We propose a method to compensate the effects of multiple carrier frequency offsets (CFOs) for orthogonal frequency division multiplexing (OFDM) systems. The multiple CFOs, we considered, are caused by several spatially separated frequency-unsynchronized transmitters as in amplify-and-forward relay communication networks or wireless ad hoc sensor networks. Despite its appealing features, OFDM is extremely sensitive to CFO, especially to multiple CFOs. At the receiver, the residual CFO which appears due to incomplete compensation of multiple CFOs destroys orthogonality among subcarriers and brings interchannel interference (ICI). In the proposed scheme, we compensate the CFOs of the received signal by the predesigned value, which results in cancellation of the opposite ICI by each other. This is based on the property that the spectrum of ICI is shown in a subcarrier has similar magnitudes and opposite signs from the fact that spectrum shapes of the ICI are of sinc function. Simulation results are shown to demonstrate the effectiveness of the proposed method.

1. Introduction

In cellular communication systems, the improvement of communication quality of subscribers who are located at the edge of the cell is important to enhance the total quality of service (QoS) [1, 2]. To achieve that, some network schemes like relay communication network and wireless sensor network are proposed. The cooperative relay communication system [3, 4] reduces a radio shadow area by installing some relays in the middle of cell. The relay communications networks are proposed to accomplish the goal for the advanced long term evolution (LTE-A) system [5].

The wireless sensor network [6, 7] consists of autonomous nodes that can be deployed in an area. Sensor nodes are generally equipped with a radio transceiver. Therefore, these nodes help to maintain strong links between transmitter and receiver. Therefore, the researchers have envisioned a wide variety of applications, such as environment monitoring, field surveillance, and structure monitoring.

In these schemes, a subscriber receives the same signal from several transmitters and sensors simultaneously, and multiple carrier offsets (CFOs) may appear at the receiver because the oscillators of transmitters and receiver have different oscillation frequencies each other. Doppler effect also brings the CFO in mobile environments.

Orthogonal frequency division multiplexing (OFDM) has many advantages such as high spectral efficiency and ability of eliminating the effects of multipath propagations [8]. Because of its advantages, OFDM or orthogonal frequency division multiple access (OFDMA) is used in several wireless communication or broadcasting systems such as IEEE 802.11n wireless local area network (WLAN) [9] and IEEE 802.16e [10] systems.

In OFDM systems, the performance depends mainly on how well the orthogonality among the subcarriers is kept at the receiver. Typically, the CFO breaks the orthogonality. The destruction of orthogonality causes interchannel interference (ICI) from adjacent subcarriers. Since the carrier frequency of the recent communication systems is on the order of GHz,
CFOs are commonly on the order of kHz with an oscillator accuracy of a few ppm. Considering the subcarrier spacing of an OFDM system, the order of the CFOs is critical enough to determine the performance of the system. Therefore, mitigating the influence is very challenging in OFDM systems.

There have been many researches to mitigate the effects of CFO in OFDM systems [11–13]. In [11], Moose proposed a method to estimate the CFO by using repetitive preambles. Then, the received signal is multiplied by the inverse rotation factor to compensate the estimated CFO. In [12], Zhao and Häggman proposed a self-cancellation scheme of ICI. The method can be regarded as a precoding approach which adopts a codeword having low ICI. Since the coding rate is less than one, the spectrum efficiency is reduced. Hou and Chen proposed a method which uses a parallel interference cancellation (PIC) scheme to eliminate the ICI in frequency domain [13]. The scheme may be considered as a frequency domain equalizer which consists of a set of prefilters for initial decision and a set of ICI cancellation filters to cancel the ICI. In the scheme, however, the performance may deteriorate severely when the initial decisions are not correct. Furthermore, the scheme increases the system complexity because the PIC needs much additional operations.

Under the condition where the multiple CFOs exist, it is difficult to apply the previously proposed scheme to eliminate the effects of multiple CFOs. The method of [14] can compensate only one CFO, so the residualICI generated by the other CFOs still exists. Furthermore, to cancel the residual ICI, the complex PIC scheme of [13] is required.

