PAPER

Preamble-Assisted Estimation for Frequency-Dependent I/Q Mismatch in Direct-Conversion OFDM Receivers

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SUMMARY In direct-conversion orthogonal frequency division multiplexing (OFDM) receivers, the impact of frequency-dependent I/Q mismatch (IQ-M) with carrier frequency offset (CFO) must be considered. A preamble-assisted estimation is developed to circumvent the frequency-dependent IQ-M with CFO. The results of a simulation and an experiment show that the proposed method could provide good estimation efficiency and enhance the system performance. Moreover, the proposed scheme is compatible with current wireless local area network standards.

key words: *carrier frequency offset, direct-conversion receiver, I/Q mismatch, orthogonal frequency division multiplexing (OFDM), wireless local area network (WLAN)*

1. Introduction

Orthogonal frequency division multiplexing (OFDM) is a spectrally efficient technique for high-speed wireless communications. OFDM systems are, however, susceptible to imperfect synchronization and non-ideal front-end effects, which cause serious performance degradation. Most OFDM systems are highly sensitive to carrier frequency offset (CFO), which is the frequency mismatch of local oscillators between the transmitter and the receiver. CFO introduces inter-carrier interference in OFDM systems due to the loss of orthogonality between sub-carriers. Recent research has also focused on developing monolithic OFDM receivers, particularly low-cost technology. Direct-conversion architecture is one potential candidate for simple integration among different architectures. However, direct-conversion receivers suffer from mismatch between the I and Q channels, such as I/Q mismatch (IQ-M) of non-ideal radio frequency (RF) circuits [1]–[3]. Specifically, IQ-M arises when the phase and gain differences between I and Q branches are not exactly 90° and 0 dB, respectively. IQ-M introduces the image interference into the desired signal which degrades the system performance.

Due to the impairment in the analog components, the low-pass filters (LPFs) of I and Q channels are not identical. The mismatched LPFs result in frequency-dependent IQ-M. Frequency-dependent IQ-M means that the imbalances can vary with frequency. In an OFDM system with frequency-dependent IQ-M, the IQ-M parameters for every sub-carrier are different. Several schemes have been proposed to estimate constant IQ-M [4]–[7]. Although these methods can work well under constant IQ-M, they do not consider non-ideal LPFs and CFO phenomenon. Therefore, these methods are not robust to frequency-dependent IQ-M and CFO. Estimation for frequency-dependent IQ-M has also been studied in the open literature [8]–[11], [14], [17]. The effect and analysis of frequency-dependent IQ-M are given in [8], [9]. Xing et al. [10] presented a method for frequency-dependent IQ-M. An optimal training sequence for frequency-dependent IQ-M has been proposed in [11]. However, these two methods have specific packet formats, making them incompatible with current wireless local area network (WLAN) standards, such as IEEE 802.11a/g [12], [13]. Cetin et al. proposed an adaptive self-calibrating image rejection receiver [14]. This method needs a long converged time. Therefore it is not suitable for packet-based WLAN systems because the packet length is always limited.

In practice, frequency-dependent IQ-M and CFO arise simultaneously. However, only some references consider the frequency-dependent IQ-M with CFO. This study mainly concentrates on the estimation of frequency-dependent IQ-M with CFO. To maintain and realize systems with imperfect front-end modules, a preamble-assisted estimation scheme is developed. Simulation results show that the performance loss is in the range of 1.6 to 1.8 dB at 10^{-4} bit error rate. Experiment results demonstrate that the proposed method could overcome joint impairments of frequency-dependent IQ-M and CFO, enabling a high performance receiver. The remainder of the paper is organized as follows. Section 2 introduces the system model. Section 3 then presents the proposed method. Simulation and experiment results are shown in Sect. 4. Conclusions are finally drawn in Sect. 5.

2. System Model

Figure 1 depicts the block diagram of a direct-conversion

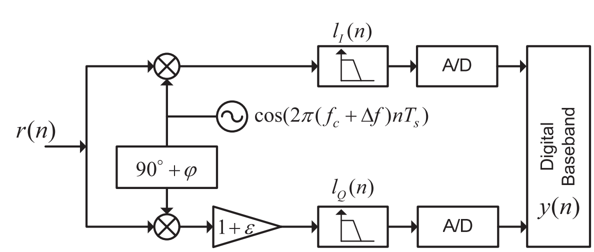

Fig. 1 Direct-conversion receiver with I/Q mismatch and CFO.

