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I/Q Linear Phase Imbalance Estimation Technique of the Wideband Zero-IF Receiver

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Abstract: The in-phase/quadrature (I/Q) imbalance encountered in the zero-IF receiver leads to incomplete image frequency suppression, which severely deteriorates image rejection ratio (IRR) of the receiver system and must be improved using additional analog or digital signal processing. The I/Q linear phase imbalance (LPI) is the key of the I/Q imbalance, which consists of the time delay deviation (TDD) and the local oscillator (LO) phase offset. TDD is negligible in most literature, but it degrades system performance largely for wideband communication systems. This paper proposes a method based on the cross-power spectrum between the I/Q signal to address the estimation problem of LPI. Compared with other conventional methods, the proposed approach calculates LPI parameters simultaneously without any additional hardware. The MATLAB simulation is utilized to evaluate the effectiveness of the presented method. Moreover, the experimental platform of detailed design demonstrates the feasibility of the proposed estimation method, and IRR of the system before and after compensation shows that LPI has been accurately estimated and eliminated with the help of an appropriate compensation structure. Both reveal that the proposed method offers an effective solution to the LPI problem.

Keywords: I/Q phase imbalance; zero-IF receiver; time delay estimation; cross-power spectrum; phase unwrapping

1. Introduction

In the past few decades, with the development of a wideband communication system, realizing the ultra-wide bandwidth is the key to realize ultra-high throughput, e.g., 5G networks [1], thus the in-phase and quadrature (I/Q) orthogonal demodulation structure [2,3] has been favored because it can relax the bandwidth and sampling rate pressure of analog-to-digital converters (ADCs). Therefore, the zero-IF architecture, which adopts the I/Q orthogonal demodulation, is getting more and more applications for its advantages of simple structure, low cost, and power consumption [4,5].

However, the mixing structure is particularly sensitive to the amplitude and phase imbalance between I and Q branches, i.e., I/Q imbalance (IQI). The IQI occurs when the I branch exhibits a different amplitude or does not have a precise phase difference from the Q branch [6]. In the actual situation, inconsistency between I/Q datapath, e.g., different devices and circuit design, makes the phase and amplitude differences between the I and Q signals [7,8]. IQI has a significant impact on the receiver system performance, which leads to the incomplete image signal rejection, and degrades the image rejection ratio (IRR) of baseband signal processing [9]. Reference [10] points out that the phase mismatch of 2 to 5 degrees can lead to the IRR of only 20–30 dB.

In the existing literature, a lot of research has been devoted to the I/Q amplitude mismatch and achieved excellent results [11,12]. The I/Q phase imbalance has become the critical problem of IQI due to its hardness for estimation. Here, the local oscillator (LO) phase offset and the time delay deviation (TDD) of I and Q signals considered in this paper are the two main parts of I/Q phase imbalance. Since they have a linear relationship with frequency, these two parts collectively compose the I/Q linear phase imbalance (LPI). Different phase offset is generally caused by the imperfection of the analog LO, and mismatch introduced by the LO phase offset is frequency-independent [13]. TDD is the result of different delays introduced by the analog components or different path lengths of I and Q branches [14,15]. The influence of TDD is similar to that of the LO phase offset, and the difference is that the mismatch brought by TDD is frequency-dependent, whereas that of the phase offset is constant over the entire signal bandwidth. LO phase offset can be considered as a constant, which is the main error considered in narrowband communication systems [16]. However, as for wideband and high-speed communication systems, the phase imbalance introduced by TDD is getting worse with the increase of frequency, which is supposed to be considered carefully.

Thus, it is necessary to estimate and eliminate LPI to improve system performance. Generally, it is hard to settle IQI via hardware effort, an effective solution without complex circuitry and expensive hardware resource is necessary to compensate for the effects of the nonideal analog components in the digital domain. At present, there is still a great deal of research devoted to the IQI problem. In [17], LA designs a filter in quadrature receivers to estimate IQI by calculating the statistic character of the I/Q signal. However, only the frequency-independent part of IQI is removed. The solution presented in [18] calculates the IQI by instantaneous power measurement of I/Q signal, but additional envelope-detector is needed, which is not practical for wideband systems because of its requirement for a very high-sampling rate ADC. The approach for the TDD and phase offset estimation in [19] is divided into two steps, which adds operation complexity. More severely, the algorithm introduces a nonlinear group delay into the system and smears the original input signal more seriously.

This paper provides new insight into I/Q phase imbalance and introduces a much simpler approach based on the cross-power spectrum [20,21] between the I/Q signal, which can simultaneously conduct TDD and LO phase offset estimation. The technique is independent of the amplitude mismatch and brings about the probability of compensating the I/Q amplitude mismatch and phase mismatch separately to maintain the various needs of all kinds of application scenarios. The compensation structure with a fractional delay (FD) filter and a delay module is applied to compensate for the estimated mismatch parameters. The accuracy and effectiveness of the proposed technique are validated through MATLAB simulations and experiments in the real-world hardware platform.

The rest of this paper is organized as follows. In Section 2, the mathematical model of the zero-IF receiver in the presence of LPI is given and the influence of LPI on the IRR is analyzed. The details of LPI estimation and compensation by the proposed method are described in Section 3. Sections 4 and 5 present the numerical results of the MATLAB simulation and hardware platform experiments to verify the presented estimation method. Finally, the conclusion is drawn in Section 6.

