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LETTER

Blind Carrier Frequency Synchronization for the Uplink of Interleaved OFDMA Systems

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SUMMARY In this letter, a novel blind CFO estimation algorithm for the uplink of an OFDMA system is proposed. The proposed method exploits the inherent redundant information in OFDMA symbols and does not require additional pilot or preamble overhead. Since it is a post-FFT estimator, it does not use filter banks to separate the desired user's signal from the others in the time domain. Hence, the subcarriers of a certain user are not restricted to be clustered in the frequency domain. Therefore, the proposed estimator can be applied to OFDMA systems with an arbitrary subcarrier assignment over the entire bandwidth, including IEEE 802.16e, to obtain sufficient frequency diversity in a frequency selective fading channel. The proposed method can be efficiently used for continuous tracking of all active users' CFOs only with two FFT windows within a single OFDM symbol. From simulation results, the performance of the proposed scheme is shown better than that uses preamble symbols.
key words: *OFDMA, blind synchronization, CFO*

1. Introduction

In an orthogonal frequency-division multiple access (OFDMA) system, multiple users transmit their own data simultaneously on each orthogonal set of subcarriers, which makes it easy to eliminate the multiuser interference (MUI) and the intercarrier interference (ICI). Thus, OFDMA has been widely accepted as a promising technique for the multiple access technique of future wireless communication systems and was adopted in IEEE 802.16e. However, an OFDMA system is much more vulnerable to carrier frequency offset (CFO) than single carrier systems. The CFO is a frequency misalignment between a transmitter and a receiver. It destroys the orthogonality between subcarriers and induces ICI, which causes MUI among simultaneously transmitting users and degrades the performance of the system. In downlink, all subcarriers received at a user have the same CFO, which can be easily estimated and recovered at the receiver. However, in uplink, the subcarriers from different users have different CFOs. So, the signal of a certain user should be separated prior to estimate the CFO of the user. To acquire this, a time-domain filter bank was used to separate several users' signals [1]. This method uses the cyclic prefix (CP) in an OFDMA symbol and does not require additional pilot or preamble symbol. But the defect of this method is that adjacent subcarriers should be allocated to each user so that frequency diversity cannot be obtained and that side-lobe leakage occurs due to the non-ideal filter bank. To solve this problem, a post-FFT estimator was proposed in [2], in which the signal of the desired user is separated after a discrete-Fourier transform (DFT) and used to estimate the CFO of the user. By doing this, no restriction is imposed on the subcarrier allocation and the frequency diversity can be achieved in a frequency selective fading channel. But this method requires several consecutive identical OFDMA symbols. In [3], a structure-based blind CFO estimation method was proposed for the uplink of an OFDMA system by investigating the algebraic structure of OFDMA signals. However, the subcarrier assignment for a user is restricted to use the periodic structure of the time-domain signal and the complexity is quite large since it requires matrix decomposition.

In this letter, a novel blind post-FFT CFO estimator is proposed for the uplink of OFDMA systems by exploiting the intrinsic phase shift between two time-shifted observation windows in a single OFDMA symbol rather than using preamble symbols. Since the proposed estimator is a post-FFT algorithm, it can support an arbitrarily interleaved subcarrier assignment. In addition, the proposed method does not require any preamble symbols. To deal with the frequency offset in uplink, two kinds of different approaches have been considered: a feedback control channel method as in [1], [2] and a signal processing method at a base station (BS) as in [4], [5]. The latter does consume feedback resource but cannot reduce ICI and MUI significantly or needs an iterative interference cancellation scheme with very high complexity. In this letter, the former approach is adopted and it will be shown that scarce control messages are sufficient even in a severely time-varying environment.

