Frontiers of Information Technology & Electronic Engineering
www.jzus.zju.edu.cn; engineering.cae.cn; www.springerlink.com
ISSN 2095-9184 (print); ISSN 2095-9230 (online)
E-mail: jzus@zju.edu.cn

Standard-independent I/Q imbalance estimation and compensation scheme in OFDM direct-conversion transceivers*

Fang-lin GU[‡1], Shan WANG[1], Wen-wu WANG[2]
[1]*College of Electronic Science and Engineering, National University of Defense Technology, Changsha 410073, China*
[2]*Department of Electronic Engineering, University of Surrey, Guildford GU2 7XH, UK*
E-mail: gu.fanglin@nudt.edu.cn; chinafir@nudt.edu.cn; w.wang@surrey.ac.uk

Received Jan. 3, 2017; Revision accepted Mar. 6, 2017; Crosschecked Mar. 5, 2018

Abstract: Direct-conversion transceivers are gaining increasing attention due to their low power consumption. However, they suffer from a serious in- and quadrature-phase (I/Q) imbalance problem. The I/Q imbalance can severely limit the achievable operating signal-to-noise ratio (SNR) at the receiver and, consequently, the supported constellation sizes and data rates. In this paper, we first investigate the effects of I/Q imbalance on orthogonal frequency division multiplexing (OFDM) receivers, and then propose a new I/Q imbalance compensation scheme. In the proposed method, a new statistic, which is robust against channel distortion, is used to estimate the I/Q imbalance parameters, and then the I/Q imbalance is corrected in the frequency domain. Simulations are performed to verify the effectiveness of the proposed method for I/Q imbalance compensation. The results show that the proposed I/Q imbalance compensation method can achieve bit error rate (BER) performance close to that in the ideal case without I/Q imbalance in additive white Gaussian noise (AWGN) or multipath environments. Furthermore, because no pilot information is required, this method can be applied in various standard communication systems.

Key words: In- and quadrature-phase (I/Q) imbalance; Orthogonal frequency division multiplexing (OFDM); Standard-independent
https://doi.org/10.1631/FITEE.1700003 **CLC number:** TN92

1 Introduction

Orthogonal frequency-division multiplexing (OFDM) is a widely recognized and standardized modulation technique for broadband wireless systems. Because of its ability to elegantly cope with a multipath environment, it has been widely used for wireless local area networks (WLANs) (IEEE, 1999), digital audio broadcasting (DAB) (Shelswell, 1995), digital video broadcasting (DVB) (Reimers, 1997), fixed wireless access (Koffman and Roman, 2002), etc. Furthermore, the lower cost and lower power consumption requirements of the fifth generation (5G) networks (Chen and Zhao, 2014) have resulted in the development of integrated, cost- and power-efficient OFDM receivers. The zero-intermediate frequency (IF) or direct-conversion receiver (Abidi, 1995), as shown in Fig. 1, is an attractive candidate, because it avoids costly IF filters and allows for easier integration than the super-heterodyne structure. However, there are also some drawbacks in using a zero-IF architecture. In a super-heterodyne architecture, the in- and quadrature-phase (I/Q) modulation and demodulation are performed in the digital domain, which results in a perfect I/Q separation. On

‡ Corresponding author
* Project supported by the National Natural Science Foundation of China (No. 61601477)
ORCID: Fang-lin GU, http://orcid.org/0000-0001-9194-280X

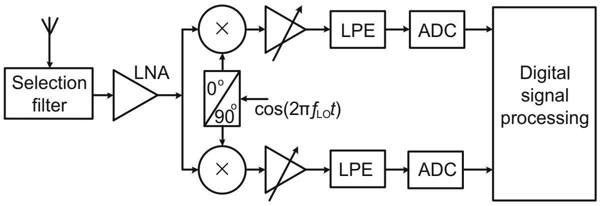

Fig. 1 Block diagram of the zero-intermediate frequency (IF) direct-conversion receiver

LPF: low-pass filter; LNA: low noise amplifier; ADC: analog-to-digital converter

the other hand, the zero-IF architecture performs I/Q modulation and demodulation in the analog domain. The non-ideal match between the I/Q components will result in I/Q imbalance and decrease the bit error rate (BER) of the systems. The impact of I/Q imbalance in OFDM systems was investigated by Hieu et al. (2007) and Lopez-Martinez et al. (2011), and the BER analysis was given by Yang et al. (2013).

Many previous studies have investigated the issue of I/Q imbalance compensation in OFDM systems. Generally, these methods can be divided into two categories. The first is the pilot-assisted method. The basic idea is that the I/Q imbalance parameters are estimated by using pilot subcarriers in the frequency domain (Tarighat et al., 2005; Sung and Chao, 2009; He et al., 2011; Wu et al., 2014). The second is the pilot-free or blind method. The main idea is that the I/Q imbalance parameters are estimated by using the statistics of the received signals without any prior pilot information (Ylamurto, 2003; Inamori et al., 2009; Anttila, 2011; Anttila and Valkama, 2013; Kim et al., 2014; Gu et al, 2016). Because the blind I/Q imbalance method does not require standard-specific structures for parameter estimation, this method is independent of the standards and can be applied in various standard applications. Hence, we focus mainly on the blind form of I/Q imbalance compensation methods in this study.

Generally, the blind form I/Q imbalance compensation methods in OFDM systems can be divided into two types: time-domain method and frequency-domain method. For the time-domain method, the idea is to estimate the I/Q imbalance parameters or to design a compensation filter by resorting to the statistics of the received signals in the time domain. For example, several adaptive filtering schemes in the time domain were proposed to eliminate the I/Q imbalance by using the circularity property of common communication signals (Anttila, 2011; Anttila and Valkama, 2013; Kim et al., 2014; Gu et al., 2016). Schemes in the frequency domain were also proposed to eliminate the I/Q imbalance by using the statistics of the received signals (Windisch and Fettweis, 2004). However, most of these methods focus only on compensating for the I/Q imbalance, without considering the distortions caused by channels. Taking the effect of the distortion caused by the channels into account, in some studies it is assumed that channel estimation can be obtained before I/Q imbalance compensation (Ylamurto, 2003). In practice, the I/Q imbalance problem occurs in the front-end of radio frequency (RF) receivers and channel estimation is usually performed after I/Q imbalance compensation. Thus, I/Q imbalance compensation may be suitable for unknown channels. Some previous research has investigated this problem. Furthermore, several adaptive filtering schemes in the frequency domain were proposed to eliminate the distortions caused by both the channels and I/Q imbalance using the least mean square (LMS) algorithm (Tandur and Moonen, 2007; Inamori et al., 2009). Nevertheless, the LMS algorithm suffers from slow convergence.