2. Model of I/Q LPI in Zero-IF Receiver

As depicted in Figure 1, the received RF analog signal is denoted as $r(t)$. The two branches of LO output used for quadrature demodulation are $\cos(\omega_c t)$ and $-g \sin(\omega_c t + \varphi)$, respectively. ω_c is the carrier frequency, φ is the LO phase offset, and g is the LO gain mismatch. After the quadrature down conversion of $r(t)$, the lowpass filters (LPFs) are added to remove the high-frequency components, and ADCs are used for the signal sampling. The time delays of I and Q branches are denoted as δ_I and δ_Q , respectively. In addition, due to the difference of I/Q path, $\delta_I \neq \delta_Q$, and $\Delta\delta = \delta_Q - \delta_I$ is the TDD between the I/Q channel.

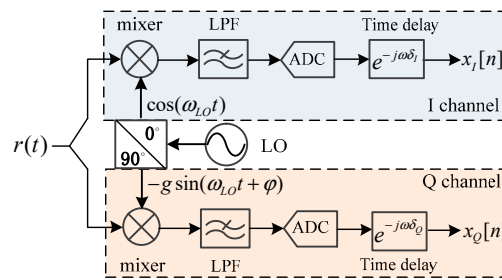


Figure 1. LPI model of the zero-IF Receiver.

The digitalized I and Q signals with the interference of I/Q imbalance are given by $x_I[n]$ and $x_Q[n]$. The I/Q amplitude mismatch is $A(\omega) = gA_Q(\omega)/A_I(\omega)$, where $A_I(\omega)$ and $A_Q(\omega)$ are the amplitude of the I/Q channel, respectively.

The received I and Q signals can be written as a complex signal $X(\omega)$ in the frequency domain [22]

$$X(\omega) = H_1(\omega)S(\omega) + H_2(\omega)S^*(-\omega) \quad (1)$$

where

$$H_1(\omega) = \frac{1 + A(\omega)e^{-j(\omega\Delta\delta + \varphi)}}{2}, H_2(\omega) = \frac{1 - A(\omega)e^{j(-\omega\Delta\delta + \varphi)}}{2} \quad (2)$$

$X(\omega)$ is composed of two parts, in which $S(\omega)$ is the frequency equivalent of the ideal baseband complex signal weighted by $H_1(\omega)$, and $S^*(-\omega)$ is the frequency equivalent of the undesired image weighted by $H_2(\omega)$.

In order to quantify the effect of image rejection, the definition of IRR is proposed, which is the power ratio of the desired signal to the undesired image signal. It follows from (2) that IRR is given by

$$IRR(\omega) = \frac{|H_1(\omega)|^2}{|H_2(\omega)|^2} \quad (3)$$

IRR is infinite when the phase and amplitude responses of the I and Q channels are equal, i.e., $\Delta\delta = 0$, $\varphi = 0$, and $A(\omega) = 1$. When this condition is not met, the image signal is not completely rejected and the I/Q signal is imbalanced. Based on Equations (2) and (3), IRR can be rewritten as

$$IRR(\omega) = \frac{|1 + A(\omega)e^{-j(\omega\Delta\delta + \varphi)}|^2}{|1 - A(\omega)e^{j(-\omega\Delta\delta + \varphi)}|^2} \quad (4)$$

IRR curves in the presence of LPI are shown in Figure 2. For simplicity, $A(\omega)$ is assumed to be 1 for the intuitive description of the influence of LPI. From the red curve in Figure 2 we can see that, if $\Delta\delta = 1$, IRR is far below 0 dB in many frequency points within Nyquist bandwidth. It shows that the IRR decreases when signal frequency increases, which means that TDD imposes a worse influence on high-frequency signals. Note that an IRR of 0 dB indicates that the power of the image signal is the same as that of the desired signal, while a negative IRR indicates the power of the image signal is higher than that of the desired signal. Hence, even $\Delta\delta = 1$ will cause the system performance to deteriorate significantly. Moreover, when $\Delta\delta = 2$, the black curve shows that IRR decreases with the increase of frequency at a faster speed than that of the red one, which means that the image signal increases with the increase of TDD at the same frequency. In addition, it can be observed that IRR is a period function when $\Delta\delta = 2$ and becomes a very small value for the normalized frequency $f = (2n + 1)/(2 \times \Delta\delta)$, where n is an integer. The straight blue line depicts that the mismatch caused by the LO phase offset is frequency-independent, where the induced IRR does not vary with frequency and is a constant proportional to the value of the LO phase offset.

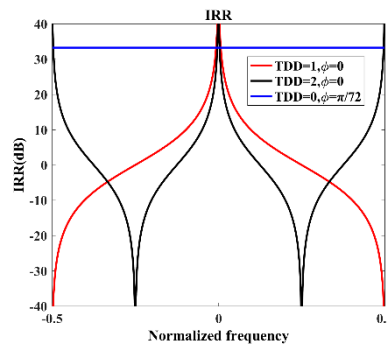


Figure 2. IRR curves with TDD and LO phase offset.

3. LPI Estimation and Compensation

Above all, TDD and LO phase offsets both have a significant influence on wideband receiver systems, and an efficient method is needed urgently to eliminate LPI. In this paper, we come up with an estimation technique based on the cross-power spectrum and the “three-point unwrapping” method.