In [14] we proposed an ICI cancellation method for a cooperative communication system where the received signal comes from two transmitters simultaneously with different CFOs. The ICI caused by multiple CFOs is to be self-cancelled by time domain processing. In the method, we assumed that the two signals have same path gains.

In this paper, further to the previous study [14], we propose a multiple-CFO compensation scheme for OFDM systems when the paths have different gains. In this condition, the compensation value for CFO, which leads to self-cancellation of ICI, should be changed according to the gain difference. To find the optimal compensation value, we define an object function which maximizes a signal to ICI power ratio (SIR) of received signal.

### 2. System Model

We consider an OFDM system with \( N \) subcarriers, where same signals are transmitted by two spatially separated transmitters simultaneously as amplify-and-forward relay communication network as shown in Figure 1 [3]. The system consists of 3 nodes, a source, a relay, and a destination. Data are transmitted from the source to the destination with the aid of the relay, and the transmission process is accomplished in two time slots. In the first time slot, the source broadcasts the signal to the relay and destination. In the second time slot, source transmits the same signal transmitted in the first time slot to the destination, and relay transmits the signal received in the first time slot to the destination at same time.

![Figure 1: An amplify-and-forward relay communication network.](image)

Then, the destination node (receiver) receives same signals simultaneously with different CFOs.

We assume that timing synchronization is perfectly achieved and the CFOs are previously estimated completely through the methods of, for example, [15, 16], by using the training sequence at the receiver. Without loss of generality, we consider only one OFDM block of signal.

After inverse discrete Fourier transform (IDFT) of data symbols and guard interval insertion, the transmitted time domain baseband signal is represented as

\[
x_n = \frac{1}{N} \sum_{k=0}^{N-1} X_k \exp\left\{\frac{2\pi nk}{N}\right\}, \quad -N_g \leq n \leq N - 1,
\]

(1)

where \( X_k \) is the data symbol transmitted to \( k \)th subcarrier and \( N_g \) is the length of guard interval. The two signals are transmitted through independent channels. The channel output for transmitter \( p \) is given as follows:

\[
y_n^p = x_n \ast h_n^p + w_n^p,
\]

(2)

where \( \ast \) denotes linear convolution operation, and \( h_n^p \) is channel impulse response (CIR) for the \( p \)th transmitter. The received signal is the sum of the two signals from two transmitters. Without timing offsets and CFOs, the time domain received signal is expressed as

\[
y_n = \sum_{p=1}^{2} y_n^p + w_n = \frac{1}{N} \sum_{p=1}^{2} \sum_{k=0}^{N-1} X_k H_k^p \exp\left\{\frac{2\pi nk}{N}\right\} + w_n,
\]

(3)

where \( H_k^p \) denotes the channel gain of \( k \)th subcarrier for the \( p \)th transmitter, and \( w_n \) is additive white Gaussian noise (AWGN) with zero mean and variance \( \sigma^2 \). After discrete Fourier transform (DFT) following guard interval removal, the \( k \)th subcarrier symbol of the received signal is given by

\[
Y_k = \sum_{p=1}^{2} \sum_{k=0}^{N-1} y_n^p \exp\left\{-j\frac{2\pi nk}{N}\right\}, \quad 0 \leq k \leq N - 1.
\]

(4)
If there are CFOs at the receiver, the received signal is given by

\[ r_n = \sum_{p=1}^{2} y_{p}^n \exp \left \{ j \frac{2\pi n \epsilon_p}{N} \right \} \]

\[ = \frac{1}{N} \sum_{p=1}^{2} \sum_{k=0}^{N-1} X_k H_k^{p} \exp \left \{ j \frac{2\pi n (k + \epsilon_p)}{N} \right \} + w_n, \]

where \( \epsilon_p \) means the normalized CFO for the \( p \)th transmitter, which is the ratio of the practical CFO \( \Delta f_p \) to the subcarrier spacing like

