Frequency-selective IQ Imbalance in zero-second-IF Transceivers for Wide-Band mmW Links

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Abstract—**This article deals with the analysis and mitigation of one of the most prominent impairments found in wideband zero-second-IF transceivers, namely frequency-selective IQ imbalance. IQ imbalance can significantly reduce the performance of a communication system if it is not appropriately compensated. This article presents a mathematical analysis, followed by a compensation technique. Compensation processing, known as predistortion, takes place at the digital baseband and it is based on test tones and spectral measurements. The compensation strategy facilitates automatic IQ imbalance compensation procedure and it requires a minimum of extra hardware. The performance of the algorithm is evaluated in a transceiver with 64-QAM modulation and 1GHz of bandwidth implemented with real hardware. The results presented in the article suggest that the use of this technique is able to mitigate the impact of the IQ imbalance effects in wideband systems.**

Index Terms—**Digital predistortion, frequency-selective IQ imbalance, transceiver, RF impairments, wideband.**

I. Introduction

The growing demand for ubiquitous broadband communication, e.g. fourth-generation (4G) wireless, has motivated the deployment of ultra high-speed communication systems. Particularly in backhauling networks, optical fiber is required to transport very high data rates. However, optical fiber exhibits important drawbacks, such as high costs, long deployment times, and low flexibility. Recently, point-to-point wireless communication systems have been proposed as an attractive alternative to optical fiber. In order to achieve data rates that are comparable to optical fiber, these communication systems demand very high bandwidth in order to transport enough data. Although the frequency spectrum is congested, the regulation of the E-band(71-76GHz, 81- 86GHz), facilitates the deployment of high-speed communication systems in which a huge amount of data can be transmitted. The European Telecommunications Standards Institute (ETSI) is carrying out a standardization process for this frequency band [1], [2].

Commercial off-the-shelf communication systems operating in the E-band support data rates of up to 2.5 Gbit/s. However, new applications demand even higher data rates, which necessitates both wide-band and high-order modulations in order to utilize the spectrum efficiently. With a signal bandwidth of 1GHz and 64-QAM modulation, the current available capacity of the E-Band could be doubled, achieving data rates of around 5 Gbits/s. The demand of high data rates and flexibility, together with the constraints on energy efficiency, cost and size, pose difficult challenges to the design and implementation of transceivers. Given these constraints the use of low-cost electronics are necessary. However, this implies physical limitations and variations of the electronic circuits, which lead to nonlinearities and impairments in the analog front-end. These impairments can play a critical role in the transceiver, since its performance can be significantly degraded [3].

Instead of tightening the specifications and tolerances of the analog circuitry, a cost-effective solution that is gaining momentum is to compensate the analog front-end impairments in the digital domain using signal processing algorithms [4].

In this article we focus on IQ imbalance impairment, which is one of the performance-critical effects of interest in zero-second-IF transceiver architectures. This impairment, caused by mismatches in the amplitude and phase responses of the I and Q signal paths, entails a degradation in the Image Rejection Ratio (IRR), which is theoretically infinite and causes interfering images at mirror frequencies. IQ imbalance encountered in narrow-band systems can be regarded as non-frequency-selective (NFS), and it is mainly caused by the local oscillators (LO) used for quadrature modulation or demodulation. However, IQ imbalance in wide-band systems may also exhibit frequency-dependent or frequency-selective (FS) behavior due to mismatches between the analog filtering paths of the I and Q components caused by finite tolerances [5], [6].

In order to digitally compensate the IQ imbalance of the modulator, the digital baseband needs to have some reference of the mismatch encountered in the analog front-end. That is why most of the contributions found in the literature addressing the IQ modulator imbalance include a feedback path with an additional ADC, so as to sample the output of the transmitter [7], [8]. It has to be taken into account that the systems presented in these articles work with relatively narrow signal bandwidths relative to the 1GHz used in the system in this article. A signal bandwidth of this order would lead to a higher sampling rate of the ADC, significantly increasing the complexity of the system.

