CFO-RESISTANT RECEIVER FOR ASYNCHRONOUS MC-DS-CDMA SYSTEMS

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ABSTRACT

In this paper we propose a new receiving algorithm for MC-DS-CDMA systems when carrier frequency offset (CFO) is significant. By exploiting the special structure of the CFO contaminated signals, the new algorithm cancels CFO completely during the despreading procedure, after which the despreaded CFO-free signal is demodulated via normal FFT-based OFDM demodulator. Guaranteeing complete CFO cancellation, this method is advantageous over the majority existing CFO-mitigation techniques that can only mitigate but not completely remove CFO. An efficient algorithm is developed, and simulations are conducted to demonstrate the performance.

***Index Terms*—** MC-DS-CDMA, carrier frequency offset, multiple access, synchronization, broadband communications

1. INTRODUCTION

Multi-carrier CDMA, which combines the OFDM-based multi-carrier transmissions and CDMA-based multi-user access, is a promising technique for future 4G broadband multi-user communication systems. The application of OFDM greatly resolves the difficulty raised by multi-path fading that is specially severe for broadband systems. The application of CDMA simplifies the multi-access and synchronization design, especially in the uplink.

There have been many different types of multi-carrier CDMA systems proposed [1]. One of them is MC-CDMA where each data symbol is spreaded into a chip sequence which is modulated onto different OFDM sub-carriers, i.e., different chip on different sub-carrier. Another major type of multi-carrier CDMA systems is the MC-DS-CDMA [2], where each OFDM-block (after IFFT and cyclic prefix) is block-wise spreaded, i.e., the OFDM-block is spreaded into multiple OFDM blocks, each multiplied with a different chip of the spreading code. A major feature of MC-DS-CDMA system is that each OFDM sub-carrier works like DS-CDMA. Specifically, if there is only one sub-carrier, then the MC-DS-CDMA reduces to a conventional DS-CDMA. One of the major advantages of MC-DS-CDMA is that each DS-CDMA signal (in each sub-carrier) of a user can be maintained orthogonal to that of all the other users, when orthogonal spreading codes are used. As a result, multi-access interference (MAI) is mostly avoided, which may greatly boost performance over conventional DS-CDMA.

Nevertheless, a major problem for MC-DS-CDMA (and in fact for all multi-carrier systems) is the loss of carrier frequency synchronization, or the residual carrier frequency offset (CFO) [3]. While in conventional DS-CDMA, CFO only makes channels time-varying, which can be conveniently dealt with by adaptive channel estimation and equalization techniques, the CFO in MC-DS-CDMA not only makes channels time-varying, but also destroys the orthogonality among sub-carriers, which causes inter-carrier interference (ICI) and multi-access interference (MAI) [3,4]. Even if orthogonal spreading codes are used, MC-DS-CDMA with CFO still suffers from MAI and ICI. As a result, MC-DS-CDMA is very sensitive to CFO, and is susceptible to CFO even more than conventional OFDM because multiple users are involved.

In MC-DS-CDMA systems, each multi-access user may suffer from a difference CFO, which makes it difficult for the receiver to remove all the CFO at the same time. Existing multi-carrier systems usually have to assume that carrier frequency can be perfectly synchronized [5,6], or at least CFO can be made small enough such that certain level of CFO mitigation is enough [7].

Note that most of the existing CFO mitigation techniques can only mitigate, not completely cancel, the CFO. As far as we know, very few can promise complete CFO cancellation by the receiver only in multi-user environment [8]. In multi-user systems, especially when the number of users is not so small, even slight CFO for each user, if left un-canceled, can be aggregated together to cause severe performance degradation. Hence approximate synchronization or mitigation may not be sufficient. Existing techniques may only be useful when the CFO is extremely small or the number of users is less enough.

In this paper, we propose a new receiving algorithm for MC-DS-CDMA systems, which exploits the special structure of CFO-contaminated signal to guarantee complete CFO cancellation. The CFO is cancelled during despreading, and thanks to the spreading codes, this procedure does not enhance noise very much. It can be implemented in a computationally efficient manner.

