A Synchronization Technique for OFDM Systems with Smart Antenna

Kyu In Lee*, Dong Han Kim*, Jae Young Ahn**, and Yong Soo Cho*
*Chung-Ang University, **ETRI**
yscho@cau.ac.kr

ABSTRACT

In this paper, a method of carrier frequency offsets (CFO) synchronization and adaptive beamforming is proposed for orthogonal frequency division multiple access (OFDMA) systems with smart antenna. An estimation method for uplink CFO is proposed for spatial division multiple access (SDMA) environments where interference signals from other users are present at the same frequency band with different CFOs. The proposed synchronization and beamforming method form a beam in the direction of a desired user with CFO compensated, while place nulls in the direction of other users with different CFOs. The proposed method does not require an additional synchronization preamble for uplink CFO or feedback loop for initial synchronization.

***Keywords* : OFDMA, SDMA, smart antenna, uplink, CFO synchronization, CFO estimation, CFO compensation, adaptive beamforming**

1. INTRODUCTION

Recently, orthogonal frequency division multiple access (OFDMA) has been widely accepted as one of the most promising radio transmission technology for the next generation cellular systems. In OFDMA systems, although the intra-cell interference between users can be ignored due to distinct allocation of subcarrier group to each user, the users at cell boundaries experience a large inter-cell interference, especially in a full-loaded cellular environment with a frequency reuse factor equal to 1 [1][2]. The OFDMA system with smart antenna, herein referred to as SA-OFDMA, was introduced to mitigate the inter-cell interference by suppressing co-channel interference as well as to provide enhanced coverage. The SA-OFDMA can also increase system capacity by multiplexing channels in the spatial dimension, e.g., spatial division multiple access (SDMA) [3][4][5]. However, the performance of a beamformer will be degraded if carrier frequency offsets (CFOs) are present in the training symbols transmitted for estimation of antenna weight vector [6][7]. In the conventional systems with smart antenna, a base station usually estimates the uplink CFO of a user by using a pre-defined synchronization preamble transmitted from the user, and sends its estimate back to the user. Then, the user, compensated with the estimated CFO, sends a training symbol so that the base station can estimate the antenna weight vector. This type of uplink CFO synchronization, however, requires a long processing time due to the feedback loop and a periodic transmission of synchronization symbols. Even in the case where the uplink CFO compensation is completed by the conventional approach, the CFO might be changed due to the feedback loop delay, especially for the user moving with a high speed, resulting in an incorrect estimation of antenna weight vector. Also, for OFDMA/SDMA systems, the interference signals existing at the same frequency band with different CFOs can affect the uplink CFO estimation in the initial synchronization process. In this paper, the uplink CFO estimation method for SA-OFDMA systems is proposed to estimate the CFO of a user in a cellular environment where interference signals from other users are present at the same frequency band with different CFOs. The proposed method achieves CFO synchronization and adaptive beamforming directly from the training sequence transmitted from a desired user. It does not require an additional synchronization preamble for uplink CFO or feedback loop for initial synchronization. In section II, the uplink CFO synchronization and beamforming method for SA-OFDMA is proposed. The performances of the proposed approach are verified by computer simulations in section III. Conclusion is made in Section IV.

2. AN UPLINK CFO SYNCHRONIZATION AND BEAMFORMING METHOD FOR SA-OFDMA

Figure 1. shows a cellular environment where many users with different CFOs are present. Here, the term, ε_m , denotes the CFO between a base station and the *m-th* user. The received signal vector, $\mathbf{Y}_{m,l,k}$, of the *m-th* user on the *k-th* subcarrier of the *l-th* OFDM symbol consists of N_{ant} elements as follows:

$$\begin{aligned}\mathbf{Y}_{m,l,k} &= \left[Y_{m,l,k}(0)\ Y_{m,l,k}(1) \cdots Y_{m,l,k}(N_{ant}-1)\right]^T \\ &= \mathbf{a}(\theta_m)\, X_{m,l,k} H_{m,l,k} \frac{\sin(\varepsilon_m \pi)}{N \sin(\varepsilon_m \pi / N)} e^{j\varepsilon_m \pi \frac{N-1}{N}} e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_m l} + \mathbf{I}_{m,l,k} + \mathbf{W}_{m,l,k},\end{aligned} \tag{1}$$

where

$$\mathbf{a}(\theta_m) = \left[1, \exp\left(-j\, {}^{2\pi}\!/\!_{\lambda}\, d \sin\theta_m\right), \cdots, \exp\left(-j\, {}^{2\pi}\!/\!_{\lambda} (N_{ant}-1)\, d \sin\theta_m\right)\right], \tag{2}$$

$$\mathbf{I}_{m,l,k} = \mathbf{a}(\theta_m) \sum_{\substack{z=0 \\ z \neq k}}^{N-1} X_{m,l,z} H_{m,l,z} \frac{\sin[\pi(z-k+\varepsilon_m)]}{N \sin[\pi(z-k+\varepsilon_m)/N]} e^{j\pi(z-k+\varepsilon_m)\frac{N-1}{N}}, \tag{3}$$

$$\mathbf{W}_{m,l,k} = \left[W_{m,l,k}(0)\ W_{m,l,k}(1) \cdots W_{m,l,k}(N_{ant}-1)\right]^T. \tag{4}$$

Here, $\mathbf{a}(\theta_m)$ denotes the steering vector of the *m-th* user. Also, $\mathbf{I}_{m,l,k}$ and $\mathbf{W}_{m,l,k}$ represent the intercarrier interference vector and noise vector of the *m-th* user on the *k-th* subcarrier of the *l-th*

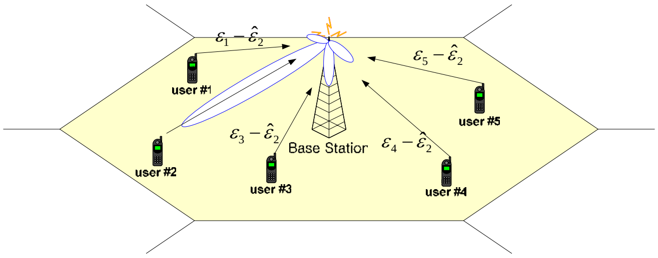
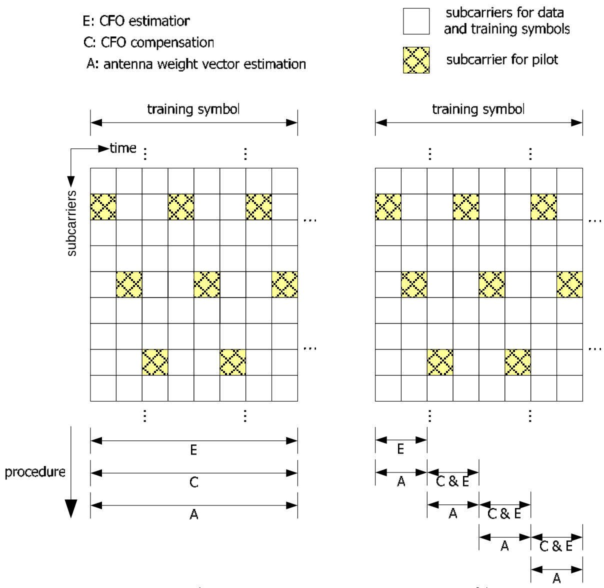