In order to avoid including an additional ADC, a different approach using test tones is proposed by [9], which employs test tones and spectral measurements. A similar approach is used in this article in order to validate the IQ imbalance compensation through measurements in a practical hardware implementation. Significantly increasing the IRR of a transceiver over the whole signal bandwidth of 1GHz goes beyond the current state of the art.

The remainder of this article is structured as follows. Section II provides a description of the system architecture. In section III the mathematical models for the quadrature imbalances of the modulator will be used to develop a compensation technique, addressing both non-frequency-selective and frequency-selective imbalance. Section IV extends the validation of the technique to a practical hardware quadrature modulator implementation. Finally, some conclusions are drawn in Section V.

II. System Architecture

In order to address new applications for the future backhauling networks, a point-to-point microwave link in the E-Band using a 64-QAM modulation with a signal bandwidth of 1GHz is considered. Figure 1 shows the proposed transceiver (TRx) architecture for a point-to-point microwave link in the E-Band. As shown, the transmitter (Tx) front-end consists of an IQ modulator that up-converts the baseband I and Q channels to an intermediate frequency (IF) of 17.5GHz. After combining the I and Q channels, the IF signal is up-converted to the E-Band by means of the millimeter-wave (mmW) mixer. Finally, the wideband mmW power amplifier (PA) is used to amplify and transmit the mmW signal. The receiver (Rx) front-end consists of a wideband Low Noise Amplifier (LNA), which receives and amplifies the signal at the E-Band. After the LNA, a first mixer down-converts the mmW signal to the same IF as in the Tx. This way, the same PLL can be re-used for the Tx and the Rx. Finally, an IQ demodulator down-converts the IF signal to 0-Hz.

This architecture presents a good balance between different design aspects, and it enables the minimization of the sampling frequency of the digital-to-analog (DAC) and analog-to-digital (ADC) converters. Nowadays we can find commercial DACs and ADCs able to provide sampling rates in the range of 2.5Gsps and even higher, which is enough for practical implementation of the zero-second-IF architecture [10].

Due to the high channel bandwidth, the architecture depicted in Figure 1 presents a good balance between the DAC and the ADC requirements and the complexity of the transceiver. The use of other architectures, such as non-zero-second-IF would require very high performance ADCs or DACs, as well as highly selective reconstruction filters to achieve a practical implementation of base-band and image rejection filters in the analog front-end.

However, the use of a zero-second-IF architecture presents well-known issues that should be addressed in order to avoid degrading the performance of the transceiver. This architecture is subject to DC offsets due to the self-mixing of the LO leakage, corruption of the signal close to DC due to flicker noise and AC coupling between different baseband components which lead to a high pass filtering of the signal. These impairments were analyzed in [10].

In order to avoid placing useful signal near DC, an alternative solution consists of splitting the signal to be transmitted in two subbands. One subband is IQ modulated using a signal bandwidth of 500 MHz and centered at 250 MHz, and the other subband uses another signal with 500 MHz of bandwidth and centred at 250 MHz. Both subbands have a 64-QAM data modulation and a raised-cosine pulse shaping with 15% roll-off. A symbol rate of 434.78MHz is used, yielding a total signal with a bandwidth of 1GHz and no information bearing components near DC. Figure 2 shows the spectrum of the transmitted signal when this configuration is considered. This architecture relaxes the performance of the ADCs and DACs, but it is at the expense of increasing the complexity of the baseband and IF sections of the front-end of the transceiver.

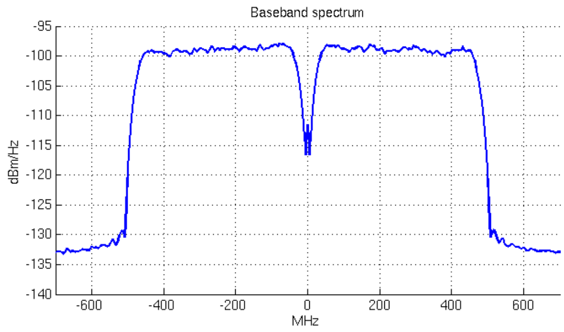

Fig. 2. Baseband spectrum of the transmitted signal

Another issue that needs to be considered in this kind of transceiver is the corruption due to IQ imbalances at both the transmitter quadrature modulator and at the receiver demodulator. The resulting system performance degradation can be significant, especially for high-order modulation schemes [11], [3].