This paper is organized as follows. In Section 2, we setup the MC-DS-CDMA system model. In Section 3, we develop the new algorithm. Then simulations are conducted in Section 4, whereas conclusions are made in Section 5.

2. MC-DS-CDMA SYSTEM MODEL

We consider the uplink of an MC-DS-CDMA system, where I users transmit to a base station. Each user, e.g., the user i, $0 \leq i \leq I-1$, has a spreading code $c_{i,g}$, where $g = 0, 1, \cdots, G-1$. Note that the spreading codes can be periodic or aperiodic. Because MC-DS-CDMA systems can use orthogonal codes such as the Walsh-Hadamard codes, we assume the code $c_{i,g}$ is periodic with period G, and G is the processing gain.

To simplify notation, we consider one symbol block only, i.e., every user transmits one symbol block in a frame. Specifically, the user i has a data block $\mathbf{b}_i = [b_i(0), \cdots, b_i(N-1)]^T$ for trans-

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mission, where N is the OFDM block length, or the FFT length. Conventional OFDM systems just use IFFT to process $\mathbf{b}_i$, and add cyclic prefix (CP) before transmission. In MC-DS-CDMA systems, the symbol block $\mathbf{b}_i$ is first (block-wise) spreaded into G blocks by $\{c_{i,g}\}$, which can be denoted as $\{\mathbf{b}_i c_{i,0}, \cdots, \mathbf{b}_i c_{i,(G-1)}\}$. Then each of the G blocks $\mathbf{b}_i c_{i,g}$ is OFDM modulated (performing N-point IFFT and adding CP with length M) and transmitted.

For the i^{th} user, the g^{th} block of the transmitted signal with CP can be express as

$$s_{i,g}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} b_i(k) c_{i,g} e^{j2\pi nk/N}, \tag{1}$$

where $n = 0, \cdots, N+M-1$. Because each block has a multiplication factor $c_{i,g}$ only, we can define

$$s_i(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} b_i(k) e^{j2\pi nk/N} \tag{2}$$

so that $s_{i,g}(n) = s_i(n) c_{i,g}$. The entire transmission frame (with G blocks) has a structure shown in Fig. 1, where a total of $G(N+M)$ samples are transmitted for N information symbols.

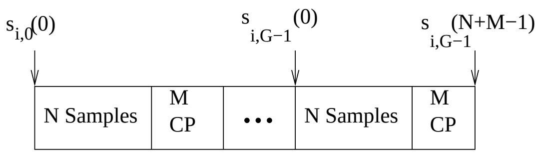

Fig. 1. Structure of MC-DS-CDMA transmission frame for a user, where one block of N information symbols is spreaded into G blocks, and each block is OFDM modulated.

Because each of the G blocks does not interference with the other blocks (thanks to the CP), we can consider in general the g^{th} block only for notational simplicity. The received signal from the i^{th} user, $r_{i,g}(n)$, can be described as the linear convolution of the channel $h_i(l)$ with $s_{i,g}(n)$,

$$r_{i,g}(n) = \sum_{\ell=0}^{L} h_i(\ell) s_{i,g}(n-\ell). \tag{3}$$

Without loss of generality, we assume all channels have order L. The overall signal received by the receiver, with delay d_i, CFO ϵ_i and initial phase ϕ_i taken into consideration, becomes

$$r_g(n) = \sum_{i=0}^{I-1} r_{i,g}(n-d_i) e^{j(\epsilon_i n + \phi_i)} + v_g(n), \tag{4}$$

where $v_g(n)$ is AWGN with zero-mean and variance σ_v^2. Note that the length of CP should satisfy $M \geq L + \max_{0\leq i \leq I-1} d_i$.

