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MEASURING DEMODULATOR IMBALANCE IN RADIO FREQUENCY RECEIVERS BY TONE TEST

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Abstract – The measurement of in-phase/quadrature (IQ) imbalance parameters of radio frequency direct downconversion receivers by tone test is considered. The receiver is excited by a radio frequency sinewave, and the digital baseband output is used as the sole basis for the determination of the IQ imbalance parameters, which are the gain imbalance, quadrature skew and local oscillator leakage. A set of parameters is estimated from the baseband output of the direct downconversion receiver and the required parameters are then calculated by a transformation. The universal software radio peripheral (USRP) is a widespread hardware for research, education, and development regarding future wireless communication receivers. A set of USRPs are tested. It is shown that for the receivers studied, gain imbalance and quadrature skew may be predicted accurately (that is, < 0.1 dB and < 1 deg, respectively) by employing baseband data covering a handful of full periods of the excitation stimuli. The work also shows that local oscillator leakage may suffer from systematic bias error of the order of 15 dB. To obtain leakage estimates with an uncertainty in the order of one dB, the measurement time has to be increased by two orders of magnitude.

Keywords: radio receiver IQ imbalance, tone test, universal software radio peripheral (USRP).

1. INTRODUCTION

The development of contemporary and future wireless MIMO communication systems puts high demands on accurate and time efficient test methods for production and product validation. Reducing the time for a test is essential and the accuracy of the applied method with respect to the length of the recorded data sequence is an important figure of merit to trade test time versus accuracy and precision of the test. Efficient measurement methods are also required as prerequisite for digital correction to combat impairments produced by the analog circuitry – in line with the dirty radio frequency (RF) paradigm [1]. Signal analysis and advanced methods for IQ imbalance compensation are given in [2, 3]. Rapid prototyping of new digital communication technologies is essential, and often relies on software defined radio implementations. An overview of signal processing challenges applying software radio principles is given in [4]. A widespread hardware platform is the available variants of the universal software radio peripheral (USRP) by Ettus Research LLC, which is becoming common technology for research, education and development [5]; see also [6–8].

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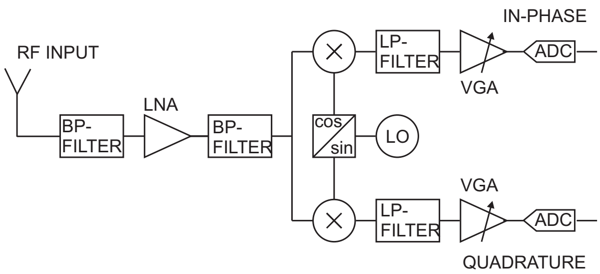

Fig. 1. Direct conversion radio receiver.

A direct conversion radio receiver is outlined in Fig. 1. An estimation method for receiver in-phase/quadrature (IQ) imbalance parameters is presented based on the IQ imbalance parameters defined in [9]. The test method is applied to IQ imbalance measurements on the USRP equipped with a FLEXRF1800 receiver, configured as a direct downconversion receiver. The imbalance in the IQ demodulator produces mirror distortion and the leakage of the local oscillator into the receiving path produces another source of distortion. The power spectral density of the received signal samples from the USRP is given in Fig. 2. The test of several units illustrates the spread in performance between different receivers, and validates the specifications given by the data sheets.

2. IQ IMBALANCE

Consider a measurement set-up where a RF receiver is directly excited by a radio frequency sine wave at F_{RF} [Hz]. The receiver outputs a time-domain digital baseband signal z_n, where n denotes the sample instants. The absolute time scale is given by n/F_{s} [s], where F_{s} [Hz] is the sampling rate of the involved analog-digital converters. The baseband output z_n is separated into its in-phase (I) and quadrature (Q) parts, that is $z_n = x_n + i y_n$ where $i = \sqrt{-1}$ is the imaginary unit. In this work, the receiver output $\{z_n\}$ is the basis for characterization of the receiver imbalance between the I and Q branches, relaxing the condition on exact knowledge of F_{RF}, F_{s} and the frequency of the local oscillator (LO) F_{LO}.