Hence, the remainder of the article will focus on the IQ imbalance of the transmitter, identifying the possible sources of this impairment and analyzing a digital signal processing based method to compensate for it.

III. IQ Imbalance Estimation and Compensation

In this section the mathematical principles behind the frequency-selective IQ imbalance encountered in real quadrature modulators are presented. From the mathematical analysis the compensation model addressing frequency-selective IQ imbalance will be derived.

A. Imbalance model

Consider the complex baseband signal

$$s(t) = s_I(t) + js_Q(t), \tag{1}$$

which in the frequency domain corresponds to

$$S(f) = S_I(f) + jS_Q(f). \tag{2}$$

In a quadrature modulation, the real $s_I(t)$ and imaginary $s_Q(t)$ parts of the complex baseband signal are, respectively, the inputs to the I and Q signal paths of the modulator. The signal $x_{tx}(t)$ at the output of a perfect modulator is given by

$$x_{tx}(t) = s_I(t)\cos(\omega_{\text{tx}}t) - s_Q(t)\sin(\omega_{\text{tx}}t), \tag{3}$$

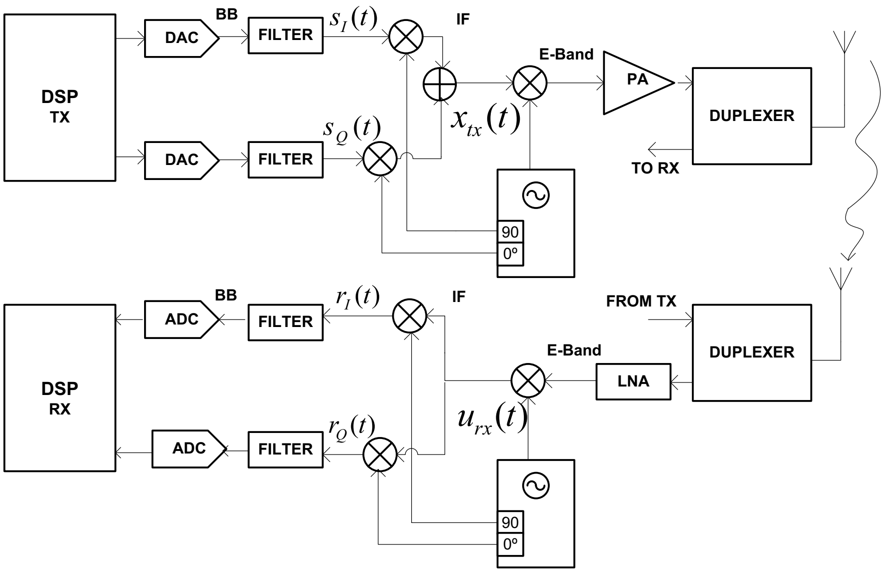

Fig. 1. Architecture of the transceiver.

which in the frequency domain should ideally correspond to a perfect single frequency translation.

In an ideal modulator, the upconverting cosine and sine oscillators have exactly the same amplitude and a phase difference of $\pi/2$. When implementing the modulator with actual electronic circuits though, they present amplitude (g_{tx}) and phase mismatches (ϕ_{tx}). Furthermore, the responses of the DACs and LPFs of both branches are not identical and the signal paths themselves may also contribute to the imbalance. Figure 3 shows the topology of a real quadrature modulator. In order to model any passband frequency response of the I and Q signal paths, bandpass filters (BPF) are defined. This approach accounts for the asymmetrical imbalance response observed during practical measurements[9].

From this point on, it is assumed that the I-datapath does not have any imbalance and $h_{tx}(t)$ represents the total relative mismatch between the I and Q paths. $h_{tx}(t)$ includes the effects of LPF, DAC, BPF and signal path imbalances[12].

