IQ Imbalance Estimation Scheme in the Presence of DC Offset and Frequency Offset in the Frequency Domain

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SUMMARY Direct conversion receivers in orthogonal frequency division multiplexing (OFDM) systems suffer from direct current (DC) offset, frequency offset, and IQ imbalance. We have proposed an IQ imbalance estimation scheme in the presence of DC offset and frequency offset, which uses preamble signals in the time domain. In this scheme, the DC offset is eliminated by a differential filter. However, the accuracy of IQ imbalance estimation is deteriorated when the frequency offset is small. To overcome this problem, a new IQ imbalance estimation scheme in the frequency domain with the differential filter has been proposed in this paper. The IQ imbalance is estimated with pilot subcarriers. Numerical results obtained through computer simulation show that estimation accuracy and bit error rate (BER) performance can be improved even if the frequency offset is small.

key words: direct conversion, OFDM, DC offset, frequency offset, IQ imbalance, pilot symbols

1. Introduction

Orthogonal frequency division multiplexing (OFDM) is one of modulation schemes for wireless communication to achieve high data transmission. The OFDM is standardized for WiMax, IEEE 802.11a/g, digital terrestrial broadcasting, and so on. At the receiving end, a direct conversion architecture has been implemented, which reduces the cost and power consumption of the receiver. However, OFDM direct conversion receivers may suffer from direct current (DC) offset, frequency offset, and IQ imbalance [1]. The frequency offset is due to the mismatch between the oscillators in the transmitter and the receiver. The frequency offset at the receiver may deteriorate the orthogonality between the subcarriers and cause intercarrier interference (ICI). The DC offset is caused by the local oscillator (LO) signal. The LO signal may mix with itself down to zero intermediate frequency (IF), resulting in the generation of the DC offset. This is known as self-mixing and is due to the finite isolation that is typical between the LO and radio frequency (RF) ports of a low-noise amplifier (LNA) or mixers. Moreover, the DC offset is also attributed to the mismatch between the mixer components [1], [2]. The IQ imbalance arises mainly from the mismatches of the components between the inphase (I) and quadrature (Q) paths. Specifically, phase mismatch occurs when the phase difference between the LO signals for the I and Q channels is not exactly 90 degrees [3]. These distortions deteriorate the quality of the demodulated signal.

Several joint compensation schemes have been presented [4]–[9]. In [4], the frequency offset and the IQ imbalance are estimated using a nonlinear least-squares scheme. This scheme requires the covariance matrix of the received samples. In [5], the IQ imbalance as well as the frequency offset and the DC offset are estimated using the maximum likelihood criterion. Although this scheme achieves a performance close to the Cramer-Rao bound, it requires a large amount of computation and channel response. In [6], a frequency offset and IQ imbalance estimation scheme is proposed on the basis of simple calculation. The scheme in [7] carries out frequency offset and IQ imbalance estimation in the time domain. The IQ imbalance estimation schemes presented in [8], [9] are conducted in the frequency domain. However, these schemes assume the absence of the DC offset.

On the other hand, in our previous research, the IQ imbalance is estimated in the presence of the DC offset and frequency offset [10]. From the output of the differential filter, the IQ imbalance is estimated from a simple equation in the time domain. However, the estimation accuracy of the IQ imbalance is significantly deteriorated when the frequency offset is small. Therefore, a novel IQ imbalance estimation scheme in the frequency domain with the differential filter is investigated in this paper. The proposed scheme uses a specific combination of symbols on symmetric pilot subcarriers. The use of orthogonality over multiple symbols on symmetric subcarriers for IQ imbalance estimation has been investigated in [3]. A similar approach has been also employed in [9] on pilot subcarriers in the absence of the DC offset and the frequency offset. This paper evaluates the IQ imbalance estimation scheme with the use of the differential filter and the existence of DC offset and frequency offset. It works well if the frequency offset is relatively small.

This paper is organized as follows. Section 2 gives the system model and Sect. 3 describes the proposed IQ imbalance model. In Sect. 4, the frequency offset estimation using the differential filter is explained. Section 5 describes the proposed IQ imbalance estimation. In Sect. 6, numerical results obtained through computer simulation are presented. Section 7 gives our conclusions.
2. System Model

2.1 Preamble Model

Figure 1 shows the IEEE 802.11a/g burst structure of the preamble signal [11],[12]. In the 802.11a/g preamble, short training sequence preamble (STSP) symbols are used for coarse frequency offset estimation and IQ imbalance estimation. Long training sequence preamble (LTSP) symbols are used for fine frequency offset estimation and channel estimation.