For a *perfect* IQ demodulator, the in-phase and quadrature

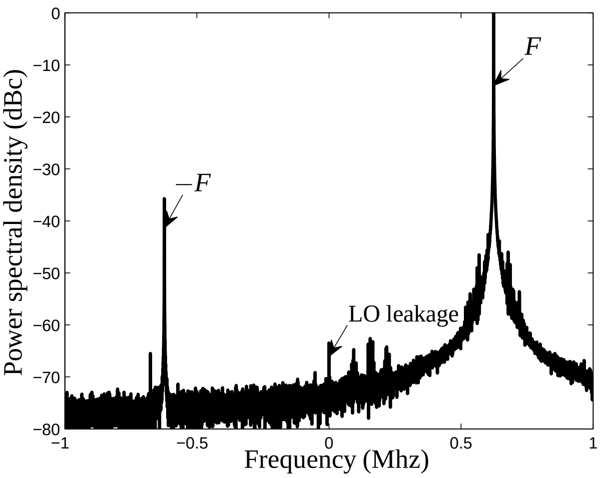

Fig. 2. Exemplary power spectral density of a USRP receiver output in the range ± 1 MHz, showing the downconverted carrier at $F = 0.6$ MHz, mirror distortion, and LO leakage.

parts are given by

$$x_n = g \sin\left(\omega_o n + \phi + \frac{\pi}{2}\right) \tag{1}$$

and

$$y_n = g \sin(\omega_o n + \phi) \tag{2}$$

for some common gain g and initial phase ϕ. The angular frequency is $\omega_o = 2\pi F/F_{\text{S}}$ with $F = F_{\text{RF}} - F_{\text{LO}}$ being the absolute frequency in Hertz of the downconverted sinewave entering the analog-digital converters (ADCs). The exact frequency values of F_{S}, F_{RF} and F_{LO} are not required for the proposed test method, which includes a method to estimate the normalized angular frequency ω_o directly from the receiver output z_n.

A *practical* receiver is never ideal, and for a RF sinewave excitation the in-phase and quadrature branches of z_n are described by

$$x_n = g_I \sin\left(\omega_o n + \phi_I + \frac{\pi}{2}\right) + c_I + v_n^I \tag{3}$$

and

$$y_n = g_Q \sin(\omega_o n + \phi_Q) + c_Q + v_n^Q \tag{4}$$

In (3)-(4), v_n^I and v_n^Q are noise terms describing thermal and quantization noise, model imperfections, and so on. The noise terms are assumed jointly uncorrelated zero mean wide sense stationary processes with equal variance σ^2. Further, g_I, g_Q are gains, ϕ_I, ϕ_Q initial phases and c_I, c_Q the DC offsets. These six parameters determine the characteristics of the IQ imbalance. They are all unknown and have to be estimated from the receiver output $\{z_n\}$.

In [9], *gain imbalance G*, *quadrature skew Q*, and *LO leakage L* are defined. The gain imbalance G is the quotient

$$G = \frac{g_I}{g_Q} \tag{5}$$

or $G_{\text{dB}} = 20\log_{10} G$. The quadrature skew is the phase difference relative $\pi/2$, that is,

$$Q = \phi_I - \phi_Q \tag{6}$$

where Q is typically displayed in degrees $Q_o = 180 \cdot Q/\pi$. Finally, L is the quotient between the total power of the DC offset to the power of the sinewaves, that is,

$$L = \frac{2}{P_s} \frac{c_I^2 + c_Q^2}{g_I^2 + g_Q^2} \tag{7}$$

where P_s is the average power of the equivalent baseband input signal. In dBs, $L_{\text{dB}} = 10\log_{10} L$. For an ideal IQ demodulator, $G_{\text{dB}} = 0$ dB, $Q_o = 0$ degrees, and $L_{\text{dB}} = -\infty$ dB, respectively.

3. IQ IMBALANCE PARAMETERS BY LEAST-SQUARES

The estimation of the required G, Q, and L from a set of recorded baseband data $\{z_n\}$ is performed in three steps. The first two steps correspond to the solution to a separable least-squares problem. First, the angular frequency of the baseband data is estimated. Second, the parameters in a vector η is found. Then, estimates of the required G, Q, and L are obtained by a proper transformation.