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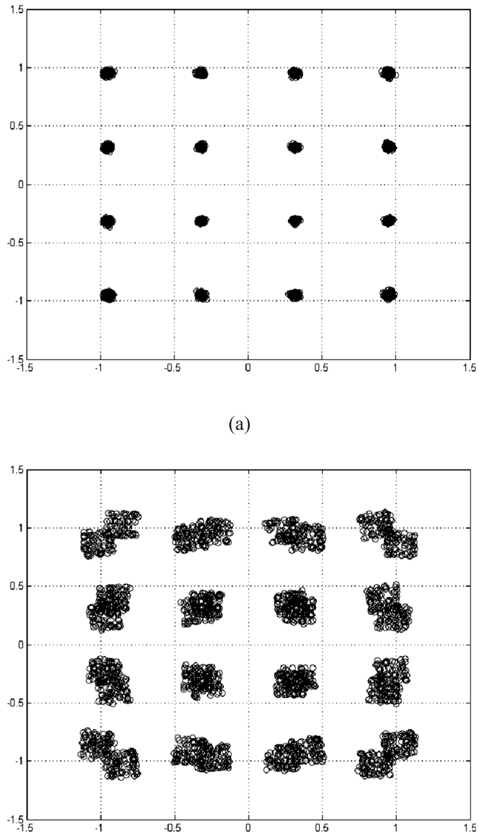


(a)

(b)

Fig. 2 16-QAM constellation. (a) Without I/Q mismatch. (b) Gain error: 1 dB, Phase error: 10°.

architecture. The received RF signal is down-converted to baseband, and then filtered out by the LPFs. Let $r(n)$ and $w(n)$ be the received signal and additive white Gaussian noise, respectively. The baseband signal with CFO Δf and IQ-M is given by [8], [10]

$$\begin{aligned} y(n) \;=\; & ((\alpha\psi(n) + \beta^*\xi(n)) \otimes r(n))\, e^{j2\pi\cdot\Delta f\cdot nT_s} \\ & + ((\beta\psi(n) + \alpha^*\xi(n)) \otimes r^*(n))\, e^{-j2\pi\cdot\Delta f\cdot nT_s} \\ & + w(n) \end{aligned} \tag{1}$$

where α and β denote the constant IQ-M parameters, and $\psi(n)$ and $\xi(n)$ represent the frequency-dependent IQ-M parameters. Note that the phase rotation is inversed in the direction between original signal and the conjugate signal if the CFO is present. The signal $r(n)$ can be further expressed as $r(n) = h(n) \otimes x(n)$, where $x(n)$ and $h(n)$ are the transmitted signal and channel impulse response, respectively. The symbol "⊗" represents the convolution operator. From Eq. (1), the received signal can be regarded as the original signal added to the conjugate signal. The constant IQ-M parameters are expressed as [5]–[7]

$$\begin{aligned} \alpha &= 0.5\left(1 + (1+\varepsilon)\, e^{-j\varphi}\right) \\ \beta &= 0.5\left(1 - (1+\varepsilon)\, e^{j\varphi}\right) \end{aligned} \tag{2}$$

where ε and φ denote the constant amplitude and phase mismatch, respectively. If neither gain nor phase error exists, α remains at unity, and β decreases to zero. Moreover, the average filter response and response mismatch are defined as [8], [10]

$$\begin{aligned} \psi(n) &= 0.5\,(l_I(n) + l_Q(n)) \\ \xi(n) &= 0.5\,(l_I(n) - l_Q(n)) \end{aligned} \tag{3}$$

where $l_I(n)$ and $l_Q(n)$ denote the LPFs of I and Q channels. If the LPFs of I and Q channels are identical, the response mismatch can be ignored. The effect of constant IQ-M on the 16-QAM constellation is depicted in Fig. 2. As a result of the IQ-M, the constellation is distorted severely. It implies that IQ-M can limit the ability of the receiver to achieve sufficient performance especially for high constellation size, e.g., 64-QAM constellation.