2. The Proposed CFO Estimation Algorithm

We consider the uplink of an OFDMA system, which has N subcarriers and P users. Each user transmits on $L = N/P$ subcarriers, which is assumed to be an integer. The L information symbols of the pth user in the nth OFDMA symbol is denoted as $[x_n^{(p)}(0), x_n^{(p)}(1), \ldots, x_n^{(p)}(L-1)]$. These symbols are mapped according to the subcarrier assignment of the system to produce the frequency-domain signal denoted as $[s_n^{(p)}(0), s_n^{(p)}(1), \ldots, s_n^{(p)}(N-1)]$. The frequency-domain symbol $s_n^{(p)}(k)$ is defined as

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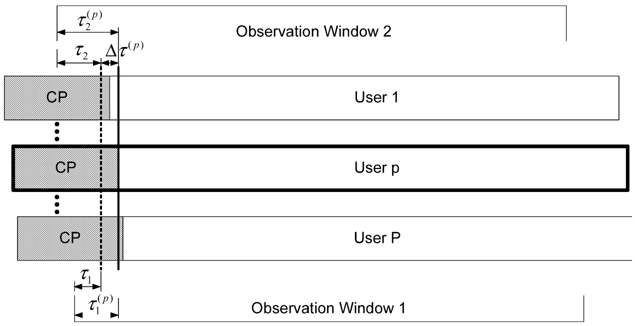

Fig. 1 Two observation windows with different delays.

$$s_n^{(p)}(k) = \begin{cases} x_n^{(p)}(l), & \text{if } k = c^{(p)}(l), \\ 0, & \text{otherwise,} \end{cases}$$

where $c^{(p)}(l)$ is the lth subcarrier index assigned to the pth user. After inverse fast Fourier transform (IFFT) operation and adding CP, the pth user's signal is transmitted to a BS. The received signal at the BS can be represented as

$$y(t) = \frac{1}{\sqrt{N}} \sum_{p=1}^{P} \sum_{k=0}^{N-1} s_n^{(p)}(k) H_n^{(p)}(k) e^{j2\pi(f_k + f_{off}^{(p)}(n))(t-\Delta t^{(p)})} + z(t), \tag{1}$$

where $H_n^{(p)}(k)$ is the channel response of the pth user at the kth subcarrier of the nth OFDMA symbol, f_k is the kth subcarrier frequency at the BS, and $z(t)$ denotes thermal noise. Also, $f_{off}^{(p)}(n)$ is the residual uplink CFO of the pth user, given as $f_{off}^{(p)}(n) = \Delta f^{(p)}(n) - f_{comp}^{(p)}(n)$, where $\Delta f^{(p)}(n)$ is the CFO due to the downlink synchronization error of the pth user and the Doppler frequency effect, and $f_{comp}^{(p)}(n)$ is the compensated frequency value on the nth OFDMA symbol at the mobile station (MS), defined as $f_{comp}^{(p)}(n) = \sum_{a=0}^{n-1} \hat{f}_{off}^{(p)}(a)$, while $\hat{f}_{off}^{(p)}(a)$ is the estimated CFO value at the ath OFDMA symbol and reported from the BS to the MS. Here, $\hat{f}_{off}^{(p)}(a)$ is zero if the estimated CFO is not reported in the ath OFDMA symbol. In systems like IEEE 802.16e, the uplink timing control is performed by using the initial and the periodic ranging operations. Here, $\Delta t^{(p)}$ denotes the residual timing offset of the pth user's signal due to the timing control error with respect to the reference timing of the BS. By sampling with period of $T = T_s/N$, where T_s is the OFDMA symbol interval without CP, we obtain two sets of discrete signals of the nth OFDMA symbol corresponding to the two observation windows as shown in Fig. 1 as

$$\begin{aligned} y_{n,i}(m) \; = \; & \frac{1}{\sqrt{N}} \sum_{p=1}^{P} \sum_{k=0}^{N-1} s_n^{(p)}(k) H_n^{(p)}(k) e^{j2\pi(k+\nu_n^{(p)})(m-\tau_i^{(p)})/N} \\ & + z_i(m), \quad i = 1, 2. \end{aligned} \tag{2}$$