Consider the least-squares criterion

$$V(\eta, \omega) = \sum_{n=1}^{N} |z_n - s_n(\eta, \omega)|^2 \tag{8}$$

where N denotes the number of complex-valued baseband samples, and $s_n(\eta, \omega)$ is a model of the receiver response as a function of the parameter values. Here the parameters are gathered in a parameter vector η and a scalar parameter ω, where the latter parameter is the angular frequency of the baseband representation of the excitation signal. The required estimated parameter values $\widehat{\eta}$ and $\widehat{\omega}$ are given by the minimizing argument

$$\widehat{\eta}, \widehat{\omega} = \arg\min_{\eta,\omega} V(\eta, \omega) \tag{9}$$

Consider,

$$s_n(\eta, \omega) = a\, e^{i\omega n} + b^* e^{-i\omega n} + c \tag{10}$$

In (10), a, b (or rather b^*, where $*$ denotes conjugate), and c are unknown complex-valued constants gathered in the parameter vector

$$\eta = \begin{pmatrix} a \\ b^* \\ c \end{pmatrix} \tag{11}$$

Using a, b, and c as parameters, the gain imbalance G in (5) and quadrature skew Q in (6) can alternatively be written as:

$$G = \frac{|a+b|}{|a-b|} \tag{12}$$

$$Q = \angle\left[\frac{a+b}{a-b}\right] \tag{13}$$

where $\angle[\cdot]$ denotes the phase angle of the complex-valued quantity within the brackets. Further, with $c = c_I + i\,c_Q$ the LO leakage in (7) can be written as:

$$L = \frac{|c|^2}{|a|^2 + |b|^2} \tag{14}$$

In summary, the IQ imbalance problem has been parameterized in the complex-valued parameters a, b, and c, gathered

Table 1. Data sheet values for AD8347 at 1.905 GHz.

	Typ (Min/Max)	
G_{dB}	$+0.3$	dB
Q_o	± 1 (± 3)	degree
L_{dB}	-60	dBm (at RFIP)
L_{dB}	-42	dBm (At IMXO/QMXO)

in the parameter vector η as in (11). The signal model that will be fitted to baseband data $\{z_1,\ldots,z_n\}$ is given by (10).

Consider N complex-valued baseband samples of the receiver output and gather them in a column vector $\mathbf{z}$, that is

$$\mathbf{z} = \begin{pmatrix} z_1 \\ \vdots \\ z_N \end{pmatrix} \tag{15}$$

We seek the parameter values of η and ω (denoted by $\widehat{\eta}$ and $\widehat{\omega}$, respectively) that minimize the least-squares criterion (8), where $s_n(\eta,\,\omega)$ is the model output given by (10). Let $N \times 3$ matrix $\mathbf{C}(\omega)$

$$\mathbf{C}(\omega) = \begin{pmatrix} e^{i\omega} & e^{-i\omega} & 1 \\ \vdots & \vdots & \vdots \\ e^{i\omega N} & e^{-i\omega N} & 1 \end{pmatrix} \tag{16}$$

The least-squares problem can be solved in two steps, where in the first step the parameter ω that enters the signal model $\mathbf{C}(\omega)\,\eta$ in a nonlinear fashion through $\mathbf{C}(\omega)$ is found by a nonlinear, but one-dimensional search for the maximum of the condensed loss function [10]

$$V(\omega) = \mathbf{z}^H\,\mathbf{C}(\omega)\,(\mathbf{C}^H(\omega)\,\mathbf{C}(\omega))^{-1}\mathbf{C}^H(\omega)\,\mathbf{z} \tag{17}$$

Once the maximizer $\widehat{\omega}$ to (17) is found, the estimate of the linear parameters follows as the least-squares solution to $\mathbf{C}(\widehat{\omega})\,\eta = \mathbf{z}$, viz.

$$\widehat{\eta} = \left(\mathbf{C}^H(\widehat{\omega})\,\mathbf{C}(\widehat{\omega})\right)^{-1}\,\mathbf{C}^H(\widehat{\omega})\,\mathbf{z} \tag{18}$$

Transform $\widehat{\eta}$ to obtain $\widehat{G}$, $\widehat{Q}$ and $\widehat{L}$ from (12), (13), and (14), respectively.