3. I/Q Mismatch Estimation

Our goal is to estimate the frequency-dependent IQ-M with CFO. The basic strategy for extracting IQ-M parameters is to employ long training symbols. The packet format used in this method is based on IEEE 802.11a/g standards. Let $x(n)$ denote a long training symbol distorted by IQ-M and CFO, as defined in Eq. (1). After N-point fast Fourier transform (FFT), the received signal is given by

$$\begin{aligned} Y_k \;&=\; \mathrm{FFT}_N\{y(n)\} \\ &= A_k\,(R_kD_k + ICI_k) + B_k\left(R^*_{-k}D_{-k} + ICI_{-k}\right) + W(k) \\ &= A_k\,(H_kX_kD_k + ICI_k) + B_k\left(H^*_{-k}X^*_{-k}D_{-k} + ICI_{-k}\right) \\ &\quad + W(k) \end{aligned} \tag{4}$$

where D_k is the distortion which is accompanied with intercarrier interference (ICI) due to CFO [15]. X_k is the data sub-carrier and H_k stands for the channel frequency response (CFR). Both X_{-k} and H_{-k} denote the frequency mirror image parts. A_k and B_k can be determined by

$$\begin{aligned} A_k &= \mathrm{FFT}_N\{\alpha\psi(n) + \beta^*\xi(n)\} \\ B_k &= \mathrm{FFT}_N\{\beta\psi(n) + \alpha^*\xi(n)\} \end{aligned} \tag{5}$$

The variables D_k, D_{-k}, ICI_k and ICI_{-k} are defined by

$$D_k = \left\{\frac{(\sin\pi\omega)}{N\,(\sin\pi\omega/N)}\right\} e^{j\pi\omega(N-1)/N} \tag{6}$$

$$D_{-k} = \left\{\frac{(\sin\pi\omega)}{N\,(\sin\pi\omega/N)}\right\} e^{j\pi(-\omega)(N-1)/N} \tag{7}$$

$$\begin{aligned} ICI_k = & \sum_{\substack{k=-K \\ l\neq k}}^{K} X_lH_l \left\{\frac{(\sin\pi\omega)}{N\,(\sin\pi\,(l-k+\omega)/N)}\right\} \\ & \cdot e^{j\pi\omega(N-1)/N} e^{-j\pi(l-k)/N} \end{aligned} \tag{8}$$

$$\begin{aligned} ICI_{-k} = & \sum_{\substack{k=-K \\ l\neq -k}}^{K} X^*_lH^*_l \left\{\frac{-(\sin\pi\omega)}{N\,(\sin\pi\,(l-k-\omega)/N)}\right\} \\ & \cdot e^{j\pi(-\omega)(N-1)/N} e^{-j\pi(l-k)/N} \end{aligned} \tag{9}$$

where ω is the normalized frequency offset, i.e., $\omega = \Delta f/(1/NT_s)$. Equation (4) clearly indicates that IQ-M results in a mutual interference between symmetric subcarriers in OFDM systems, as shown in Fig. 3. Let $x(n)$ and $x(n+N_l)$ represent two consecutive long training symbols, as defined in Eq. (1), where N_l is the samples of a long training symbol. In practice, IQ-M not only introduces unwanted image interference into the desired signal, but also restricts the accuracy of CFO estimation. A basic strategy for computing the CFO is to employ two repeated training symbols [15]. However, the estimation error increases when the gain or phase error is not zero. In [16], a pseudo-CFO (P-CFO) algorithm, which rotates three training symbols by adding extra frequency offset into the received sequence, is developed to improve CFO estimation. Since the P-CFO method can estimate the frequency offset more accurately than the two-repeat preamble-based scheme, it is adopted to estimate the CFO value in this work. The two training symbols are then rotated by the estimated CFO. Because the long training symbols are periodic, the received signals, $r(n)$ and $r(n+N_l)$, can be replaced by $r(n)$. Therefore, the following equations hold:

$$y_1(n) = \frac{x(n)e^{-j(2\pi\Delta f nT_s)} - x(n+N_l)e^{-j(2\pi\Delta f(n+N_l)T_s)}}{e^{-2j(2\pi\Delta f nT_s)} - e^{-2j(2\pi\Delta f(n+N_l)T_s)}}$$

