2007 International Symposium on Nonlinear Theory and its Applications
NOLTA'07, Vancouver, Canada, September 16-19, 2007

Frequency Offset Compensation Scheme Using Soft Interference Cancellation in Reverse Link of MIMO–OFDMA Systems

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Abstract—In the reverse link of multiple input multiple output orthogonal frequency division multiple access (MIMO–OFDMA) systems, the amounts of carrier frequency offsets (CFOs) are different in each user. CFO spoils the orthogonality among subcarriers, and causes notable degradation in demodulation performance. Thus, we must compensate for CFO accurately. To compensate for CFO of every transmitted signals, the superposed received signals must be separated into each user's component before CFO compensation. The proposed system achieves this by the use of soft parallel interference cancellation (PIC). By using a soft PIC, the received signals can be separated before CFO compensation. Simulation results can show the proposed system effectively compensates for CFO, especially in a high E_b/N_0 region.

1. Introduction

Multiple input multiple output orthogonal frequency division multiple access (MIMO–OFDMA) attracts much attention as a multiple access technique because of its inherent high spectral efficiency and robustness against frequency selective fading channels [1–3]. In the reverse link of MIMO–OFDMA systems, the amounts of carrier frequency offsets (CFOs) are different in each user because every user uses their own local oscillators. CFO spoils the orthogonality among subcarriers, and results in both intercarrier interference (ICI) and multiple access interference (MAI). ICI and MAI significantly degrade the demodulation performance [4–6]. Thus, CFO must be compensated for in MIMO–OFDMA systems. However, CFO compensation schemes proposed in orthogonal frequency division multiplexing (OFDM) systems cannot be applied directly.

In this paper, we propose a CFO compensation scheme using soft interference cancellation in the reverse link of MIMO–OFDMA systems. CFO compensation must be carried out before fast Fourier transform (FFT) processing because the ICI/MAI is generally caused at FFT processing. This implies that the receiver must operate user separation against the time domain superposed and multiplexed OFDM signals like a noise waveform. However, to compensate for CFOs of every transmitted time domain signal, the superposed received signals must be separated into each user's component before CFO compensation. The proposed system can achieve signal separation before CFO compensation by the use of a soft parallel interference cancellation (PIC). By using soft interference cancellation, the received signals can be separated before CFO compensation. Simulation results can show the proposed system effectively compensates for CFO, especially in a high E_b/N_0 region.

2. System Model

We explain a MIMO–OFDMA system considered in this paper. In the system, K users exist and each user has N_p transmit antennas. First, information bits of each user are divided into N_p substreams. Each substream is convolutionally encoded and interleaved. The interleaved encoded sequence is mapped to symbols according to a modulation scheme, and allocated to utilizable subcarriers. Every user generates N_p OFDM signals by using inverse fast Fourier transform (IFFT) and adding guard interval. Finally, every user transmits N_p OFDM signals using N_p transmit antennas, simultaneously.

At the transmission, since every user uses their own local oscillators, carrier frequency cannot be completely synchronized among users. Therefore, the amount of CFO in each received signal is different from others. At the receiver, there are N_q receive antennas. Received signals from multiple users are superposed at each receive antenna. The received signal vector is given by

$$\mathbf{r}(t) = [r_1(t)\; r_2(t)\; \cdots\; r_q(t)\; \cdots\; r_{N_q}(t)]^{\dagger}, \tag{1}$$

$$r_q(t) = \sum_{k=1}^{K}\sum_{p=1}^{N_p}\sum_{l=1}^{L} h_{p,q}^{k}(l)s_p^k(t-l+1)e^{j2\pi\Delta f^k t} + \eta_q(t), \tag{2}$$

where $(\cdot)^{\dagger}$ is the transpose of a vector, $h_{p,q}^k(l)$ is the lth channel response between the pth transmit antenna of the kth user and the qth receive antenna, $s_p^k(t)$ is the kth user's transmit signal of the pth transmit antenna, Δf^k is the normalized CFO (NCFO) of the kth user which is normalized by a subcarrier spacing, and $\eta_q(t)$ is the additive white Gaussian noise (AWGN) of the qth receive antenna. As described in Eq. (2), the receiver receives multiple OFDM signals simultaneously at each receive antenna. This greatly complicates the CFO compensation in MIMO–OFDMA systems because antenna separation as well as user separation is required.