As in conventional MC-DS-CDMA demodulator, for each block we remove CP and consider the samples $r_g(n), n = M, \cdots, N+M-1$, which can be put into a vector $\mathbf{r}(g) = [r_g(M), \cdots, r_g(N+M-1)]^T$. Then we have

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \begin{bmatrix} h_i(L) & \cdots & h_i(0) & & \\ & \ddots & & \ddots & \\ & & h_i(L) & \cdots & h_i(0) \end{bmatrix} \times \begin{bmatrix} s_i(M-d_i-L) \\ \vdots \\ s_i(N+M-d_i-1) \end{bmatrix} + \mathbf{v}(g), \tag{5}$$

where $\mathbf{E}_i(g) = e^{j[\epsilon_i(N+M)g+\phi_i]} c_{i,g} \mathrm{diag}\{e^{j\epsilon_i M}, \cdots, e^{j\epsilon_i(N+M-1)}\}$ is the $N \times N$ diagonal CFO matrix, and noise vector $\mathbf{v}(g) = [v_g(M), \cdots, v_g(N+M-1)]^T$. Because of the CP, we can rewrite (5) in matrix form as

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \tilde{\mathbf{H}}_i \mathbf{s}_i(d_i) + \mathbf{v}(g), \tag{6}$$

where the channel matrix $\tilde{\mathbf{H}}_i$ is an $N \times N$ circulant matrix, whose first row (row $k=0$) is $[h_i(0), \mathbf{0}_{N-L-1}, h_i(L), \cdots, h_i(1)]$, and the subsequent k^{th} row is a $(k-1)$-step right cyclic shift of the first row. For example, the second row ($k=1$) is $[h_i(1), h_i(0), \mathbf{0}_{N-L-1}, h_i(L), \cdots, h_i(2)]$. In (6), the symbol vector $\mathbf{s}_i(d_i) = [s_i(M-d_i), \cdots, s_i(N+M-d_i-1)]^T$.

Now consider the symbol vector $\mathbf{s}_i(d_i)$. We can substitute the last $M-d_i$ symbols with their equivalent symbols (because of the CP), i.e., replace $s_i(N+l)$ with $s_i(l)$, from which we can rewrite $\mathbf{s}_i(d_i)$ as $\mathbf{s}_i(d_i) = [s_i(M-d_i), \cdots, s_i(N-1), s_i(0), \cdots, s_i(M-d_i-1)]^T$. Then, by rearranging the order of the entries of $\mathbf{s}_i(d_i)$ and switching correspondingly the columns of $\tilde{\mathbf{H}}_i$, we can change (6) into

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \mathbf{H}_i \mathbf{s}_i + \mathbf{v}(g), \tag{7}$$

where $\mathbf{s}_i = [s_i(0), \cdots, s_i(N-1)]^T$ and $\mathbf{H}_i$ is an $N \times N$ circulant matrix

$$\mathbf{H}_i = \begin{bmatrix} \mathbf{0}_{M-d_i-L} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i-1} \\ \mathbf{0}_{M-d_i-L+1} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i-2} \\ \vdots & & \vdots \\ \mathbf{0}_{M-d_i-L-1} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i} \end{bmatrix}. \tag{8}$$

Note that the rows of (8) are the right cyclic shift of its first row. An important feature of the model (7) is that the delay d_i is contained in $\mathbf{H}_i$ only, whereas the CFO ϵ_i is contained in the diagonal CFO matrix $\mathbf{E}_i(g)$ only. Because the channel matrix $\mathbf{H}_i$ is independent of CFO, once CFO is mitigated, d_i will just introduce phase shifts to the frequency domain channels after FFT in OFDM demodulation, which is easy to deal with.