The STSP symbols consist of 12 subcarrier signals, which are repeated with a period of 0.8 μs (\(= T_{\text{DFT}}/4 = 3.2/4\)), where \(T_{\text{DFT}}\) is the discrete Fourier transform (DFT) and inverse DFT (IDFT) period. On the other hand, the LTSP symbols consist of 52 subcarrier signals, which have two periods of 3.2 μs (\(= T_{\text{DFT}}\)). During the period of the LTSP, the OFDM training symbols for channel estimation are transmitted [11],[12]. Therefore, frequency offset needs to be estimated appropriately during the period of coarse estimation.

2.2 Subcarrier Allocation

In IEEE 802.11a/g, four subcarriers in one OFDM symbol during a data period are dedicated to pilot symbols [11],[12]. These pilot symbols are transmitted on the subcarrier numbers of \(-21, -7, 7, 21\) as shown in Fig. 2. Moreover, in IEEE 802.11a/g, the DC subcarrier is not used to avoid interference from the DC offset. Although the subcarriers do not interfere with one another, if frequency offset exists, the orthogonality between the subcarriers and the DC offset is deteriorated. A high-pass filter (HPF) can be used to eliminate the static DC offset without removing the received signal.

3. Influence of IQ Imbalance

3.1 Cause of IQ Imbalance

The direct conversion architecture is suitable for mobile terminals since it does not require costly IF filters and allows for single-chip integration. However, because of the absence of the digital IF components, IQ demodulation is handled in the RF domain. Therefore, the direct conversion receiver introduces the IQ imbalance in the mixers, as shown in Fig. 3. In the figure, the phase mismatch and the gain mismatch are represented as \(\theta\) and \(\beta\), respectively. Specifically, phase mismatch occurs when the phase difference between the local oscillator’s signals for the I and Q channels is not exactly 90 degrees. Gain imbalance refers to the gain mismatch in the path of the I and Q signals [3].

3.2 IQ Imbalance Model with Frequency Offset

Assuming that the \(k\)th sample of the OFDM preamble in the time domain is \(s(k)\), a received signal only with the frequency offset, \(r(k)\), is expressed as

\[
r(k) = s(k) \exp\left(\frac{2\pi\alpha k}{N}\right) + \omega(k),
\]

where \(\alpha\) is the frequency offset normalized by subcarrier separation, \(N\) is the number of samples for DFT, and \(\omega(k)\) is the \(k\)th additive white gaussian noise (AWGN) sample with zero mean and variance \(\sigma_n^2\). In this paper, due to the symmetry of the upper and lower paths, the I-phase local signal, \(L_I\), and the Q-phase local signal, \(L_Q\), are assumed to be as follows:

**I component:** \(L_I(t) = (1 + \beta) \cos(2\pi f_c t - \theta/2)\),

**Q component:** \(L_Q(t) = -(1 - \beta) \sin(2\pi f_c t + \theta/2)\),

where \(f_c\) is the carrier frequency. These local signals are multiplied by the received signal. By applying a low pass filter (LPF), the baseband signals, \(\hat{r}_I(k)\) and \(\hat{r}_Q(k)\), with IQ imbalance are obtained. The \(k\)th digitized signal with a sampling interval of \(T_s\) is given by

\[
\hat{r}(k) = \hat{r}_I(k) + j\hat{r}_Q(k),
\]

where

\[
(1 + \beta) \cos(2\pi f_c t - \theta/2)
\]

\[
(1 - \beta) \cos(2\pi f_c t + \pi/2 + \theta/2) = -(1 - \beta) \sin(2\pi f_c t + \theta/2)
\]
\[
\hat{r}_t(k) = (1 + \beta) \left\{ r_t(k) \cos \left( \frac{\theta}{2} \right) - r_Q(k) \sin \left( \frac{\theta}{2} \right) \right\}, \quad (3)
\]
\[
\hat{r}_Q(k) = (1 - \beta) \left\{ r_Q(k) \cos \left( \frac{\theta}{2} \right) - r_t(k) \sin \left( \frac{\theta}{2} \right) \right\}, \quad (4)
\]

where \( r_t(k) \) and \( r_Q(k) \) are the I component and the Q component of \( r(k) \), respectively. Hence, the complex baseband signal \( \hat{r}(k) \) is