3.1. LPI Extraction by the Cross-Power Spectrum

The single tone is injected in the transmitted signal and received by the receiver, which is used as the test tone for the phase imbalance estimation, then the test I/Q signal can be represented as

$$x_I[n] = A_I(\omega_0) \cos(\omega_0 n) + w_I[n] \quad x_Q[n] = A_Q(\omega_0) \sin(\omega_0(n - \Delta\delta) - \varphi) + w_Q[n] \quad (5)$$

where ω_0 is the angular frequency of the test tone. $w_I[n]$ and $w_Q[n]$ are the noise part of the I/Q signal, respectively, which are assumed to be uncorrelated to each other.

Because of the orthogonal relationship between the I/Q signal, the original phase deviation, i.e., $\pi/2$ needs to be neutralized before the estimation of LPI. We choose Hilbert transform [23] to shift the Q signal $x_Q[n]$ by $\pi/2$, which is denoted as $x'_Q[n]$ and drawn by the black dotted in Figure 3. The amplitude of $x'_Q[n]$ is equal to $x_Q[n]$ due to the all-pass quality of Hilbert transform in the pass-band.

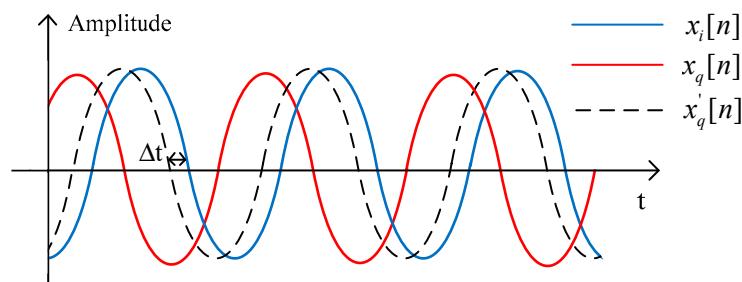


Figure 3. Procedure of Hilbert transform.

From Figure 3, the test Q signal after the Hilbert transform is a phase-mismatched equivalent of the I signal, and the mismatch is denoted as Δt , i.e., the time-domain equivalent of LPI between I/Q channel. Then, the cross-power spectrum calculation can be performed on these two signals to extract LPI.

The Discrete Fourier transform (DFT) of $x_I[n]$ and $x'_Q[n]$ are $X_I[k]$ and $X'_Q[k]$, respectively, and the cross-power spectrum between $X_I[k]$ and $X'_Q[k]$ can be calculated as

$$\begin{aligned} G_{x_I x_Q}[k] &= X_I[k] X_Q^*[k] \\ &= A[k] G_{x_I x_I}[k] e^{j(\frac{2\pi}{N} k \Delta\delta + \varphi)} + G_{w_I w_Q}[k] \end{aligned} \quad (6)$$

where $X_Q^*[k]$ is the conjugate of $X_Q[k]$, and $G_{x_I x_I}[k]$ represents the auto-power spectrum of the I signal, and N is the DFT length. $G_{w_I w_Q}[k]$ denotes the cross-power spectrum between noise $w_I[n]$ and $w_Q[n]$, which is zero because of the uncorrelation between the two parts of noise. The cross-power spectrum between baseband signals and their noise portion are assumed to be uncorrelated, hence their cross-power spectrums are both zero. Therefore,

$$\begin{aligned} G_{x_I x_Q}[k] &= X_I[k] X_Q^*[k] \\ &= A[k] G_{x_I x_I}[k] e^{j(\frac{2\pi}{N} k \Delta \delta + \varphi)}, k > 0 \end{aligned} \quad (7)$$

On account of the different signs of phase characteristic in the positive and negative frequency part of $G_{x_I x_Q}[k]$, only the positive frequency part ($k > 0$) of the spectrum is taken when calculating TDD and LO phase offset, for simplicity.

The phase characteristic of $G_{x_I x_Q}[k]$ can be calculated as

$$\Delta\theta[k] = \arg\left(\frac{G_{x_I x_Q}[k]}{|G_{x_I x_Q}[k]|}\right) = (2\pi/N)\Delta\delta k + \varphi \quad (8)$$

where $\Delta\theta[k]$ is the LPI factor between $X_I[k]$ and $X_Q^*[k]$, which includes the unknown TDD $\Delta\delta$ and LO phase offset φ . $\Delta\theta[k]$ can be interpreted as a linear function of the k th spectral line with slope $(2\pi/N)\Delta\delta$ and intercept φ . Therefore, TDD $\Delta\delta$ can be estimated from the gradient of $\Delta\theta[k]$, then the LO phase offset φ can be calculated from the intercept of $\Delta\theta[k]$ -axis.

It should be highlighted that the cross-power spectrum is divided by its modulus in Equation (8), thus the influence of amplitude mismatch is neutralized and does not affect the accuracy of LPI calculation.

Due to the periodicity of the tangent function, the LPI factor $\Delta\theta[k]$ wraps between $-\pi$ to π . Therefore, in order to calculate the gradient and intercept accurately, $\Delta\theta[k]$ should be unwrapped in advance. Here, the “Three-point unwrapping” method is proposed to address the problems of phase unwrapping and calculation of $\Delta\delta$ and φ .