In this study, we investigate the problem of estimation and compensation of I/Q imbalance in OFDM systems when no knowledge of the multipath channel is available. Our basic idea is to extend the method in Windisch and Fettweis (2004) to the case where the distortion caused by the channel is considered. First, we analyze the distortions caused by the multipath channel and I/Q imbalance on the received OFDM signal. Because OFDM is able to deal with a multipath environment, the distortion caused by the channel will lead only to the change in the amplitude of the signal on each subcarrier, but not to crosstalk between the mirror subcarrier. Hence, the statistics in Windisch and Fettweis (2004), which is related only to the crosstalk between mirror subcarriers, can also be used for the estimation of I/Q imbalance parameters. Then we propose an I/Q imbalance compensation scheme for an OFDM system, in which the distortion caused by the channel is considered. The method is implemented in two steps. The first step is to estimate the I/Q imbalance parameters by evaluating the statistics and then to correct the distortion in the received signal caused by I/Q imbalance with the parameter estimates. In the second step, the signal corrected is used for channel estimation and equalization.

2 Problem formulation

2.1 System model

For the convenience of presentation, we will drop the sample time index n in this study, except for a particular declaration. However, it should be kept in mind that the investigated OFDM symbols are generally time-variant. In OFDM systems, a block of data is transmitted as an OFDM symbol. Assuming a symbol size equal to N (N is a power of 2), the transmitted block of data is denoted as

$$\boldsymbol{s} = [s(1), s(2), \cdots, s(N)]^{\mathrm{T}}, \tag{1}$$

where $[\cdot]^{\mathrm{T}}$ is the transpose operation. Each block passes through the inverse discrete Fourier transform (IDFT) operation

$$\tilde{\boldsymbol{s}} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{s}, \tag{2}$$

where $(\cdot)^{\mathrm{H}}$ denotes the conjugate transpose operation and $\boldsymbol{F}$ is the unitary discrete Fourier transform (DFT) matrix of size N.

A cyclic prefix (CP) of length P is added to each transformed block of data and then the data is transmitted through the channel. A finite impulse response (FIR) model with $(L+1)$ taps is assumed for the channel, i.e., $\boldsymbol{h} = [h_0, h_1, \cdots, h_L]^{\mathrm{T}}$ with $L \leq P$, to preserve the orthogonality between subcarriers. At the receiver, the received samples corresponding to the transmitted block $\tilde{\boldsymbol{s}}$ are collected into a vector, after discarding the received CP samples. The received block of data before being distorted by I/Q imbalances can be written as

$$\tilde{\boldsymbol{y}} = \boldsymbol{H}^{\mathrm{c}}\tilde{\boldsymbol{s}} + \tilde{\boldsymbol{w}}, \tag{3}$$

where

$$\boldsymbol{H}^{\mathrm{c}} = \begin{bmatrix} h_0 & h_1 & \cdots & h_L & 0 & 0 & \cdots & 0 \\ 0 & h_0 & h_1 & \cdots & h_L & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & \cdots & 0 & h_0 & h_1 & h_2 & \cdots & h_L \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ h_2 & \cdots & h_L & 0 & \cdots & 0 & h_0 & h_1 \\ h_1 & h_2 & \cdots & h_L & 0 & \cdots & 0 & h_0 \end{bmatrix}$$

is an $N \times N$ circulant matrix and $\tilde{\boldsymbol{w}}$ is the additive white noise at the receiver. It is known that $\boldsymbol{H}^{\mathrm{c}}$ can be diagonalized by the DFT matrix as $\boldsymbol{H}^{\mathrm{c}} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{\Lambda}\boldsymbol{F}$, where $\boldsymbol{\Lambda} = \mathrm{diag}\ (\boldsymbol{\lambda})$ and vector $\boldsymbol{\lambda}$ is related to $\boldsymbol{h}$ via

$$\boldsymbol{\lambda} = \sqrt{N}\boldsymbol{F}^{\mathrm{H}} \begin{bmatrix} \boldsymbol{h} \\ \boldsymbol{0}_{(N-(L+1))\times 1} \end{bmatrix}. \tag{4}$$

Then Eq. (3) gives

$$\tilde{\boldsymbol{y}} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{\Lambda}\boldsymbol{F}\tilde{\boldsymbol{s}} + \tilde{\boldsymbol{w}} = \boldsymbol{F}^{\mathrm{H}}\mathrm{diag}\{\boldsymbol{\lambda}\}\boldsymbol{F}\tilde{\boldsymbol{s}} + \tilde{\boldsymbol{w}}. \tag{5}$$

As shown in Fig. 2, the received block of data $\tilde{\boldsymbol{y}}$, after being distorted by I/Q imbalances, is given by

$$\tilde{\boldsymbol{z}} = \mu\tilde{\boldsymbol{y}} + \nu\tilde{\boldsymbol{y}}^*, \tag{6}$$

where $(\cdot)^*$ denotes complex conjugate operation. The distortion parameters μ and ν are related to the amplitude and phase imbalances between the I and Q branches in the RF/analog demodulation process through a simplified model as follows:

$$\begin{cases} \mu = \cos(\theta/2) + \mathrm{j}\alpha\sin(\theta/2), \\ \nu = \alpha\cos(\theta/2) - \mathrm{j}\sin(\theta/2), \end{cases} \tag{7}$$

where α and θ are the amplitude imbalance and phase between the I and Q branches, respectively. The values of α and θ are not known at the receiver because they are caused by manufacturing inaccuracies in the analog components.