In ideal MC-DS-CDMA systems without CFO, the sample vectors (7) become $\mathbf{r}(g) = \sum_{i=0}^{I-1} e^{j\phi_i} c_{i,g} \mathbf{H}_i \mathbf{s}_i + \mathbf{v}(g)$ (which is the g^{th} block of the received signal). Then conventional demodulator performs FFT of $\mathbf{r}(g)$, which diagonalizes $\mathbf{H}_i$ into $\mathrm{diag}\{H_i(0), \cdots, H_i(N-1)\}$. The signals in the k^{th} sub-carrier becomes $w_{k,g} = \sum_{i=0}^{I-1} e^{j\phi_i} c_{i,g} H_i(k) b_i(k)$, based on which despreading is conducted to estimate symbol $\hat{b}_i(k) = \sum_{g=0}^{G-1} w_{k,g} c_{i,g}^* / [G e^{j\phi_i} |c_{i,g}|^2 H_i(k)]$. As can be seen, orthogonal spreading codes such as Walsh-Hadamard codes can be used because the code orthogonality is preserved even if there is delay mismatch.

However, if different user's signal suffers from different CFO ϵ_i, then there is no such easy way of demodulation. Specifically, the presence of $\mathbf{E}_i(g)$ prevents the diagonalization of $\mathbf{H}_i$ by conducting FFT on $\mathbf{r}(g)$. Therefore, we need to look for new ways to cancel the CFO matrices $\mathbf{E}_i(g)$.

3. MC-DS-CDMA RECEIVER WITH CFO CANCELLATION

In this section, we present a new receiving algorithm with the capability of complete CFO cancellation. We assume that the receiver knows the delay d_i, CFO ϵ_i, initial phase ϕ_i, which can be estimated conveniently by training or blindly [9].

3.1. Basic idea

From (7), we can see that the received sample vectors $\mathbf{r}(g)$, $0 \le g \le G-1$, are different in the CFO matrices $\mathbf{E}_i(g)$ only, but contain the same $\mathbf{H}_i$ and $\mathbf{s}_i$. This observation serves as our ground for removing CFO. Stacking together all available G vectors, we have

$$\begin{bmatrix} \mathbf{r}(0) \\ \vdots \\ \mathbf{r}(G-1) \end{bmatrix} = \sum_{i=0}^{I-1} \begin{bmatrix} \mathbf{E}_i(0) \\ \vdots \\ \mathbf{E}_i(G-1) \end{bmatrix} \mathbf{H}_i \mathbf{s}_i + \begin{bmatrix} \mathbf{v}(0) \\ \vdots \\ \mathbf{v}(G-1) \end{bmatrix}, \tag{9}$$

which for notational simplicity can be defined as

$$\mathbf{y} = \sum_{i=0}^{I-1} \mathbf{A}_i \mathbf{H}_i \mathbf{s}_i + \mathbf{u}. \tag{10}$$

Note that $\mathbf{y}$ has dimension $GN \times 1$, and $\mathbf{A}_i$ has dimension $GN \times N$.

Our basic idea of removing CFO is thus to design an $N \times GN$ CFO-cancellation matrix $\mathbf{X}_i$ for each user i, such that

$$\mathbf{X}_i \mathbf{A}_k = \delta_{i,k} \mathbf{I}_N = \begin{cases} \mathbf{I}_N, & \text{if } i = k \\ \mathbf{0}_{N\times N}, & \text{if } i \ne k \end{cases} \tag{11}$$

A direct solution to (11) is

$$\mathbf{X}_i = \left[\mathbf{0}_{N\times(i-1)N},\ \mathbf{I}_N,\ \mathbf{0}_{N\times(I-i)N}\right] \left[\mathbf{A}_0, \cdots, \mathbf{A}_{I-1}\right]^+, \tag{12}$$

where $(\cdot)^+$ denotes pseudo-inverse.

After obtaining the matrix $\mathbf{X}_i$, we can apply it on $\mathbf{y}$ to get $\mathbf{z}_i = \mathbf{X}_i \mathbf{y}$, where $i = 0, \cdots, I-1$. If (11) is perfectly satisfied, then $\mathbf{z}_i = \mathbf{H}_i \mathbf{s}_i + \mathbf{X}_i \mathbf{u}$. Because $\mathbf{H}_i$ is $N \times N$ circulant, FFT can be conducted on $\mathbf{z}_i$ to detect the signal $b_i(k)$ of each user i. Note that CFO is completely removed, and this approach has combined the CFO cancellation with despreading, which is performed before the FFT-based OFDM demodulation.

