UNIVERSITY OF WESTMINSTER

WestminsterResearch
http://www.wmin.ac.uk/westminsterresearch

Adaptive compensation of analog front-end I/Q mismatches in digital receivers.

Ediz Cetin
Izzet Kale
Richard Morling

Cavendish School of Computer Science

Copyright © [2001] IEEE. Reprinted from IEEE International Symposium on Circuits and Systems (ISCAS 2001), pp. 370-373.

This material is posted here with permission of the IEEE. Such permission of the IEEE does not in any way imply IEEE endorsement of any of the University of Westminster's products or services. Internal or personal use of this material is permitted. However, permission to reprint/republish this material for advertising or promotional purposes or for creating new collective works for resale or redistribution must be obtained from the IEEE by writing to pubs-permissions@ieee.org. By choosing to view this document, you agree to all provisions of the copyright laws protecting it.

The WestminsterResearch online digital archive at the University of Westminster aims to make the research output of the University available to a wider audience. Copyright and Moral Rights remain with the authors and/or copyright owners. Users are permitted to download and/or print one copy for non-commercial private study or research. Further distribution and any use of material from within this archive for profit-making enterprises or for commercial gain is strictly forbidden.

Whilst further distribution of specific materials from within this archive is forbidden, you may freely distribute the URL of WestminsterResearch. (http://www.wmin.ac.uk/westminsterresearch).

In case of abuse or copyright appearing without permission e-mail wattsn@wmin.ac.uk.

ADAPTIVE COMPENSATION OF ANALOG FRONT-END I/Q MISMATCHES IN DIGITAL RECEIVERS

Ediz Çetin, Izzet Kale and Richard C. S. Morling

University of Westminster, Department of Electronic Systems,
Applied DSP and VLSI Research Group,
London W1M 8JS, United Kingdom

ABSTRACT

I and Q Channel phase and gain mismatches are of great concern in communications receiver design. In this paper we analyse the effects of I and Q channel mismatches and propose a low-complexity blind adaptive algorithm to minimize this problem. The proposed solution consists of two, 2-tap adaptive filters, arranged in Adaptive Noise Canceller (ANC) set-up, with the output of one cross-fed to the input of the other. The system works as a de-correlator eliminating I and Q mismatch errors.

1. INTRODUCTION

With the scaling of microelectronic circuits more and more of the analog functions are replaced with their digital counterparts. Low cost and low power consumption in the handsets are the driving forces and they make the analog front-ends the bottleneck in future transceiver designs. Both low cost and low power are closely linked to the trend of full integration. However, many receivers still rely on the classic well-known narrowband analog heterodyne approach. [1],[2] This approach relies on analog off-chip components implemented in different technologies such as SiGe, GaAs, BiCMOS and CMOS [2], [3] hence, hindering the goal of achieving higher levels of integration. The homodyne/zero-IF receiver architecture eliminates many off-chip components required by the superheterodyne receiver. However, zero-IF receivers suffer from DC-offsets and I & Q channel mismatches [4]. The low-IF architecture alleviates all the above problems. It enables high performance and high levels of integration at the same time. The front-end of the low-IF receiver consists of a coarse RF filter, quadrature mixer and an ADC. The precision with which the I and Q paths can be matched determines to what extend the image signal can be suppressed. Low-IF receivers require image rejection in the 60-90 dB range [5]. A typical quadrature downconverter with a phase error of 3° results in image signal suppression of only 26 dB [5]. In this paper we will analyse the effects of the phase and gain mismatch on the analog front-end's performance. We will propose a feasible and simple adaptive (blind) correction scheme that can minimise this problem.

