LETTER

Joint CFO and DOA Estimation Based on MVDR Criterion in Interleaved OFDMA/SDMA Uplink

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SUMMARY This letter deals with joint carrier frequency offset (CFO) and direction of arrival (DOA) estimation based on the minimum variance distortionless response (MVDR) criterion for interleaved orthogonal frequency division multiple access (OFDMA)/space division multiple access (SDMA) uplink systems. In order to reduce the computational load of two-dimensional searching based methods, the proposed method includes only once polynomial CFO rooting and does not require DOA paring, hence it raises the searching efficiency. Several simulation results are provided to illustrate the effectiveness of the proposed method.

key words: *carrier frequency offset, direction of arrival, MVDR, interleaved OFDMA/SDMA*

1. Introduction

As technology advances, the interleaved orthogonal frequency division multiple access/space division multiple access (OFDMA/SDMA) has drawn substantial research interests, and it has shown great gains in capacity and spectral efficiency. However, the multiuser frequency offset problem must be solved first and can be rather challenging due to the coexistence of multiple carrier frequency offsets (CFO) between the carrier frequency of the received signal of each user and the oscillator at base station (BS) [1]. Besides, high accuracy of direction of arrival (DOA) estimation is a key requirement for cellular networks. The interleaved OFDMA/SDMA can improve the spectral efficiency and provide high spatial diversity in communications [2], [3]. Hence, the CFO and DOA issues for interleaved OFDMA/SDMA uplink are more challenging and difficult.

Several methods have also been proposed to estimate CFO and DOA for OFDMA uplink systems. Among them, minimum variance distortionless response (MVDR) [4] and multiple signal classification (MUSIC) methods [5] are widely used two-dimensional (2D) searching methods. It is noted that the estimate accuracy of the 2D-MUSIC and 2D-MVDR strictly depends on the number of search grids used during the peaks searching process, which is time consuming and the required number of search grids is not clear to determine. The study [6] uses two consecutive one-dimensional (1D) estimation of signal parameters via rotational invariance techniques (ESPRIT) [7] methods in conjunction with the beamforming process to estimate the DOA and CFO of the multiuser signals. A Hierarchical-ESPRIT is proposed in [8] for joint CFO and DOA estimation. These parameters are estimated in a hierarchical tree structure, meanwhile, performing eigenvalue decomposition (EVD) is required. A reduced-dimension Capon (RD-Capon) method is proposed in [9] for direction of departure and DOA estimation under a bistatic MIMO radar. It also can be applied for CFO and DOA estimation, which can reduce the computational cost by replacing the 2D searching with one-dimensional search. However, it also requires a one-dimensional global spectrum searching for CFO estimation, the complexity and estimation accuracy strictly depend on the grid size used during the search. It is time consuming and the search grid is not clear.

In order to obtain more accuracy CFO and DOA estimation, this letter presents an efficient method based on MVDR criterion without global spectrum searching for an interleaved OFDMA/SDMA uplink system. Although it is similar to that the method which described in [9]. The proposed method makes use of the signal structure by estimating the CFO and DOA. Of particular importance, it only requires once polynomial CFO rooting procedure with the averaging diagonal submatrices of correlation matrix to reduce the estimation error caused by small data blocks, hence it raises the estimate accuracy. Since it computes the roots directly from the MVDR cost function, the resulting solution has large CFO estimate error at low signal-to-noise ratio (SNR). In the DOA estimation, via Lagrange multiplier optimization, the auto-paired array steering vectors of users can be estimated with closed-form expressions. In order to mitigate the DOA wrapping problem, this letter also merges into a phase unwrapping procedure to correct phase angle to improve DOA estimate. Meanwhile, it does not require DOA paring, hence it also raises the searching efficiency. Finally, the simulation results are included to demonstrate the effectiveness of the proposed method.

2. Background

2.1 Signal Model

Consider the interleaved uplink of an OFDMA/SDMA system, where the BS is equipped with M uniformly spaced antennas. Assume that there are N subcarriers which are divided into Q subchannels with each subchannel having $P = N/Q$ subcarriers and the subcarriers assigned to differ-

Manuscript received July 15, 2023.
Manuscript revised September 27, 2023.
Manuscript publicized October 26, 2023.