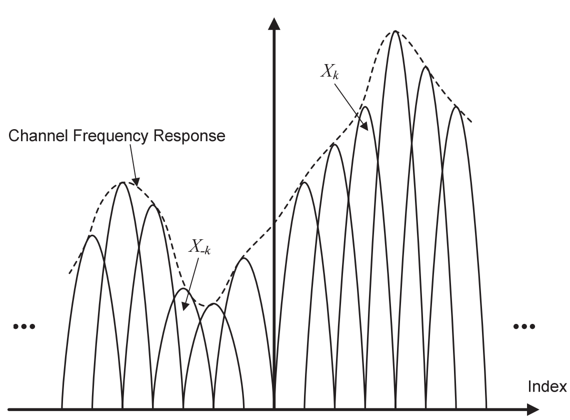


Fig. 3 Mutual interference due to I/Q mismatch.

$$= (\beta\psi(n) + \alpha^*\xi(n)) \otimes r^*(n) \tag{10}$$

$$y_2(n) = \frac{x(n)^* e^{-j(2\pi\Delta f nT_s)} - x(n+N_l)^* e^{-j(2\pi\Delta f(n+N_l)T_s)}}{e^{-2j(2\pi\Delta f nT_s)} - e^{-2j(2\pi\Delta f(n+N_l)T_s)}}$$

$$= ((\alpha\psi(n) + \beta^*\xi(n)) \otimes r(n))^* \tag{11}$$

The IQ-M parameters can be obtained as follows.

$$\frac{B_k}{A^*_{-k}} = \frac{\mathrm{FFT}_N\{y_1(n)\}}{\mathrm{FFT}_N\{y_2(n)\}} \tag{12}$$

Equations (10) and (11) indicate that the two training symbols are multiplied by the estimated CFO and then are subtracted to eliminate the image interference. Once the IQ-M parameters are estimated, the effect of IQ-M could be corrected by the frequency-domain compensation. With the estimated IQ-M parameters, the frequency domain compensation is given by

$$Z_k = \mathrm{FFT}_N\left\{x(n)e^{-j(2\pi\Delta f nT_s)}\right\} - \frac{B_k}{A^*_{-k}}\mathrm{FFT}_N\left\{x^*(n)e^{-j(2\pi\Delta f nT_s)}\right\}$$

$$= \underbrace{\left(A_k - \frac{B_k B^*_{-k}}{A^*_{-k}}\right)}_{compensation\ gain} H_k X_k \tag{13}$$

The compensation gain could be balanced by the channel equalizer. Moreover, the proposed method could be directly applied to IEEE 802.11a/g standards because it does not require any special packet format.

In the open literature [9], [11], [17], most of them consider the IQ-M only. Therefore, these methods may not be suitable for joint CFO and IQ-M. In other words, these methods can tolerate slight CFO value. By contrast, the proposed method is capable of tolerating large CFO value. In the condition of slight CFO value, the inter-carrier interference is smaller than large CFO value. For instance, if the estimated CFO value is within ±2 ppm, the existing methods [9], [11], [17] for frequency-dependent IQ-M could be adopted. In the frequency domain, phase tracking is still performed. Since some rotation operations (CFO estimation and compensation) are eliminated, this approach can maintain the system performance with reasonable complexity.

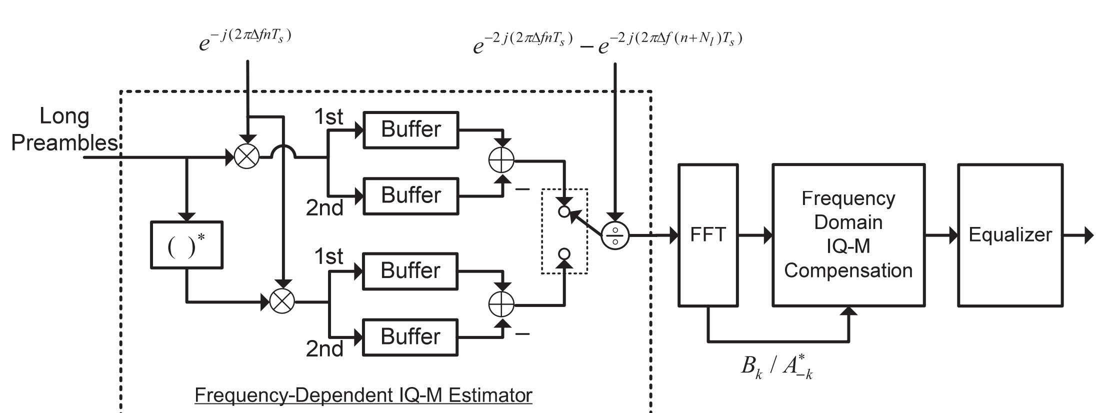


Fig. 4 The proposed frequency-dependent IQ-M estimation architecture.