\[ \epsilon_p = \frac{\Delta f_p}{T_s}, \]

where \( T_s \) represents useful duration of an OFDM block. From (4) and (5), we can find the frequency domain signal expressed as

\[ R_k = \sum_{p=1}^{2} X_k H_k^{p} \frac{\sin \pi \epsilon_p}{N \sin \left ( \pi \epsilon_p / N \right )} \exp \left \{ j \frac{\pi \epsilon_p (N - 1)}{N} \right \} \]

\[ + \sum_{p=1}^{2} \sum_{l \neq k}^{N-1} \alpha^p_{k,l} X_l H_l^{p} + W_k. \quad (7) \]

The first term on the right-hand side of (7) represents the effects of the CFO on the desired \( k \)th subcarrier symbol, which are attenuation in magnitude and rotation in phase of the symbol. The last term is the DFT of the received AWGN. The second term means ICI introduced into the \( k \)th subcarrier from the other subcarriers. The factor \( \alpha^p_{k,l} \) may be regarded as an interference coefficient which is related to the amplitude of interference into the \( k \)th subcarrier from the \( l \)th subcarrier by the CFO and expressed as

\[ \alpha^p_{k,l} = \frac{\sin \pi (l - k + \epsilon_p)}{N \sin \left ( \pi (l - k + \epsilon_p) / N \right )} \exp \left \{ j \frac{\pi (l - k) \epsilon_p (N - 1)}{N} \right \}. \]

This is a constant determined by the DFT size \( N \), the subcarrier distance \( (l - k) \), and CFO \( \epsilon_p \). As we know, \( \alpha^p_{k,l} \) is maximum for \( |l - k| = 1 \) and decreases as \( |l - k| \) increases.

Figure 2 shows the subcarrier spectrum of the received signals from two transmitters with different CFOs. As shown in \( k \)th subcarrier frequency, there are ICI components from both received signals of transmitters 1 and 2 as represented in (7).

### 3. Proposed Self-Cancellation Scheme

In this paper, we assume that all the CFOs are known to the receiver and smaller than the subcarrier spacing (this means that the coarse estimation and correction of a CFO is not necessary or is done already [17]).

For a single CFO case, one step of time domain CFO compensation is sufficient if the CFO estimate is known a priori. The compensation is accomplished by shifting the carrier frequency of the receiver by the CFO. Although all the CFOs are known, however, it is impossible to compensate the received signal completely for a two-CFO case by simply matching the carrier frequencies.

For example, if we compensate for the CFO \( \epsilon_1 \), the received signal (5) becomes

\[ r'_n = r_n \exp \left \{ -j \frac{2\pi n \epsilon_1}{N} \right \}. \]

Then, after DFT processing, the \( k \)th subcarrier symbol is expressed as

\[ R^1_k = X_k H_k^1 + X_k H_k^2 \frac{\sin \pi \epsilon_d}{N \sin \left ( \pi \epsilon_d / N \right )} \exp \left \{ j \frac{\pi \epsilon_d (N - 1)}{N} \right \} \]

\[ + \sum_{l=0}^{N-1} \sum_{l \neq k} \alpha^{2,1}_{k,l} X_l H_l^1 + W_k, \]

\[ \text{residual ICI} \quad (10) \]

where \( \epsilon_d = \epsilon_2 - \epsilon_1 \) and the interference coefficient \( \alpha^{2,1}_{k,l} \) is

\[ \alpha^{2,1}_{k,l} = \frac{\sin \pi (l - k + \epsilon_d)}{N \sin \left ( \pi (l - k + \epsilon_d) / N \right )} \exp \left \{ j \frac{\pi (l - k) \epsilon_d (N - 1)}{N} \right \}. \]

The first and second terms of the right-hand side of (10) are the desired symbol where the second is attenuated in magnitude and rotated in phase by the CFO difference. The third term represents the residual ICI, and we can see that there are still ICI components caused by the CFO difference \( \epsilon_d \). This is represented graphically in Figure 3. Therefore, to
remove the effects of CFOs completely, additional process is required. There are literatures which compensate for one CFO in time domain and adopt the (PIC) scheme in frequency domain to cancel the residual ICI [13].