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DOI: 10.1587/transfun.2023EAL2064

ent users are interleaved over the whole bandwidth. The kth user is assigned to the q_k subchannel which is a subset of P subcarriers with the index set $\{q_k, Q+q_k, \cdots, (P-1)Q+q_k\}$. After passing through the multipath channel and removing the cyclic prefix, the received signal from the kth user is $r_k(n) = \sum_{p=0}^{P-1} X_k(p)H_k(p)e^{j2\pi n(pQ+q_k+\varepsilon_k)/N}$, where $0 \le n \le N-1$ and $\varepsilon_k \in (-0.5, 0.5)$ denotes the kth user's CFO normalized by subcarrier spacing $2\pi/N$. $X_k(p)$ is a set of P data streams of the kth user and $H_k(p)$ represents the samples from $\bar{H}_k(\delta)$ at $\delta = pQ + q_k$. The channel frequency response of the δth subcarrier for the kth user is $\bar{H}_k(\delta) = \sum_{l=0}^{L-1} h_{k,l}e^{-j2\pi l\delta/N}$, where L is the channel length. The channel taps $h_{k,l}$ can be expressed as $h_{k,l} = \beta_{k,l}e^{-j2\pi d\sin(\theta_{k,l})}$ and $\beta_{k,l}$ is statistically independent Gaussian random variables with zero mean and an exponentially decaying power profile [1], $E\{\beta_{k,l}^2\} = \alpha_l e^{-l/5}$, $0 \le l \le L-1$, where α_l is the normalized factor used to set the channel power to unity [10]. $d = 0.5\lambda$ is the distance between two adjacent antennas and λ is the wavelength of the impinging signal. Due to the multipath propagation near the kth user, its contribution to the array is modeled as a superposition of L local scattered rays [11]. We assume that the incident angles of the kth user is $\theta_{k,l}$, which can be distributed uniformly within a certain DOA region [12], i.e., $\theta_{k,l} \in (\theta_k - \Delta\theta_{k,l}, \theta_k + \Delta\theta_{k,l})$, where θ_k and $\Delta\theta_{k,l}$ represent the line of sight angle of the kth user and the angular spread due to local scattering, respectively. The received signal is $y(n) = \sum_{k=1}^{K} r_k(n) + z(n)$ at the reference antenna element, where $r_k(n) = \sum_{p=0}^{P-1} X_k(p)H_k(p)e^{j2\pi np/P}e^{j2\pi n\omega_k/P}$, $\omega_k = (q_k + \varepsilon_k)/Q$ denotes the effective CFO of the kth user, and $z(n)$ is the additive white Gaussian noise with zero mean and variance σ_n^2. It is noted that the received signal set has a special periodic feature with every P samples [1], i.e., $r_k(n+vP) = e^{j2\pi v\omega_k}r_k(n)$, where v $(0 \le v \le Q-1)$ is an integer.

In one OFDMA block at the mth antenna, $\{y_m(n)\}_{n=0}^{N-1}$ can be arranged in to a $Q \times P$ matrix $\bar{\mathbf{Y}}_m$ as follows

$$\bar{\mathbf{Y}}_m = \bar{\mathbf{C}}_m(\omega,\theta)\mathbf{S}_f + \bar{\mathbf{Z}}_m, m = 1, 2, \cdots, M \tag{1}$$

where $\bar{\mathbf{C}}_m(\omega,\theta) = [\bar{\mathbf{c}}_m(\omega_1,\theta_1), \bar{\mathbf{c}}_m(\omega_2,\theta_2), \cdots, \bar{\mathbf{c}}_m(\omega_K,\theta_K)]$ is an $Q \times K$ matrix with $\bar{\mathbf{c}}_m(\omega_k,\theta_k) = \mathbf{b}(\omega_k)a_m(\theta_k), k = 1, 2, \cdots, K$. $\mathbf{b}(\omega_k) = [1, e^{j2\pi\omega_k}, \cdots, e^{j2\pi(Q-1)\omega_k}]^T$ is the CFO vector and $a_m(\theta_k) = \sum_{l=0}^{L-1} e^{-j2\pi d\sin(\theta_{k,l})}$. $\mathbf{S}_f = \mathbf{D} \odot \mathbf{S}$, $(\bullet)^T$ is the transpose operation, and $\odot$ denotes the Hadamard product. $\mathbf{D} = [\mathbf{d}_1, \mathbf{d}_2, \ldots, \mathbf{d}_K]^T$ with $\mathbf{d}_k = [1, e^{j2\pi\omega_k/P}, \cdots, e^{j2\pi(P-1)\omega_k/P}]^T$ and $\mathbf{S} = [\mathbf{s}_1, \mathbf{s}_2, \cdots, \mathbf{s}_K]^T$ with $\mathbf{s}_k = [X_k(0)H_k(0), X_k(1)H_k(1), \cdots, X_k(P-1)H_k(P-1)]^T$. $\bar{Z}_m$ is the $Q \times P$ noise matrix. Then, the uth data blocks can be rearranged as an $QM \times P$ matrix and is given by