Figure 4 shows the frequency-dependent IQ-M estimation architecture. The estimator begins to work when long training symbols (preambles) arrive. Firstly, the long preambles are rotated by the estimated CFO. To reduce the image interference due to the IQ-M effect, the first training symbol should subtract the second training symbol. At the same time, the conjugate preamble follows the same procedure described above. After calculating the difference between the first and the second preambles, a complex divider is used to eliminate the phase rotation. Finally, these estimated parameters are applied to compensate for the frequency-dependent IQ-M in the frequency domain.

4. Simulation and Experiment Results

On both the simulation and experiment results, the CFO compensation scheme in [16] is applied. A typical OFDM system for wireless LAN is adapted as the referred specification to evaluate the performance [12], [13]. Similar to IEEE 802.11a/g [12], [13], the parameters of the system employed are as follows: FFT size is 64; the cyclic prefix (CP) contains 16 samples; system bandwidth is 20 MHz; the carrier frequency is 2.4 GHz. The packet format is shown in Fig. 5. The preamble includes two parts. The first part is composed of 10 repetitions of a short training signal with a symbol period of 0.8 μs. The second part of the preamble consists of one cyclic prefix with 1.6 μs and two successive long preambles. The period of a long preamble is 3.2 μs.

The number of taps of frequency-selective fading is of order 8, simulating as an i.i.d. complex Gaussian random variable. All data are modulated by 16-QAM and 64-QAM. The CFO is set at 50 ppm, and the constant IQ-M parameters are 1 dB gain error and 10° phase error. Two-tap FIR filters are adopted to model the frequency-dependent IQ-M [10]. These two LPFs are modeled as

$$\begin{aligned} l_I(n) &= 0.1\delta(n) + \delta(n-1) \\ l_Q(n) &= 0.01\delta(n) + 0.9\delta(n-1) \end{aligned} \tag{14}$$

The Z-transform results of these two LPFs are $L_I(z) = 0.1 + z^{-1}$ and $L_Q(z) = 0.01 + 0.9z^{-1}$, respectively. Figures 6 and 7 show the bit error rate (BER) and packet error rate (PER) performance. In these two figures, the following scenarios are considered. "Multipath only" legend refers to the system with perfect RF front-end and "without comp" means that the estimation scheme is not applied to the system with frequency-dependent IQ-M. "FD IQ-M" legend means that the estimation scheme is applied to the system with frequency-dependent IQ-M and "constant IQ-M" means that the estimation scheme is applied to the system with constant IQ-M. Figures 6 and 7 show that the degradation in BER and PER owing to IQ-M was significant, particularly for 64-QAM modulation. After compensation, the performance under the condition of constant IQ-M was close to the case with multipath only. The SNR loss of the proposed scheme for frequency-dependent IQ-M is in the range of 1.6 to 1.8 dB at 10^{-4} BER. This finding implies that the proposed method is robust to the conditions of both frequency-dependent IQ-M and CFO in fading envi-

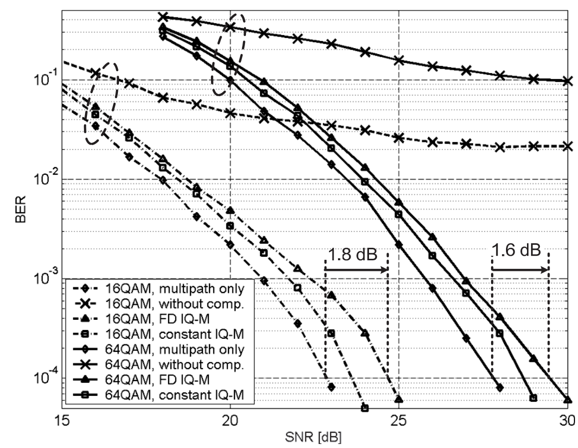


Fig. 6 BER performance of 16-QAM and 64-QAM modulation.