Here, $\nu_n^{(p)} = f_{off}^{(p)}(n)\, T_s$ is the normalized residual frequency offset and $\tau_i^{(p)} = \tau_i + \Delta\tau^{(p)}$ is the timing delay for each observation window including the intentional delay for the ith window, τ_i, and the timing offset of the pth user, $\Delta\tau^{(p)} = \Delta t^{(p)}/T$. In Fig. 1, the dashed line denotes the reference timing of the BS receiver and the solid line denotes the exact timing of the pth user. The above two sets of the received time-domain signals can be written in more compact form by using matrix expression as

$$\mathbf{y}_{n,i} = \sum_{p=1}^{P} e^{-j2\pi\nu_n^{(p)}\tau_i^{(p)}/N} \mathbf{\Gamma}(\nu_n^{(p)}) \mathbf{W} \mathbf{E}(\tau_i^{(p)}) \tilde{s}_n^{(p)} + \mathbf{z}_i, \; i = 1, 2, \tag{3}$$

where $\mathbf{\Gamma}(\nu_n^{(p)}) = \mathrm{diag}\{[\exp(j2\pi\nu_n^{(p)} m/N)]_{m=0}^{N-1}\}$, $\mathbf{E}(\tau) = \mathrm{diag}\{[\exp(-j2\pi\tau k/N)]_{k=0}^{N-1}\}$, $\tilde{s}_n^{(p)} = ([H^{(p)}(k) s_n^{(p)}(k)]_{k=0}^{N-1})^T$, and $\mathbf{W}$ is the $N \times N$ IDFT matrix defined as

$$\mathbf{W} = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & e^{j2\pi\frac{1}{N}} & \dots & e^{j2\pi\frac{(N-1)}{N}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{j2\pi\frac{(N-1)}{N}} & \dots & e^{j2\pi\frac{(N-1)(N-1)}{N}} \end{bmatrix}. \tag{4}$$

Here, $[b(m)]_{m=0}^{N-1}$ denotes a $1 \times N$ vector whose mth element is $b(m)$. Note that $\mathbf{\Gamma}(\nu_n^{(p)})$ and $\mathbf{E}(\tau)$ represent intrinsic phase shifts caused by the CFO and the timing delay, respectively. After DFT, we obtain the frequency-domain signal $\mathbf{Y}_{n,1}^{(p)}$ of the pth user corresponding to the observation window 1 as

$$\mathbf{Y}_{n,1}^{(p)} = \mathbf{F}^{(p)} \mathbf{W}^H \mathbf{y}_{n,1}, \tag{5}$$

where $\mathbf{F}^{(p)}$ is a $L \times N$ matrix to select L subcarriers assigned to the pth user as

$$\left[F^{(p)}\right]_{l,k} = \begin{cases} 1, & \text{if } k = c^{(p)}(l), \\ 0, & \text{otherwise.} \end{cases}$$

Similarly, we obtain the frequency-domain signal according to the observation window 2 after DFT. Note that there exists additional phase shifts in $\mathbf{y}_{n,2}^{(p)}$ according to $\mathbf{E}(\tau_0)$ and $e^{-j2\pi\nu_n^{(p)}\tau_0/N}$ compared to $\mathbf{y}_{n,1}^{(p)}$ due to the timing difference $\tau_0 = \tau_2^{(p)} - \tau_1^{(p)}$ and the frequency offset. By compensating the phase shift, $\mathbf{E}(\tau_0)$, we obtain another set of frequency-domain signal as

$$\mathbf{Y}_{n,2}^{(p)} = \mathbf{F}^{(p)} \mathbf{E}^H(\tau_0) \mathbf{W}^H \mathbf{y}_{n,2}. \tag{6}$$