$$\begin{aligned}\mathbf{Y}(u) &= [\bar{\mathbf{Y}}_1^T(u), \bar{\mathbf{Y}}_2^T(u), \cdots, \bar{\mathbf{Y}}_M^T(u)]^T \\ &= \mathbf{C}(\omega,\theta)\mathbf{S}_f(u) + \mathbf{Z}(u)\end{aligned} \tag{2}$$

where $\mathbf{C}(\omega,\theta) = [\mathbf{c}(\omega_1,\theta_1), \mathbf{c}(\omega_2,\theta_2), \cdots, \mathbf{c}(\omega_K,\theta_K)]$ is the $QM \times K$ matrix with $\mathbf{c}(\omega_K,\theta_K) = \mathbf{a}(\theta_k) \otimes \mathbf{b}(\omega_k), k = 1, 2, \cdots, K$. $\mathbf{a}(\theta_k) = [a_1(\theta_k), a_2(\theta_k), \cdots, a_M(\theta_K)]^T$ is the steering vector, $\mathbf{Z}(u) = [\bar{\mathbf{Z}}_1^T(u), \bar{\mathbf{Z}}_2^T(u), \cdots, \bar{\mathbf{Z}}_M^T(u)]^T$, and $\otimes$ denotes the Kronecker product. The ensemble correlation matrix of (2) is $\mathbf{R} = E\{\mathbf{Y}(u)\mathbf{Y}^H(u)\} = \mathbf{C}(\omega,\theta)E\{\mathbf{S}_f(u)\mathbf{S}_f^H(u)\}\mathbf{C}^H(\omega,\theta) + \sigma_n^2\mathbf{I}_{QM}$, where $\mathbf{I}_{QM}$ is the $QM \times QM$ identity matrix, $E\{\bullet\}$ is the expectation operator, and $(\bullet)^H$ represents the conjugate transpose. The $QM \times QM$ correlation matrix of $\mathbf{Y}(u)$ under U blocks at the antenna array is $\hat{\mathbf{R}} = (1/UP)\sum_{u=1}^{U} \mathbf{Y}(u)\mathbf{Y}^H(u)$.

2.2 2D-MVDR and 2D-MUSIC Methods

According to (2), the conventional 2D-MVDR [4] and the 2D-MUSIC [5] can be invoked to estimate CFO and DOA. The 2D-MVDR pseudo-spectrum is defined by

$$F_{\text{2D-MVDR}}(\omega,\theta) = [\mathbf{c}^H(\omega,\theta)\hat{\mathbf{R}}^{-1}\mathbf{c}(\omega,\theta)]^{-1} \tag{3}$$

where $(\bullet)^{-1}$ represents the matrix inversion operator. The CFO and DOA are estimated by detecting the K largest peaks in the spectrum. Also, the search ranges of ω and θ are $[-0.5/Q, (0.5+Q)/Q]$ and $[-90^\circ, 90^\circ]$, respectively.

Next, the 2D-MUSIC estimates noise subspace $\hat{\mathbf{U}}_n$ using EVD of $\hat{\mathbf{R}}$ and then the estimates of CFO and DOA are taken as those $\{\omega,\theta\}$ that give the smallest value of $\mathbf{C}^H(\omega,\theta)\hat{\mathbf{U}}_n = \mathbf{0}$. These values of $\{\omega,\theta\}$ result in a vector $\mathbf{c}(\omega,\theta)$ farthest away from the noise subspace and as orthogonal to the noise subspace as possible. This is done by finding the K peaks in the 2D-MUSIC pseudo-spectrum defined by

$$F_{\text{2D-MUSIC}}(\omega,\theta) = [\mathbf{c}^H(\omega,\theta)\hat{\mathbf{U}}_n\hat{\mathbf{U}}_n^H\mathbf{c}(\omega,\theta)]^{-1} \tag{4}$$

Here we have the K largest peaks of (4) taken as the estimates of CFO and DOA for the users.