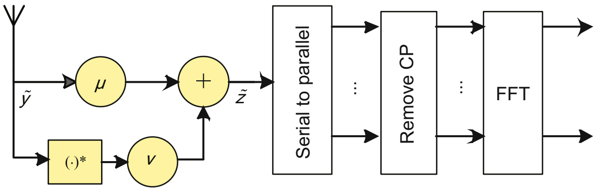

Fig. 2 Block diagram representation of an OFDM system with I/Q imbalance (CP: cyclic prefix)

2.2 Effect of I/Q imbalance on the OFDM transceiver

Remember that the DFT of the complex conjugate of a sequence is related to the DFT of the original sequence through a mirrored relation (assuming $1 \leq n \leq N$ and $1 \leq k \leq N$):

$$\tilde{y}(n) \xrightarrow{\mathrm{DFT}} y(k), \tag{8}$$

$$\tilde{y}^*(n) \xrightarrow{\mathrm{DFT}} y^*(N-k+2). \tag{9}$$

For notational simplicity, we denote this operation by the superscript '†', i.e., for a vector $\tilde{\boldsymbol{y}}$ of size N,

$$\begin{aligned} \boldsymbol{y} =& [y(1), y(2), \cdots, y(N/2), \\ & y(N/2+1), y(N/2+2), \cdots, y(N/2)]^{\mathrm{T}}. \end{aligned} \tag{10}$$

Now, from Eq. (3), we have

$$\tilde{\boldsymbol{y}}^* = (\boldsymbol{H}^{\mathrm{c}})^* \tilde{\boldsymbol{s}}^* + \tilde{\boldsymbol{w}}^*, \tag{11}$$

where $(\boldsymbol{H}^{\mathrm{c}})^*$ is a circulant matrix defined in terms of $\boldsymbol{h}^*$. Then in a similar way, it is easy to derive

$$(\boldsymbol{H}^{\mathrm{c}})^* = \boldsymbol{F}^{\mathrm{H}} \mathrm{diag}(\boldsymbol{\lambda}^\dagger) \boldsymbol{F}. \tag{12}$$

Substituting Eq. (12) into Eq. (11) results in

$$\tilde{\boldsymbol{y}}^* = \boldsymbol{F}^{\mathrm{H}} \mathrm{diag}(\boldsymbol{\lambda}^\dagger) \boldsymbol{s}^\dagger + \tilde{\boldsymbol{w}}^*. \tag{13}$$

Consider a receiver that applies DFT operation to the received block of data, as performed in a standard OFDM receiver. Applying the DFT matrix to Eq. (6), i.e., setting $\boldsymbol{z} = \boldsymbol{F}\tilde{\boldsymbol{z}}$, after substituting Eqs. (5) and (13) into Eq. (6), we have

$$\boldsymbol{z} = \mu \mathrm{diag}(\boldsymbol{\lambda}) \boldsymbol{s} + \nu \mathrm{diag}(\boldsymbol{\lambda}^\dagger) \boldsymbol{s}^\dagger + \boldsymbol{w}, \tag{14}$$

where $\boldsymbol{w}$ is a transformed version of the original noise vector $\tilde{\boldsymbol{w}}$. As seen from Eq. (14), vector $\boldsymbol{z}$ is no longer related to the transmitted block $\boldsymbol{s}$ through a diagonal matrix, as is the case in an OFDM system with ideal I and Q branches. For simplicity of presentation, we discard the samples corresponding to the subcarriers 1 and $N/2+1$, i.e., $z(1)$ and $z(N/2+1)$ (Note that in standardized OFDM systems such as IEEE 802.11a, these two subcarriers do not carry any information due to implementation issues. Sending zeros on these two subcarriers relaxes the implementation requirements on the receiver analog filters and DC offset). We define two new vectors:

$$\begin{cases} \bar{\boldsymbol{s}} = [s(2), \cdots, s(N/2), s^*(N/2+2), \cdots, s^*(N)]^{\mathrm{T}}, \\ \bar{\boldsymbol{z}} = [z(2), \cdots, z(N/2), z^*(N/2+2), \cdots, z^*(N)]^{\mathrm{T}}. \end{cases} \tag{15}$$

The second-half elements in $\bar{\boldsymbol{s}}$ and $\bar{\boldsymbol{z}}$ are conjugated based on the structure of Eq. (14). Using Eqs. (15) and (14) leads to

$$\bar{\boldsymbol{z}} = \bar{\boldsymbol{\Lambda}} \bar{\boldsymbol{s}} + \bar{\boldsymbol{w}}, \tag{16}$$

where

$$\bar{\boldsymbol{\Lambda}} = \begin{bmatrix} \mu\lambda(2) \cdots & 0 & 0 & \cdots \nu\lambda^*(N) \\ \vdots & \vdots & \vdots & \vdots \\ 0 & \cdots \mu\lambda(N/2) & \nu\lambda^*(N/2+2) \cdots & 0 \\ 0 & \cdots \nu^*\lambda(N/2) & \mu^*\lambda^*(N/2+2) \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ \nu^*\lambda(2) \cdots & 0 & 0 & \cdots \mu^*\lambda^*(N) \end{bmatrix}$$

and $\bar{\boldsymbol{w}}$ is related to $\boldsymbol{w}$ in a manner similar to Eq. (15). Note that matrix $\bar{\boldsymbol{\Lambda}}$ in the above equation is not a diagonal matrix, as is the case in Eq. (5), although it collapses to a diagonal matrix by setting ν equal to zero. Eq. (16) can be reduced to 2×2 decoupled sub-equations, for $k = \{2, 3, \cdots, N/2\}$:

$$\bar{\boldsymbol{z}}(k) = \boldsymbol{\Gamma}(k) \bar{\boldsymbol{s}}(k) + \bar{\boldsymbol{w}}(k), \tag{17}$$

where

$$\bar{\boldsymbol{z}}(k) = \begin{bmatrix} z(k) \\ z^*(N-k+2) \end{bmatrix},$$

$$\bar{\boldsymbol{s}}(k) = \begin{bmatrix} s(k) \\ s^*(N-k+2) \end{bmatrix},$$

$$\boldsymbol{\Gamma}(k) = \begin{bmatrix} \mu\lambda(k) & \nu\lambda^*(N-k+2) \\ \nu^*\lambda(k) & \mu^*\lambda^*(N-k+2) \end{bmatrix},$$

and $\bar{\boldsymbol{w}}(k)$ is the corresponding noise on subcarrier k defined from noise vector $\boldsymbol{w}$ in a manner similar to Eq. (15). The objective is to recover $\bar{\boldsymbol{s}}(k)$ from $\bar{\boldsymbol{z}}(k)$ in Eq. (17) for $k = \{2, 3, \cdots, N/2\}$, or equivalently, $\bar{\boldsymbol{s}}$ from $\bar{\boldsymbol{z}}$ in Eq. (16).