A potential problem is whether (11) can have exact solutions. Another problem is about the computational complexity of (12), which requires large matrix inversion. In the next section, we derive an algorithm that calculates the matrices $\mathbf{X}_i$ more efficiently, during which we also show that CFO is indeed completely removed.

3.2. Element-wise derivation of $\mathbf{X}_i$

Consider the matrices $\mathbf{X}_i$ and $\mathbf{E}_i(g)$ which have dimensions $N \times GN$ and $N \times N$, respectively. Define the m^{th} row of $\mathbf{X}_i$ as $[\mathbf{x}_i^m(0), \cdots, \mathbf{x}_i^m(G-1)]$, where each $\mathbf{x}_i^m(g)$ is a $1 \times N$ vector. We also define $[\mathbf{x}_i^m(g)]_\ell$ as the ℓ^{th} entry of the $1 \times N$ vector $\mathbf{x}_i^m(g)$. To satisfy (11), each row ℓ of $\mathbf{X}_i$ must satisfy

$$\sum_{g=0}^{G-1} \mathbf{x}_i^\ell(g) \mathbf{E}_k(g) = \delta_{i,k}\ \mathbf{e}_\ell^T \tag{13}$$

where

$$\mathbf{e}_\ell^T = [0, \cdots, 0, 1, 0 \cdots, 0]^T \tag{14}$$

is a unit vector with 1 in the ℓ^{th} entry, $0 \le \ell \le N-1$. Considering that all CFO matrices $\mathbf{E}_k(g)$ are diagonal, (13) can be changed to

$$\sum_{g=0}^{G-1} [\mathbf{x}_i^\ell(g)]_\ell [\mathbf{E}_k(g)]_{\ell,\ell} = \delta_{i,k} \tag{15}$$

where $[\mathbf{E}_k(g)]_{\ell,\ell}$ is the $(\ell,\ell)^{th}$ entry of the $N \times N$ diagonal matrix $\mathbf{E}_k(g)$,

$$[\mathbf{E}_k(g)]_{\ell,\ell} = e^{j(\epsilon_k M + \phi_k)} e^{j\epsilon_k(N+M)g} c_{i,g} e^{j\epsilon_k \ell}. \tag{16}$$

In addition, we need to have $\left[\mathbf{x}_i^\ell(g)\right]_m = 0$ for all $m \ne \ell$. Note that

Therefore, the ℓ^{th} row of the matrix $\mathbf{X}_i$ have only G non-zero entries, which must satisfy (15). Define these G entries as vector

$$\mathbf{f}_i(\ell) = \begin{bmatrix} [\mathbf{x}_i^\ell(0)]_\ell \\ \vdots \\ \left[\mathbf{x}_i^\ell(G-1)\right]_\ell \end{bmatrix}. \tag{17}$$

Then (15) becomes

$$\mathbf{B}\mathbf{f}_i(\ell) = \mathbf{e}_i e^{-j[\epsilon_i(M+\ell)+\phi_i]}, \tag{18}$$

where the $I \times G$ matrix

$$\mathbf{B} = \begin{bmatrix} c_{0,0} & \cdots & c_{0,G-1} e^{j\epsilon_0(N+M)(G-1)} \\ \vdots & & \vdots \\ c_{I-1,0} & \cdots & c_{I-1,G-1} e^{j\epsilon_{I-1}(N+M)(G-1)} \end{bmatrix}. \tag{19}$$

Note that $[\mathbf{B}]_{m,n} = c_{m,n} e^{j\epsilon_m(N+M)n}$, where $0 \le m \le I-1$, $0 \le n \le G-1$. In practice, we have $G \ge I$, so (18) always have exact solutions (thanks to the spreading codes)

$$\mathbf{f}_i(\ell) = \mathbf{B}^{-1} \mathbf{e}_i e^{-j[\epsilon_i(M+\ell)+\phi_i]} \tag{20}$$

Note that $\mathbf{e}_i$ in (18) and (20) is an $I \times 1$ unit vector (has value 1 in the i^{th} entry and zero elsewhere).

