear model in(6) allows other matrix system rearrangement, which is

$$\mathbf{Y}_n = \mathbf{H}D_n(\mathbf{A})\mathbf{B}^T, \quad n = 1, 2, \ldots, N \tag{7}$$

where $\mathbf{Y}_n, n = 1, 2, \ldots, N$, can be regarded as the constructing signal of Eq.5. In the presence of noise, the received signal model becomes $\tilde{\mathbf{Y}}_n = \mathbf{H}D_n(\mathbf{A})\mathbf{B}^T + \mathbf{W}_n$, $n = 1, 2, \ldots, N$, where $\mathbf{W}_n$ is the received noise corresponding to the nth subcarrier.

3 Multiple-Invariance ESPRIT (MI-ESPRIT) Algorithm for Frequency Offset Estimation

Multiple-invariance ESPRIT algorithm [6] was proposed to handle arrays composed of multiple identical subarrays. Multiple-invariance ESPRIT, as a generalization of ESPRIT, employs the property of multiple shift invariance to enhance the performance of the ESPRIT algorithm.

According to Eq.7, we form the following matrix

$$\mathbf{Y} = \begin{bmatrix} \mathbf{Y}_1 \\ \mathbf{Y}_2 \\ \vdots \\ \mathbf{Y}_N \end{bmatrix} = \begin{bmatrix} \mathbf{H}D_1(\mathbf{A}) \\ \mathbf{H}D_2(\mathbf{A}) \\ \vdots \\ \mathbf{H}D_N(\mathbf{A}) \end{bmatrix} \mathbf{B}^T = \begin{bmatrix} \mathbf{H} \\ \mathbf{H}\mathbf{\Phi} \\ \vdots \\ \mathbf{H}\mathbf{\Phi}^{N-1} \end{bmatrix} \mathbf{B}^T \tag{8}$$

where $\mathbf{\Phi} = \text{diag}\{e^{j2\pi\Delta f}, e^{j2\pi(\frac{1}{N}+\Delta f)}, \ldots, e^{j2\pi(\frac{P-1}{N}+\Delta f)}\}$ is the rotation matrix. According to multi-invariance property in Eq.8, we can use multiple-invariance ESPRIT [6] to estimate frequency offset. For Eq.8, $\mathbf{R}_Y = \mathbf{Y}\mathbf{Y}^H$. We denote the matrix containing the eigenvectors $\{\mathbf{f}_p\}_{p=1}^{P}$ associated with the P largest eigenvalues of $\mathbf{R}_Y$ by $\mathbf{E}$

$$\mathbf{E} = \begin{bmatrix} \mathbf{H} \\ \mathbf{H\Phi} \\ \vdots \\ \mathbf{H\Phi}^{N-1} \end{bmatrix} \mathbf{T} \tag{9}$$

where $\mathbf{T}$ is a P × P full-rank matrix. According to (9), we define $\mathbf{E}_1$ and $\mathbf{E}_2$

$$\mathbf{E}_1 = \begin{bmatrix} \mathbf{H} \\ \mathbf{H\Phi} \\ \vdots \\ \mathbf{H\Phi}^{N-2} \end{bmatrix} \mathbf{T}; \quad \mathbf{E}_2 = \begin{bmatrix} \mathbf{H\Phi} \\ \mathbf{H\Phi}^2 \\ \vdots \\ \mathbf{H\Phi}^{N-1} \end{bmatrix} \mathbf{T} \tag{10}$$

According to Eq.10,

$$\mathbf{E}_2 = \begin{bmatrix} \mathbf{H} \\ \mathbf{H\Phi} \\ \vdots \\ \mathbf{H\Phi}^{N-2} \end{bmatrix} \mathbf{\Phi T} = \begin{bmatrix} \mathbf{H} \\ \mathbf{H\Phi} \\ \vdots \\ \mathbf{H\Phi}^{N-2} \end{bmatrix} \mathbf{TT}^{-1}\mathbf{\Phi T} = \mathbf{E}_1\mathbf{T}^{-1}\mathbf{\Phi T} \tag{11}$$