Figure 1: An uplink cellular environment with different CFOs

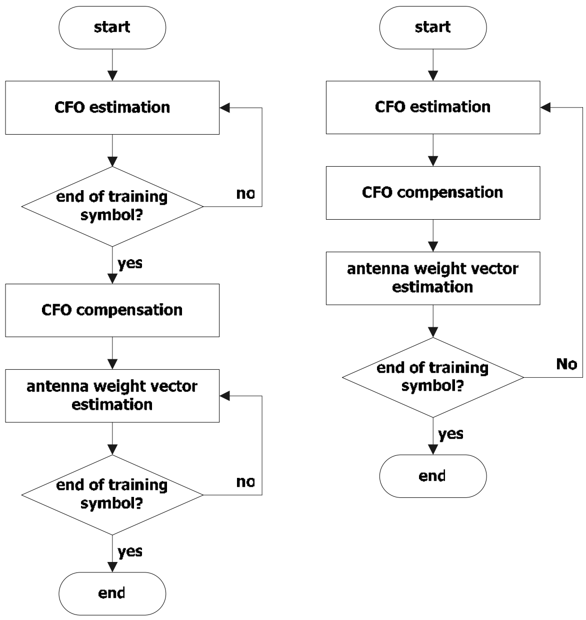
OFDM symbol, respectively. The term, λ, d, and θ_m denote the wavelength, the distance between antenna elements, and the angle of the *m-th* user's plane wave incident on the array, respectively. $H_{m,l,k}$ represents the channel frequency response of the *m-th* user on the *l-th* OFDM symbol of the *k-th* subcarrier, respectively. From (1), one can see that the post-FFT OFDM signal received at the base station contains the amplitude attenuation, phase rotation as well as intercarrier interference due to the CFO of the *m-th* user, ε_m [7]. Since the received signal on the *k-th* subcarrier of the *l-th* OFDM symbol at the base station can be represented as the sum of signals transmitted from all users in a cell, i.e., $\mathbf{Y}_{l,k}=\sum_{m=0}^{M-1}\mathbf{Y}_{m,l,k}$, the estimation of antenna weight vector for the *m-th* user is influenced not only by the CFO of the *m-th* user but also the CFOs of other users. Joint estimation of the CFO and antenna weight vector for each user is desirable in this situation. However, it is difficult to come up with an exact solution for the joint estimation of the uplink CFO and unbiased antenna weight vector in OFDM/SDMA systems. Instead, this paper proposes a sequential approach where the uplink CFO synchronization and adaptive beamforming are performed sequentially using the training sequence transmitted from the desired user.

Figure 2(a) and Figure 3(a) show the procedure and the corresponding flowchart of the proposed method, respectively. In this method, the CFO of the desired user is first estimated with the corresponding training sequence in a cellular environment where interference signals from other users are present at the same frequency band with different CFOs. Then, the estimated CFO of the desired user is used to compensate the CFO of the received signal. Note that the CFO compensation of the desired user, however, may change the CFOs of other users in the received signal. Finally, the antenna weight vector is estimated by using the minimum mean square error (MMSE) criterion such that the array can form a beam in the direction of the desired user with CFO compensated, while placing nulls in the direction of other users with different CFOs as shown in Figure 1. Notice that both uplink CFO synchronization and adaptive beamforming are achieved in the proposed approach directly from the training sequence. An additional synchronization preamble for uplink CFO or feedback loop for initial synchronization is not required. Since the CFO compensation or adaptive beamforming can be made by existing methods, we will focus, in this paper, on the CFO estimation method of the desired user in a cellular environment where interference signals from other users exist at the same frequency band with different CFOs.



(a) (b)

Figure 2 : Procedure for uplink CFO synchronization and beamforming



(a) (b)

Figure 3 : Flowchart for uplink CFO synchronization and beamforming

A typical structure of training symbols for uplink CFO synchronization and adaptive beamforming in SA-OFDMA systems is shown in Figure 2. The training symbols shown in Figure 2. have the same format as the reverse link training

(RLT) sequence used for adaptive antenna system (AAS) mode in IEEE 802.16e. It is composed of 8 consecutive bins, each consisting of 9 subcarriers (one for pilot subcarrier and eights for training sequence) [1][2].