The lth component $\bar{Y}_{n,1}^{(p)}(l)$ of $\mathbf{Y}_{n,1}^{(p)}$ can be given as

$$\begin{aligned} \bar{Y}_{n,1}^{(p)}(l) \; = \; & e^{-j2\pi\left(\nu_n^{(p)}\tau_1^{(p)} + \tau_1^{(p)} c^{(p)}(l)\right)/N} H_n^{(p)}\left(c^{(p)}(l)\right) \\ & \cdot x_n^{(p)}(l)\alpha(\nu_n^{(p)}) + ICI_{n,1}^{(p)}(l) + MUI_{n,1}^{(p)}(l) \\ & + Z_{n,1}^{(p)}(l), \end{aligned} \tag{7}$$

where

$$\alpha(\nu) = \frac{sin(\pi\nu)}{N sin(\pi\nu/N)} \cdot e^{j\pi\nu(N-1)/N}, \tag{8}$$

$ICI_{n,1}^{(p)}(l)$ is the self interference from the subcarriers assigned to the pth user as

$$ICI_{n,1}^{(p)}(l) = \sum_{\substack{l'=0 \\ l'\neq l}}^{L-1} e^{-j2\pi\nu_n^{(p)}\tau_1^{(p)}/N} e^{-j2\pi\tau_1^{(p)}c^{(p)}(l')/N} \cdot H_n^{(p)}\left(c^{(p)}(l')\right) x_n^{(p)}(l')\alpha\left(\nu_n^{(p)}+c^{(p)}(l')-c^{(p)}(l)\right), \tag{9}$$

$MUI_{n,1}^{(p)}(l)$ is the interference from the subcarriers assigned to the other users represented as

$$MUI_{n,1}^{(p)}(l)=\sum_{\substack{p'=1 \\ p'\neq p}}^{P}\sum_{l'=0}^{L-1} e^{-j2\pi\tau_1^{(p')}\left(\nu_n^{(p')}+c^{(p')}(l')\right)/N} \cdot H_n^{(p')}\left(c^{(p')}(l')\right)x_n^{(p')}(l')\alpha\left(\nu_n^{(p')}+c^{(p')}(l')-c^{(p)}(l)\right), \tag{10}$$

and $Z_{n,1}^{(p)}(l)$ denotes thermal noise. Also, the lth component of $\mathbf{Y}_{n,2}^{(p)}$ can be represented as

$$\bar{Y}_{n,2}^{(p)}(l) = e^{-j2\pi\left(\nu_n^{(p)}\tau_2^{(p)}+(\tau_2^{(p)}-\tau_0)c^{(p)}(l)\right)/N}H_n^{(p)}\left(c^{(p)}(l)\right) \cdot x_n^{(p)}(l)\alpha(\nu_n^{(p)})+ICI_{n,2}^{(p)}(l)+MUI_{n,2}^{(p)}(l)+Z_{n,2}^{(p)}(l). \tag{11}$$

We can see that the exponential terms in (7) and (11) are the same except $e^{j2\pi\nu_n^{(p)}\tau_0/N}$. Thus, if the interference and noise terms are negligible, the residual uplink CFO of the pth user can be obtained as

$$\hat{\nu}_a^{(p)} = \frac{N}{2\pi\tau_0} \arg \sum_{n=a-M+1}^{a} \mathbf{Y}_{n,2}^{(p)H}\mathbf{Y}_{n,1}^{(p)}, \tag{12}$$

where M is the number of OFDMA symbols used for the CFO estimation in each reporting time. The estimated CFO is then reported to the pth MS and used for the compensation at the transmitter of the MS. During the CFO compensation, the ICI and MUI terms in (7) and (11) can be kept negligible so that (12) is not much influenced by them, if the estimation and the compensation works well. The receiver structure of the proposed scheme is shown in Fig. 2. As can be seen in the figure, the receiver supports all active users' CFO tracking tasks simultaneously only with two FFT operations. This is because the proposed algorithm needs only the time difference τ_0 between the two observation windows to estimate the CFO and the exact values of $\tau_i^{(p)}$ of each user is not required. Thus, the complexity of the proposed scheme is surely within the range of practical implementation. One thing we might consider is the relationship between τ_0 and the estimation performance. As is expected, the performance of the proposed CFO estimator gets better as τ_0 increases. This comes from the fact that the phase rotation caused by a given CFO $\nu_n^{(p)}$ becomes larger as τ_0 increases, which increases the difference between $\mathbf{Y}_{n,1}^{(p)}$ and $\mathbf{Y}_{n,2}^{(p)}$ and improves the performance of the proposed scheme. However, if τ_0 is set too large, the OFDMA symbol from observation window 2 is corrupted by the previous symbol in a frequency selective fading channel and the performance suffers from ISI. Thus, it is reasonable to set τ_2 such that $T_{cp} - (\tau_2 + \Delta\tau)$ remains greater than the largest delay of the channel, where T_{cp} is the duration of CP and $\Delta\tau$ is the maximum timing offset between users.