Define $\mathbf{\Psi} = \mathbf{T}^{-1}\mathbf{\Phi T}$, and Eq.11 becomes $\mathbf{E}_2 = \mathbf{E}_1\mathbf{\Psi}$, then $\mathbf{\Psi} = \mathbf{E}_1^{+}\mathbf{E}_2$. Because $\mathbf{\Psi}$ and $\mathbf{\Phi}$ has the same eigenvalues, we can use eigenvalue decomposition (EVD) for $\mathbf{\Psi}$ to get $\widehat{\mathbf{\Phi}}$, here we have

$$e^{j2\pi\Delta\hat{f}} = \frac{tr(\hat{\mathbf{\Phi}})}{\sum_{n=0}^{P-1} e^{j\frac{2\pi}{N}n}} \tag{12}$$

where *tr*(.) denotes the sum of the elements of the principal diagonal of the matrix. The CFO $2\pi\Delta\hat{f}$ can be calculated from (12).

In contrast to ESPRIT, our algorithm has a heavier computational load, which is usually dominated by formation of the covariance matrix and calculation of EVD. The major computational complexity of our algorithm is $O(\mathrm{KI}^2\mathrm{N}^2 + \mathrm{I}^3\mathrm{N}^3 + \mathrm{P}^3)$, while ESPRIT [5] requires $O(\mathrm{KIN}^2 + \mathrm{N}^3 + \mathrm{P}^3)$.

4 Simulation Results

The noisy received signal model becomes $\tilde{\mathbf{Y}}_n = \mathbf{H}D_n(\mathbf{A})\mathbf{B}^T + \mathbf{W}_n, \quad n = 1, 2, \ldots, N$, where $\mathbf{W}_n$ is the received noise. The Signal-Noise-Ratio (SNR) is defined as

$$\mathrm{SNR} = 10\log_{10}\frac{\sum_{n=1}^{N}\left\|\mathbf{H}D_n(\mathbf{A})\mathbf{B}^T\right\|_F^2}{\sum_{n=1}^{N}\|\mathbf{W}_n\|_F^2}dB$$

We present Monte Carlo simulations that are to assess the CFO estimation performance of our proposed algorithm. The number of Monte Carlo trials is 1,000. We consider OFDM system with N subcarriers, and a CP with eight sampling intervals is used in this simulation.

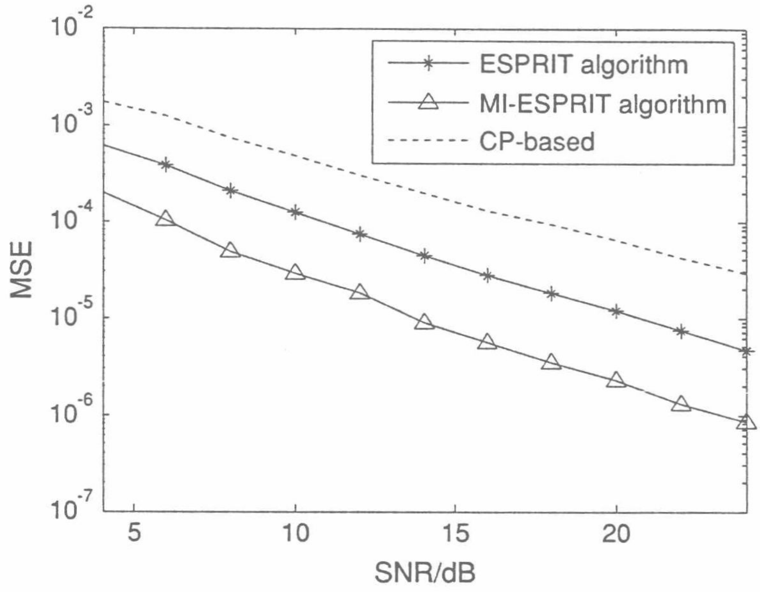

Fig. 1 CFO estimation performance comparison

The channel is modeled as an FIR filter of length $L_m = 4(L_m < L)$. $\Delta\omega = 2\pi\Delta f$ is fixed at 0.4ω, where $\omega = 2\pi/N$ is the subcarrier frequency spacing. To quantify the performance of the CFO estimation, the mean square error (MSE) of the estimates is used, and it is defined as

$$MSE = \frac{1}{M}\sum_{m=1}^{M}\left(\frac{\Delta\hat{f}^m - \Delta f}{1/N}\right)^2$$

where $\Delta\hat{f}^m$ is the estimated CFO of the$m$th Monte Carlo test, M is the number of Monte Carlo trials, Δf is the perfect CFO.