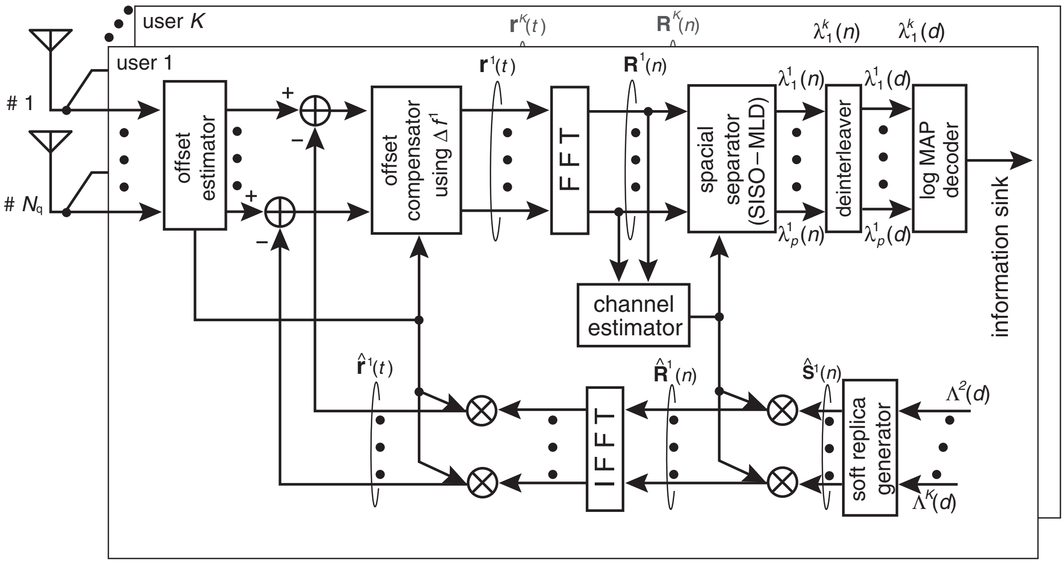

Figure 1: Proposed receiver.

In MIMO–OFDMA systems, we must separate the received signal into each user's signal before the CFO compensation because every signal has different amount of CFO. Furthermore, the CFO compensation must be carried out before FFT processing because the ICI/MAI is generally caused at FFT processing. This implies that the receiver must operate user separation against the time domain superposed and multiplexed OFDM signal. This can be achieved with the aid of the principle of PIC. Thus, we consider to use PIC in this paper. By employing soft PIC, the received signals can be separated before CFO compensation as well as FFT processing.

3. Proposed Scheme

In general, the interference cancellation is carried out by removing the replica of undesired signals from the received signal. To generate the replica signals in MIMO–OFDMA systems, the antenna separation is required which can be achieved after FFT processing. Thus, FFT processing is required before CFO compensation after all. This is contradictory to the requirement for CFO compensation. To mitigate this contradiction, therefore, iteration of the cancellation and FFT processing with CFO compensation is required in the proposed system.

The diagram of the proposed receiver is depicted in Fig. 1. First, the received signal at each receive antenna is distributed into K user-branches. In the ith branch ($i = 1, 2, \ldots, K$), NCFO of the ith user are estimated as $\Delta\hat{f}^i$. There is no signal from a soft replica generator block at first. Thus, each signal is directly fed to an offset compensator block. In the ith branch, CFO compensation is carried out using $\Delta\hat{f}^i$. The compensated signal vector $\mathbf{r}^i(t)$ in the ith branch is given by

$$\mathbf{r}^i(t) = [r_1^i(t)\; r_2^i(t)\; \cdots\; r_q^i(t)\; \cdots\; r_{N_q}^i(t)]^\dagger, \tag{3}$$