The imbalanced baseband equivalent signal at the output of the transmitter is given by

$$x_{tx}(t) = C_1(t) * s(t) + C_2(t) * s^*(t) \tag{4}$$

where

$$C_1(t) = \frac{\delta(t) + h_{tx}(t) g_{\mathrm{tx}} e^{j\phi_{\mathrm{tx}}}}{2} \tag{5a}$$

$$C_2(t) = \frac{\delta(t) - h_{tx}(t) g_{\mathrm{tx}} e^{j\phi_{\mathrm{tx}}}}{2}. \tag{5b}$$

In the literature there are studies in which only the IQ imbalance produced by the LOs is considered, yielding the non-frequency-selective IQ imbalance model[13], where

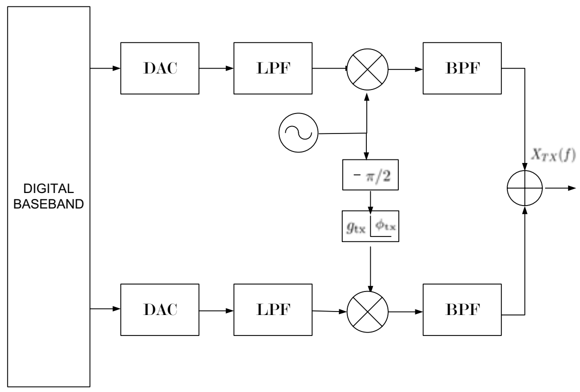

Fig. 3. Architecture of the quadrature modulator.

$$C_1 = \frac{1 + g_{\mathrm{tx}} e^{j\phi_{\mathrm{tx}}}}{2} \tag{6a}$$

$$C_2 = \frac{1 - g_{\mathrm{tx}} e^{j\phi_{\mathrm{tx}}}}{2}. \tag{6b}$$

The non-frequency-selective model, therefore, can be seen as a special case of the frequency-selective model when $h_{tx}(t) = \delta(t)$. In a wideband system context, the overall IQ imbalance can vary over the whole frequency band[14]. Consequently, the frequency-selective behavior has to be taken into account and therefore equations (5) are the ones used to model the IQ imbalance of the system presented in this article.

Equation (4) can be rewritten in the frequency domain as

$$X_{TX}(f) = C_1(f)S(f) + C_2(f)S^*(-f). \tag{7}$$

In (7) only the term relating to $S(f)$ is desired, whereas the term related to $S^*(-f)$ is considered an undesired signal.

The quality of the Tx can be quantified in terms of the Image Rejection Ratio, which measures the relative power between the desired signal and the image.

$$\mathrm{IRR_{tx}} = \frac{|C_1(t)|^2}{|C_2(t)|^2} \tag{8}$$

When the cascaded gain and phase responses of the I and Q channels are equal, i.e. $H_{TX}(f) = 1$, $g_{tx} = 0$ and $\phi_{tx} = \pm\pi/2$, the IRR is infinite. When this is not the case, the channels are imbalanced and the image is not completely rejected.

B. IQ imbalance compensation

The mathematical model in III-A is used to develop the FS IQ imbalance compensator. The compensation processing, referred as a predistortion, takes place at the digital baseband, in the digital front-end of the transmitter and it is the last step before digital-to-analog conversion.

The aim of the IQ predistorter is to remove the term related to $S^*(-f)$ in equation (7) as much as possible, which leads to an increase in the IRR factor.

Equation (4) represents a widely-linear transformation of the baseband input signal of the IQ modulator. Therefore, a natural form for the compensator is another widely-linear transformation[15]. Assuming FIR filters, the precompensated signal $s_{pre}(t)$ is given by

$$s_{pre}(t) = w_1(t) * s(t) + w_2(t) * s^*(t), \tag{9}$$

which in the frequency domain corresponds to

$$S_{pre}(f) = W_1(f)S(f) + W_2(f)S^*(-f). \tag{10}$$

$W_1(f)$ and $W_2(f)$ denote the frequency responses of two precompensation filters that operate on the baseband signal to be transmitted, $S(f)$, and its frequency reversed complex conjugate, $S^*(-f)$. An option is to filter only the complex signal, arbitrarily choosing $W_1(f) = 1$ and yielding

$$s_{pre}(t) = s(t) + w_2(t) * s^*(t), \tag{11}$$

and in the frequency domain

$$S_{pre}(f) = S(f) + W_2(f)S^*(-f). \tag{12}$$

This compensation structure is given in Figure 4.