3. Simulation Results

For computer simulation, the number of subcarriers, N, is set to 1024 and the number of users, P, is set to 8. Also, the system bandwidth is 10 MHz and the carrier frequency is 2 GHz. For each user, $L = 128$ uniformly spaced subcarriers are allocated. The channel is assumed to be a 63-tap multipath channel with an exponentially decaying power delay profile with decay factor of 0.2. The CP length is assumed to be 64 samples. Figure 3 shows the performances of the proposed CFO estimator in (12) and the conventional method using preambles [2]. We assume that 2 consecutive preambles are transmitted once in every 50 OFDMA symbols and that the CFO of a target user is ν and the other users' CFOs are uniformly distributed

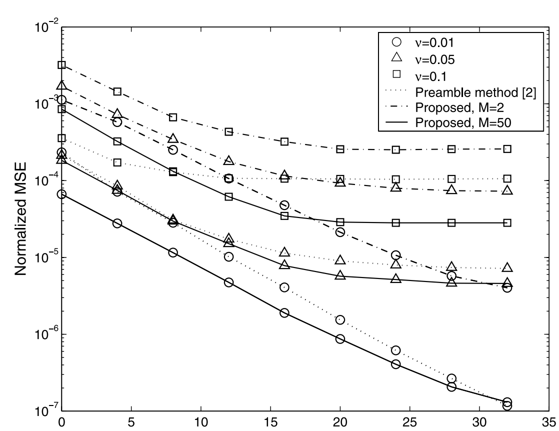

Fig. 3 Normalized MSE of the proposed blind CFO estimator in a dispersive channel with uniformly distributed other users in $[-\nu, \nu]$.

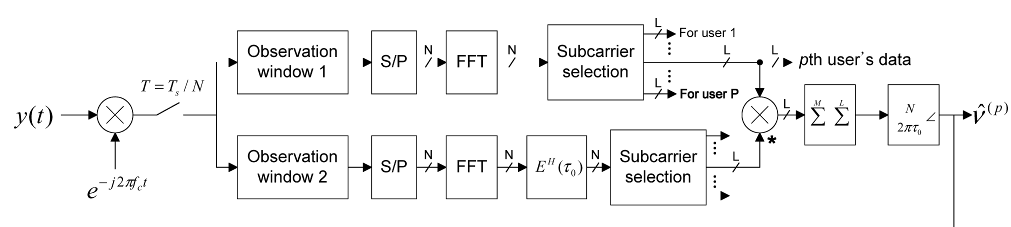

Fig. 2 Uplink receiver structure for the proposed frequency synchronization scheme.

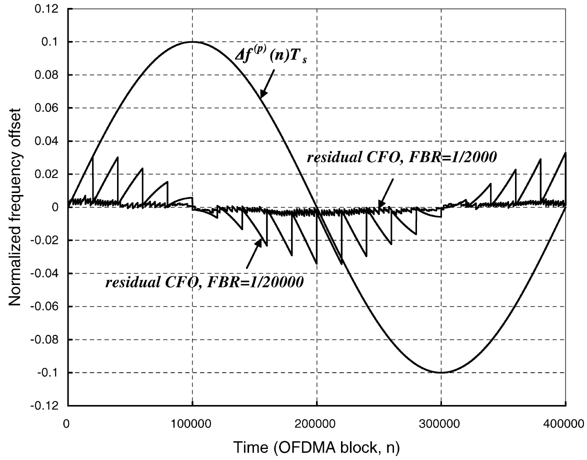

Fig. 4 Residual CFO of the proposed tracking scheme as time passes at SNR 25dB in a rapidly time-varying CFO environment.