\[
\hat{r}(k) = \hat{r}_t(k) + j\hat{r}_Q(k)
\]

\[
= \left\{ \cos \left( \frac{\theta}{2} \right) + j\beta \sin \left( \frac{\theta}{2} \right) \right\} \{ r_t(k) + j r_Q(k) \}
+ \left\{ \beta \cos \left( \frac{\theta}{2} \right) - j \sin \left( \frac{\theta}{2} \right) \right\} \{ r_t(k) - j r_Q(k) \}
\]

\[
= \left\{ \cos \left( \frac{\theta}{2} \right) + j\beta \sin \left( \frac{\theta}{2} \right) \right\} r(k)
+ \left\{ \beta \cos \left( \frac{\theta}{2} \right) - j \sin \left( \frac{\theta}{2} \right) \right\} r^*(k)
\]

where * denotes complex conjugate. From Eq. (5), the received signal with the IQ imbalance is given as

\[
\hat{r}(k) = \phi r(k) + \psi^* r^*(k) + \delta(k),
\]

where

\[
\phi = \cos \left( \frac{\theta}{2} \right) + j\beta \sin \left( \frac{\theta}{2} \right), \quad (7)
\]

\[
\psi = \beta \cos \left( \frac{\theta}{2} \right) + j \sin \left( \frac{\theta}{2} \right), \quad (8)
\]

and \( \delta(k) \) is the DC offset that occurs at the mixer.

4. Frequency Offset Estimation Using Differential Filter

In this frequency offset estimation scheme, the received signal with IQ imbalance is substituted into the differential filter used to eliminate the residual DC offset that passes through the HPF. The \( k \)-th output, \( \hat{d}_{\text{SP}}(k) \), after the differential filter is

\[
\hat{d}_{\text{SP}}(k) = \hat{r}_{\text{SP}}(k) - \hat{r}_{\text{SP}}(k - 1)
= \phi \{ r_{\text{SP}}(k) - r_{\text{SP}}(k - 1) \}
+ \psi^* \{ r_{\text{SP}}^*(k) - r_{\text{SP}}^*(k - 1) \}
+ \Delta \delta(k, k - 1), \quad k \geq 1,
\]

where

\[
\hat{r}_{\text{SP}}(k) = \phi r_{\text{SP}}(k) + \psi^* r_{\text{SP}}^*(k), \quad (10)
\]

\( r_{\text{SP}}(k) \) is the \( k \)-th signal with the frequency offset in the STSP period, and \( \Delta \delta(k, k - 1) \) is the difference between the \( k \)-th and \( (k - 1) \)-th residual DC offsets. In IEEE 802.11a/g standards, the coarse frequency offset estimation is carried out in STSP and the fine frequency offset estimation is carried out in LTSP [11], [12]. In this paper, the frequency offset estimation is calculated from auto-correlation value of STSP and LTSP received signals with IQ imbalance and frequency offset. From Eq. (9), the estimated frequency offset with STSP, \( \hat{\alpha}_{\text{ST}} \), is given as

\[
\hat{\alpha}_{\text{ST}} = \frac{4}{2\pi} \arg \left\{ \sum_{k=0}^{N-1} \hat{d}_{\text{SP}}^*(k) \hat{d}_{\text{ST}}(k + N) \right\}, \quad (11)
\]

where \( k = N \) corresponds to the time index of the first symbol of \( t_5 \) and \( k = \frac{9N}{4} - 1 \) corresponds to the time index of the last symbol of \( t_4 \). Here, the STSP symbols from \( t_5 \) to \( t_{10} \) are used for frequency offset estimation. The auto-correlations between \( t_1 \) and \( t_4 \) are not used because of possible gain shift of the LNA [13]. The coarse frequency offset value obtained from STSP is used for compensation in LTSP. The estimated frequency offset with LTSP, \( \hat{\alpha}_{\text{LP}} \), is then given as

\[
\hat{\alpha}_{\text{LP}} = \frac{1}{2\pi} \arg \left\{ \sum_{k=1}^{N-1} \hat{d}_{\text{LP}}^*(k) \hat{d}_{\text{SP}}(k + N) \right\}, \quad (12)
\]

where \( \hat{d}_{\text{LP}}^*(k) \) is the \( k \)-th output of the differential filter in LTSP. As a result, the fine frequency estimation \( \hat{\alpha} \) is expressed as

\[
\hat{\alpha} = \hat{\alpha}_{\text{ST}} + \hat{\alpha}_{\text{LP}}. \quad (13)
\]

The estimated frequency offset obtained from Eqs. (11) and (12) is deteriorated by the IQ imbalance because the frequency offset estimation in the time domain is carried out in the presence of IQ imbalance. However, MSE of frequency offset estimation influenced by the IQ imbalance is less than \( 10^{-3} \) of the square of the frequency offset [14]. Thus, the IQ imbalance is neglected at this stage for estimation of the frequency offset.