In [14], we proposed method, to compensate the received signal for the average value of the two CFOs in time domain. Then, the ICIs are almost cancelled themselves and any additional process may not be required in frequency domain.

Since the shape of the spectrum of a subcarrier is of the sinc function, the sign of the sidelobes appeared alternately as the subcarrier index increases or decreases. This is the key motivation point to propose our method.

If we compensate the received signal with the mid value of \( \epsilon_1 \) and \( \epsilon_2 \), the ICI components from lth (\( l \neq k \)) subcarriers of the two transmitters have similar magnitude with opposite signs as shown in Figure 4. Then, the ICIs are cancelled each other.

If we rotate the received signal by the inverse of the phase caused by the average CFO \( \epsilon_m = (\epsilon_1 + \epsilon_2)/2 \), then (5) becomes

\[
\begin{align*}
    r_m^n &= r_n \cdot \exp \left\{ -j \frac{2\pi n \epsilon_m}{N} \right\} \\
    &= \frac{1}{N} \sum_{p=1}^{N-1} \sum_{k=0}^{N-1} X_k H_k^p \exp \left\{ j \frac{2\pi n (k + \epsilon_p - \epsilon_m)}{N} \right\} + w_m^n, \\
\end{align*}
\]

where \( w_m^n = w_n \cdot \exp[-j2\pi n \epsilon_m/N] \). The kth subcarrier output of the DFT is expressed as

\[
R_k^m = \sum_{p=1}^{2} X_k H_k^p \sin \left[ \frac{\pi (\epsilon_p - \epsilon_m)}{N} \right] \frac{N}{\sin \left[ \frac{\pi (\epsilon_p - \epsilon_m)}{N} \right]} \exp \left\{ j \frac{\pi (\epsilon_p - \epsilon_m)(N-1)}{N} \right\} \\
+ \sum_{p=1}^{N-1} \sum_{l=0}^{N-1} a_{kl}^{pm} X_l H_l^p + W_k^m, \\
\]

where, as shown in Figure 4, the interference coefficient has the property of

\[
\alpha_{kl}^{1m} = -\alpha_{kl}^{2m}, \quad \text{for } l \neq k. \tag{14}
\]

Therefore, if we do not take the channel gain into account, the ICI components in (13) are cancelled each other. Despite the preceding at the transmitter, which makes the transmitted signal undergo flat fading, the two paths may have different path gains due to the limited transmission power. That means \( H_k = g H_1^k \), where the \( g \) is a ratio of two path gains.

In this case, (14) does not hold anymore as illustrated in Figure 5 and needs to select new CFO compensation value. To obtain an optimal compensation value with different path gains, we define signal to interference power ratio (SIR) as an object function to be maximized as follows:

\[
\text{argmax}_{\epsilon_c} J_k(\epsilon_c) = \text{argmax}_{\epsilon_c} \frac{E \left[ \sum_{p=1}^{2} \left( \alpha_{kl}^{1m} H_k^p X_k^2 \right) \right]}{E \left[ \sum_{p=1}^{2} \sum_{l=0}^{N-1} \left( \alpha_{kl}^{2m} H_l^p X_l^2 \right) \right]}. \tag{15}
\]

There is a function of interference coefficient only which is determined by CFOs. Therefore, we can easily find the compensation value once the CFOs are known a priori.

4. Simulation Results

Through this section, we verify the effectiveness of the proposed ICI cancellation method. The following results are based on the modulation of quadrature phase shift keying (QPSK) and DFT size \( N \) of 128. All the CFOs are assumed to be estimated a priori.