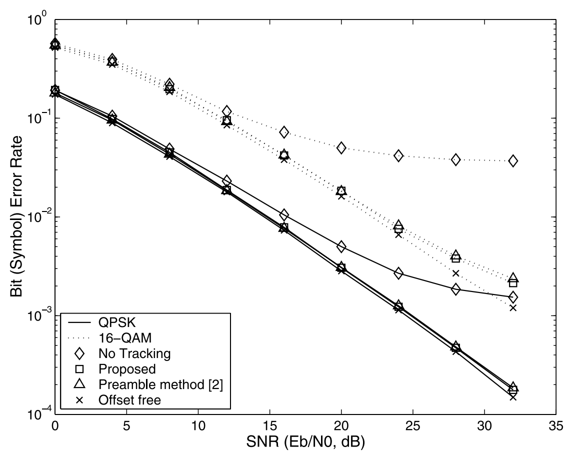

Fig. 5 Performance comparison for QPSK and 16-QAM system, where all 8 active users' normalized CFOs are time-varying.

in $[-\nu, \nu]$. In the OFDMA system used in the simulation, $\nu = 0.1$ corresponds to approximately 250 km/h of mobile speed. Also, the maximum timing offset, $\Delta\tau$, is set to 5 samples by assuming up to 250 km/h of mobile speed and a timing control feedback in every 2 seconds in a system with 10 MHz bandwidth. For the proposed estimator, the time delay τ_0 between the two observation windows is set to 50 samples. It is shown in Fig. 3 that the preamble method is better than the proposed estimator, when only two symbols ($M = 2$) are used for the frequency estimation in (12). However, the proposed estimator is able to use all the transmitted symbols within a reporting period. When M is 50, the proposed estimator shows better performance than the preamble method. In Figs. 4 and 5, the CFO tracking performance of the proposed scheme is shown in a time varying CFO environment. We assume the normalized CFOs of all 8 active users including the target user vary sinusoidally (i.e., the Doppler frequency (mobility) varies sinusoidally) with period of 400000 OFDMA symbols (40 seconds) and amplitude of 0.1 ($\nu = 0.1$). According to the sinusoidally varying mobility, the timing offsets without a timing control also vary up to ± 15 samples. Figure 4 shows that even in that environment, a feedback rate (FBR) of 1/20000, which indicates one control message is transmitted in every 20000 OFDMA symbols, is sufficient to suppress the residual CFO within about 3% of the subcarrier spacing. The same FBR is assumed for the periodic ranging for the time synchronization. Here, the timing estimation in a BS is assumed to be perfect and only the delay effect due to the given FBR is taken into account. In Fig. 5, the bit error rate performances for QPSK and 16-QAM modulations are shown with the same simulation conditions used in Fig. 4. As can be seen in Fig. 5, the performance of the proposed scheme is close to the CFO free case without additional preamble overhead.

4. Conclusion

The proposed CFO estimator uses the intrinsic phase shift within an OFDMA symbol by using the two sets of samples from two observation windows after DFT and user separation. The advantage is that it does not require additional preamble symbols and that it allows an arbitrarily interleaved subcarrier assignment. The proposed blind estimator can use the all transmitted symbols and shows better or as good performance as the preamble-based method. Also, it was shown that using the proposed estimation algorithm for the continuous CFO tracking can provide reliable communication even in an environment of rapid CFO variation. The proposed scheme can be adopted to improve the uplink performance of interleaved OFDMA systems such as IEEE 802.16e by simply sending control messages with a long period without changing the physical layer frame structure.

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