We compare our proposed algorithm against ESPRIT method and ML method (CP-based) [6]. According to Eq.4, we get

$$\begin{aligned}\mathbf{X}_E &= [\mathbf{X}_1 \quad \mathbf{X}_2 \quad \cdots \quad \mathbf{X}_I] \\ &= \mathbf{A}\left[\mathrm{diag}(\mathbf{h}_1)\mathbf{B}^T \quad \mathrm{diag}(\mathbf{h}_2)\mathbf{B}^T \quad \cdots \quad \mathrm{diag}(\mathbf{h}_I)\mathbf{B}^T\right]\end{aligned}$$

The matrix **A** is with Vandermonde characteristic, and ESPRIT algorithm can be used for CFO estimation.

Figure1 shows the CFO estimation performance comparison with N= 32, P = 20, K = 100, and I = 4, where N, I, K and P are the number of the total subcarriers, antennas, signal blocks, and subcarriers which are used to transmit data, respectively. From Fig1., we find that in contrast to ESPRIT method and CP-based method, CFO estimation performance of our proposed algorithm is improved. Our algorithm, which employs the multi-invariance property, has the better capability to suppress noise than single-invariance ESPRIT.

Figure2 presents the mean square error performance of the CFO estimation in different data block numbers. N= 32, P = 20 and I = 4 are used in Fig.2. From Fig.2 the CFO estimation performance of our proposed algorithm is improved with K increasing.

Figure3 shows the performance of the CFO estimator with different numbers of receive antennas. N= 32, P = 20 and K= 100 in this simulation. From Fig3, the CFO estimation