2. EFFECTS OF QUADRATURE MIXER PHASE AND GAIN MISMATCHES

Low-IF architecture has evolved from the marriage of the traditional heterodyne approach with the homodyne. With the low-IF receiver the incoming signal at f_{RF} is directly quadrature downconverted to a low-IF, f_{IF} not much greater than the signal bandwidth. Hence this architecture is insensitive to parasitic baseband signals like dc-offset voltages and self-mixing products that would otherwise be generated if the signal was to be downconverted to the baseband as in the case of the zero-IF receiver [6]. Thus the need to do image signal suppression at high frequencies with a high quality RF filter is eliminated. A broadband RF filter, as used in a homodyne receiver, will suffice.

With the low-IF topology the image and the wanted signal are both downconverted to opposite IF frequencies rather than being superimposed [6]. However, finite matching between the mixers and quadrature generator for the LO make it impossible to create a perfect single positive frequency. Therefore, although in theory the need to do image signal suppression at high frequencies is eliminated, in real life it is still required due to the imbalances between the *I* and *Q* channels [6]. The precision at which the *I* and *Q* paths are matched determines the image signal suppression performance of the low IF receiver.

Low-IF receivers require image rejection in the 60-80 dB range. Image rejection mixer performance may fall short for some standards' requirements. A typical quadrature downconverter with a phase error of 3° results in image signal suppression of 26 dB [5]. Figure below depicts a typical low IF architecture [6].

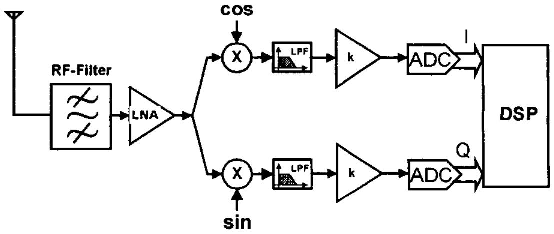


Figure 1 A Typical Low IF receiver[6]

The following is an outline of an analysis to determine the effects of gain and phase mismatches in a quadrature down-conversion architecture. The matching error is modeled as shown in Figure 2 [7].

0-7803-6685-9/01/$10.00©2001 IEEE

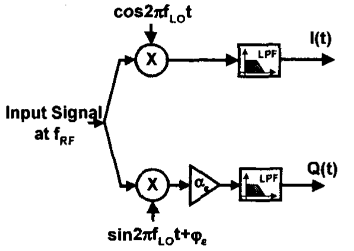


Figure 2 I and Q Phase and Gain Mismatch Model

The phase mismatch between the LO inputs of the mixers is defined as φ_ε. The gain mismatch between the I and Q channels is on the other hand defined by α_ε. These imbalances can cause the generation of an image signal that can in return limit the useful dynamic range of the receiver.

Using the model depicted in Figure 2 and input signal $x(t) = 2\cos 2\pi f_{RF} t$, we can write:

$$I(t) = (2\cos 2\pi f_{RF} t) \times \cos 2\pi f_{LO} t \tag{7}$$

$$Q(t) = 2\cos 2\pi f_{RF} t \times \alpha_\varepsilon \sin(2\pi f_{LO} t + \varphi_\varepsilon) \tag{8}$$

using product formulas:

$$I(t) = [\cos 2\pi(f_{RF} + f_{LO})t + \cos 2\pi(f_{RF} - f_{LO})t] \tag{9}$$

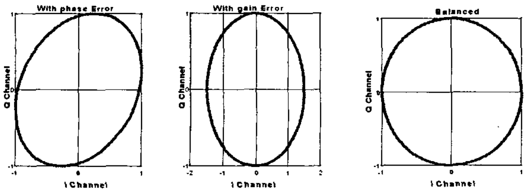
$$Q(t) = \alpha_\varepsilon[-\sin(2\pi(f_{RF} - f_{LO})t - \varphi_\varepsilon) + \sin(2\pi(f_{RF} + f_{LO}) + \varphi_\varepsilon)] \tag{10}$$