Note that Eq. (17) can be converted into the following form:

$$\boldsymbol{z}(k) = \boldsymbol{\Pi} \boldsymbol{\Lambda}(k) \boldsymbol{s}(k) + \boldsymbol{w}(k), \tag{18}$$

where

$$\boldsymbol{\Pi} = \begin{bmatrix} \mu & \nu \\ \nu^* & \mu^* \end{bmatrix}$$

and

$$\boldsymbol{\Lambda}(k) = \begin{bmatrix} \lambda(k) & 0 \\ 0 & \lambda^*(N-k+2) \end{bmatrix}.$$

As shown in Eq. (18), I/Q imbalance parameters $\boldsymbol{\Pi}$ and channel coefficients $\boldsymbol{\Lambda}$ are independent. Hence, I/Q imbalance compensation and channel estimation can be implemented separately in two stages. The details will be given in Section 3.

2.3 Noise statistics

The time domain samples of the additive white Gaussian noise (AWGN) are independent, identically distributed complex Gaussian random variables, with mean zero and variance σ_z^2 in the real and imaginary parts. Thus, the real and imaginary parts of $w(k)$, $\Re(w(k))$ and $\Im(w(k))$, are both Gaussian random variables with mean zero and variance σ_w^2/N. The real parts of $w(k)$ at different subcarriers are independent, so are the imaginary parts, i.e., $E[\Re(w(k_1))\Re(w(k_2))] = \sigma_w^2/N \cdot \delta(k_1 - k_2)$ and $E[\Im(w(k_1))\Im(w(k_2))] = \sigma_w^2/N \cdot \delta(k_1 - k_2)$, where $E[\cdot]$ denotes the expectation operator, and $\delta(x)$ is 1 for $x = 0$ and 0 otherwise. Additionally, the real and imaginary parts of $w(k)$ at the same or different subcarriers are also independent, i.e., $E[\Re(w(k_1))\Im(w(k_2))] = 0$.

3 I/Q imbalance compensation scheme

The proposed I/Q imbalance compensation approach involves the following steps: (1) evaluating the associated statistics of the observed signals and deriving the I/Q imbalance parameters; (2) compensating for the I/Q imbalance in the frequency domain using the I/Q imbalance estimates on a designed OFDM receiver architecture (Fig. 3).

3.1 I/Q imbalance parameter estimation

Here, we assume that $E[s(k)s(N-k+2)] = 0$ holds at the examined subcarrier index k; i.e., $s(k)$ and $s(N-k+2)$ are uncorrelated and have zero mean. This assumption is realistic at least for pairs of data-subcarriers if a proper source and channel coding are applied. Analyzing the cross-correlation term yields

$$\begin{aligned}
&E[z(k)z(N-k+2)] = \mu\nu|\lambda(k)|^2 E[s(k)s^*(k)] \\
&+ |\mu|^2\lambda(k)\lambda(N-k+2)E[s(k)s(N-k+2)] \\
&+ |\mu|^2\lambda(k)\lambda(N-k+2)E[s^*(k)s^*(N-k+2)] \\
&+ \mu\nu|\lambda(N-k+2)|^2 E[s(N-k+2)s^*(N-k+2)] \\
&+ \mu\nu\left\{E[w(k)w^*(k)] + E[w(N-k+2)w^*(N-k+2)]\right\}.
\end{aligned} \tag{19}$$

Because $E[w(k)w^*(k)] = \sigma^2/N$, Eq. (19) can be simplified as

$$\begin{aligned}
&E[z(k)z(N-k+2)] \approx \mu\nu\Big\{|\lambda(k)|^2 E[s(k)s^*(k)] \\
&+ |\lambda(N-k+2)|^2 E[s(N-k+2)s^*(N-k+2)]\Big\}.
\end{aligned} \tag{20}$$

On the other hand, the second expectation term to be analyzed can be written as

$$\begin{aligned}
&E[|z(k) + z^*(N-k+2)|^2] \\
&= [|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu)]|\lambda(k)|^2 E[s(k)s^*(k)] \\
&+ [\nu(\mu^* + \nu)]\lambda(k)\lambda(N-k+2)E[s(k)s(N-k+2)] \\
&+ [\mu(\mu + \nu^*)]\lambda^*(k)\lambda^*(N-k+2) \\
&\cdot E[s^*(k)s^*(N-k+2)] \\
&+ [|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu)]|\lambda(N-k+2)|^2 \\
&\cdot E[s(N-k+2)s^*(N-k+2)] \\
&+ [|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu)] \\
&\cdot \left\{E[w(k)w^*(k)] + E[w(N-k+2)w^*(N-k+2)]\right\}.
\end{aligned} \tag{21}$$

Similarly, Eq. (21) can be simplified as

$$\begin{aligned}
&E[|z(k) + z^*(N-k+2)|^2] = [|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu)] \\
&\cdot \Big\{|\lambda(k)|^2 E[s(k)s^*(k)] + |\lambda(N-k+2)|^2 \\
&\cdot E[s(N-k+2)s^*(N-k+2)]\Big\}.
\end{aligned} \tag{22}$$