As a summary, by solving (20) we obtain the G non-zero entries of the ℓ^{th} row of $\mathbf{X}_i$. Doing (20) for all the N rows $0 \le \ell \le N-1$, we obtain the CFO mitigation matrix $\mathbf{X}_i$ for the user i. By $\mathbf{X}_i \mathbf{y}$ we conduct both CFO removal and despreading for the signal of user i. Because (11) can be accurately satisfied, we have

$$\mathbf{z}_i = \mathbf{X}_i \mathbf{y} = \mathbf{H}_i \mathbf{s}_i + \mathbf{X}_i \mathbf{u}. \tag{21}$$

Performing FFT on $\mathbf{z}_i$, we can detect the symbols $b_i(k)$ transmitted by the user i just as conventional single-user OFDM. This procedure can be repeated for every user i, $0 \le i \le I-1$.

For each user i, we need to compute a unique matrix $\mathbf{X}_i$, the computational complexity of which is $O(GI^2N)$ for each user, or $O(GI^3N)$ for all the I users. After (20), the rest demodulation and detection has complexity of $O(GN + N\log N)$ only for each user, or $O(GIN + IN\log N)$ for all the users. On the other hand, the inverse of $\mathbf{B}$ in (20), which is the most complex one, just needs to be calculated once for all the users and for all the OFDM frames if CFOs are constant and spreading codes are periodic. In this case, the complexity of (20) becomes in fact negligible compared with $O(GIN + IN\log N)$. Note that the conventional MC-DS-OFDM receiver has complexity $O(GIN + GN\log N)$ for all the I users. Therefore, if $G \gg I$, our algorithm may be even more efficient.

4. SIMULATIONS

In order to evaluate the performance of our algorithm, we simulated a system with two to four multiple access users and one base-station receiver. The parameters we used were MC-DS-CDMA with $N = 32$, QPSK, $G = 16$ orthogonal Walsh-Hadamard codes. The CFO and delay were randomly generated for each user. Randomly generated channel with order $L = 3$ were used. 10000 runs of the program were conducted to derive the average symbol error rate (SER) under various signal-to-noise ratio (SNR) or various CFO.

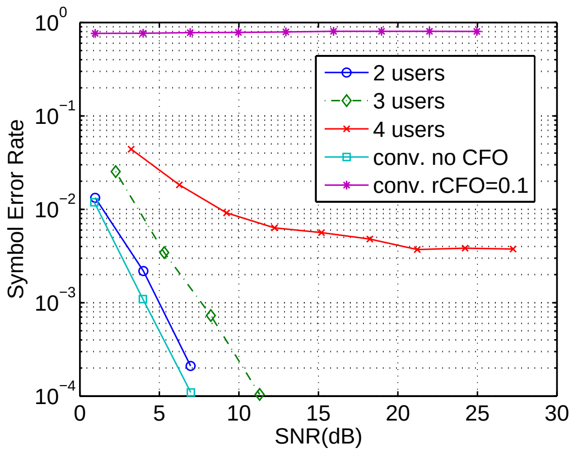

Fig. 2. SER vs. SNR for our new algorithm (with 2 to 4 users), and conventional MC-DS-CDMA without CFO or with CFO (with 2 users)

First, we studied the performance of our algorithm in combating CFO when different number of users $I = 2, 3, 4$ were presented. We set the relative CFO (rCFO) between the two users when $I = 2$ to be 0.1, which was $\epsilon_1 - \epsilon_0 = 0.1$, or specifically, $\epsilon_0 = 0.1$ and $\epsilon_1 = 0.2$. For $I = 3$ and $I = 4$, the rCFO were 0.1, 0.2 and 0.1, 0.2, 0.3, respectively, compared with ϵ_0. The simulation results are shown in Fig. 2. Under CFO, our algorithm can support a reasonable number of users, while the conventional method fails. In particular, our method does not lose much performance compared with the ideal no-CFO case, which can be seen from $I = 2$ cases.