The CFO of the *m-th* user is estimated in the frequency domain by

$$\hat{\varepsilon}_m = \frac{1}{2\pi\left(1+N_G/N\right)} \arg\left(\sum_{l=0}^{L-2}\sum_{k=0}^{K-1} \mathbf{P}_{m,l,k}^H \cdot \mathbf{P}_{m,l+1,k}\right) \tag{5}$$

where

$$\begin{aligned} \mathbf{P}_{m,l,k} &= \left[P_{m,l,k}(0)\ P_{m,l,k}(1) \cdots P_{m,l,k}(N_{ant}-1)\right]^T \\ &= X_{m,l,k}^* \cdot \mathbf{Y}_{l,k}, \end{aligned} \tag{6}$$

Here, $\hat{\varepsilon}_m$, K, and L represent the estimated value of CFO for the *m-th* user, the number of subcarriers assigned per bin, and the number of OFDM symbols for training, respectively. Also, N and N_G denote the FFT size and the length of cyclic prefix, respectively. The operator $\mathbf{A}^H$ denotes the Hermition of vector $\mathbf{A}$.

If the number of training sequence is large enough, i.e., L and K are large enough, the sum of inner product of $\mathbf{P}_{m,l,k}$ and $\mathbf{P}_{m,l+1,k}$ in (5) can be approximated as $E\left[\mathbf{P}_{m,l,k}^H \cdot \mathbf{P}_{m,l+1,k}\right]$. Here, it is assumed that the training sequence of one user is orthogonal to the ones of other users. That is, the training sequence between the m_1*-th* user and the m_2*-th* user satisfies the orthogonal property, $E\left[X_{m_1,l,k} \cdot X_{m_2,l,k}^*\right] = 0$ and $E\left[X_{m_1,l,k} \cdot X_{m_1,l,k}^*\right] = 1$. Using this property, the term $E\left[\mathbf{P}_{\tilde{m},l,k}^H \cdot \mathbf{P}_{\tilde{m},l+1,k}\right]$ for the $\tilde{m}$*-th* user can be derived as

$$\begin{aligned}
&E\left[\mathbf{P}_{\tilde{m},l,k}^H \cdot \mathbf{P}_{\tilde{m},l+1,k}\right] \\
&= E\left[\left(X_{\tilde{m},l,k}^* \cdot \sum_{m=0}^{M-1} \mathbf{Y}_{m,l,k}\right)^H \cdot \left(X_{\tilde{m},l+1,k}^* \cdot \sum_{m=0}^{M-1} \mathbf{Y}_{m,l+1,k}\right)\right] \\
&= E\left[\left(X_{\tilde{m},l,k}^* \cdot \mathbf{Y}_{\tilde{m},l,k}\right)^H \cdot \left(X_{\tilde{m},l+1,k}^* \cdot \mathbf{Y}_{\tilde{m},l+1,k}\right)\right. \\
&\quad \left. + \left(X_{\tilde{m},l,k}^* \cdot \sum_{\substack{m=0 \\ m\neq\tilde{m}}}^{M-1} \mathbf{Y}_{m,l,k}\right)^H \cdot \left(X_{\tilde{m},l+1,k}^* \cdot \sum_{\substack{m=0 \\ m\neq\tilde{m}}}^{M-1} \mathbf{Y}_{m,l+1,k}\right)\right] \\
&= E\left[N_{ant} \cdot \left|H_{\tilde{m},l,k}\right|^2 \cdot \alpha\left(\varepsilon_{\tilde{m}}\right)^2 \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_{\tilde{m}}}\right. \\
&\quad + \left(X_{\tilde{m},l,k} \cdot X_{\tilde{m},l+1,k}^* \cdot \mathbf{N}_{\tilde{m},l,k}^H \cdot \mathbf{N}_{\tilde{m},l+1,k}\right) \\
&\quad + \left(X_{\tilde{m},l,k}^* \cdot \mathbf{N}_{\tilde{m},l,k}\right)^H \cdot \left(\mathbf{a}\left(\theta_{\tilde{m}}\right) \cdot \left|X_{\tilde{m},l+1,k}\right|^2 \cdot H_{\tilde{m},l+1,k} \alpha\left(\varepsilon_{\tilde{m}}\right) \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_{\tilde{m}}(l+1)}\right) \\
&\quad + \left(X_{\tilde{m},l+1,k}^* \cdot \mathbf{N}_{\tilde{m},l+1,k}\right) \cdot \left(\mathbf{a}\left(\theta_{\tilde{m}}\right) \cdot \left|X_{\tilde{m},l,k}\right|^2 \cdot H_{\tilde{m},l,k} \cdot \alpha\left(\varepsilon_{\tilde{m}}\right) \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_{\tilde{m}} l}\right)^H \\
&\quad + \left(\sum_{\substack{m=0 \\ m\neq\tilde{m}}}^{M-1} \mathbf{a}\left(\theta_m\right) \cdot X_{\tilde{m},l,k}^* \cdot X_{m,l,k} \cdot H_{m,l,k} \cdot \alpha\left(\varepsilon_m\right) \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_m l} + X_{\tilde{m},l,k}^* \cdot \mathbf{N}_{m,l,k}\right)^H \\
&\quad \left. \cdot \left(\sum_{\substack{m=0 \\ m\neq\tilde{m}}}^{M-1} \mathbf{a}\left(\theta_m\right) \cdot X_{\tilde{m},l+1,k}^* \cdot X_{m,l+1,k} \cdot H_{m,l+1,k} \cdot \alpha\left(\varepsilon_m\right) \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_m (l+1)} + X_{\tilde{m},l+1,k}^* \cdot \mathbf{N}_{m,l+1,k}\right)\right] \\
&= N_{ant} \cdot \left|H_{\tilde{m},l,k}\right|^2 \cdot \alpha\left(\varepsilon_{\tilde{m}}\right)^2 \cdot e^{j2\pi\left(\frac{N+N_G}{N}\right)\varepsilon_{\tilde{m}}},
\end{aligned} \tag{7}$$