3. Proposed Method

Although one can utilize 2D spectrum searching scheme to minimize the function (3) and (4), it is of computational inefficiency. In order to reduce the computational complexity, we herein adopt a MVDR criterion that makes use of polynomial rooting. First, the matrix $\hat{\mathbf{R}} \in \mathbb{C}^{QM \times QM}$ can be partitioned into $M \times M$ block matrices $[\hat{\mathbf{R}}_{i,h}]_{i=1,h=1}^{M}$ and the dimension of submatrix $\hat{\mathbf{R}}_{i,h}$ is $Q \times Q$ for $i, h = 1, 2, \cdots, M$. It is noted that the diagonal submatrices $\{\hat{\mathbf{R}}_{m,m}\}_{m=1}^{M}$ only contain CFO information. We further take advantage of the submatrices of $\hat{\mathbf{R}}$, only the diagonal submatrices $\{\hat{\mathbf{R}}_{m,m}\}_{m=1}^{M}$ are utilized. However, to reduce the estimation error caused by noise, the following averaging procedure is performed

$$\bar{\mathbf{R}} = \frac{1}{M}\sum_{m=1}^{M}\hat{\mathbf{R}}_{m,m} \simeq \mathbf{B}(\omega)E\{\mathbf{S}_f\mathbf{S}_f^H\}\mathbf{B}^H(\omega) + \sigma_n^2\mathbf{I}_Q \tag{5}$$

from which the CFO can be calculated from each submatrix $\hat{\mathbf{R}}_{m,m}$. $\mathbf{I}_Q$ denotes $Q \times Q$ identity matrix and $\mathbf{B}(\omega) = [\mathbf{b}(\omega_1), \mathbf{b}(\omega_2), \cdots, \mathbf{b}(\omega_K)]$. The cost function of CFO estimate for 1D-MVDR is $F(\omega) = [\mathbf{b}^H(\omega)\bar{\mathbf{R}}^{-1}\mathbf{b}(\omega)]^{-1}$. When $\mathbf{b}(\omega)$ is orthogonal to $\bar{\mathbf{R}}$, the denominator of $F(\omega)$ is zero and $F(\omega)$ trends to be infinite. For $F(\omega)$, we can redefine a new cost function as $G(\omega) = \left|\mathbf{b}^H(\omega)\bar{\mathbf{R}}^{-1}\mathbf{b}(\omega)\right|$, which can be viewed as the null spectrum of 1D-MVDR. It is interested to find a polynomial whose roots correspond to the minimum of $G(\omega)$. Let $z = e^{j2\pi\omega}$ and $\mathbf{b}(z) = [1, z, \cdots, z^{Q-1}]^T$, then $G(\omega)$ can be rewritten as

$$G(z) = \left|\mathbf{b}^T(z^{-1})\bar{\mathbf{R}}^{-1}\mathbf{b}(z)\right| \tag{6}$$

The nulls of $G(\omega)$ are due to the root of $G(z)$ that lie on the unit circle. The 1D-MVDR constructs $G(z)$ using the appropriate $\bar{\mathbf{R}}^{-1}$, and computes the K roots inside and closest to the unit circle. Then,

$$\hat{\omega}_k = angle\{\hat{z}_k\}/2\pi, k = 1, 2, \cdots, K \tag{7}$$

As shown in [9], the DOA estimation can be achieved by the following optimization problems

$$\min_{\theta}\ \mathbf{a}^H(\theta)\bar{\mathbf{E}}(\hat{\omega})\mathbf{a}(\theta), s.t.\ \mathbf{e}_l^T\mathbf{a}(\theta) = 1 \tag{8}$$