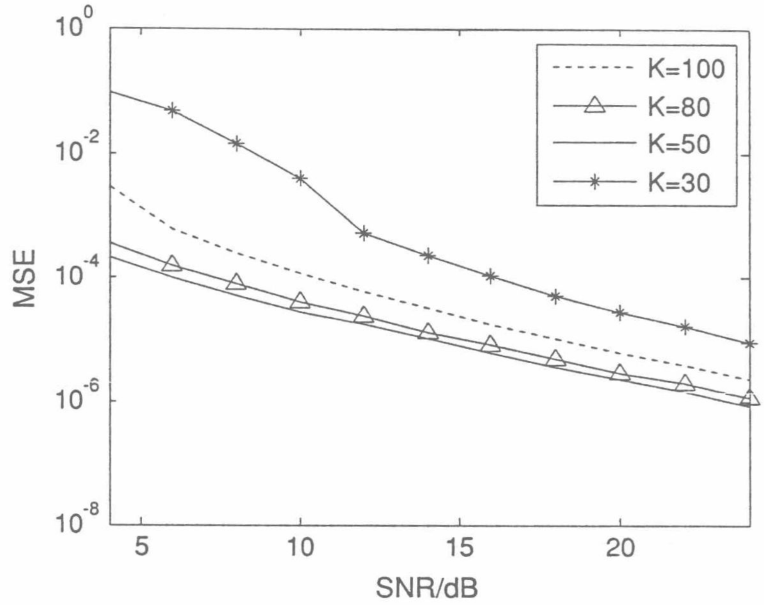

Fig. 2 CFO estimation performance at different values of K

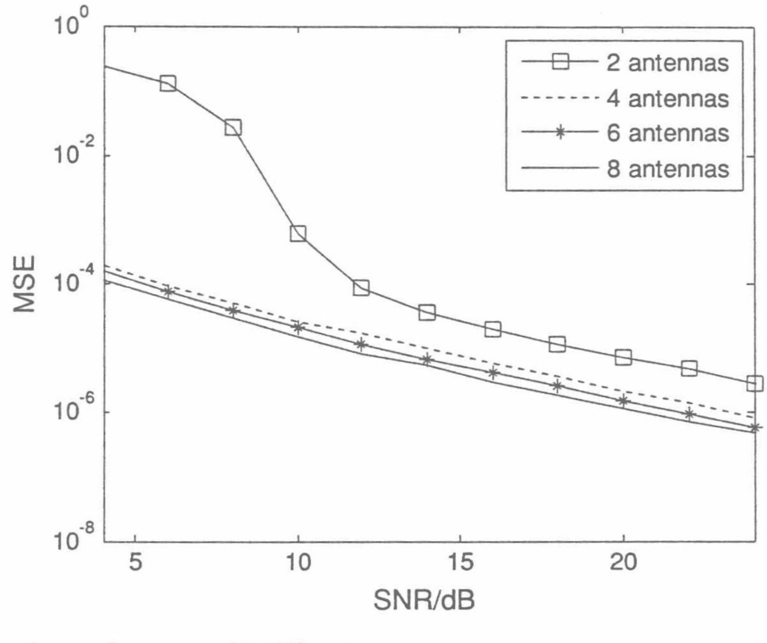

Fig. 3 CFO estimation performance with different antennas

performance of our proposed algorithm is improved with the number of antennas increasing. Multiple receive antennas improve CFO estimation performance because of receive diversity gain. However, when I > 4, the increment in diversity gain is often slowed down with the number of antennas increasing.

Figure4 presents the performance of the CFO estimator with N= 32, I = 4, K = 100 and different values of P. From Fig.4, we find that CFO estimation performance degrades with P increasing. Since the rank of the matrix$\mathbf{A} \in \mathbb{C}^{N\times P}$ increases while the P indicates increment, then CFO estimation performance gets worsen.

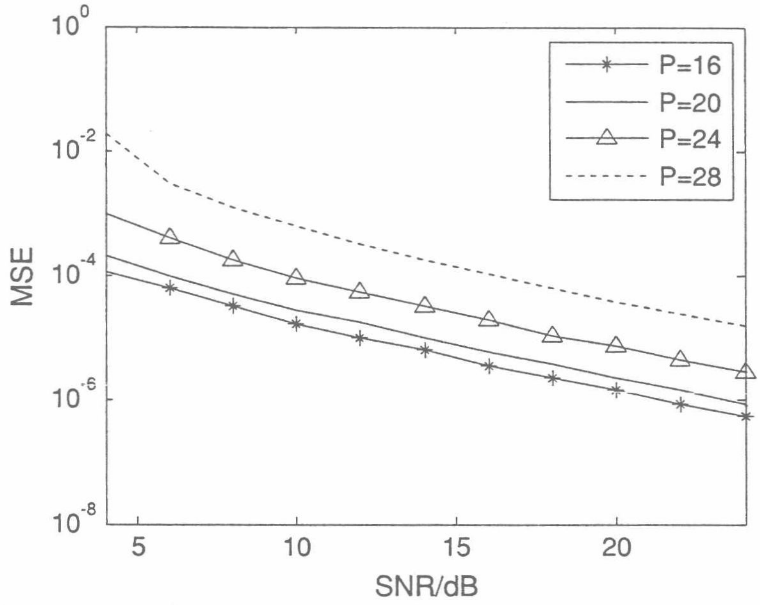

Fig. 4 CFO estimation performance with different values of P

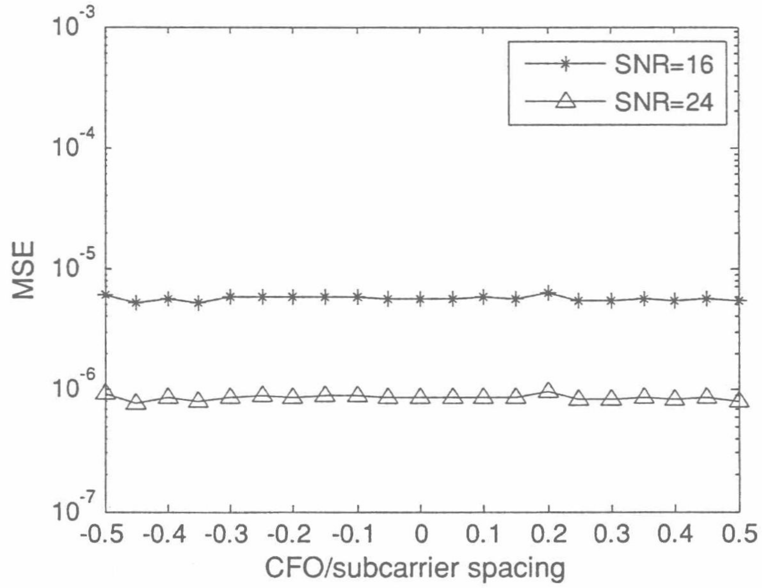

Fig. 5 CFO estimation performance with different CFOs

Figure 5 shows the performance of CFO estimator with different CFOs. N $=$ 32, P $=$ 20, I $=$ 4, K $=$ 100 in this figure, and CFO varies from -0.5ω to 0.5ω. From Fig. 5, we find that our proposed algorithm has very close CFO estimation performance for different CFOs.