where

$$\alpha\left(\varepsilon_m\right) = \frac{\sin\left(x\pi\right)}{N\sin\left(x\pi/N\right)} \cdot e^{jx\pi\frac{N-1}{N}}. \tag{8}$$

Note that the phase of $E\left[\mathbf{P}_{\tilde{m},l,k}^H \cdot \mathbf{P}_{\tilde{m},l+1,k}\right]$ is given by $2\pi\left(N+N_G\right)\varepsilon_{\tilde{m}}/N$, which is independent of the CFOs of other users except the *m-th* user. Then, the CFO of the *m-th* user can be estimated by extracting the phase component, $2\pi\left(N+N_G\right)\varepsilon_{\tilde{m}}/N$, from (7). From this result, one can see that the CFO of the desired user is correctly estimated by (5) in the initial synchronization stage where the interference signals from other users exists at the same frequency band with different CFOs.

After successful uplink CFO estimation of the *m-th* user, the CFO of the *m-th* user is compensated at all antennas of the base station using the conventional method as follows:

$$\mathbf{y}_n' = \mathbf{y}_n \cdot e^{-j2\pi\hat{\varepsilon}_m n/N} \tag{9}$$

where $\mathbf{y}_n = \left[y_n(0)\ y_n(1) \cdots y_n(N_{ant}-1)\right]^T$ and $\mathbf{y}_n' = \left[y_n'(0)\ y_n'(1) \cdots y_n'(N_{ant}-1)\right]^T$ represent the time-domain received signal vector and its CFO compensated version, respectively. Notice that the CFO compensation of the *m-th* user with $\hat{\varepsilon}_m$ may change the CFOs of other users in the received signal. In the proposed approach, the interference signals from other users with different CFOs are eliminated by placing nulls in those directions while the CFO compensated signal of the desired user is obtained by forming a beam in the direction of desired user. Various beamforming techniques based on training sequence such as SMI, LMS, and RLS can be applied at this stage.