Merging Eqs. (20) and (22) results in the backbone equation of the proposed scheme for imbalance parameter estimation:

$$\begin{aligned}
\varUpsilon &= \frac{E[z(k)z(N-k+2)]}{E[|z(k) + z^*(N-k+2)|^2]} \\
&\approx \frac{\mu\nu}{|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu)}.
\end{aligned} \tag{23}$$

Substituting Eq. (7) into Eq. (23), it is straightforward to derive

$$\begin{aligned}
|\mu|^2 + |\nu|^2 + 2\Re(\mu\nu) &= 1 + \alpha^2 + 2\alpha \\
&= (1 + \alpha)^2.
\end{aligned} \tag{24}$$

Hence, Eq. (23) can be simplified as

$$\varUpsilon = \frac{\alpha - \mathrm{j}[(1-\alpha^2)/2]\sin\theta}{(1+\alpha)^2}. \tag{25}$$

Eq. (25) implies that the real and imaginary parts of the left and right sides should be identical and thus, the amplitude imbalance parameter can be obtained by

$$\left\{\begin{aligned}
\hat{\alpha}_1 &= \frac{1 - 2\Re(\varUpsilon) - \sqrt{1 - 4\Re(\varUpsilon)}}{2\Re(\varUpsilon)}, \\
\hat{\alpha}_2 &= \frac{1 - 2\Re(\varUpsilon) + \sqrt{1 - 4\Re(\varUpsilon)}}{2\Re(\varUpsilon)}.
\end{aligned}\right. \tag{26}$$

Given that α is close to 0, $\hat{\alpha}_1$ is the desired solution, whereas $\hat{\alpha}_2$ is the mirror solution. Similarly, the phase imbalance parameter can be derived:

$$\hat{\theta} = \arcsin\left(-\frac{2(1+\hat{\alpha})^2}{1-\hat{\alpha}^2}\Im(\varUpsilon)\right). \tag{27}$$

Therefore, the I/Q imbalance parameters can be estimated with only the received signals and without a need for pilot signal.

3.2 I/Q imbalance compensation scheme

Using the I/Q imbalance estimates $\hat{\alpha}$ and $\hat{\theta}$, it is easy to obtain the parameters $\hat{\mu}$ and $\hat{\nu}$. Thus, by using the estimates $\hat{\mu}$ and $\hat{\nu}$, we can correct the I/Q imbalance corrupted frequency domain vector $\boldsymbol{z}$ as

$$\hat{\boldsymbol{y}} = \frac{\hat{\mu}^*\boldsymbol{z} - \hat{\nu}\boldsymbol{z}^\dagger}{|\hat{\mu}|^2 - |\nu|^2}, \tag{28}$$

where $\hat{\boldsymbol{y}}$ is an estimate of the correct frequency domain vector $\boldsymbol{y}$. With the corrected frequency domain vector $\hat{\boldsymbol{y}}$, the transmitted signal can be recovered by implementing an equalization operation (Fig. 3) at the receiver.

Based on the above analysis, we propose a new OFDM receiver architecture as depicted in Fig. 3. At the receiver, the received data is first transformed into a parallel form, before removing the cyclic prefix and applying the fast Fourier transform (FFT) operation. Subsequently, several frames of the OFDM symbols are stored and used to calculate the relevant statistic $\varUpsilon$. With the obtained statistic $\varUpsilon$, I/Q imbalance parameters α and θ can be derived according to Eqs. (26) and (27). Finally, using the derived I/Q imbalance parameters α and θ, which are equivalent to μ and ν, respectively, we can remove the effect of I/Q imbalance on the received data with Eq. (28). In addition, the I/Q imbalance is introduced by the analog circuits for up/down conversion, and does not vary with time. Once the I/Q imbalance is corrected, there is no need to deal with the I/Q imbalance periodically. In a practical receiver, the expectation terms of Eq. (23) must be replaced by sample-based approximations. This can be done by an averaging operation over multiple pairs of uncorrelated subcarriers. Furthermore, based on the assumption that the channel is static or quasi-static, an averaging over time is reasonable. Hence, the estimation of the statistic $\varUpsilon$ can be formally written as

$$\varUpsilon = \frac{1}{M}\sum\nolimits_{m\in\mathcal{M}} \frac{\sum_{k\in\mathcal{N}} z_m(k)z_m(N-k+2)}{\sum_{k\in\mathcal{N}} |z_m(k)+z_m^*(N-k+2)|^2}, \tag{29}$$

where $\mathcal{M}$ denotes the chosen subset of M sample time indices, and $\mathcal{N}$ denotes the chosen subset of N subcarrier indices. Obviously, the accuracy of the estimation will be affected by the number of incorporated sample pairs $M \cdot N$. An increased subcarrier block size N raises the computational demand at each time instant m, whereas an increased temporal block size M raises the duration of the parameter estimation. Hence, the proposed parameter estimation allows for a tradeoff among accuracy, computational effort, and measurement time.

3.3 Effect of timing offset

Assuming that the timing offset is smaller than the guard interval, if the timing offset τ is present, the received signal with both the timing offset and I/Q imbalance becomes $z_m^{(t)}(k) = \mu y'_m(k) + \nu y'_m(k)(N-k+2)$, where $y'_m(k)$ is the received signal affected by the timing offset without I/Q imbalance and is given by $y'_m(k) = y_m(k)\mathrm{e}^{-\mathrm{j}2\pi k\tau/N}$. We can thus obtain the received signal with both timing offset and I/Q imbalance as

$$z_m^{(t)}(k) = [\mu y_m(k) + \nu y_m^*(N-k+2)]\,\mathrm{e}^{-\mathrm{j}2\pi k\tau/N}. \tag{30}$$