Figure 6 presents the performance of the CFO estimator with K$=$ 100, I $=$ 4 and different N. The value of N varies, and $P = 5N/8$ in this figure. From Fig. 6, we find that CFO estimation performance degrades with N increasing. N is selected 32, 64, and 128. P also proportionally enlarges when N is being added, then the rank of the matrix $\mathbf{A} \in \mathbb{C}^{N\times P}$ also increases. As a result, the estimation accuracy of matrix $\mathbf{A}$ degrades and then CFO estimation performance aggravates.

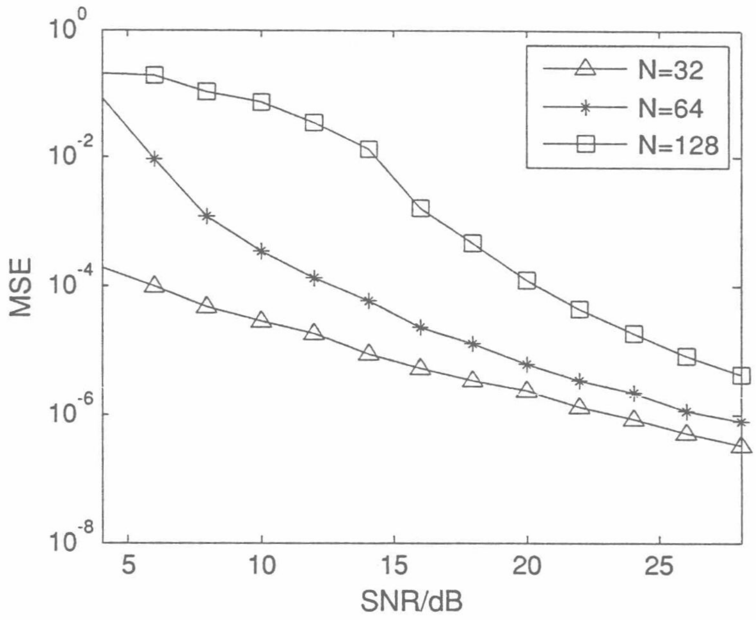

Fig. 6 CFO estimation performance with different subcarriers

5 Conclusions

In this paper, we address the problem of CFO estimation for OFDM communications systems with multiple antennas. We reconstruct the received signal to form data model with multiple-invariance property, so multiple-invariance ESPRIT algorithm for carrier frequency offset estimation is proposed in this paper. This algorithm has improved CFO estimation compared to ESPRIT method and ML method.

Acknowledgements This work is supported by China NSF Grants (60801052), Ph. D. programs foundation of ministry of education of China (No. 200802871056), the startup funding of Nanjing University of Aeronautics and Astronautics (S0583-041) and Jiangsu NSF Grants (BK2007192). The authors wish to thank the anonymous reviewers for their valuable suggestions on improving this paper.

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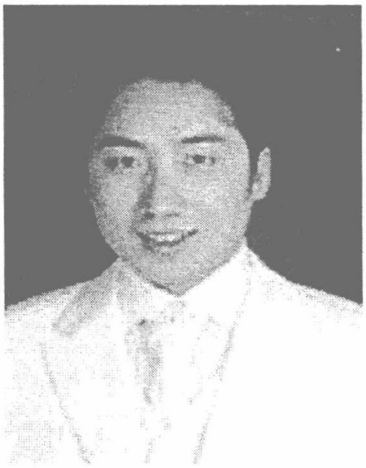
Author Biographies



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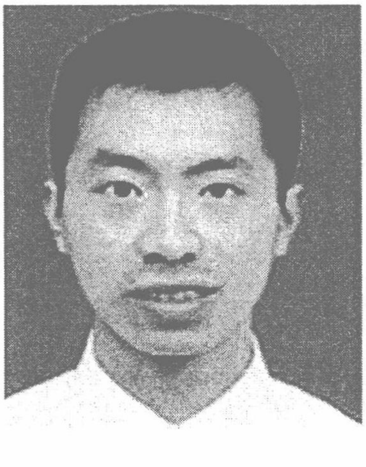

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