An iterative version of the proposed method is shown in Figure 2(b) and Figure 3(b). In this approach, we first estimate the CFO of the desired user by (5) with $L=2$ and form a coarse beam in the direction of the desired user using the first two training symbols. Next, we compensate the next two symbols with the CFO estimated from the first two symbols, which will change the CFO's of other users. Then, the CFOs of other users are reduced by placing nulls in the directions of other users. This process is repeated until the end of training symbols. As the number of iterations increases, the weights of the beamformer will converge to the ideal ones for CFO-free uplink situation. This approach requires less memory, delay, and complexity at the expense of possible performance degradation.

3. SIMULATION

In this section, the performances of the proposed uplink CFO synchronization and adaptive beamforming method are evaluated with following parameter set: carrier frequency= 2GHz, bandwidth= 20MHz, FFT size= 2048, guard interval= 512, and the number of antennas= 12. Uncoded QPSK and NLMS are used for modulation and beamforming algorithm, respectively. ITU-R models are used for multipath fading channel. Minimum resource for transmission of data and training symbol is assumed 5 bins for each user. As shown in Figure 2, the training sequences transmitted over 8 OFDM symbol periods are used for CFO synchronization and beamforming. Four users are accommodated in the same frequency band for one cell (SDMA=4). Here, it is assumed that the CFOs of four users in the same band are 0.3 for target user and -0.25, -0.2, -0.15 for three other users. Note that the CFOs of interferers have an opposite sign to that of target user, implying that CFOs of interference signals will be increased after CFO compensation. The same set of CFOs, given above, is allocated to the users assigned to different frequency bands. Figure 4. shows the CFO estimation performances, obtained by (5), for OFDMA/SDMA systems. A mean square error (MSE) of $4 \star 10^{-4}$ in CFO can be achieved for the desired user at the SNR of -2dB when Pedestrian channel A/B and Vehicular channel A are used.

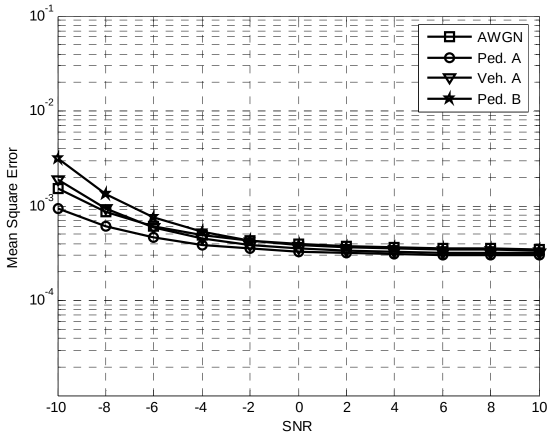


Figure 4 : CFO estimation performances of the proposed method for OFDMA/SDMA systems

Figure 5. shows the bit error rate (BER) performances of SA-OFDMA when the proposed method for CFO synchronization and adaptive beamforming is used. Here, the dotted lines represent the ideal cases where no CFO is present in uplink. About 3 dB performance degradation at 10^{-5} occurs when Pedestrian channel A is used. The broken lines represent the cases where CFO synchronization process is not performed during training period. Here, in order to clearly see the behavior of adaptive beamformer when CFOs are not compensated, CFOs are not applied during the period of data transmission.