Therefore, the timing offset causes a phase rotation in the frequency domain. Since

$$z^{(t)}(N-k+2) = [\mu y_m(N-k+2) + \nu y_m^*(k)]\,\mathrm{e}^{-\mathrm{j}2\pi k\tau/N},$$

using $z_m^{(t)}(k)$ to calculate the statistic $\varUpsilon$, we can obtain

$$E[z^{(t)}(k)z^{^{(t)}}(N-k+2)] = E[z(k)z(N-k+2)]. \tag{31}$$

Similarly, since

$$\begin{aligned} & z^{(t)}(k) + z^{(t)*}(N-k+2) \\ & = [z(k) + z^*(N-k+2)]\,\mathrm{e}^{\mathrm{j}2\pi k\tau/N}, \end{aligned}$$

we can obtain

$$\begin{aligned} & E[|z^{(t)}(k) + z^{(t)*}(N-k+2)|^2] \\ =& E[(z^{(t)}(k) + z^{(t)*}(N-k+2)) \\ & \cdot (z^{(t)}(k) + z^{(t)}*(N-k+2))^*] \\ =& E[|z(k) + z^*(N-k+2)|^2]. \end{aligned} \tag{32}$$

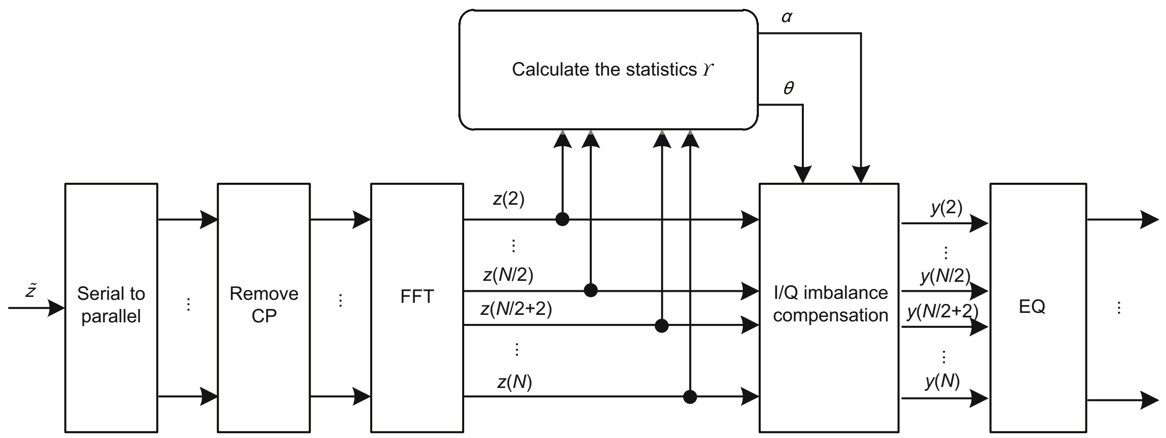

Fig. 3 Framework of the proposed I/Q imbalance compensation scheme in OFDM direct-conversion transceivers (CP: cyclic prefix; EQ: equalization)

Hence, the numerator and denominator used to calculate the statistic Υ are the same as those in Eq. (23) without timing offset. Therefore, the considered timing offset does not affect the imbalance estimates in the proposed method.

4 Simulations and analysis

In this section, simulations are employed to demonstrate the performance of the proposed method. We first evaluate the precision of the I/Q imbalance parameter estimation. Then the BER performance of the proposed I/Q imbalance compensation scheme is investigated.

In the simulations, the tested methods are applied to the OFDM systems with 256-point FFT/inverse FFT (IFFT) (Koffman and Roman, 2002). The number of subcarriers is 256 and the cyclic prefix has 64 samples (1/4 of an OFDM symbol). The symbol duration including the guard interval is 80 μs, the sampling period is 1/4 μs, and the subcarrier spacing is 15.625 kHz. The pilots in the OFDM systems consist of scattered pilots and continual pilots. The continual pilot cells are located every 10 symbols, while the locations of the scattered pilot subcarriers are 41, 66, 91, 116, 142, 167, 192, and 217, respectively, and are the same in different OFDM symbols. The continual pilot cells are used to evaluate the channel response, whereas the scattered pilot carriers are used to trace the phase response. The gain mismatch is set from 0.01 to 0.08, and the phase mismatch is set from 0° to 10° (Dawkins, 2002). Performance evaluation is based on the normalized mean-squared error (NMSE) of the I/Q imbalance parameter estimation and the uncoded BER. The constellation scheme adopted here is 16-quadrature amplitude modulation (QAM). Finally, every result below is obtained by averaging 500 tests.

4.1 I/Q imbalance parameter estimation

First, we evaluate the I/Q imbalance parameter estimation performance in the case with ideal carrier frequency synchronization and time synchronization.

Fig. 4 shows the NMSE of the estimated I/Q imbalance parameters at different signal-to-noise ratios (SNRs) on AWGN and multipath channels. The gain and phase imbalances are assumed to be 0.05 and 10°, respectively, and the number of OFDM symbols used for I/Q imbalance estimation is 50. The multipath channel has six paths, and the power and delay are $[-4.3145, -4.3074, -6.9822, -13.8955, -37.1012, -17.6018]^{\mathrm{T}}$ dB and $[0, 2, 4, 7, 11, 16]^{\mathrm{T}}$ samples, respectively. Simulation results show that the proposed method performs well on both AWGN and multipath channels. Moreover, it can be observed that the SNR has little influence on the estimation accuracy of the I/Q imbalance parameters.

Fig. 5 shows the NMSE of the estimated I/Q imbalance parameters with different OFDM symbol lengths on AWGN and multipath channels. The gain and phase imbalances are also assumed to be 0.05 and 10°, respectively, and the SNR is set as 15 dB. The multipath channel considered here is the same as that used in the last experiment. Simulation results show that the estimation precision becomes higher with the increase in the length of the OFDM symbols, while the improvement in estimation precision is limited when the length of the OFDM symbols is larger than 100.