Substituting f_{IF} for $(f_{RF} - f_{LO})$ into equations (9) and (10), and low-pass filtering we have:

$$I(t) = \cos 2\pi f_{IF} t \tag{11}$$

$$Q(t) = -\alpha_\varepsilon \sin(2\pi f_{IF} t - \varphi_\varepsilon) \tag{12}$$

The new I and Q values of Equations (11) and (12) can be pictorially expressed as in Figure 3 below. Figure 3(a) depicts the effects of phase, 3(b) depicts the effects of gain imbalance on the quadrature mixer. Figure 3(c) displays the error free balanced system



(a) (b) (c)

Figure 3 (a) Phase, (b) Gain imbalance (phase error of 15° and gain error of 6dB) (c) ideal

As it can be seen from figure 3(a), the phase error rotates the quadrature axis distorting the quadrature mixers coordinate system. Figure 3(b) illustrates the effect of a gain mismatch between the I and Q channels. Circle is now turned into an ellipse.

Equations (11) and (12) can be combined as:

$$\begin{aligned} I(t) + jQ(t) &= \cos 2\pi f_{IF} t - j\alpha_\varepsilon \sin(2\pi f_{IF} t - \varphi_\varepsilon) \\ &= \tfrac{1}{2}\left[e^{j2\pi f_{IF} t}(1 - \alpha_\varepsilon e^{-j\varphi_\varepsilon}) + e^{-j2\pi f_{IF} t}(1 + \alpha_\varepsilon e^{j\varphi_\varepsilon})\right] \end{aligned} \tag{13}$$

From equation (13) the desired signal can be considered as $e^{-j2\pi f_{IF} t}$ and the image response as $e^{j2\pi f_{IF} t}$. If α_ε=1 and φ_ε=0, the desired term becomes $e^{-j2\pi f_{IF} t}$ and the image becomes zero as expected. In general, the corresponding amplitudes of the signal and image are $1+\alpha_\varepsilon e^{j\varphi\varepsilon}$ and $1-\alpha_\varepsilon e j^{-\varphi\varepsilon}$, respectively. A_{sig} is the amplitude of the desired signal and A_{img} is the amplitude of the image signal. The image amplitude relative to the desired signal amplitude can be written in decibels as: [7]

$$10\log\left(\frac{A_{sig}}{A_{img}}\right)^2 = 10\log\left(\frac{1+\alpha_\varepsilon^2 - 2\alpha_\varepsilon\cos(\varphi_\varepsilon)}{1+\alpha_\varepsilon^2 + 2\alpha_\varepsilon\cos(\varphi_\varepsilon)}\right) \tag{14}$$

Figure 4 depicts the calculation of α_ε and φ_ε values for given required image rejection ratios.

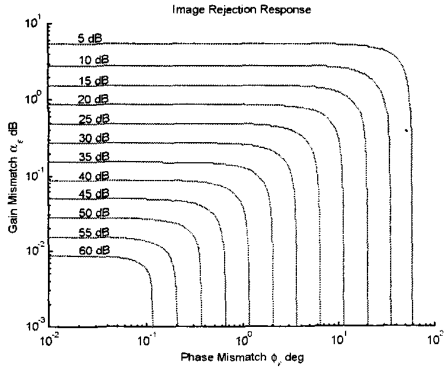


Figure 4 Image Reject ratio as a function of α_ε and φ_ε

As it can be seen from the above figure in order to achieve 30 dB image suppression gain and phase errors must be constrained to 0.5dB and 0.1degrees. Designing Quadrature mixers with such accuracy is not a trivial task.

3. SOLUTION TO GAIN AND PHASE MISMATCH ERRORS

Different solutions have been reported in the literature tackling the problem of I and Q channel imbalances [8]-[10]. The majority of these solutions rely on a pilot tone to adjust the phase and gain errors. For wideband receivers this is a tedious process since more than one test tone is required to cover the input bandwidth of the receiver. Also, test tone generation is not an error free process. It is prone to errors generating erroneous results and leading to wrong calibration values.