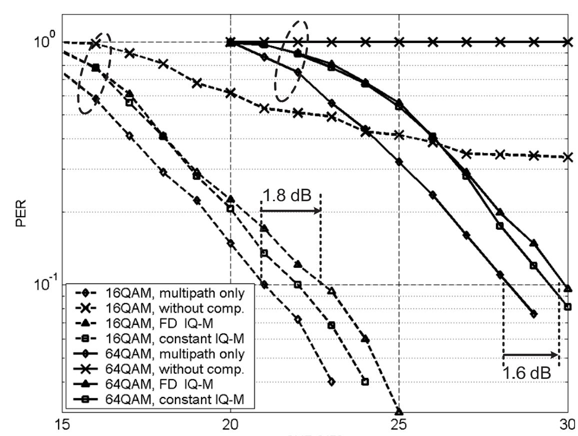


Fig. 7 PER performance of 16-QAM and 64-QAM modulation.

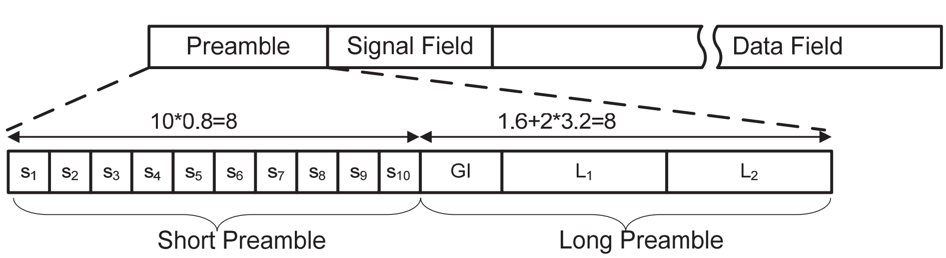


Fig. 5 The packet format.

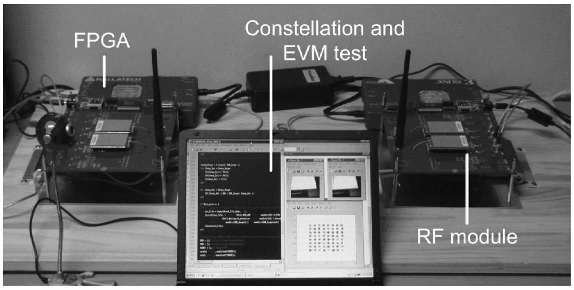


Fig. 8 Experiment setup for verification.

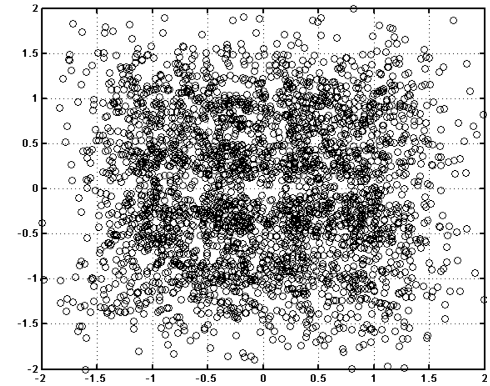
Table 1 Experiment parameters.

Parameters	Value
RF carrier frequency	2.4 GHz
RF power level	−40 dBm
ADC/DAC resolution	14 bits
(I)FFT	64 points
Modulation	16 QAM

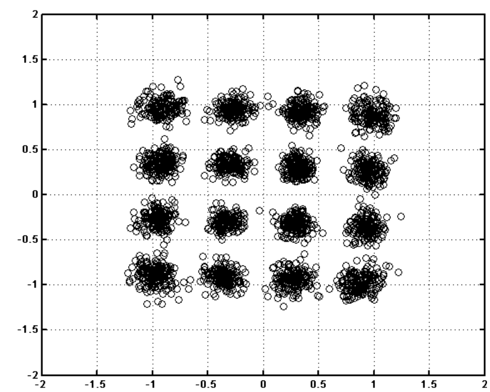
ronments.