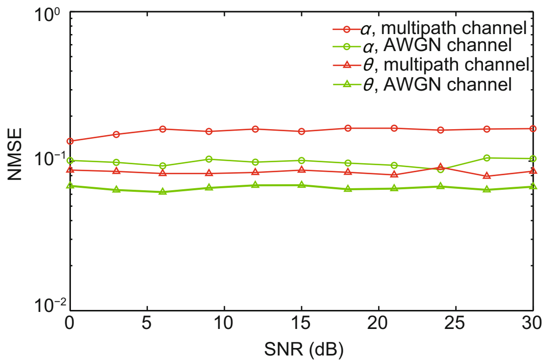

Fig. 4 Normalized mean-squared error (NMSE) of the estimated I/Q imbalance parameters versus the SNR of received signals (α: gain imbalance; θ: phase imbalance)

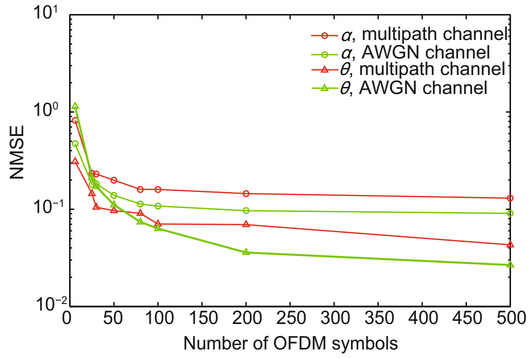

Fig. 5 Normalized mean-squared error (NMSE) of the estimated I/Q imbalance parameters versus the number of symbols (α: gain imbalance; θ: phase imbalance)

4.2 BER performance evaluation

Second, we evaluate the BER performance of the I/Q imbalance compensation scheme in the case of ideal carrier frequency synchronization and time synchronization. Besides, no channel coding is considered in the simulations.

Fig. 6 shows the BER performance of the compensated received signal at different SNRs on the AWGN channel. The gain imbalance is assumed as 0.05 and 0.08, respectively, and the phase imbalance is 5°. The number of OFDM symbols used for I/Q imbalance estimation is 50. Simulation results show that: (1) if the I/Q imbalance is uncompensated, system performance will be severely degraded; (2) the BER performance of the OFDM system deteriorates with an increase in the gain imbalance without I/Q imbalance compensation; (3) the BER performance of the OFDM system can be close to that achieved in the case with no I/Q imbalance by applying I/Q imbalance compensation, and different gain imbalance factors have little effect on BER performance. Fig. 7 shows the BER performance of the compensated received signal at different SNRs on the multipath channel. The gain imbalance is assumed to be 0.05 and 0.08, respectively, and the phase imbalance is assumed to be 5°. The number of OFDM symbols used for I/Q imbalance estimation is 50. Simulation results show that: (1) the BER performance of the OFDM system deteriorates with an increase in the gain imbalance without I/Q imbalance compensation; (2) the OFDM system can also achieve BER performance close to that in the case with no I/Q imbalance by applying I/Q imbalance compensation when the SNR is less than 30 dB. However, compared with the no I/Q imbalance case, there is still a gap even using the proposed compensation scheme when the SNR is larger than 30 dB. The main reason is that when the SNR is low, SNR is the main factor hindering the improvement of BER performance, while the residual I/Q imbalance is the main factor hindering BER performance improvement when SNR is high.

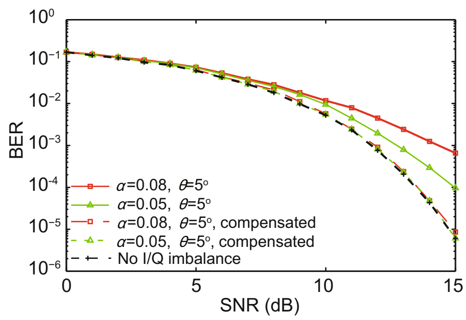

Fig. 6 Bit error rate (BER) of the compensated received signal at different SNRs in the AWGN with different gain imbalances (α: gain imbalance; θ: phase imbalance)

Fig. 8 shows the BER performance of the compensated received signal at different SNRs on the AWGN channel. The gain imbalance is assumed to be 0.05, while the phase imbalance is assumed to be 2° and 10°, respectively. The number of OFDM symbols used for I/Q imbalance estimation is 50. Fig. 9 shows the BER performance of the compensated received signal at different SNRs on the multipath channel. From the simulation results, we can draw similar conclusions to the case with different gain imbalance factors.

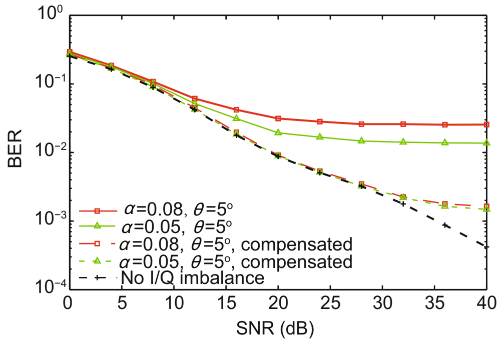

Fig. 7 Bit error rate (BER) of the compensated received signal at different SNRs in the multipath channel with different gain imbalances (α: gain imbalance; θ: phase imbalance)

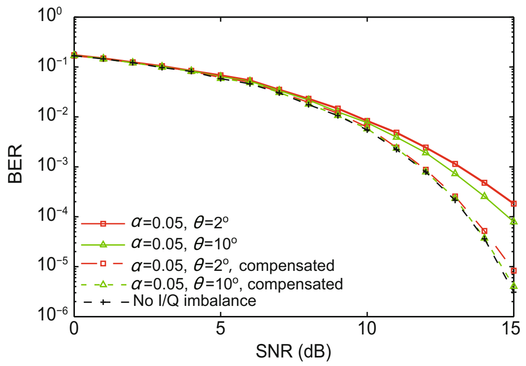

Fig. 8 Bit error rate (BER) of the compensated received signal at different SNRs in the AWGN with different phase imbalances (α: gain imbalance; θ: phase imbalance)

5 Conclusions

In this paper, we have investigated the I/Q imbalance compensation issue in OFDM direct-conversion receivers. First, we showed that the I/Q imbalance in OFDM direct-conversion receivers is the same as introducing interference from mirror subcarriers, which will destroy the orthogonality between the subcarriers and result in poor BER