Our novel approach differs from other approaches in that it does not make use of a test tone. Instead we make use of blind adaptive algorithm to combat the mismatch errors. We also do not try to re-orthogonalise the I and Q channels by means of Gram Schmidt [9] or related technique. We instead show that there exists a correlation between the I and Q channels if there are phase and gain errors. In a fully matched system the I and Q channels are uncorrelated since they are orthogonal. However, in the presence of analog impairments this orthogonal relationship no longer exists. This results in a cross-correlation between the two channels. This is depicted in Figure 5. Depending on the size of mismatch the level of cross correlation is determined.

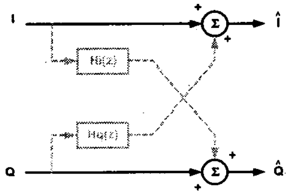


Figure 5 Signal leakage model due to phase and gain errors.

Hence $\hat{I}$ and $\hat{Q}$ can be expressed as:

$$\begin{bmatrix} \hat{I} \\ \hat{Q} \end{bmatrix} = \begin{bmatrix} 1 & Hq(z) \\ Hi(z) & 1 \end{bmatrix} \times \begin{bmatrix} I \\ Q \end{bmatrix} \tag{15}$$

As shown in Figure 6 below, cross correlation increases with the phase and gain errors. Phase error is varied from 22.5 to 0°, and gain error is varied between 6 to 0 dBs. As it can bee seen from the graph in the ideal case there is no cross correlation between the I and Q channels.

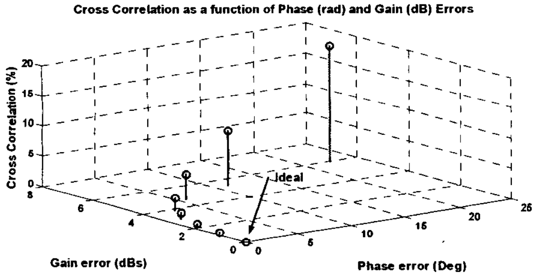


Figure 6 Cross Correlation as a function of phase and gain errors

The proposed algorithm tries the model $Hq(z)$ and $Hi(z)$. Hence doing so it de-correlates the two signals, $\hat{I}$ and $\hat{Q}$. Figure 7 below depicts our proposed solution that deals with the I and Q channel imbalances.

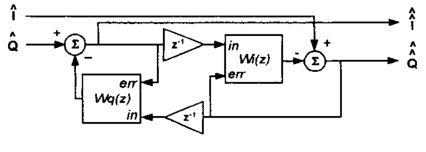


Figure 7 Proposed LMS based algorithm

As depicted in Figure 7 above, the proposed solution consists of two 2-tap adaptive filters, arranged in Adaptive Noise Canceller (ANC) set-up, with the output of one cross-fed to the input of the other. The unit delay is required in each adaptive filter to compensate for the delay in the feedback path. Adaptive system in steady state can be expressed as [11]:

$$\begin{bmatrix} 1 & Wi(z)z^{-1} \\ Wq(z)z^{-1} & 1 \end{bmatrix} \times \begin{bmatrix} \hat{\hat{I}} \\ \hat{\hat{Q}} \end{bmatrix} = H(z) \times \begin{bmatrix} I \\ Q \end{bmatrix} \tag{16}$$

where

$$H(z) = \begin{bmatrix} 1 & Hq(z) \\ Hi(z) & 1 \end{bmatrix} \tag{17}$$

This equation can be solved for $\hat{\hat{I}}$ and $\hat{\hat{Q}}$ to give:

$$\begin{bmatrix} \hat{\hat{I}} \\ \hat{\hat{Q}} \end{bmatrix} = \frac{1}{1-Wq(z)Wi(z)z^{-2}} \times \begin{bmatrix} 1 & -Wi(z)z^{-1} \\ -Wq(z)z^{-1} & 1 \end{bmatrix} \times H(z) \times \begin{bmatrix} I \\ Q \end{bmatrix} \tag{18}$$