3.2. “Three-Point Unwrapping” Method

The unwrapping of $\Delta\theta[k]$, which is finished by adding multiples of 2π to the original data, can be represented as

$$\widetilde{\Delta\theta}[k_i] = \Delta\theta[k_i] + m_i \cdot 2\pi, \quad i = 1, 2, \dots, G \text{ and } m_i \in N, \quad (9)$$

where $\widetilde{\Delta\theta}[k_i]$ denotes the unwrapped version of $\Delta\theta[k_i]$, m_i are integers, and G represents the number of frequency components contained in the input signal. To estimate the slope, $G = 3$ at least. We suppose that $(k_1, \Delta\theta[k_1])$, $(k_2, \Delta\theta[k_2])$, and $(k_3, \Delta\theta[k_3])$ are three sets of wrapped $\Delta\theta[k]$ samples. The slope of $\Delta\theta[k]$ can be obtained by using the first two wrapped samples, which is

$$(2\pi/N)\Delta\delta = (\Delta\theta(k_2) - \Delta\theta(k_1) + m_2 \cdot 2\pi) / (k_2 - k_1) \quad (10)$$

Equation (10) is based on the assumption that the first phase deviation $\Delta\theta[k_1]$ does not need to unwrap, i.e., $m_1 = 0$. Since $(k_1, \Delta\theta[k_1])$ and $(k_2, \Delta\theta[k_2])$ have been available by measurement, the unwrapping factor m_2 is the only remaining variate to estimate. Now that the three sets of points are in one line,

$$\frac{\widetilde{\Delta\theta}[k_3] - \widetilde{\Delta\theta}[k_1]}{k_3 - k_1} = \frac{\widetilde{\Delta\theta}[k_2] - \widetilde{\Delta\theta}[k_1]}{k_2 - k_1}. \quad (11)$$

By inserting Equation (9) into Equation (11) and simplifying, we can obtain

$$(k_3 - k_1)m_2 = C + (k_2 - k_1)m_3 \quad (12)$$

where

$$C = 1/2\pi(k_2 - k_1)(\Delta\theta[k_3] - \Delta\theta[k_1]) \\ - 1/2\pi(k_3 - k_1)(\Delta\theta[k_2] - \Delta\theta[k_1]).$$

From Equation (12), it can be confirmed that $(k_3 - k_1)m_2 - C$ is multiples of $(k_2 - k_1)$, and

$$(k_3 - k_1)m_2 = C(\text{mod}(k_2 - k_1)) \quad (13)$$

where $\text{mod}(\cdot)$ represents the remainder operator. Equation (13) can be solved by a linear modular equation solver [24], which can be written as

$$m_2 = m_{2,0} + (H(k_2 - k_1)/D) \quad (14)$$

The factor D is the greatest common divisor (GCD) of $(k_3 - k_1)$ and $(k_2 - k_1)$. $H = 0, 1, \dots, D - 1$, and $m_{2,0} = A(C/D)(\text{mod}(k_2 - k_1))$. The integer A is the coefficient generated by the extended Euclid algorithm [25], and satisfies the formula $D = A(k_3 - k_1) + B(k_2 - k_1)$, and B is an integer.

Eventually, the unwrapping factor m_2 can be calculated by Equation (14), which is inserted into Equation (10) and $\varphi = \Delta\theta[k_1] - (2\pi/N)\Delta\delta k_1$ can be obtained.

Based on the analysis above, TDD $\Delta\delta$ and LO phase offset φ between $x_I[n]$ and $x_Q[n]$ can be calculated by the following steps:

1. Apply Hilbert transform on $x_Q[n]$ and obtain $x'_Q[n]$.
2. Calculate FT of $x_I[n]$ and $x'_Q[n]$, which are $X_i[k]$ and $X'_q[k]$.
3. Calculate the cross-power spectrum $G_{x_I x'_Q}[k]$ and obtain the LPI factor $\Delta\theta[k]$.
4. TDD $\Delta\delta$ and LO phase offset φ can be calculated from $\Delta\theta[k]$ by the "Three-point unwrapping" method.

3.3. Fractional-Delay Filter Design with Farrow Structure

As a matter of fact, TDD may not be an integer, and the compensation will be implemented by the FD filter in this paper. Farrow structure [26] is an efficient realization structure for variable FD filters, and the structure uses spline fitting or polynomial approximation method to further decompose the filter coefficients into multiple sub-filter banks. Supposing that TDD $\Delta\delta = p$, the impulse response of the FD filter is $h_{FD}[n]$ with the order R , and each coefficient of $h_{FD}[n]$ is further decomposed into L th sub-filters, which is represented by

$$h_{FD}[n] = \sum_{l=0}^{L-1} c_l[n] p^l \quad (15)$$

where $c_l[n]$ represents the l th order real-valued coefficients.

Therefore, the frequency-domain expression for $h_{FD}[n]$ is given by

$$H_{FD}[z, p] = \sum_{n=0}^{R-1} h[n, p] z^{-n} = \sum_{n=0}^{R-1} \left(\sum_{l=0}^{L-1} c_l[n] p^l \right) z^{-n} \\ = \sum_{n=0}^{R-1} \left(\sum_{l=0}^{L-1} c_l[n] z^{-n} \right) p^l = \sum_{n=0}^{R-1} C_l[z] p^l \quad (16)$$

where $C_l[z] = \sum_{l=0}^{L-1} c_l[n] z^{-n}$.