In a related context, maximization of (17) around an initial value with the aid of the scalar bounded nonlinear function minimization routine *fminbnd* provided by Matlab was considered in [11] – an approach that is employed in this work as well. Another implementation is provided by the so called, seven-parameter fit in [12] (i.e., one may note that η and ω contain seven real-valued parameters, in total). One may note that similar estimation problems arise for other engineering applications, such as impedance measurements [13] and the estimation of particle size and velocity in laser anemometry [14, 15]. The seven-parameter fit and the method in [14] are both nonlinear least-squares types of methods requiring some interpolation or iterations to provide the required estimates.

4. EXPERIMENTAL EVALUATION

The FLEXRF family of daughter-boards that are widely used with the USRP platform is designed around the AD8347 direct conversion quadrature demodulator. Its specification regarding the imbalance is given in Tab. 1. The USRP receivers were excited by a $F_{\mathrm{RF}} = 882.3$ Mhz sinewave at 12 dBm by a HP8656B, followed by a Mini-Circuits FK3000 frequency doubler, Mini-Circuits SHP-900 high-pass filter, Mini-Circuits attenuators (3×10 dB and 2×5 dB), and a ZAPD-30 splitter for simultaneous excitation of the two inputs of the USRP. Measured input signal level is -60 dBm. No internal downconversion was employed in the USRP. For each USRP receiver, 100 k of baseband data was collected. The sampling rate of the USRP ADCs is $F_{\mathrm{S}} = 4$ MHz, followed by a digital filtering and subsampling. With the $F_{\mathrm{LO}} = 1764$ MHz, we obtain normalized angular frequency $\omega_o = 0.15$.

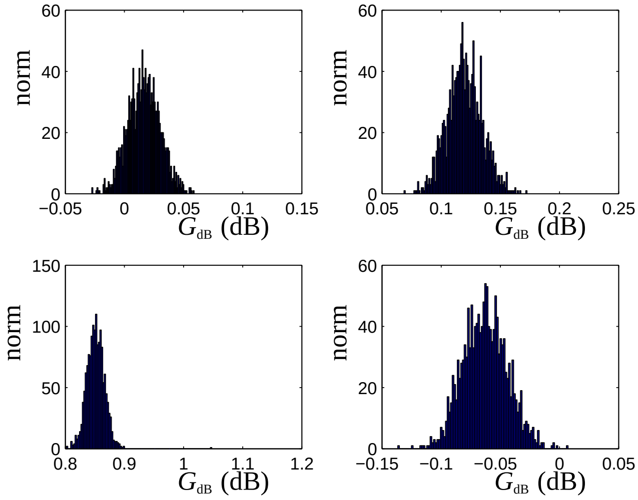

Fig. 3. USRP results based on 1500 nonoverlapping segments, each of length $N = 64$. Histograms of the estimated $\widehat{G}_{\mathrm{dB}}$. USRP #1: channel 1 (top left), channel 2 (top right), USRP #2: channel 1 (down left), channel2 (down right).

In the experiment, the baseband output from a USRP receiver was divided into nonoverlapping blocks of length $N = 64$ samples. The imbalance test was employed. The results based on the statistics employing 1500 nonoverlapping segments of data are shown in Figs. 3–5. We note that three out of the four receivers have performance according to the data-sheet of the AD8347 direct conversion quadrature demodulator, whereas the fourth unit has outlier performance. The estimated values for gain imbalance and quadrature skew are accurate, that is, within a 0.1 dB and and 1 deg, respectively. The estimate of the LO leakage on the other hand has a broad spread, and does not provide reliable estimates of the leakage. To reduce the spread, the number of samples has to be increased quite significantly. Other experiments, not included herein, indicate that the measurement record has to be increased by two orders of magnitude to provide estimates within 1 dB. It is also indicated that for short data records there is a significant bias or systematic error. For the receivers considered the bias is in order of 15 dB.