Following our initial assumption that I and Q are uncorrelated signals such that:

$$E[I(n) \times Q(n-k)] = 0, \quad \forall k, \tag{19}$$

where E[.] denotes expectation. Based on the orthogonal principle, the filter $Wi(z)$ decorrelates its error signal $\hat{\hat{Q}}$ with the input signal $\hat{\hat{I}}(z^{-1})$; while the filter $Wq(z)$ decorrelates its input signal $\hat{\hat{Q}}(z^{-1})$ with the error signal $\hat{\hat{I}}$. Thus, the output signals $\hat{\hat{I}}$ and $\hat{\hat{Q}}$ are statistically uncorrelated that is,

$$E\left[\hat{\hat{I}}(n) \times \hat{\hat{Q}}(n-k)\right] = 0, \quad \forall k, \tag{20}$$

From (17) and (18), it is clear that the choice of the following solution

$$\begin{bmatrix} Wi(z)z^{-1} \\ Wq(z)z^{-1} \end{bmatrix} = \begin{bmatrix} Hi(z) \\ Hq(z) \end{bmatrix} \tag{21}$$

yields,

$$\begin{bmatrix} \hat{\hat{I}}(n) \\ \hat{\hat{Q}}(n) \end{bmatrix} = \begin{bmatrix} I(n) \\ Q(n) \end{bmatrix} \tag{22}$$

which satisfies (20). Therefore, the filter $Wi(z)$ identifies the crosstalk between the I and Q whereas $Wq(z)$ identifies the crosstalk between the Q and I channels. Hence the crosstalk effect is eliminated.

4. SIMULATION RESULTS

I and Q channels in GSM when plotted one against the other result in a circle. Any deviations from the orthogonality results in this circle turning into an ellipse. Shown in Figures 8, 9 and 10 are the results of the application of the adaptive system to quadrature mixer problem in the presence of Additive White Gaussian Noise (AWGN) and phase and gain errors.

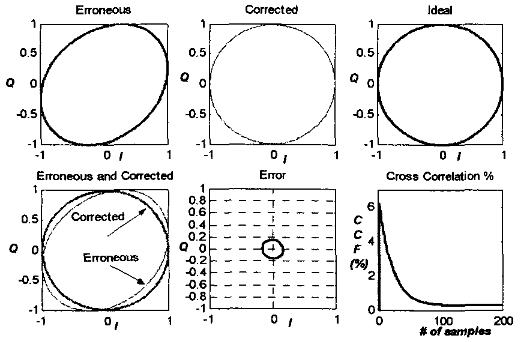


Figure 8 Simulation Results

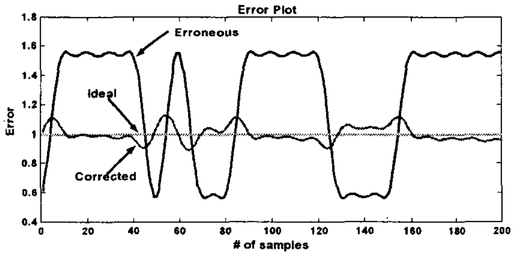
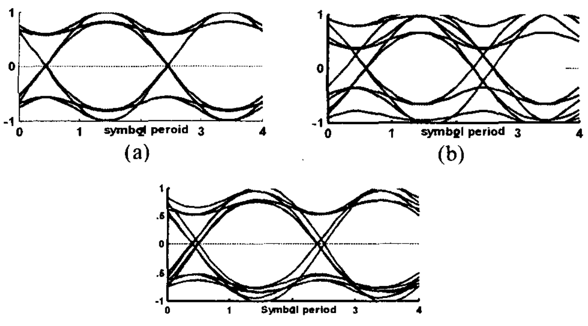


Figure 9 Simulation Results



(a)

(b)

(c)