5. IQ Imbalance Estimation

5.1 Conventional Scheme

In the conventional scheme, IQ imbalance estimation is carried out in the time domain as shown in Fig. 4 [10]. The IQ imbalance is estimated from the outputs of the differential filter. From Eq. (9), the 3 preamble symbols repeated in \( N/4 \) samples in the STSP can be expressed as

\[
\hat{d}_{\text{SP}} \left( k - \frac{N}{4} \right) = \hat{r}_{\text{SP}} \left( k - \frac{N}{4} \right) - \hat{r}_{\text{SP}} \left( k - \frac{N}{4} - 1 \right)
\]

![Fig. 4](image-url) Receiver architecture of conventional scheme.
The following equation is derived.

\[
\hat{d}_S(k) = \hat{r}_S(k) - \hat{r}_S(k-1)
\]

Here, with the assumption of small \( \varepsilon \)

\[
\hat{d}_S(k) = \phi \hat{r}_S(k) - \hat{r}_S(k-1)
\]

\[
\approx \phi \hat{r}_S(k) + \psi \hat{d}_S(k)
\]

where \( \phi \) and \( \psi \) are approximated as

\[\phi = \cos\left(\frac{\theta}{2}\right) + j\beta \sin\left(\frac{\theta}{2}\right) \approx 1 + j\beta \frac{\theta}{2},\]

\[\psi = \beta \cos\left(\frac{\theta}{2}\right) + j \sin\left(\frac{\theta}{2}\right) \approx \beta + j \frac{\theta}{2},\]

using the first-order approximation of the Taylor expansion. Thus, Eq. (19) becomes

\[
\frac{\beta - j^2 \frac{\theta}{2}}{1 - j\beta \frac{\theta}{2}} \approx \varepsilon Q + j e Q.
\]

The estimated \( \hat{\beta} \) and \( \hat{\theta} \) can then be calculated as follows.

\[
\hat{\beta} \approx \frac{2e I}{2 - e Q \theta},
\]

\[
\hat{\theta} \approx \frac{e I^2 + e Q^2 - 1 - \sqrt{(e I^2 + e Q^2 - 1)^2 + 4e_Q^2}}{e Q}.
\]

To obtain \( \varepsilon \) in Eq. (19), \( \alpha \) in Eq. (18) is substituted with the

value estimated in Sect. 4. However, the disadvantage of this scheme is that the IQ imbalance estimation is deteriorated when the frequency offset is small as both the numerator and denominator approaches to 0 in Eq. (19) [10].

5.2 Proposed Scheme

5.2.1 Influence of Differential Filter

Aforementioned in Sect. 4, the differential filter is used to cut the residual DC offset. The IQ imbalance is estimated by the pilot symbols in the data period, which passes through the differential filter as shown in Fig. 5. The phase and amplitude responses of the received symbols are affected due to the differential filter. The output from the differential filter in the frequency domain is

\[D(l) = H_{DF}(l)\tilde{R}(l),\]

where \( D(l), H_{DF}(l), \) and \( \tilde{R}(l) \) are the output of the differential filter, the frequency response of the differential filter, and the received signal with IQ imbalance on \( l \)th subcarrier, respectively. The channel response on the \( l \)th subcarrier is given as

\[H_{DF}(l) = 1 - \exp\left(-j\frac{2\pi l}{N}\right), \quad (l = -N/2, \ldots, N-1).\]

The frequency responses of the differential filter outputs are compensated from Eq. (26).