Now the signal $s_{pre}(t)$ is the input of the imbalanced quadrature modulator. The frequency domain representation of the output of the modulator is now given by

$$X_{TXpre}(f) = C_1(f)S_{pre}(f) + C_2(f)S^*_{pre}(-f). \tag{13}$$

By substituting the expression of $S_{pre}(f)$ in the above equation, the output of the precompensated, imbalanced quadrature modulator is now written as

$$X_{TXpre}(f) = S(f)[C_1(f) + W_2^*(-f)C_2(f)] + \quad \text{(14a)}$$
$$S^*(-f)[C_2(f) + W_2C_1(f)]. \quad \text{(14b)}$$

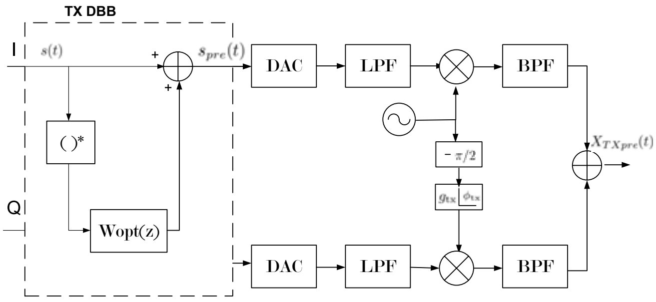

Fig. 4. IQ imbalance precompensation model.

In order to eliminate the undesired image component in (14), the frequency response of $W_2(f)$ should be given by

$$W_2(f) = \frac{-C_2(f)}{C_1(f)}. \tag{15}$$

Equation (15) suggests that the imbalance extraction technique needs to estimate only the ratio between the desired and the image signal. Therefore, it is shown that the spectral analysis of the signal, after it has passed through the imperfect modulator, will render enough information to obtain the frequency response of the compensation filter[12].

In order to measure the ratio in (15), a tone is passed through the imbalanced modulator and the spectral measurements of the desire and its mirror signals are taken. In order to take these measurements, an option proposed by [9] is to use a previously balanced demodulator as the measurement device. For the compensation technique of this article, however, a Power Spectrum Analyzer (PSA) placed at the output of the transmitter has been used to measure the ratio.

Using a single tone as a test signal will only provide the information regarding the imbalance ratio at that frequency. Therefore, we propose using a collection of tones in order to characterize the IQ imbalance of the whole band of interest.

As a final stage of the process, the frequency response of the compensation filter has to be translated to filter coefficients to be implemented in the digital domain of the modulator. The compensator filter can have real coefficients when the gain imbalance is an odd function and the phase imbalance is even. Whenever this condition is not fulfilled, a complex filter is required[12].

IV. Performance evaluation

This section evaluates the effectiveness of the proposed imbalance extraction and compensation technique.

A. Measurement setup

Figure 5 shows the setup used to implement and test the imbalance extraction and compensation developed in this article.

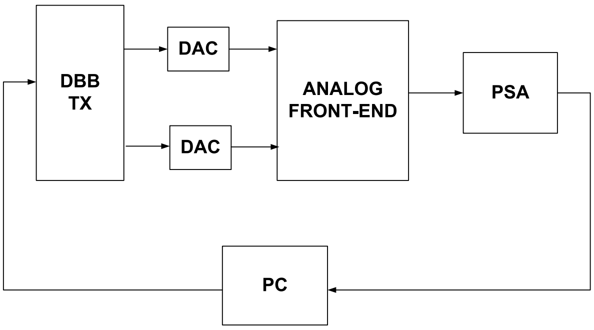

Fig. 5. Simulation setup

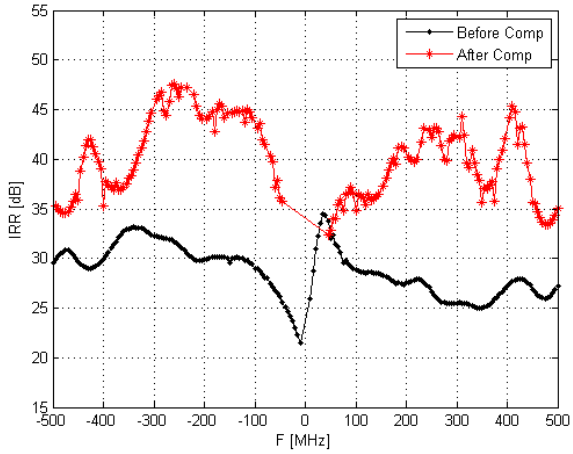

Fig. 6. IRR before and after IQ imbalance compensation.