Finally, the TDD compensation is conducted by convolution in time-domain, which is given by

$$x_{Q_TD}[n] = x_Q[n] * h_{FD}[n] \quad (17)$$

In addition, it should be noted that the FD filter is a finite impulse response (FIR) filter, which introduces additional time delay in the compensation channel, hence it is necessary to add a delay module to the counterpart channel to balance the influence of the FD filter.

In practice, the compensation channel can be selected between I/Q channel according to the sign of TDD, i.e., a positive TDD indicates that the Q samples lag behind the I samples in the time domain, then a delay compensator needs to be implemented on I channel, and vice versa.

For the LO phase offset compensation, the phase relationship between the two branches of LO outputs can be changed directly in the MATLAB simulation. In practice, phase shifter [27] with appropriate frequency resolution is adopted to complete the fine phase tuning. The main idea is to adjust the phase relationship between I/Q channel according to the obtained LO phase offset. The detailed operation will be specified in Section 5.

4. Simulation Verification

In this section, the proposed method is evaluated by the simulations in the MATLAB with the instantaneous bandwidth of $[-300 \text{ MHz}, 300 \text{ MHz}]$. The TDD and LO phase offsets of the receiver system are estimated by the test three-tone signal. The system sampling rate is $f_s = 1 \text{ GHz}$, with the sampling period $T_s = 1/f_s$, and the signal-to-noise ratio (SNR) is 35 dB. Suppose that TDD $\Delta\delta = 3T_s$, and LO phase offset $\varphi = \pi/36$. The test signal is set as a three-tone signal, and the frequencies of the test tones are 1100 MHz, 1200 MHz, 1250 MHz, respectively. The frequency of LO is set to 1 GHz. Then, the cross-power spectrum is calculated and the phase spectrum obtained by Equation (8) is shown in Figure 4, and three sets of the input frequency and corresponding phase value are marked.

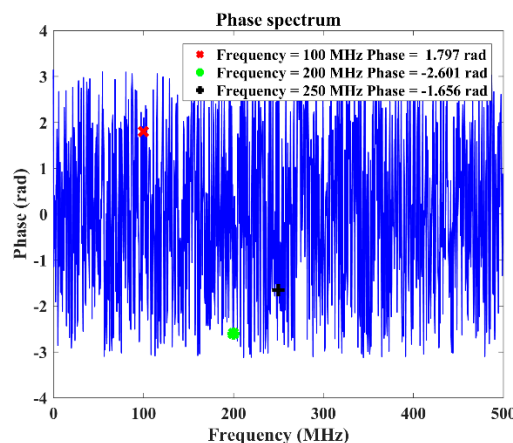


Figure 4. Phase spectrum of cross-spectrum.

Then, TDD and LO phase offsets can be obtained by the method above, which are $\Delta\delta = 3.00037$ and $\varphi = 0.08818$, respectively. The obtained parameters can be used to adjust the set TDD and LO phase offsets. The measured IRRs before and after compensation are calculated and shown in Figure 5. From the inspection of Figure 5a, it is shown that IRR is well below 0 dB to many frequencies within the Nyquist bandwidth. The IRR lobe is shifted from zero frequency due to the existence of LO phase offset. A significant observation obtained from Figure 5b is that the IRR is enhanced to the range of values larger than about 25 dB in most of the bandwidth after the TDD estimation, which is shown by the black IRR curve. After that, the LO phase offset compensation yields the blue IRR curve of around 65 dB. The rising trend indicates the correctness of the method. Note that the fluctuation of IRR curves can be attributed to the slight influence of the noise interference on the estimation results.

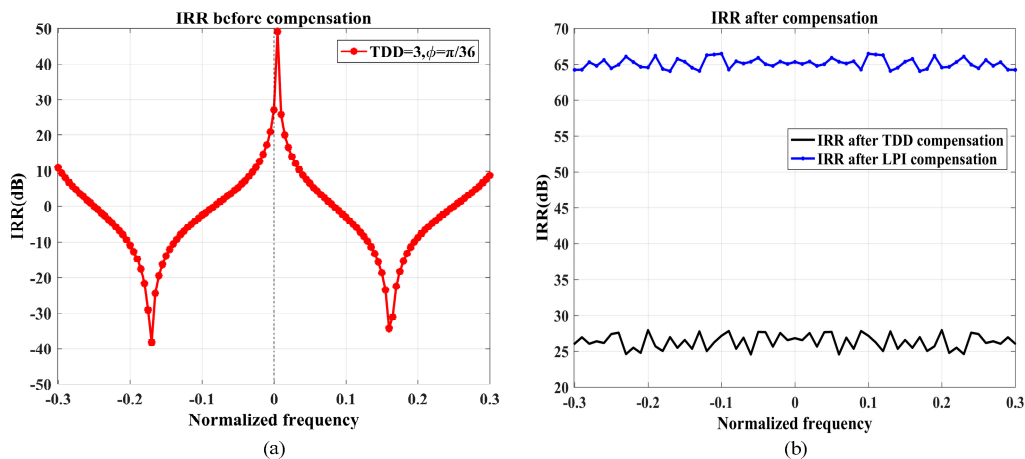


Figure 5. IRR curves: (a) IRR before any compensation and (b) IRR after compensation.