5.2.2 IQ Imbalance Estimation without Frequency Offset

In the proposed scheme, the pilot subcarriers in the data period are used for IQ imbalance estimation. If the frequency offset does not exist, the \( l \)th received symbol in the frequency domain after DFT, \( \tilde{R}(l) \), is given as

\[\tilde{R}(l) = \phi(l)T(l) + \psi(-l)T^*(-l),\]

with

\[
\text{Fig. 5} \quad \text{Receiver architecture of proposed scheme.}
\]
In Eq. (27),

$$\phi(l) = \phi H(l), \quad (29)$$

$$\psi(l) = \psi H(l), \quad (30)$$

and $H(l)$ is the channel response of the $l$th subcarrier. The channel response is assumed to be constant during one OFDM symbol interval. From Eq. (27), the symbol on the $l$th subcarrier OFDM symbol is affected by the symbol on the $(l-1)$th subcarrier due to the IQ imbalance. To estimate the IQ imbalance, the pilot symbols shown in Table 1 are transmitted. The IQ imbalance is estimated from the $m$th and $(m+1)$th consecutive OFDM symbols.

The received two $l$th pilot symbols on $m$th and $(m+1)$th consecutive OFDM symbols, $P_m(l)$ and $P_{m+1}(l)$, are written as

$$\hat{P}_m(l) = \phi(l)P_m(l) + \psi^*(l)P_m^*(-l), \quad (31)$$

$$\hat{P}_{m+1}(l) = \phi(l)P_{m+1}(l) + \psi^*(l)P_{m+1}^*(-l). \quad (32)$$

The mirror subcarriers of Eqs. (31) and (32) are also written as

$$\hat{P}_m(-l) = \phi(-l)P_m(-l) + \psi^*(-l)P_m^*(-l), \quad (33)$$

$$\hat{P}_{m+1}(-l) = \phi(-l)P_{m+1}(-l) + \psi^*(-l)P_{m+1}^*(-l). \quad (34)$$

By substituting the values of the pilot symbols from Table 1 into Eqs. (31)–(34), $\phi$ and $\psi$ are calculated as

$$\psi^*(-l) = \frac{\hat{P}_m(l) + \hat{P}_{m+1}(l)}{2}, \quad (35)$$

$$\phi(l) = \frac{\hat{P}_m(l) - \hat{P}_{m+1}(l)}{2}, \quad (36)$$

$$\phi(-l) = \frac{\hat{P}_m(-l) + \hat{P}_{m+1}(-l)}{2}, \quad (37)$$

$$\psi^*(l) = \frac{\hat{P}_m(-l) - \hat{P}_{m+1}(-l)}{2}. \quad (38)$$

From Eqs. (35) to (38), it is given as

$$\frac{\psi^*(l)}{\phi^*} = \frac{\psi^*(-l) + \psi(l)}{\phi^*(-l) + \phi(l)}, \quad \text{for } l \in \mathbb{N}_p. \quad (39)$$

From Eqs. (23), (24) and (39), $\hat{\beta}$ and $\hat{\theta}$ are calculated. In the data period, the received signal is compensated with the estimations of $\phi$ and $\psi$ given in Eqs. (7) and (8). The received symbol after IQ imbalance compensation, $\hat{R}(l)$, is expressed as

$$\hat{R}(l) = \phi^* \hat{R}(l) - \psi^* \hat{R}(-l) \frac{|\phi|^2 - |\psi|^2}{|\phi|^2 + |\psi|^2}, \quad \text{for } l \in \mathbb{N}_d \cup \mathbb{N}_p. \quad (40)$$

If the IQ imbalance is compensated completely, Eq. (40) is given as

$$\hat{R}(l) = H(l)T(l), \quad \text{for } l \in \mathbb{N}_d \cup \mathbb{N}_p. \quad (41)$$

The compensated symbol shown in Eq. (41) contains the channel response on the $l$th subcarrier. From the estimated channel response on each pilot subcarrier, the channel response of the other subcarriers is compensated with the 1st order interpolation.

5.2.3 IQ Imbalance Estimation in the Presence of Frequency Offset

In the time domain, the frequency offset causes additional phase rotation in the data period. The frequency offset is estimated and compensated in the time domain and IQ imbalance estimation is carried out in the frequency domain as shown in Fig. 5. From Eq. (6), the $d$th received signal after frequency offset compensation in the time domain, $\hat{r}(k)$, is expressed as

$$\hat{r}(k) = \phi r(k) + \psi^* r^*(k) \exp \left(-\frac{2\pi (\alpha + \hat{\theta})}{N} k \right). \quad (42)$$