5. CONCLUSIONS

We have proposed a least-squares approach to determine the IQ imbalance of a direct conversion receiver. By a reparametrization of the problem, a least-squares problem with six real-valued parameters that enters the problem in a linear fashion is obtained, with only the angular frequency of the carrier entering the problem in a nonlinear fashion.

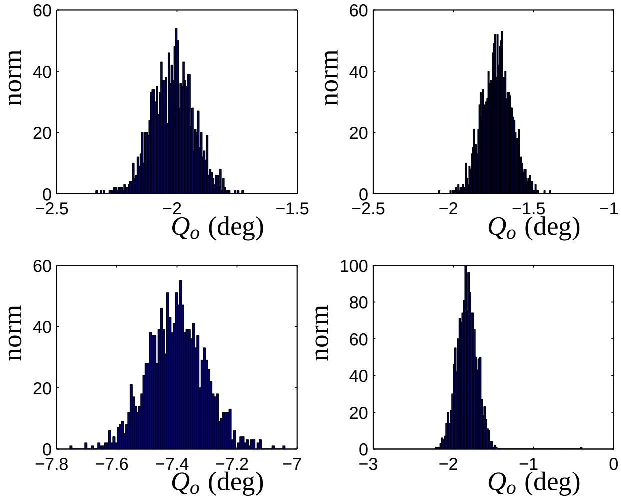

Fig. 4. USRP results based on 1500 nonoverlapping segments, each of length $N = 64$. Histograms of the estimated $\widehat{Q}_o$. USRP #1: channel 1 (top left), channel 2 (top right), USRP #2: channel 1 (down left), channel2 (down right).

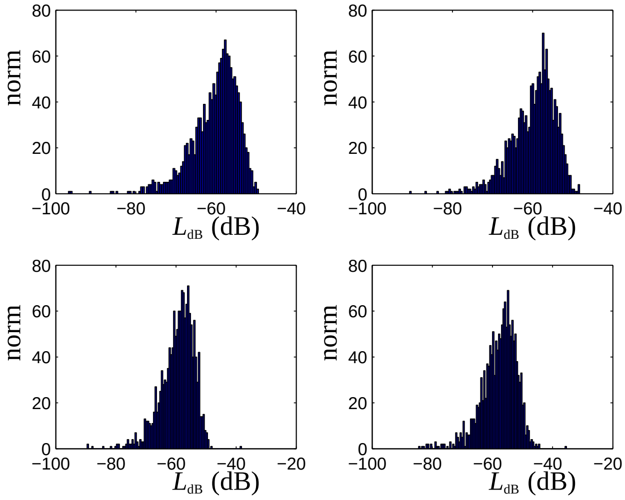

Fig. 5. USRP results based on 1500 nonoverlapping segments, each of length $N = 64$. Histograms of the estimated $\widehat{L}_{\text{dB}}$. USRP #1: channel 1 (top left), channel 2 (top right), USRP #2: channel 1 (down left), channel2 (down right). To compare the obtained LO-leakage values with the datasheet figures provided in Tab. 1, we note that the AD8347 input power at RFIP equals $P_{\text{RFIP}} = P_{\text{RF}} + 13$ dBm, taking the gain of the FLEXRF1800-internal MGA-82563 amplifier into account. Accordingly, the leakage reads $L_{\text{dBm}} = -47 + L_{\text{dB}}$ dBm at RFIP of the AD8377, yielding leakage values below -100 dBm at RFIP because of the enabled internal automatic DC offset calibration provided by the USRP.

Experimental data from four different USRP receivers has been collected. It has been shown that gain imbalance and quadrature skew are accurately estimated by employing data only covering a handful of full periods of the carrier, which highlights the practical relevance of the derived test method. By examining samples of the USRP receiver, we have not only validated its data-sheet performance but also observed some outlier performance of a sample receiver. The evaluation of USRP receivers has an interest of its own owing to the widespread use of the USRPs for research, education, and development. A detailed study of the derived method can be found in [16].

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