where $\bar{\mathbf{E}}(\hat{\omega}) = [\mathbf{I}_M \otimes \mathbf{b}(\hat{\omega})]^H\hat{\mathbf{R}}^{-1}[\mathbf{I}_M \otimes \mathbf{b}(\hat{\omega})]$ and $\mathbf{e}_1 = [1, 0, \cdots, 0]^T$. Due to we have acquired the estimated $\{\hat{\omega}_k\}_{k=1}^K$ information, if we take them into (8), then auto-paired steering vectors can be easily obtained by means of Lagrange multiplier method, i.e.,

$$\hat{\mathbf{a}}(\theta_k) = \frac{\bar{\mathbf{E}}^{-1}(\hat{\omega}_k)\mathbf{e}_1}{\mathbf{e}_1^T\bar{\mathbf{E}}^{-1}(\hat{\omega}_k)\mathbf{e}_1}, k = 1, 2, \cdots, K \tag{9}$$

and $\{\hat{\theta}_k\}_{k=1}^K$ information can be drawn from that steering vector via least squares (LS) principle. The detailed procedures are shown as below. Let $\mathbf{g}_k^{wrap} = angle\{\hat{\mathbf{a}}(\theta_k)\}$ be the phase of normalized $\hat{\mathbf{a}}(\theta_k)$, where $angle\{\bullet\}$ is to get the phase angles for each element of complex array. When $d \le 0.5\lambda$, the unwrapped phases are given by

$$\begin{cases} \bar{\mathbf{g}}_k^{unwrap}(1) = \mathbf{g}_k^{wrap}(1), if\ m-1 = 0 \\ \bar{\mathbf{g}}_k^{unwrap}(m) = \bar{\mathbf{g}}_k^{unwrap}(m-1) + \boldsymbol{\delta}_k(m-1), if\ m-1 \neq 0 \end{cases} \tag{10}$$

where $\mathbf{g}_k^{wrap}(m)$ is the mth element of $\mathbf{g}_k^{wrap}$, $m = 1, 2, \cdots, M$ and $\boldsymbol{\delta}_k(m-1)$ is expressed as [13]

$$\boldsymbol{\delta}_k(m-1) = \begin{cases} \mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1), if\ \left|\mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1)\right| \le \pi \\ \mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1) - 2\pi, if\ \mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1) > \pi \\ \mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1) + 2\pi, if\ \mathbf{g}_k^{wrap}(m) - \mathbf{g}_k^{wrap}(m-1) < -\pi \end{cases} \tag{11}$$

The unwrapped phase of steering vector is $\mathbf{g}_k^{unwrap} = \mathbf{p}\sin\theta_k$, where $\mathbf{p} = [0, -2\pi\Delta/\lambda, \cdots, -2\pi\Delta(M-1)/\lambda]^T$. Then, the LS fitting is $\mathbf{p}v_k = \mathbf{g}_k^{unwrap}$. And, the LS solution for $\hat{v}_k$ is given by

$$\hat{v}_k = (\mathbf{p}^T\mathbf{p})^{-1}\mathbf{p}^T\mathbf{g}_k^{unwrap} \tag{12}$$

Finally, the CFO and DOA are estimated via

$$\hat{\varepsilon}_k = \hat{\omega}_k Q - q_k, k = 1, 2, \cdots, K \tag{13}$$

$$\hat{\theta}_k = \sin^{-1}(\hat{v}_k), k = 1, 2, \cdots, K \tag{14}$$

In summary, the detailed steps of the proposed method are summarized as follows:

Step 1. Calculate $\hat{\mathbf{R}}$.

Step 2. Construct $\bar{\mathbf{R}}$ by (5) and calculate $\bar{\mathbf{R}}^{-1}$.

Step 3. Construct a polynomial $G(z)$ by (6). Then, rooting the polynomial and using (7) to get $\hat{\omega}_k, k = 1, 2, \cdots, K$.

Step 4. Calculate $\hat{\mathbf{R}}^{-1}$ and substitute into $\bar{\mathbf{E}}(\hat{\omega})$ of (8).

Step 5. Using (10) to obtain $\bar{\mathbf{g}}_k^{unwrap}$ for each user. Then execute (12) to obtain $\hat{g}_k, k = 1, 2, \cdots, K$.

Step 6. Converting $(\hat{\omega}_k, \hat{v}_k)$ to $(\hat{\varepsilon}_k, \hat{\theta}_k)$ using (13) and (14).