The received symbol with the frequency offset in the frequency domain, $\hat{R}(l)$, is then given as

$$\hat{R}(l) = \sum_{k=0}^{N-1} \hat{r}(k) \exp \left(-\frac{2\pi l}{N} k \right),$$

$$= \phi R(l) + \frac{\psi^*}{N} \sum_{k=0}^{N-1} R^*(-l) \exp \left(-\frac{2\pi (\alpha + \hat{\theta})}{N} k \right) \exp \left(-\frac{2\pi (\alpha + \hat{\theta})}{N} k \right). \quad (43)$$

where $R(l)$ is the received signal on $l$th subcarrier given by

$$R(l) = H(l)T(l). \quad (44)$$

From Eq. (43), it is shown that all the subcarriers cause ICI to the $l$th subcarrier which deteriorates the accuracy of IQ imbalance estimation. Moreover, the averaging does not ignore the accuracy since the ICI components are affected by the frequency offset.

6. Simulation Results

6.1 Simulation Conditions

The MSE of the IQ imbalance estimation is evaluated through computer simulation. The simulation conditions

<table>
<thead>
<tr>
<th>Table 1</th>
<th>Pilot subcarriers.</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l = -21$</td>
<td>$l = -7$</td>
</tr>
<tr>
<td>$2m^{th}$ symbol</td>
<td>1</td>
</tr>
<tr>
<td>$(2m + 1)^{th}$ symbol</td>
<td>1</td>
</tr>
</tbody>
</table>
are shown in Table 2. Information bits are modulated with QPSK on each subcarrier. The number of DFT/IDFT points is set to 64 while 52 subcarriers are used for the LTSP symbols, which follows the IEEE 802.11a/g standard. The 1st order butterworth filter is employed as the Rx HPF. The cut-off frequency of the received HPF is set to 10 [kHz]. The DC offset is set to 10 [dB] [13]. The normalized frequency offset is from 0.1 to 0.5. The mismatch of amplitude is set to from 0 to 0.1 and the mismatch of phase is set to from 0 to 10 [degrees] [7].

### 6.2 Channel Models

In this paper, AWGN and multipath channel models are employed. The exponential model is recommended as the multipath fading channel in IEEE 802.11. This model represents a real world scenario, in which the positions of the reflectors generate paths that have longer delay [15]. As shown in Fig. 6, the path delay profile for this model has the form: 

\[ P[\tau] = \frac{1}{\tau_d \epsilon} \exp(-\tau / \tau_d) \]

where \( \tau_m \) is the maximum delay. The path delay profile for the exponential model, the maximum excess delay is given by

\[ \tau_m = \frac{A \times \tau_d}{10 \times \log_{10}(\epsilon)} \]  

![Fig. 6 Path delay profile for an exponential channel.](image)

6.3 Normalized MSE Performance vs. Frequency Offset

Figures 7 and 8 show the normalized MSE performance of gain and phase mismatch estimation, respectively, when the frequency offset \( \alpha \) is varied. The phase mismatch \( \beta \) is set to 0.05 and the phase mismatch \( \theta \) is set to 5 [degrees]. \( E_b/N_0 \) is set to \{10, 15, or 20\} [dB]. In those figures, ‘Conventional scheme’ refers to the IQ imbalance estimation scheme in the time domain as shown in Sect. 5.1 [10]. It is clear that the proposed scheme is superior to the conventional scheme when the frequency offset \( \alpha \) is less than 0.2. The frequency offset can also be compensated in the analog domain with automatic frequency control, and the residual frequency offset remains small [18], [19]. Hence, such values seem to be appropriate. If the residual frequency offset is large, the time domain IQ imbalance scheme shown in Sect. 5.1 should be used [10]. Either frequency or time domain IQ estimation scheme can be selected depending on the amount of frequency offset. In both figures, the fluctuation in the MSE

![Fig. 7 Normalized MSE performance of gain mismatch estimation (\( \beta=0.05, \theta=5 \) [degrees]).](image)

![Fig. 8 Normalized MSE performance of phase mismatch estimation (\( \beta=0.05, \theta=5 \) [degrees]).](image)
of the proposed scheme for different frequency offset is due to the effect of the frequency offset in the second term of the right side of Eq. (43). This tendency can be observed without ICI ($R(l) = 0$, for $l \in \mathbb{N}_0$) in the numerical results. This is because the frequency offset rotates the phase of $\psi$ in the second term of Eq. (43).