5. Hardware Platform Experiment and Results

In this section, the proposed method will be validated using the two-PXIe-slot receiver platform developed by the authors. The block diagram of this receiver platform is shown in Figure 6. The main parts are the RF signal receiving module, baseband signal processing module, and bus interface module, respectively. The input RF signal is demodulated by a quadrature demodulator after the user-configurable gain and filtering. Then, the demodulated I/Q signal is filtered by two LPFs and separately sampled. The digitalized I and Q signals are sent to the field programmable gate array (FPGA) for signal processing and analysis.

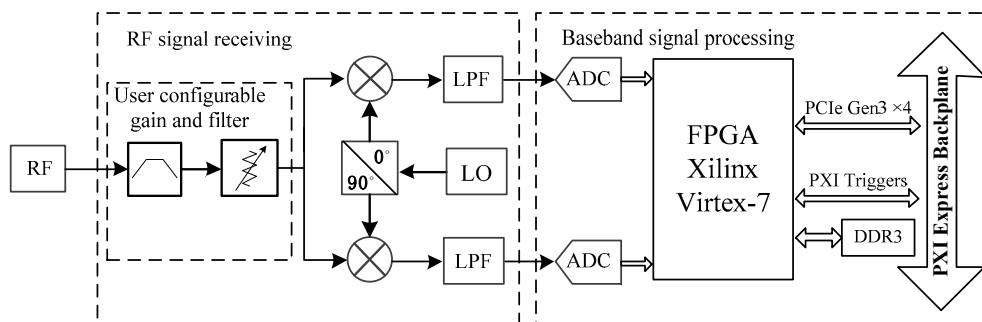


Figure 6. Block diagram of the experimental platform.

The system consists of two printed circuit boards (PCBs), which are a radio frequency signal receiving board (Board I) and a baseband signal processing board (Board II), as shown in Figure 7. In Board I, Port A is used for transmitting digital data and commands, and Port B is used to output the demodulated I/Q signal to Board II. The platform adopts the quadrature demodulator chip LTC5586 to complete demodulation. The ultra-wide IF bandwidth of more than 1GHz makes the LTC5586 particularly suited for the demodulation of ultra-wideband signals.

In Board II, the chip AD9691 is a dual, 14-bit, 1.25 GSPS ADC with two cores sampling analog I and Q signals by the sampling rate of 1.25 GSPS, respectively. The digitalized I/Q signal is transmitted to FPGA I for signal processing, and the processing results are sent to the host computer for display. The main function of FPGA II is PCI Express communication between FPGA I and the host computer via the PXIe interface.

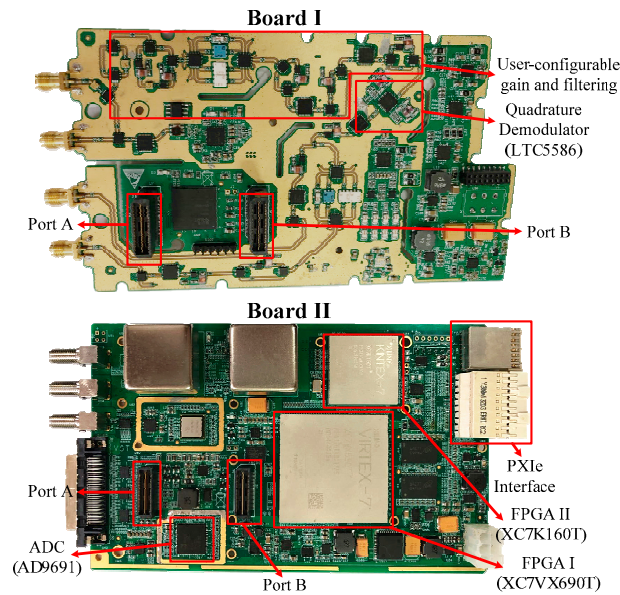


Figure 7. Receiver prototype.

The LPI estimation and compensation presented in this paper are implemented in LTC5586 and FPGA I, which are depicted in Figure 8. LO phase offset is tuned by the phase shifter register embedded in chip LTC5586, which allows for fine-tuning of the phase offset between I/Q channel over a range from -2.5 to 2.5 degrees with a resolution of around 0.05 degrees. In practice, the phase adjustment value is generally controlled by the digital control word (DCW), which is transformed by the estimated LO phase offset according to the DCW calculation rule in LTC5586. The obtained DCW is then sent to the corresponding register address via the SPI interface to control the value of I/Q phase shift by the red feedback path as shown in Figure 8.