Figure 6 shows the SIR curves of (15) with a different \( g \) where the CFOs of two signals are \([\epsilon_1, \epsilon_2] = [0, 0.15] \); that is, the CFO difference \( \epsilon_d \) is 0.15. In the case of a small \( g \), the ICI which comes from the second signal can be negligible, and the SIR depends mostly on its denominator value. Therefore, the maximum value of (15) is seen when the \( \epsilon_c \) gets close to \( \epsilon_1 \). In the same vein, when the \( g \) is close to 1, the SIR is maximized where \( \epsilon_c \) approaches \((\epsilon_1 + \epsilon_2)/2 \) because that is the same case as explained in (12) to (14).

Figure 7 illustrates the curves which link between the optimal CFO compensation values that maximize the (15) with a different \( \epsilon_d \) as \( g \). Similar to Figure 5, if the \( g \) approaches 1, every curve converges to each of \( \epsilon_c \approx (\epsilon_1 + \epsilon_2)/2 \). On the contrary, when the \( g \) comes close to 0.1, the curves converge to each of \( \epsilon_c \approx \epsilon_1/10 \).

Figures 8 and 9 show comparisons of the bit error rate (BER) performance of proposed method with the mid-value compensation scheme of [14].

Figure 8 represents the BER under the condition when \( \epsilon_d = 0.1 \) and \( g = [0.5, 0.75] \). The performance gets better by approximately 0.63 and 0.18 dB where \( g = 0.5 \) and \( g = 0.75 \), respectively, than the mid-value compensation scheme when using the proposed method at BER of 10^{-5}. As the result shows, the performance improvement is greater where the \( g \) is small. It is because the smaller \( g \) means the greater difference between two path gains, and the greater difference
Figure 4: The subcarrier spectrum of two OFDM signals with compensation of $\epsilon_m = (\epsilon_1 + \epsilon_2)/2$.

Figure 5: The spectrum of the subcarriers of the two OFDM signals with different CFOs and different path gains.

Figure 6: SIR curves of (15) where $[\epsilon_1, \epsilon_2] = [0, 0.15]$, $\epsilon_d = 0.15$.

Figure 7: The curves of optimal CFO compensation value $\epsilon_c$ which maximize (15) with difference $\epsilon_d$.

brings a longer distance between the optimal value and the mid-value of two CFOs.

Figure 9 indicates the BER under the condition when $\epsilon_d = 0.2$ which is relatively large. Performance results of proposed method are better than the mid-value compensation scheme by about 2.3 dB and 0.7 dB at BER of $10^{-5}$. The performance improvements are generally greater than the case of $g = 0.5$.

From Figures 6 and 9, we have verified that it is possible to effectively estimate the optimal CFO compensation value in any CFOs and path gains by using the object function of (15).

In Figures 10 and 11, the graphs show the BER performance comparison between the proposed scheme with the method that compensates only by $\epsilon_1$ and PIC-combined method under the condition of AWGN channel. The PIC-combined method is expressed as

$$y'_n = y_ne^{-j(2\pi\epsilon_1n/N)}.$$

(16)
Middle value, $J_k(\epsilon_c)$, ($g = 0.5$) Maximize $\epsilon_c = (\epsilon_1 + \epsilon_2)/2$, ($g = 0.5$)

Maximize $J_k(\epsilon_c)$, ($g = 0.75$)

Middle value, $\epsilon_c = (\epsilon_1 + \epsilon_2)/2$, ($g = 0.75$)

Figure 8: BER performance comparison of the proposed optimal CFO compensation value that maximizes $J_k(\epsilon_c)$ with mid-value of $\epsilon_c = (\epsilon_1 + \epsilon_2)/2$ in $\epsilon_d = 0.1$ and $g = \{0.5, 0.75\}$.

Figure 9: BER performance comparison of the proposed optimal CFO compensation value that maximizes $J_k(\epsilon_c)$ with mid-value of $\epsilon_c = (\epsilon_1 + \epsilon_2)/2$ in $\epsilon_d = 0.2$ and $g = \{0.5, 0.75\}$.