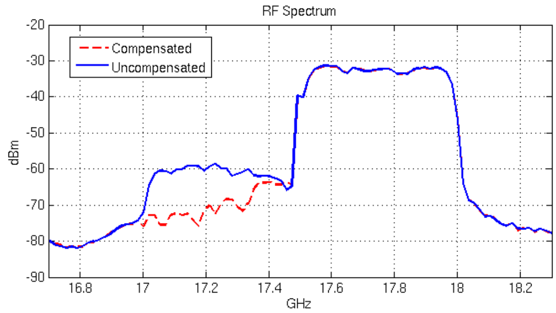

Fig. 7. RF spectrum of the negative subband before and after compensation.

The digital baseband processing consists of a Virtex-7 FPGA prototyping board (VC707). The FPGA is programmed in order to generate data bits and transform them into 64-QAM IQ symbols. These symbols are pulse shaped by a RRC filter and a mixer modulates each digital subband to subcarriers at 250MHz and -250MHz. The final stage of digital baseband processing is the IQ imbalance compensation filter.

At the output of the FPGA, a couple of high speed DACs are responsible for converting the complex baseband signal to the analog domain. The DAC board used in the prototype consists of an FMC230 board. The DACs work at a sampling rate of 2.4GHz.

At the analog front-end the signal is first passed through the analog reconstruction filters to remove the signal replicas at multiples of the digital sampling rate. Finally, the IQ modulator translates the baseband signal to 17.5GHz.

At the output of the analog front-end a PSA measures the relative power between the desire and the image components of the test signal.

Software run on a PC is used to perform the translation from the extracted measurement from the PSA to the actual coefficients of the filter. Moreover, the PC produces the appropriate control and configuration signals and sends them to the FPGA through an Ethernet connection.

B. Results

1) IRR: Figure 6 shows the obtained IRR curves as a function of the frequency. The curve labeled *'Before Comp'* represents the IRR of the analog front-end without any compensation. It is shown that the transmitter analog front-end model has a frequency-dependent IQ imbalance, with the resulting uncompensated IRR varying between 23 and 34dBs. The curve labeled *'After Comp'* shows the IRR after the compensation has been applied. The resulting IRR after compensation was tested by retransmitting the same tones but this time pre-filtered by the compensation filter. The first important observation is that the compensation technique is able to attenuate the image signal between 35 and 47dBs.

Similar conclusions can be taken from Figures 7 and 8, which depict the obtained transmitter spectrum at IF. As explained in Section II, two subbands with 500MHz of bandwidth, are transmitted, at -250MHz and 250MHz, which in IF correspond to 17.25GHz and 17.75GHz, respectively. The mirror frequency component of the signal at 17.75GHz is located in 17.25GHz, i.e. exactly where the other subband is placed. Similarly, the image of the signal at 17.25 is hidden by the other subband. Therefore, we are unable to see the mirror components when both subbands are transmitted.

In order to evaluate the impact of the IQ imbalance compensation, one subband is transmitted each time, allowing the corresponding image component to be seen. Figure 7 shows the subband at 17.75GHz together with its mirror frequency before and after compensation has been applied. It is shown that the image component is well attenuated when the compensation filter is active, reaching an IRR of almost 40dB and leaving the mirror component a at almost the level of the noise floor. The same conclusions can be obtained for Figure 8, which represents the equivalent case when the subband at 17.25GHz is transmitted.