Figure 10 Eye Diagrams (a) Original, (b) With phase and gain error and (c) Corrected

As it can be seen from the simulation results of Figure 8, the I and Q channels were correlated by 6% this was reduced all the way down to 0.12% improvement of 97.6%. The ellipse is transformed into a circle. The plot entitled Error depicts the difference between the ideal and corrected result. The maximum amplitude error on the processed Q signal is 0.15 and 0.14 on the I signal. Another important measure is the value of I^2+Q^2. In a balanced system it should result in value of 1. Error plot of Figure 9 depicts this for ideal, erroneous and corrected systems. Ideal is 1, erroneous varies between 0.56 and 1.56 hence maximum deviation from the ideal is 56%. In the case of corrected system output values vary between 0.89 and 1.12 hence maximum variance from the ideal is 0.12 resulting in maximum variance of 12%, improvement of fivefold. Figure 10 depicts the eye diagrams for three cases. As it can be seen the output of the adaptive system matches the original eye diagram.

5. CONCLUSIONS

In this paper we have presented the design of a blind adaptive system to combat the phase and gain errors generated by quadrature mixer based digital receivers. We have reported on the performance of our adaptive cancellation scheme through simulation results that demonstrate substantial improvements. The algorithm is extremely simple to implement and lends itself to efficient real-time realisation.

6. REFERENCES

[1] Shen D. H., Chien Meen Hwang, Brice B. Lusignan and bruce A wooley, "A 900MHz RF Front-End with Integrated Discrete-Time Filtering", *IEEE Journal of Solid State Circuits*, vol. 31, no.12, December 1996

[2] Razavi B., "Recent Advances in RF Integrated Circuits", *IEEE Communications Magazine*, pp.36-43, December 1997

[3] Gray P.R. and R.G. Meyer, "Future Directions in Silicon ICs for RF Personal Communications," *Proc. of IEEE Custom Integrated Circuits Conference (CICC)*, pp.91-94, May 1995.

[4] Behzad R, "Design Considerations for Direct-Conversion Receivers", *IEEE Trans. On Circuits and Systems-II: Analog and Digital Signal Processing*, vol.44, No, 6, pp. 428-435, June 1997.

[5] Jan Crols and M.J. Steyaert, "Low-IF Topology for High Performance Analog Front Ends of Fully Integrated Receivers", *IEEE Trans. on Circuits and systems- II: Analog and Digital Signal Processing*, vol. 45, no.3, pp. 269-282 March 1998

[6] J. Crols and M. Steyaert, "A Single-Chip 900 MHz CMOS Receiver Front End with a High-Performance Low-IF Topology," *IEEE J. Solid-State Circuits*, vol. 30, pp. 1483–1492, Dec. 1995.

[7] Jacques C. Rudell at al , "A 1.9MHz Wide-band IF Double Conversion CMOS Receiver for Cordless Telephone Applications", *IEEE Journal of solid state circuits*, vol. 32, no. 12, pp.2071-2088 December 1997

[8] Gray P.R. and R.G. Meyer, "Future Directions in Silicon ICs for RF Personal Communications," *Proc. of IEEE Custom Integrated Circuits Conference (CICC)*, pp.91-94, May 1995.

[9] Churchill F.E., G.W. Ogar and B.J. Thompson, "The Correction of I and Q Errors in a Coherent Processor", *IEEE Trans. on Aerospace and Electronic Systems*.vol. AES-17, no.1, pp. 131-137, January 1981.

[10] Tsui J B.Y, " *Digital Microwave Receivers-Theory and Concepts*", Artech House, 1989, ISBN: 0-89006-339-7.

[11] Mirchandani G. Zinser R. L. and J. B. Evans "A New Adaptive Noise Cancellation Scheme in the Presence of Crosstalk", *IEEE Trans. On Circuits and Systems—II: Analog and Digital Signal Processing*, vol 39, no.10, pp. 681-694, October 1992.