4. Computational Complexity Analysis

In this section, the number of complexity multiplications (CM) of the 2D-MVDR [4], 2D-MUSIC [5], Hierarchical-ESPRIT [8], RD-Capon [9], and the proposed method are evaluated. The proposed advantage is the reduction of computational complexity by reducing the number of the search while maintaining comparable performance. Assuming K users, M antennas, Q subchannels, U OFDMA data blocks, and P subcarriers for each test, the computational complexities of calculating $\hat{\mathbf{R}}$, $\hat{\mathbf{R}}^{-1}$, EVD, and finding root for an $QM \times QM$ correlation matrix require approximately $(QM)^2PU$, $2(QM)^3$, $12(QM)^3$, and $8(QM)^3$ CM [14], respectively. For 2D-MUSIC and 2D-MVDR, let the searching numbers of CFO and DOA are denoted as F_ω and F_θ, respectively, which depend on the searching grid size within search range $\{\omega, \theta\}$. In addition, the searching number of CFO for RD-Capon is also denoted as F_ω. Briefly, the required CM are listed in Table 1.

5. Simulation Results

The simulation results were used to compare the performance of the 2D-MVDR [4], 2D-MUSIC [5], Hierarchical-ESPRIT [8], RD-Capon [9], and the proposed method in this section. Consider a uniform linear array with half-wavelength spacing at BS. Unless stated otherwise, the number of antennas and the number of users on the BS are $M = 12$ and $K = 8$, respectively. The total number of subcarriers in the OFDMA/SDMA system is $N = 1024$, which is divided into $Q = 32$ subchannels and each user is allocated to $P = 32$ subcarriers. The CFO and DOA of the eight users are denoted as $\varepsilon = \{-0.4325, -0.3461, -0.2812, -0.1294, 0.1851, 0.2319, 0.3256, 0.4131\}$ and $\theta = \{-50.23^\circ, -35.41^\circ, -20.91^\circ, -5.64^\circ,$

Table 1 Computational complexity analysis.

Estimators	Computational complexities
Hierarchical-ESPRIT	$12Q^3 + Q^2MPU + 14K^3 + 3(Q-1)K^2 + K[12M^3 + M^2QPU + 3(M-1) + 14] + K[12Q^3 + Q^2MPU + 3(Q-1) + 14]$
2D-MVDR	$2(QM)^3 + (QM)^2PU + F_\omega F_\theta[(QM)^2 + QM]$
2D-MUSIC	$12(QM)^3 + (QM)^2(QM-K) + (QM)^2PU + F_\omega F_\theta[(QM)^2 + QM]$
RD-Capon	$2(QM)^3 + (QM)^2PU + F_\omega[(QM)^2M + (QM)M^2 + 2M^3] + K[(QM)^2M + (QM)M^2 + 2M^3 + 10M + 16]$
Proposed	$2(QM)^3 + (QM)^2PU + 10Q^3 + K[(QM)^2M + (QM)M^2 + 2M^3 + 3M + 2]$

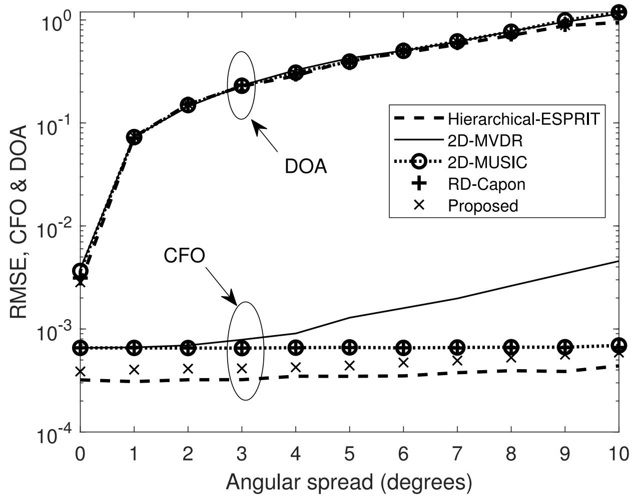

Fig. 1 RMSE of CFO and DOA estimation versus angular spread.