$$r_q^i(t) = r_q(t)e^{-j2\pi\Delta\hat{f}^i t} \tag{4}$$

$$= \sum_{k=1}^{K}\sum_{p=1}^{N_p}\sum_{l=1}^{L} h_{p,q}^k(l)s_p^k(t-l+1)e^{j2\pi\delta f^{k,i}t} + \eta_q^{'}(t), \tag{5}$$

where $\delta f^{k,i} = \Delta f^k - \Delta\hat{f}^i$, and $\eta_q^{'}(t) = \eta_q(t)e^{-j2\pi\Delta\hat{f}^i t}$. An N-point FFT is applied to Eq. (5) for demodulation, where N denotes the total number of FFT point. The resultant signal vector is expressed as Eq. (6). In Eqs. (7) and (8) [6], $\mathcal{F}$ is the FFT operator, $S_p^{i,k}(n)$ $(n = 1, 2, \ldots, N)$ is the kth user's signal on the nth subcarrier from the pth transmit antenna, $H_{p,q}^{i,k}(n)$ is the channel response of the kth user's signal on the nth subcarrier from the pth transmit antenna to the qth receive antenna, and $\aleph_q^i(n)$ is the AWGN on the nth subcarrier of the qth receive antenna in the frequency domain.

In Eq. (5), only the CFO of the ith user's signal is compensated for when $\delta f^{k,i}$ $(k = i)$ becomes zero. As a result, CFO of the kth user can be compensated for. However, from Eq. (8), the signals from other users still have CFO

$$\mathbf{R}^i(n) = [R_1^i\; R_2^i\; \cdots\; R_q^i\; \cdots\; R_{N_q}^i]^\dagger, \tag{6}$$

$$R_q^i(n) = \mathcal{F}\left(r_q^i(t)\right) \tag{7}$$

$$= \sum_{p=1}^{N_p}\left(S_p^{i,i}(n)H_{p,q}^{i,i}(n) + \sum_{\substack{k=1\\k\neq i}}^{K} S_p^{i,k}(n)H_{p,q}^{i,k}(n)\frac{\sin\left(\pi\delta f^{k,i}\right)}{\sin\left(\frac{\pi\delta f^{k,i}}{N}\right)}e^{\frac{j\pi\delta f^{k,i}(N-1)}{N}}\right) + \aleph_q^i(n). \tag{8}$$

$\delta f^{k,i}$ ($k \neq i$) because CFO compensation using $\Delta \hat{f}^i$ overcorrects or undercorrects the CFOs of others. Thus, FFT-processed signal in Eq. (8) still has MAI components.

The signals of Eq. (6) are separated at a spacial separator block. In this paper, we use a soft input soft output maximum likelihood detection (SISO–MLD) as the spacial separation scheme [7]. To separate the superposed signal, we use the channel response matrix. The estimated channel matrix is expressed by

$$\hat{\mathbf{H}}(n) = \begin{bmatrix} \hat{H}_{1,1}(n) & \hat{H}_{2,1}(n) & \dots & \hat{H}_{N_q,1}(n) \\ \hat{H}_{1,2}(n) & \hat{H}_{2,2}(n) & \dots & \hat{H}_{N_q,2}(n) \\ \vdots & \vdots & \ddots & \vdots \\ \hat{H}_{1,N_p}(n) & \hat{H}_{2,N_p}(n) & \dots & \hat{H}_{N_q,N_p}(n) \end{bmatrix}. \tag{9}$$

SISO–MLD outputs log likelihood ratio (LLR) using the channel matrix given in Eq. (9). The resultant LLR of bit in the ith branch is calculated, and expressed by

$$\lambda_p^k(n) = \ln \frac{\sum_{v=0}^{2^{U-1}-1} \exp\left(- \frac{||\mathbf{R}^i(n) - \hat{\mathbf{H}}(n)\mathbf{S}^0_{rep}||^2}{2\sigma_n^2} \right)}{\sum_{v=0}^{2^{U-1}-1} \exp\left(- \frac{||\mathbf{R}^i(n) - \hat{\mathbf{H}}(n)\mathbf{S}^1_{rep}||^2}{2\sigma_n^2} \right)}, \tag{10}$$