Figure 8 displays the experiment setup for verification. Table 1 lists the experiment parameters. The in-house RF front-ends are used in the transmitter as well as in the receiver. Therefore, there is also IQ-M at the transmitter part. Some compensation methods for transmitter IQ-M are available in the open literature [5], [18]. Since our goal is to estimate the receiver IQ-M with CFO, the pre-compensation scheme is applied to reduce the transmitter IQ-M effect. The basic concept is that a training sequence is utilized to estimate the IQ-M. After calculating the IQ-M parameters, the transmitter can compensate for the transmitter IQ-M using the baseband processing. The CFO is set at 50 ppm. The constant IQ-M parameters at the receiver part are 1 dB gain error and 10° phase error. At the receiver part, we assume the frequency-dependent IQ-M. The impulse responses of these two LPFs are $l_I(n) = 0.1\delta(n) + \delta(n-1)$ and $l_Q(n)$= $0.01\delta(n) + 0.9\delta(n-1)$, respectively. In the verification platform, the functions implemented as the baseband processing include timing synchronization, CFO estimation, IQ-M estimation, channel estimation, phase tracking, and the forward error correction (FEC) decoder. The proposed method is mapped onto the FPGA chips (Xilinx XtremeDSP, Virtex-II) with on-board 14-bit digital-to-analog converters (DACs) to convert the digital data into analog signals. At the transmitter part, the transmitted data are produced using MATLAB and then these data are stored in memory. The signals are then transmitted via an in-house RF front-end. After down-converting the RF signals to baseband at the receiver, analog signals are fed into 14-bit analog-to-digital converters (ADCs). The proposed algorithm then processes the down-converted signals.

Figure 9 shows the constellation performance before and after compensation. The performance improvement is clearly seen because the constellation becomes more distinguishable after compensation. After IQ-M compensation,



(a)



(b)

Fig. 9 Measurement of constellation diagram: (a) Before compensation. (b) After compensation.

the remaining IQ-M parameters are less than 3% (0.13 dB) gain error and 2° phase error. In the proposed design, only two long training symbols are acquired to estimate the IQ-M parameters. If there are more training symbols than currant standards, the accuracy of the IQ-M parameters could be improved further. The error vector magnitude (EVM) after compensation is 8.6% (−21 dB), meeting the standard (IEEE 802.11a/g) allowed 11.2% (−19 dB). The image rejection ratio (IRR) as a function of the mismatch is given by [19]

$$\mathrm{IRR\,(dB)} = 10 \cdot \log\left[\frac{1+(1+\varepsilon)^2+2(1+\varepsilon)\cos(\varphi)}{1+(1+\varepsilon)^2-2(1+\varepsilon)\cos(\varphi)}\right] \quad (15)$$

where ε and φ represent the amplitude and phase mismatch, respectively. After compensation, the remaining gain error and phase error are ∼3% and ∼2°. An IRR of ∼31 dB is achieved. Figure 10 shows the estimated CFR. The CFR after compensation becomes more smoother than the distorted CFR. In conjunction with the proposed scheme, accuracy CFR could be calculated. The proposed method is compatible with current standards. In contrast, recently proposed methods [10], [11] are incompatible with current standards. In addition, the proposed method can estimate the frequency-dependent IQ-M with CFO, whereas reference

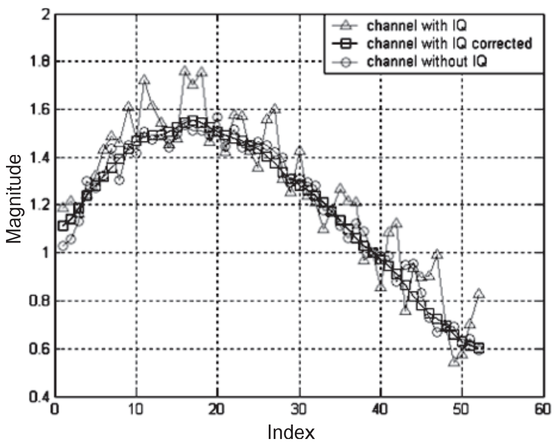


Fig. 10 Channel frequency response.

methods [9], [11], [14], [17] consider frequency-dependent IQ-M only.

5. Conclusions

This study develops a preamble-assisted method for combating the frequency-dependent IQ-M with CFO in direct-conversion OFDM receivers. The frequency-dependent IQ-M with CFO can be estimated by taking advantage of the relationship between desired sub-carriers and image sub-carriers. Both simulation and experiment results indicate that the proposed method can meet system requirements by preventing frequency-dependent IQ-M from significantly degrading the performance. Furthermore, this method is compatible with current standards because it does not require special packet formats.

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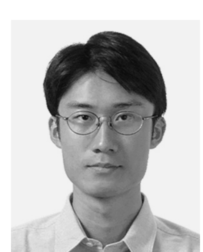

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