Then the DFT output becomes

$$Y'_k = X_kH_k \left\{ 1 + g \frac{\sin \left[ \frac{\epsilon_d/N}{N} \right]}{N \sin \left[ \frac{\epsilon_d/N}{N} \right]} e^{i\pi r_c (N-1)/N} \right\}$$

$$+ \sum_{l=0}^{N-1} \left( a_{l,k}^{2,1} \right) H_lX^2_l + W_k', \quad (17)$$

where $\epsilon_d = \epsilon_2 - \epsilon_1$. The PIC is the method which eliminates the residual ICI terms in (17), as follows:

$$\tilde{Y}_l = \text{dec}(Y_l),$$

$$\tilde{Y}_k = Y_k + \sum_{l=-K}^{K} \tilde{Y}_l a_{l,k}, \quad (18)$$

where the $\text{dec}()$ represents the decision function and the $K$ means a range of cancellation. In this paper, we set the $K$ to 5.

Figure 10 illustrates the performances in the case of $\epsilon_d = 0.05$ and $g = \{1, 0.5\}$. In this case, every scheme has similar performances. It is difficult to show outstanding abilities to compensate or cancel the ICIs because the small $\epsilon_d$ brings less ICIs. To be specific, at the BER of $10^{-5}$, the performance of proposed method is better by about 0.25 dB and 0.16 dB than $\epsilon_1$ compensation scheme and the case of [14], respectively.

Figure 11 is the case of $\epsilon_d = 0.2$. Generally, the performance of case of $g = 0.5$ is better than case of $g = 1$. The reason is that the residual ICI of the case of $g = 0.5$ is less than the case of $g = 1$. Even though this residual ICI is eliminated by PIC or cancelled out itself, it is impossible to be cancelled out perfectly. The case of $g = 0.5$, therefore, shows a tendency to have a better performance than the case of $g = 1$.

Figures 12 and 13 show performances in the multipath channel which is nonfading and 20-path randomly generated. In a transmitter, we adopted a precoding scheme for channel equalization, and then the frequency selectivity of a channel can be considered as a flat-fading.

In Figure 12, the graph represents the performance where $\epsilon_d = 0.05$ and $g = \{1, 0.5\}$. Generally, three schemes
show similar performances. Unlike AWGN case which shows almost the same performance regardless of $g$, however, the case of $g = 1$ has better performance than the case of $g = 0.5$. This phenomenon can be interpreted such that the power of the received signal in the case of $g = 1$ is larger than the case of $g = 0.5$.

Figure 11: BER performance comparison of the proposed method with CFO compensation by only $\epsilon_1$ and [14] where $\epsilon_d = 0.2$ and $g = \{0.5, 1\}$ in AWGN channel.

Figure 12: BER performance comparison of the proposed method with CFO compensation by only $\epsilon_1$ and [14] where $\epsilon_d = 0.05$ and $g = \{0.5, 1\}$ in multipath channel.

Figure 13: BER performance comparison of the proposed method with CFO compensation by only $\epsilon_1$ and [14] where $\epsilon_d = 0.2$ and $g = \{0.5, 1\}$ in multipath channel.

In Figure 13, performance is shown in case of $\epsilon_d = 0.2$. The case of $g = 1$ has better performance than the case of $g = 0.5$ in common with the case of $\epsilon_d = 0.05$. The proposed method is about 2 dB and 2.7 dB better than the PIC scheme and $\epsilon_1$ compensation scheme, respectively, where $g = 1$ at the BER of $10^{-5}$. In the case of $g = 1$, the proposed scheme is about 2.2 dB and 2.8 dB better than the PIC scheme and $\epsilon_1$ compensation scheme, respectively.

5. Conclusions

In this paper, we propose a simple ICI self-cancellation method for OFDM systems with multiple CFOs resulting from the discrepancies of oscillator frequency of the two transmitters and receiver. In the previous work of author, we compensate the CFO of the received signal with the mid-value of two CFOs. To deal with the received signal with different path gains, in this paper, we use the compensation value maximizing the SIR in time domain processing. Simulation results show that the proposed method can effectively cancel the ICI introduced by multiple CFOs.

Conflict of Interests

The authors declare that there is no conflict of interests regarding the publication of this paper.