$$U = N_p \log(Q) \tag{11}$$

$$\mathbf{S}^0_{rep} = \text{map}(\text{s2p}([\text{bin}(v)_{1:u-1}\ 0\ \text{bin}(v)_{u:U-1}])), \tag{12}$$

$$\mathbf{S}^1_{rep} = \text{map}(\text{s2p}([\text{bin}(v)_{1:u-1}\ 1\ \text{bin}(v)_{u:U-1}])), \tag{13}$$

where $|| \cdot ||^2$ denotes the Euclidean norm, σ_n^2 is the noise variance, Q is the number of constellation points, map$(\cdot)$ indicates the mapping, s2p$(\cdot)$ is the serial to N_p parallel conversion, bin$(\cdot)$ is the binary representation of v. Furthermore, bin$(v)_{x:y}$ denotes the elements from the xth to the yth of bin(v). By calculating Eq. (10) in every branch, the proposed receiver obtains the estimates of all users' bits in symbols. However, the estimated symbols are affected by MAI. Thus, to reduce the effect of MAI, error-correction decoding with deinterleaving is carried out. As a result, we can obtain the LLR of the code bits, and can generate replica signals of all users.

At the iterative process, we can use the replica signals. The replica signals for the kth user is generated from the decoded bits in the ith branch ($i = k$). At the soft replica generator, the estimated symbols in the frequency domain are generated through convolutional encoding, interleaving, and symbol mapping. For example, when the modulation scheme is QPSK, the soft symbol $\hat{S}_p(d)$ is calculated as

$$\hat{S}_p(d) = \frac{1}{\sqrt{2}} \left(\tanh\left[\frac{\Lambda_p^k(d+1)}{2} \right] + j \tanh\left[\frac{\Lambda_p^k(d)}{2} \right] \right), \tag{14}$$

where $\Lambda_p^k(d)$ denotes the kth user's *a posterior* LLR of the dth code bit and d is the index of the code bit after deinterleaving. Then the estimated symbols are multiplied by the channel response $\hat{H}_{p,q}(n)$ before IFFT because the channel response is obtained in the frequency domain. The resultant signal vector is given by

$$\hat{\mathbf{R}}^i(n) = [\hat{R}_1^i(n)\ \hat{R}_2^i(n)\ \cdots\ \hat{R}_q^i(n)\ \cdots\ \hat{R}_{N_q}^i(n)]^\dagger, \tag{15}$$

$$\hat{R}_q^i(n) = \sum_{p=1}^{N_p} \hat{S}_p(n)\hat{H}_{p,q}(n). \tag{16}$$

After IFFT, the signals of Eq. (15) are multiplied by CFO information; the resultant signal becomes

$$\hat{\mathbf{r}}^i(t) = [\hat{r}_1^i(t)\ \hat{r}_2^i(t)\ \cdots\ \hat{r}_q^i(t)\ \cdots\ \hat{r}_{N_q}^i(t)]^\dagger, \tag{17}$$

$$\hat{r}_q^i(t) = \mathcal{F}^{-1}\left(\hat{R}_q^i(n)\right) e^{j2\pi\Delta\hat{f}^k t}, \tag{18}$$

where $\mathcal{F}^{-1}$ is the IFFT operator. By subtracting the replica signals from the received signal, we can separate received signals before FFT.

Finally, it should be noted that the replicas at the first iteration are inaccurate, because the decoded bits are affected by residual MAI. Thus, we must do the cancellation iteratively. The resultant signal vector is expressed as

$$\mathbf{r}^{'i}(t) = \mathbf{r}(t) - \hat{\mathbf{r}}^i(t). \tag{19}$$

By cancelling interference signals iteratively, the proposed scheme eliminates MAI components before CFO compensation, and thus the frequency offset compensator block works effectively. As a result, the proposed scheme eliminates residual ICI/MAI components.