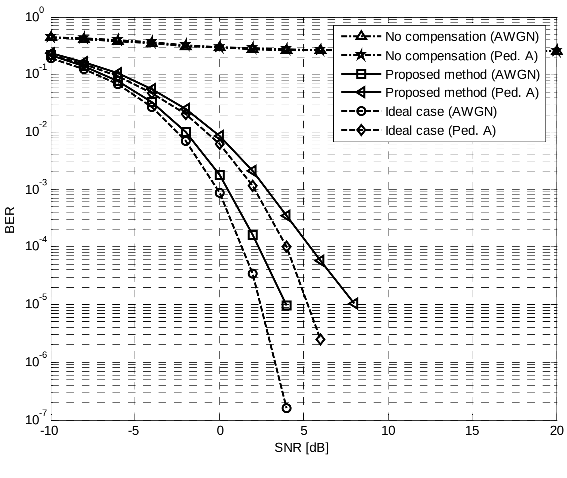


Figure 5 : BER performances of the proposed method for OFDMA/SDMA systems

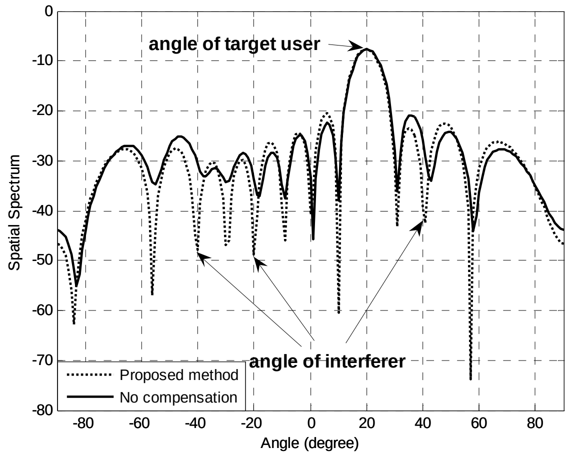


Figure 6 : Beam pattern of the proposed method for OFDMA/SDMA systems

From this figure, one can see that the error floor phenomena occur due to the incorrect estimation of antenna weight vector during training period when CFOs are present. The solid lines represent the cases where the proposed method is applied. The proposed method performs only about 2 dB lower than the ideal cases at the BER of 10^{-5} in both AWGN and Pedestrian channel A.

Figure 6. shows the corresponding beam patterns when Pedestrian channel A is used. From this figure, one can see that the beamformer cannot form nulls in the direction of interference signals if the uplink CFO synchronization is not achieved before the estimation of antenna weight vector. In this case, the beamformer cannot eliminate interference signals, leading to error floor phenomenon as shown in Figure 5. On the other hand, the proposed method can successfully place nulls in the direction of interference signals, reducing the BER close to ideal case.

4. CONCLUSTION

In this paper, the uplink CFO synchronization and adaptive beamforming methods were proposed for OFDMA/SDMA systems. The proposed CFO estimation method was shown to provide the correct CFO estimate of the desired user even when interference signals from other users are present at the same frequency band with different CFOs. It was also shown that the proposed synchronization and beamforming methods can form a beam in the direction of the desired user with CFO compensated, while place nulls in the direction of other users with different CFOs. The proposed method does not require an additional synchronization preamble for uplink CFO or feedback loop for initial synchronization.

REFERENCES

[1] IEEE P802.16-REVd/D5-2004, *Air Interface for Fixed Broadband Wireless Access Systems*, May 2004.

[2] IEEE P802.16e/D5, *Air Interface for Fixed and Mobile Broadband Wireless Access Systems*, Sep. 2004.

[3] B. D. Van Veen and K. M. Buckly, "Beamforming: a versatile approach to spatial filtering," *IEEE Aerosp. Electron. Syst. Mag.*, vol. 5, pp. 4-24, 1988.

[4] J. C. Liberti, Jr. and T. S. Rappaport, *Smart Antennas for Wireless Communications: IS-95 and Third Generation CDMA Applications*, Prentice-Hall Inc., 1999.
[5] N. Geng, I. Viering, and M. Kiessling, “Multi-user MIMO-OFDM cell throughput under real-world propagation conditions,” *Proc. IEEE VTC*, vol. 2, pp 879-883, Sept. 2002.
[6] A. Kannan, T.P. Krauss, and M.D. Zoltowski, “Separation of cochannel signals under imperfect timing and carrier synchronization,” *IEEE Trans. Veh. Tech.*, vol. 50, no. 1, pp. 79-96, Jan. 2001.
[7] T. Pollet, M. V. Bladel, and M. Moeneclaey, “BER sensitivity of OFDM systems to carrier frequency offset and Wiener phase noise,” *IEEE Trans. Comm.*, vol. 43, no, 234, pp. 191-193, Feb./Mar./Apr. 1995.