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References

[1] Y. Fan and J. Thompson, “MIMO configurations for relay channels: theory and practice,” IEEE Transactions on Wireless Communications, vol. 6, no. 5, pp. 1774–1786, 2007.

[2] H. Bölcskei, R. U. Nabar, Ö. Yılmaz, and A. J. Paulraj, “Capacity scaling laws in MIMO relay networks,” IEEE Transactions on Wireless Communications, vol. 5, no. 6, pp. 1433–1444, 2006.

[3] C. Conne and I.-M. Kim, “Outage probability of multi-hop amplify-and-forward relay systems,” IEEE Transactions on Wireless Communications, vol. 9, no. 3, pp. 1139–1149, 2010.

[4] A. El-Keyi and B. Champagne, “Adaptive linearly constrained minimum variance beamforming for multiuser cooperative relaying using the kalman filter,” IEEE Transactions on Wireless Communications, vol. 9, no. 2, pp. 641–651, 2010.

[5] S. Stefania and T. Issam, LTE—The UMTS Long Term Evolution, John Wiley & Sons, 2009.

[6] A. Arora, R. Ramnath, E. Ertin et al., “ExScal: Elements of an extreme scale wireless sensor network,” in Proceedings of the 11th IEEE International Conference on Embedded and Real-Time Computing Systems and Applications (RTCSA ’05), pp. 102–108, August 2005.

[7] G. Barrenetxea, F. Ingelrest, G. Schaefer, and M. Vetterli, “The hitchhiker’s guide to successful wireless sensor network deployments,” in Proceedings of the 6th ACM Conference on Embedded Networked Sensor Systems (SenSys ’08), pp. 43–56, November 2008.

[8] J. A. C. Bingham, “Multicarrier modulation for data transmission: an idea whose time has come,” IEEE Communications Magazine, vol. 28, no. 5, pp. 5–14, 1990.

[9] IEEE Std. 802.11n-2009, “Part II: wireless LAN medium access control (MAC) and Physical Layer (PHY) specifications—amendment 5: enhancements for higher throughput,” Tech. Rep. ct-502, 2009.

[10] IEEE Std. 802.16e-2005, Part 16: Air Interface for Fixed and Mobile Broadband Wireless Access Systems, 2005.

[11] P. H. Moose, “Technique for orthogonal frequency division multiplexing frequency offset correction,” IEEE Transactions on Communications, vol. 42, no. 10, pp. 2908–2914, 1994.

[12] Y. Zhao and S.-G. Håggman, “Intercarrier interference self-cancellation scheme for OFDM mobile communication systems,” IEEE Transactions on Communications, vol. 49, no. 7, pp. 1185–1191, 2001.

[13] W.-S. Hou and B.-S. Chen, “ICI cancellation for OFDM communication systems in time-varying multipath fading channels,” IEEE Transactions on Wireless Communications, vol. 4, no. 5, pp. 2100–2110, 2005.

[14] Y.-J. Won and B.-S. Seo, “Compensation of multiple carrier frequency offsets in amplify-and-forward cooperative networks,” in Proceedings of the IEEE 24th Annual International Symposium on Personal, Indoor, and Mobile Radio Communications (PIMRC ’13), pp. 159–163, London, UK, September 2013.

[15] M. J. F. G. Garcia and J. M. Páez-Borrero, “Tracking of time misalignments for OFDM systems in multipath fading channels,” IEEE Transactions on Consumer Electronics, vol. 48, no. 4, pp. 982–989, 2002.

[16] A. Filippi, S. A. Husen, and P. van Voorhuisen, “Effects of time synchronization on OFDM systems over time-varying channels,” in Proceedings of the IEEE 17th International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC ’06), pp. 1–5, September 2006.

[17] B.-S. Seo, S.-C. Kim, and J. Park, “Fast coarse frequency offset estimation for OFDM systems by using differentially modulated subcarriers,” IEEE Transactions on Consumer Electronics, vol. 48, no. 4, pp. 1075–1081, 2002.