Table 1: Simulation parameters

Number of users	4
Number of transmit antennas	2
Number of receive antennas	2
Modulation scheme	QPSK
Number of subcarriers	256
Number of FFT points	256
Length of guard interval	64 [sample]
Packet structure	60 OFDM symbol + 8 pilot OFDM symbol
Normalized Doppler frequency	0.001
Error correcting code	convolutional code
Code rate	1/2
Constraint length	7
Decoder algorithm	log MAP algorithm
Interleaver	random interleaver

4. Simulation Results

We evaluate the performance of the proposed scheme by computer simulation. Table 1 shows the simulation parameters. In this simulation, 64 subcarriers are assigned to each user. To investigate the performance of the proposed scheme under the worst case scenario, interleaved allocation [8] is used as the subcarrier allocation method because it causes a great deal of MAI. The channel model

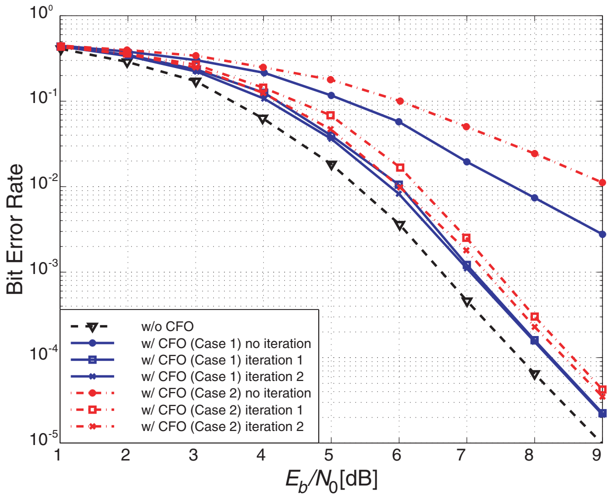

Figure 2: BER performance of the proposed scheme.

used in this evaluation is three-sample-spaced exponentially decaying 10-ray Rayleigh fading. Channel response is estimated, and linearly interpolated using preamble and postamble in a packet. In this simulation, we assume that CFOs are known at the receiver.

Fig. 2 shows the bit error rate (BER) performances of the proposed scheme in the cases with and without CFO. Dashed line indicates the BER performance in the case without CFO. Solid lines and dashed dotted lines indicate the BER performances in the case with CFO. The BER curves are obtained by averaging every user's BERs. We compare the BER performances in the case with CFO and that in the case without CFO. Here, the case without CFO is equivalent to the case when the CFO is perfectly compensated for in the case with CFO. Therefore, the case without CFO is the ideal case regarding the CFO compensation in the proposed scheme.

In the evaluation, we employ two combinations of CFOs: $(\Delta f^1 = 0.1, \Delta f^2 = -0.15, \Delta f^3 = 0.1, \Delta f^4 = -0.15)$ for Case 1, and $(\Delta f^1 = 0.2, \Delta f^2 = -0.1, \Delta f^3 = 0.2, \Delta f^4 = -0.1)$ for Case 2. From the comparison, the performance loss due to Case 1 CFO at the second iteration is about 0.4 [dB], and the performance loss due to Case 2 CFO at the second iteration is 0.6 [dB] both at BER of 10^{-4}. Thus, the BER performance of the proposed scheme in the case with CFO approaches that in the case without CFO in a high E_b/N_0 region. Since the proposed scheme attains the improvement in the precision of the soft replica signal, the proposed scheme can compensate for CFO effectively. On the other hand, in the region of low E_b/N_0, the performance loss is relatively large depending on the combination of CFOs. Since the proposed system is affected by large MAI in such an E_b/N_0 region, bit errors cannot be sufficiently corrected even with the error-correction code. As a result, the precision of replicas is not greatly improved with the progress of iteration.

5. Conclusions

In this paper, we have proposed a frequency offset compensation scheme using parallel interference cancellation in the reverse link of MIMO–OFDMA systems. To compensate for CFO in the reverse link of MIMO–OFDMA, signals from different users must be separated before CFO compensation and FFT. This is because the amount of CFO of each subcarriers is different from others. We achieve this by using the principle of PIC.

In our proposed scheme, we first decode the received signal, and generate the soft replica signals of every user. Next, by subtracting the soft replica signals from the multiplexed received signal, we can separate the received signal into each user's signal before FFT, and thus, we can compensate for CFO individually. By cancelling interference signals iteratively, the proposed scheme eliminates residual ICI/MAI. Simulation results can show the proposed scheme effectively compensates for CFO.

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