Next, in Fig. 3, we show that our algorithm can cancel all CFO, so its performance is reliable for whatever (even large) CFO. This clearly shows the advantage of complete CFO cancellation over certain level of mitigation only.

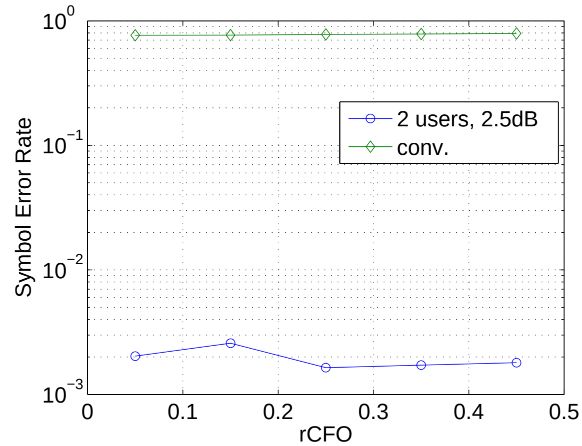

Fig. 3. SER vs. rCFO for 2 users at SNR=2.5dB. rCFO= $\epsilon_1 - \epsilon_0 = 0.1$. The new algorithm has CFO-independent performance, while conventional method fails.

5. CONCLUSIONS

In this paper, we proposed a new receiving algorithm for MC-DS-CDMA systems, whose major advantage is to completely cancel CFO. In our algorithm, the despreading and CFO removal are jointly performed by an efficient algorithm, which is then followed by the conventional FFT-based OFDM demodulation conducted user by user. Simulations show that this method has a superior performance which is independent of CFO, even when CFO is very large.

6. REFERENCES

[1] S. Hara and R. Prasad, "Overview of multi-carrier CDMA," *IEEE Commun. Mag.*, vol. 14, no. 12, pp. 126-133, Dec. 1997.

[2] V. M. DaSilva and E. S. Sousa, "Performance of orthogonal CDMA codes for quasi-synchronous communication system," *Proc. of IEEE ICUPC'93*, pp. 995-999, Ottawa, Canada, Oct. 1993.

[3] H. Steeudam and M. Moeneclaey, "The effect of carrier frequency offsets in down-link and up-link MC-DS-CDMA," *IEEE J. Select. Areas Commun.*, vol. 19, no. 12, pp. 2528-2536, Dec. 2001.

[4] F. Ng and X. Li, "Cooperative STBC-OFDM transmissions with imperfect synchronization in time and frequency," *IEEE 39th Asilomar Conf. on Signals, Systems and Computers*, Pacific Grove, CA, Oct. 30-Nov. 2, 2005.

[5] S. Barbarossa, M. Pompili and G. B. Giannakis, "Channel-independent synchronization of orthogonal frequency division multiple access systems," *IEEE J. Select. Areas Commun.*, vol. 20, no. 2, pp. 474-486, Feb. 2002.

[6] X. Cai, S. Zhou and G. B. Giannakis, "Group orthogonal multi-carrier CDMA," *IEEE Trans. Commun.*, vol. 52, no. 1, pp. 90-99, Jan. 2004.

[7] S.-H. Tsai, Y.-P. Lin and C.-C. J. Kuo, "An approximately MAI-free multi-access OFDM system in carrier frequency offset environment," *IEEE Trans. Signal Processing*, vol. 53, no. 11, pp. 4339-4353, Nov. 2005.

[8] X. Li and F. Ng, "Using cyclic prefix to mitigate carrier frequency and timing asynchronism in cooperative OFDM transmission," *IEEE 40th Asilomar Conf. on Signals, Systems and Computers*, Pacific Grove, CA, Oct. 2006.

[9] F. T. Chien, C.-H. Hwang and C.-C. Kuo, "Blind adaptive frequency offsets and channel estimations of MC-DS-CDMA systems with correlated fading," *Proc. of the SPIE*, vol. 5100, pp. 157-168, July 2003.