6.4 Normalized MSE Performance vs. Gain Mismatch and Phase Mismatch

Figure 9 shows the normalized MSE performance of gain mismatch estimation with the DC offset and the frequency offset when the gain mismatch value is varied. The phase mismatch $\theta$ is set to 5 [degrees] and the frequency offset $\alpha$ is set to 0.001. $E_b/N_0$ is set to $\{10, 15, 20\}$ [dB]. The MSE performance of the proposed scheme is superior than the conventional scheme. It can be seen from Fig. 9 that the normalized MSE performance improves as the gain mismatch increases. This is because the MSE is normalized by the gain mismatch $\beta$. The normalized MSE of the proposed scheme is reduced by a factor of about $10^{-100}$ as compared to the conventional scheme.

Figure 10 shows the normalized MSE performance of phase mismatch estimation with the DC offset and the frequency offset when the phase mismatch value is varied. The phase mismatch $\beta$ is set to 0.05 and the frequency offset $\alpha$ is set to 0.001. $E_b/N_0$ is set to $\{10, 15, 20\}$ [dB]. In the conventional scheme, the MSE performance is deteriorated in the small frequency offset region and exceeds $(2\pi)^2$ for any amount of the phase mismatch. Thus, the MSE curve of the conventional scheme is set to $(2\pi)^2$ and normalized, here. The MSE performance of the proposed scheme is again superior than the conventional scheme. It can be seen from Fig. 10 that the normalized MSE curves of both the proposed and conventional schemes improve as the phase mismatch increases. This is because the MSE is normalized by the phase mismatch $\theta$. The normalized MSE of the proposed scheme reduces by a factor of about $10^{10}$ as compared to the conventional scheme.

6.5 BER Performance vs. Frequency Offset

Figures 11 and 12 show the BER performance versus the frequency offset $\alpha$ with the AWGN and indoor office JTC models, respectively. The gain mismatch $\beta$ is set to 0.05, the phase mismatch $\theta$ is set to 5 [degrees], and the frequency offset $\alpha$ ranges from 0.001 to 0.3. In Fig. 11, $E_b/N_0$ is set to $\{10, 15, 20\}$ [dB]. In Fig. 12, $E_b/N_0$ is set to $\{20, 30, 40\}$ [dB].

As shown in these figures, the BER curve for the conventional scheme decreases as the frequency offset $\alpha$ increases. This is because IQ imbalance estimation does not work well when the frequency offset is small as mentioned in Sect. 5.1 [10]. On the other hand, the BER curve of the proposed scheme is deteriorated as the frequency offset $\alpha$ grows. This is again due to the ICI caused by the frequency offset. From those figures, the proposed scheme exhibits superior estimation performance as compared with the conventional scheme when the frequency offset is small.
Fig. 12 BER vs. normalized frequency offset $\alpha$ (Indoor, 64QAM, $\beta=0.05$, $\theta=5$ [degrees]).

Fig. 13 BER vs. $E_b/N_0$ (AWGN, 64QAM, $\beta=0.05$, $\theta=5$ [degrees]).

Fig. 14 BER vs. $E_b/N_0$ (Indoor, 64QAM, $\beta=0.05$, $\theta=5$ [degrees]).

As shown in Fig. 13, the BER curve of the conventional scheme actually makes the BER worse because of inaccurate estimation of the IQ imbalance. The proposed IQ imbalance estimation improves the BER performance. However, the BER performance is deteriorated as the frequency offset $\alpha$ increases. Moreover, as shown in Fig. 14, the BER curve of the proposed scheme at $\alpha=0.05$ is worse than that of the conventional scheme when the $E_b/N_0$ is more than 30 [dB]. This is because the estimated value in Eq. (43) suffers from the phase rotation due to the frequency offset.

7. Conclusion

In this paper, a low-complexity IQ imbalance estimation scheme in the presence of the DC offset and the frequency offset has been proposed. The conventional scheme uses the preamble signals in the time domain. However, the BER performance is deteriorated when the frequency offset is small. In the proposed IQ imbalance estimation scheme, the pilot subcarriers in the frequency domain are employed. The numerical results obtained through computer simulation shows that the proposed scheme works well when the frequency offset is small. The proposed scheme improves the accuracy of estimation by a factor of $10^1$ for the gain mismatch and $10^1$ for the phase mismatch with the small frequency offset. Thus, depending on the amount of frequency offset, the IQ imbalance estimation scheme in the time domain or the frequency domain can be utilized accordingly.

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