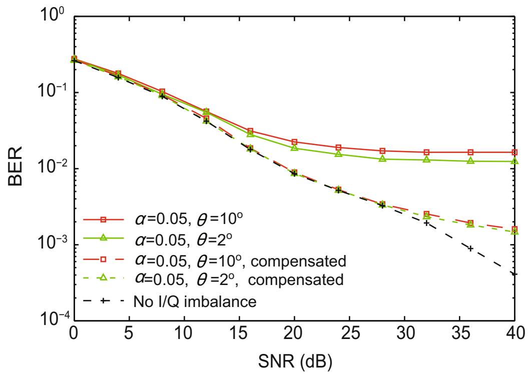

Fig. 9 Bit error rate (BER) of the compensated received signal at different SNRs in the multipath channel with different phase imbalances (α: gain imbalance; θ: phase imbalance)

performance. Second, we designed a new statistic related only to the I/Q imbalance, and further estimated the I/Q imbalance parameters. In this way, the distortions introduced by the I/Q imbalance and the channel can be independently corrected by I/Q imbalance compensation and equalization. Simulation results showed that the proposed method can achieve good performance in AWGN and multipath environments. Specifically, the proposed I/Q imbalance compensation method can achieve BER performance close to that in the ideal case with no I/Q imbalance.

References

Abidi AA, 1995. Direct-conversion radio transceivers for digital communications. *IEEE J Sol-State Circ*, 30(12): 1399-1410. https://doi.org/10.1109/4.482187

Anttila L, 2011. Digital Front-End Signal Processing with Widely-Linear Signal Models in Radio Devices. PhD Thesis, Tampere University of Technology, Tempere, Finland.

Anttila L, Valkama M, 2013. Blind signal estimation in widely-linear signal models with fourth-order circularity: algorithms and application to receiver I/Q calibration. *IEEE Signal Process Lett*, 20(3):221-224. https://doi.org/10.1109/LSP.2012.2230625

Chen S, Zhao J, 2014. The requirements, challenges, and technologies for 5G of terrestrial mobile telecommunication. *IEEE Commun Mag*, 52(5):36-43. https://doi.org/10.1109/MCOM.2014.6815891

Dawkins MT, 2002. Up-Integration in Radio-Frequency Tuners for Digital Terrestrial Television. PhD Thesis, University of London, UK.

Gu F, Wang S, Wei J, et al., 2016. Higher-order circularity based I/Q imbalance compensation in direct-conversion receivers. IEEE Int Conf on Vehicular Technology, p.1-6. https://doi.org/10.1109/VTCFall.2016.7881033

He L, Ma S, Wu Y, et al., 2011. Pilot-aided IQ imbalance compensation for OFDM systems operating over

doubly selective channels. *IEEE Trans Signal Process*, 59(5):2223-2233.
https://doi.org/10.1109/TSP.2011.2112649

Hieu NT, Ryu HG, Wang CX, et al., 2007. The impact of the I/Q mismatching errors on the BER performance of OFDM communication systems. IEEE Int Conf on Communications, p.5423-5427.
https://doi.org/10.1109/ICC.2007.898

IEEE, 1999. Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) specifications: high-speed physical layer in the 5 GHz band. IEEE Std 802.11a-1999.

Inamori M, Bostamam AM, Sanada Y, et al., 2009. IQ imbalance compensation scheme in the presence of frequency offset and dynamic DC offset for a direct conversion receiver. *IEEE Trans Wirel Commun*, 8(5):2214-2220.
https://doi.org/10.1109/TWC.2009.080139

Kim S, Yoon D, Park H, et al., 2014. A fast and precise blind I/Q mismatch compensation for image rejection in direct-conversion receiver. *ETRI J*, 36(1):12-21.
https://doi.org/10.4218/etrij.14.0113/0223

Koffman I, Roman V, 2002. Broadband wireless access solutions based on OFDM access in IEEE 802.16. *IEEE Commun Mag*, 40(4):96-103.
https://doi.org/10.1109/35.995857

Lopez-Martinez FJ, Martos-Naya E, Paris JF, et al., 2011. Exact closed-form BER analysis of OFDM systems in the presence of IQ imbalances and ICSA. *IEEE Trans Wirel Commun*, 10(6):1914-1922.
https://doi.org/10.1109/TWC.2011.032411.101356

Reimers U, 1997. DVB-T: the COFDM-based system for terrestrial television. *Electron Commun Eng J*, 9(1):28-32.
https://doi.org/10.1049/ecej:19950309

Shelswell P, 1995. The COFDM modulation system: the heart of digital audio broadcasting. *Electron Commun Eng J*, 7(3):127-136.
https://doi.org/10.1049/ecej:19950309

Sung K, Chao C, 2009. Estimation and compensation of I/Q imbalance in OFDM direct-conversion receivers. *IEEE J Sel Top Signal Process*, 3(3):438-453.
https://doi.org/10.1109/JSTSP.2009.2020241

Tandur D, Moonen M, 2007. Joint adaptive compensation of transmitter and receiver IQ imbalance under carrier frequency offset in OFDM-based systems. *IEEE Trans Signal Process*, 55(11):5246-5252.
https://doi.org/10.1109/TSP.2007.898788

Tarighat A, Bagheri R, Sayed AH, 2005. Compensation schemes and performance analysis of IQ imbalances in OFDM receivers. *IEEE Trans Signal Process*, 53(8): 3257-3268. https://doi.org/10.1109/TSP.2005.851156

Windisch M, Fettweis G, 2004. Standard-independent I/Q imbalance compensation in OFDM direct-conversion receivers. Int OFDM Workshop, p.57-61.

Wu F, Li Y, Zhao M, 2014. Estimation of TX I/Q imbalance at the RX side with RX I/Q imbalance and carrier frequency offset for OFDM systems. IEEE Int Conf on Globecom, p.960-965.
https://doi.org/10.1109/GLOCOMW.2014.7063557

Yang L, Panta K, Armstrong J, 2013. Impact of timing jitter and I/Q imbalance in OFDM systems. *IEEE Commun Lett*, 17(2):253-256.
https://doi.org/10.1109/LCOMM.2012.121912.122280

Ylamurto TM, 2003. Frequency domain IQ imbalance correction scheme for orthogonal frequency division multiplexing (OFDM) systems. IEEE Wireless Communications Networking Conf, p.20-25.
https://doi.org/10.1109/WCNC.2003.1200313