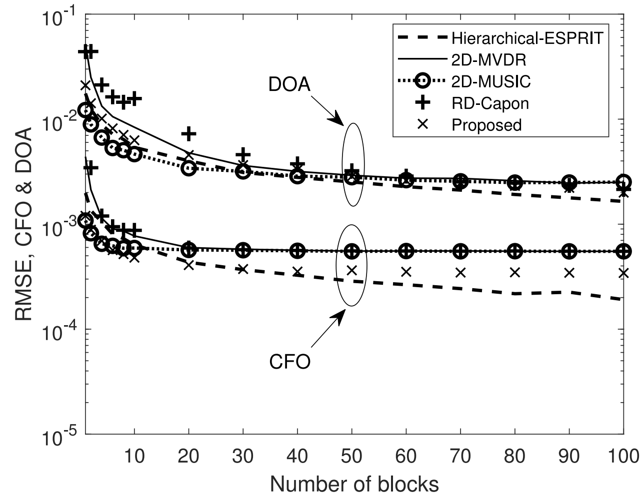

Fig. 2 RMSE of CFO and DOA estimation versus number of blocks.

$10.52^\circ, 25.99^\circ, 40.37^\circ, 55.19^\circ\}$, respectively. It is assumed that $\Delta\theta_{k,l}$ is uniformly distributed over the angular spread interval $[-\Delta\theta_k, \Delta\theta_k]$. All OFDMA signals were generated with binary phase-shift-keying modulation and the average received signal power from all users is the same. Each user transmits signal to BS through independent multipath channel and $L = 10$. The CFO searching grid of 2D-MVDR, 2D-MUSIC, and RD-Capon is set to $\mu_\omega = 10^{-4}$; their DOA searching grid is set to $\mu_\theta = 10^{-2}$. The input SNR of the kth user is defined as $\mathrm{SNR} = 10\log E\{r_k^2(n)\}/\sigma_n^2$. As indices of evaluation, the root mean square error (RMSE) of CFO and DOA are defined as $\frac{1}{\Pi K}[\sum_{\rho=1}^{\Pi}\sum_{k=1}^{K}(\varepsilon_k - \hat{\varepsilon}_k^{\rho})^2]^{0.5}$ and $\frac{1}{\Pi K}[\sum_{\rho=1}^{\Pi}\sum_{k=1}^{K}(\theta_k - \hat{\theta}_k^{\rho})^2]^{0.5}$ with $\Pi = 100$ is the number of runs, respectively.

Figure 1 shows RMSE of CFO and DOA versus angular spread $\Delta\theta_k$ under SNR = 10dB and $U = 50$. Local scattering may be viewed as a form of model error, which results the DOA estimation performance of all methods gradually decreases when faced with local scatters. It is noted that these methods are not specifically designed to estimate DOA in local scattering environments. However, the BS is deployed above the surrounding scatters (several tens of meters high) in a macrocell environment. Hence, the received signals at the BS come from the scattering process in the vicinity of the mobile station. The multipath components at the BS are thus restricted to a small angular range [15]. Particularly, in typical urban and rural area propagation environment, the impinging angle of received waves can be approximated by a single main DOA [16]. In following simulations, we assume that the angular spread of each user is set as $\Delta\theta_k = 0^\circ$ in order to facilitate comparison of the DOA estimation. Figure 2 shows RMSE of CFO and DOA versus the number of blocks under SNR = 10 dB. We first observe the CFO estimation performance. When the block number $U < 10$, however, the proposed method has similar estimation performance as the 2D-MUSIC and has better estimation than the 2D-MVDR, RD-Capon, and Hierarchical-ESPRIT. As U becomes larger, the proposed method provides superior performance compared to the 2D-MVDR, 2D-MUSIC, RD-Capon. Since the proposed method is not limited to the searching grid, it can provide excellent resolution capabilities in CFO and DOA estimation under the whole search range. In general, since the Hierarchical-ESPRIT is one of the signal subspace processing methods and is not limited by the size of the search grid, its estimation performance is slightly better than that of the proposed method. Clearly, this figure shows that the convergence speeds of all methods have the same fast and stable under different parameters. Figure 3 presents RMSE of CFO and DOA estimation versus input SNR under $U = 50$. In particular, the proposed method provides superior CFO estimation performance compared to the other methods under SNR $\geq$ 15 dB as shown in Fig. 3. For CFO estimation, the resolution threshold performance of a root-MVDR in the proposed method is better than the searching-based MVDR, but for the noises dominant situation the appearance of both methods' threshold performance is opposite. It is noted that in the low SNR environments, to find out the roots'