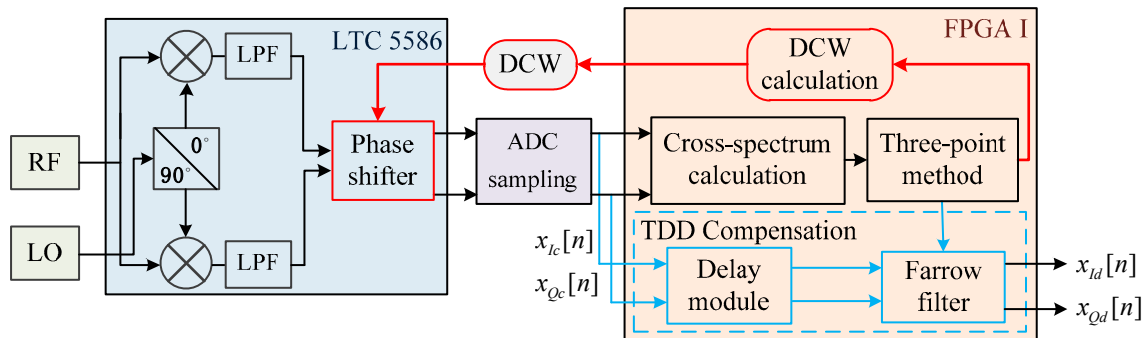


Figure 8. Block diagram of the compensation.

After the LO phase offset is eliminated, the compensated I/Q data $x_{Ic}[n]$ and $x_{Qc}[n]$ are transmitted to the TDD compensation unit as shown in the blue dotted square in Figure 8. As stated in Section 3.3, the TDD compensation unit consists of two parts, i.e., the integer part and fraction part. On the one hand, the TDD part is usually a fraction, and it is adjusted by the FD filter, which is realized by the Farrow filter as stated in Section 3.3.

The order of FD filter $R = 50$, and the order of the sub filters in Farrow structure $L = 3$. According to the minimax design criterion, the FD filter is designed and the bandwidth of the filter is set to 0.8π . Because the design of the FD filter is not the focus of this paper, full details of the filter design can be found in [28]. The magnitude and phase delay responses of the FD filter are shown in Figure 9. The figure indicates that the filter has a flat unity gain for different delay parameters and the phase delay is very stable within 0.8π . Therefore, this FD filter meets the design requirements.

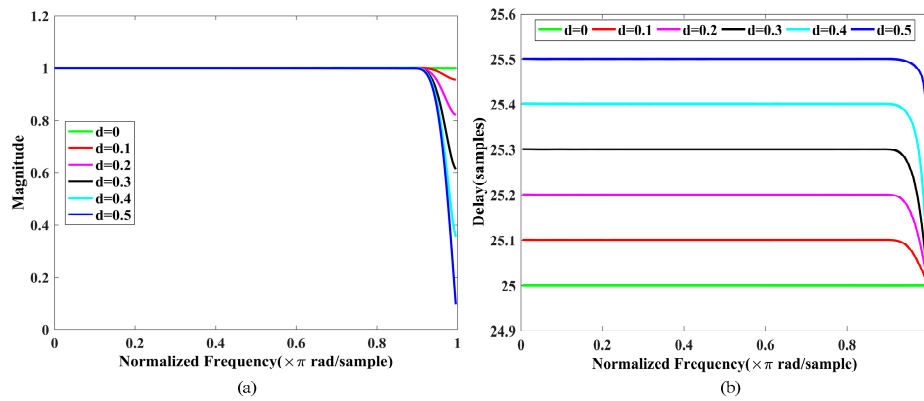


Figure 9. Frequency response of the FD filter: (a) magnitude-frequency response and (b) phase-frequency response.

On the other hand, the Delay module is used to compensate for the time delay introduced by the FD filter. The delay is equal to the length of the FD filter, which is an integer, i.e., 50 in the experiment. The I and Q signals after LPI compensation are denoted as $x_I^d[n]$ and $x_Q^d[n]$.

Derived as above, the method presented in this paper is independent of the amplitude mismatch. Methods for correcting the amplitude mismatch are already known in the art, where the Frequency Sampling Method (FSM) is an efficient technique [29,30]. FSM estimates the amplitude-frequency response of the filter based on its sampled complex frequency response and the corresponding Inverse Discrete Fourier transform (IDFT).

For the simple presentation of the LPI analysis, we apply FSM to design the compensation filter and compensate for the amplitude-frequency response mismatch between the I/Q channel, which is not detailed because it is not the focus of the paper. The design details can be found in [30]. Therefore, it should be noted that the experiment on the platform is based on the premise of no amplitude mismatch.

A three-tone signal is set as the input to the receiver, whose frequencies are 1.06 GHz, 1.1 GHz, and 1.15 GHz, respectively. The frequency of LO is set to 1 GHz; then, the input signal is converted to the quadrature demodulator LTC5586, and the frequencies of the output signal are 60 MHz, 100 MHz, and 150 MHz. The complex frequency spectrum of the demodulated three-tone signal is shown in Figure 10.

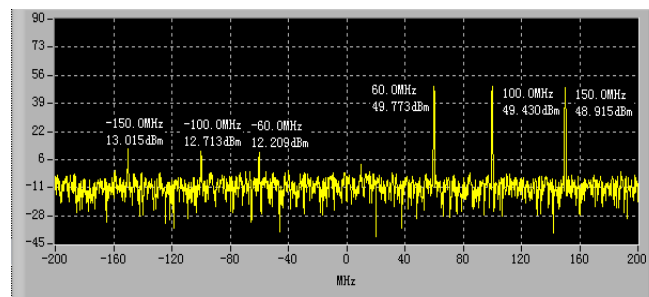


Figure 10. Frequency spectrum of the three-tone signal before LPI compensation.

After lowpass filtering and sampling, the estimation and compensation procedures are performed in FPGA I as shown in Figures 7 and 8. The frequency spectrums after LPI compensation are shown in Figure 11. It is evident that the image frequency due to LPI has been degraded to the noise floor.