2) Stability: Figure 9 and 10 show the gain and phase imbalance functions over the frequency band of interest. The plots include 6 measurements in each figure, taken in 6 different days, in which measurement conditions, such as temperature, where different. It is shown that, despite this,

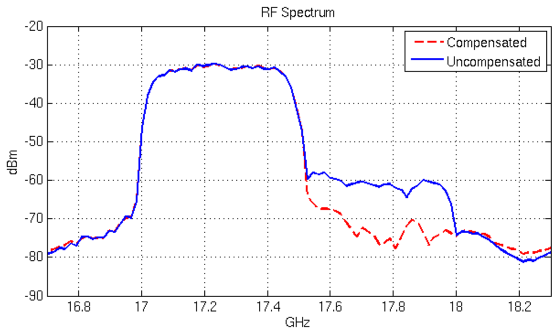

Fig. 8. RF spectrum of the positive subband before and after compensation..

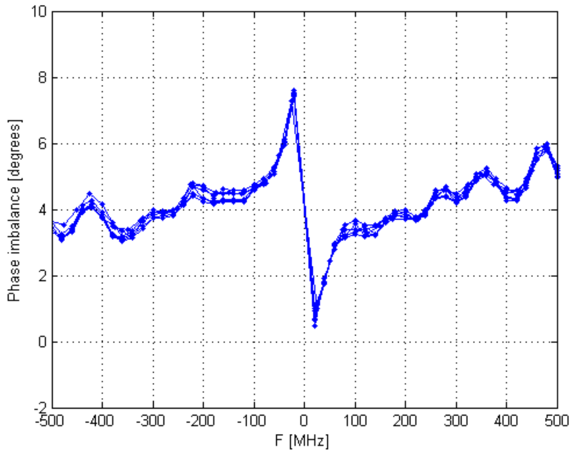

Fig. 9. Phase Imbalance.

the curves are almost identical. The similarity of the curves proves that the gain and phase imbalance of the system vary very slowly over time. This suggests that the compensation procedure can be performed as a rapid calibration routine that can be done from time to time, without the need to calibrate the system each time it has to switch to its normal operation.

Figures 9 and 10 are also used to evaluate the symmetry of the gain and phase imbalance functions. They show that the gain imbalance does not match an even function, nor does the phase imbalance match an odd one. This confirms the use of BPF mismatches in the imbalance model, as well as employing complex coefficients in the compensation filter[12].

V. Conclusion

This article deals with the analysis and mitigation of one of the most prominent impairments found in zero-second-IF transceivers. A theoretical analysis of the frequency-selective IQ imbalance has been presented, together with a compensation technique based on test signals. The compensation processing, referred to as a predistortion, takes place at the digital baseband, in the digital front-end of the transmitter. The proposed technique provides a practical and effective solution to the real-life circuit implementation problems of modern communication transceivers, one that can be performed from time to time, since it has been shown that quadrature imbalances tend to vary slowly over time.

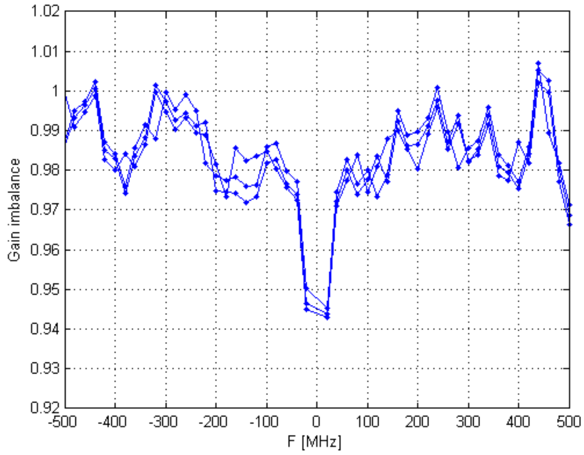

Fig. 10. Gain Imbalance.

The functionality and performance of the algorithm were verified with laboratory radio signal measurements, in a zero-second-IF transceiver using 64-QAM with a signal bandwidth of 1GHz. It was demonstrated that the system exhibits good performance under variable system conditions. The results show that the proposed technique provides efficient and robust digital signal processing based solutions for the analog implementation impairments. Moreover, the presented configuration requires that minimal extra hardware be added to the front-end. The proposed technique, therefore, facilitates the implementation of more flexible, low-cost radio devices for wideband wireless systems.

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