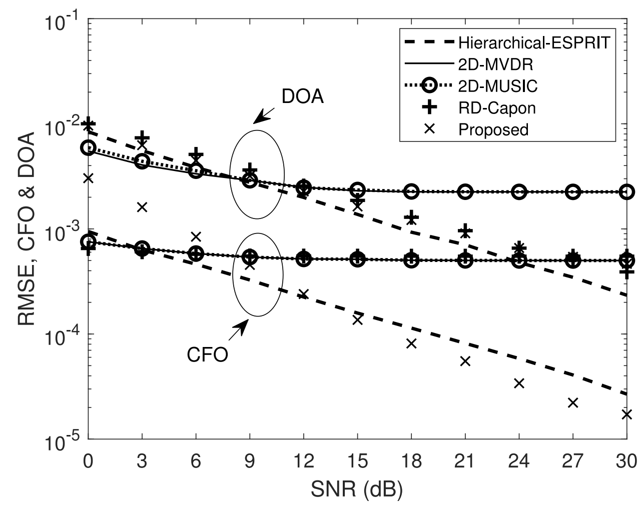

Fig. 3 RMSE of CFO and DOA estimation versus SNR.

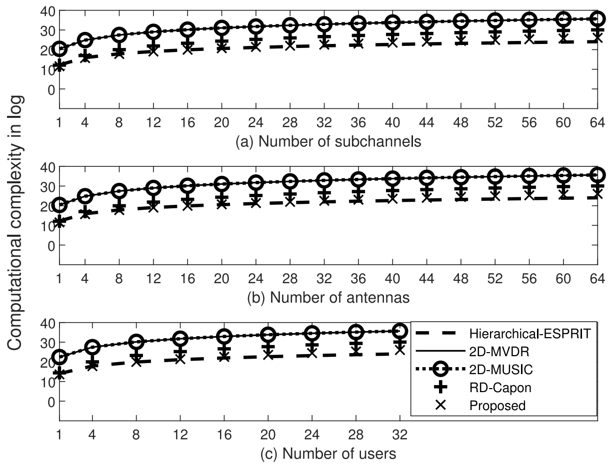

Fig. 4 Computational complexity versus (a) Q (b) M (c) K.

precise location becomes more ambiguously for CFO estimation by using the root-MVDR approach in the proposed method. Observe that the proposed method has acceptable DOA estimate performance and has very close DOA estimate performance to the Hierarchical-ESPRIT. Figure 4 shows the computational complexity comparison of the methods. For CFO estimation, the searching grid size of 2D-MVDR, 2D-MUSIC, and RD-Capon is set to $\mu_\omega = 10^{-4}$, then estimating CFO requires calculation of $F_\omega = [(1 + Q^{-1})/\mu_\omega] + 1$ times. For DOA estimation, the searching grid size of 2D-MVDR and 2D-MUSIC is set to $\mu_\theta = 10^{-2}$, then estimating DOA require calculations of $F_\theta = (180/\mu_\theta) + 1$ times. Assume that $N = 1024$, $P = N/Q$, and $U = 100$ in Table 1. For the sake of convenience, let $Q = M$ and $K = 0.5Q$. Figure 4(a) shows number of CM in logarithmic scale versus number of subchannels Q. Figure 4(b) shows number of CM in logarithmic scale versus number of antennas M. Figure 4(c) depicts number of CM in logarithmic scale versus number of users K. It can be seen from these figures that the proposed method has significantly reducing the computational complexity compared to the 2D-MVDR, 2D-MUSIC, and RD-Capon.

6. Conclusion

A joint CFO and DOA estimation method is presented in the interleaved OFDMA/SDMA uplink. Of particular importance, it does not perform EVD to obtain subspace and offer high searching efficiency than the 2D-MVDR, 2D-MUSIC, and RD-Capon, while maintaining comparable estimate performance with the Hierarchical-ESPRIT. Simulation results have demonstrated the effectiveness of the proposed method.

Acknowledgments

The authors wish to thank the anonymous reviewers for their helpful reviews and suggestions.

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