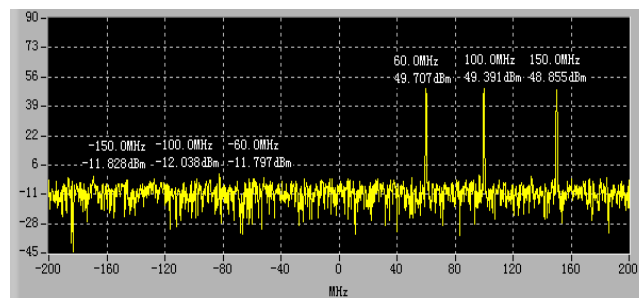


Figure 11. Frequency spectrum of the three-tone signal after LPI compensation.

The measured IRRs of the receiver system before and after LPI compensation are shown in Figure 12. As shown in Figure 12, the instantaneous bandwidth of $[-200\text{ MHz}, 200\text{ MHz}]$, and the IRR values in most of the bandwidth are well below 40 dB before LPI compensation, whereas the IRR values after LPI compensation have been enhanced significantly to the value range of greater than 60 dB, which indicates the improvement of the image frequency suppression and verifies the accuracy of the algorithm.

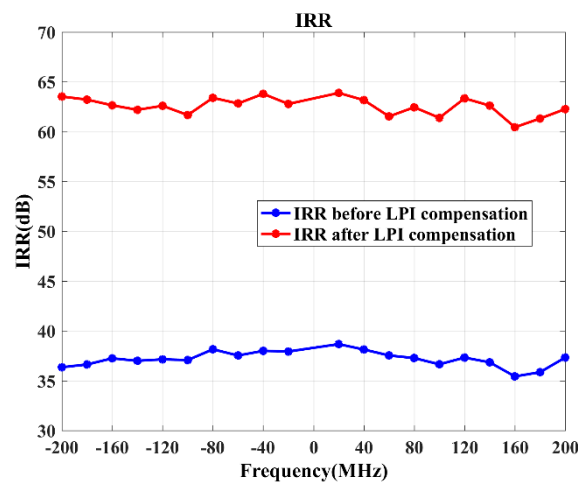


Figure 12. IRR curves before and after LPI compensation.

6. Conclusions

This paper proposes a novel estimation method for I/Q phase imbalance in the wideband zero-IF receiver. The LPI considered in this paper accounts for the frequency-dependent TDD part and frequency-independent LO phase offset part. The paper shows that LPI seriously deteriorates IRR of the system as a result of incomplete suppression of the image frequency, which needs to be solved. The method presented in this paper applies cross-power spectrum and “Three-point unwrapping” method to estimate LPI parameters simultaneously. The simulation results show the large improvement of IRR after compensating LPI obtained from the proposed method. Furthermore, the experiment results from the platform reveal that the proposed method has decreased the image frequency to the noise floor and achieves an IRR of more than 60 dB for the Nyquist bandwidth eventually, with the help of the compensation structure. Therefore, it can be concluded that the proposed approach is an efficient solution to address the LPI of I/Q signal. In addition, the implementation only requires three test tones without any additional hardware, and it is a significant feature to guarantee that the proposed method outperforms the conventional methods concerning the more general application, especially wideband communication systems where LPI compensation is needed urgently. For all these reasons, the algorithms presented in this paper facilitate the implementation of the wideband zero-IF receiver.

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Nomenclature

$r(t)$	received RF analog signal;
ω_c	carrier frequency;
φ	LO phase offset;
δ_I	time delays of the I branch;
δ_Q	time delays of the Q branch;
φ	LO phase offset;
$\Delta\delta$	TDD between I/Q channel;
$x_I[n]$	digitalized I signal;
$x_Q[n]$	digitalized Q signal;
$A_I(\omega)$	amplitude of the I channel;
$A_Q(\omega)$	amplitude of the Q channel;
$A(\omega)$	amplitude mismatch of the I/Q channel;
$X(\omega)$	complex signal model of received;
$S(\omega)$	frequency equivalent of the ideal baseband complex signal;
$S^*(-\omega)$	frequency equivalent of the image frequency;
$w_I[n]$	noise of I channel;
$w_Q[n]$	noise of Q channel;
$x'_Q[n]$	Hilbert transform of the Q signal $x_Q[n]$;
$X_I[k]$	Discrete Fourier transform of $x_I[n]$;
$X'_Q[k]$	Discrete Fourier transform of $x'_Q[n]$;
$X_Q^*[k]$	conjugate of $X'_Q[k]$;
$G_{x_I x_I}[k]$	auto-power spectrum of $X_I[k]$;
$G_{w_I w_Q}[k]$	cross-power spectrum between noise;
$G_{x_I x_Q}[k]$	cross-power spectrum between $X_I[k]$ and $X'_Q[k]$;
$\Delta\theta[k]$	LPI factor between $X_I[k]$ and $X'_Q[k]$;
m_i	unwrapping factor of the i th frequency point;
G	number of frequency components contained in the input signal;
$h_{FD}[n]$	impulse response of the FD filter;
R	the order of the FD filter;
L	the order of the sub-filters of the FD filter;
T_s	sampling period of the system;
f_s	sampling rate of the system.

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