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ADAPTIVE DIGITAL RECEIVERS FOR ANALOG FRONT-END MISMATCH CORRECTION

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***Abstract*- Phase and gain mismatches between the I and Q analog signal processing paths of a quadrature receiver are responsible for the generation of image signals which limit the dynamic range of a practical receiver. In this paper we analyse the effects these mismatches and propose a low-complexity blind adaptive algorithm to minimize this problem. The proposed solution is based on two, 2-tap adaptive filters, arranged in Adaptive Noise Canceller (ANC) set-up. The algorithm lends itself to efficient real-time implementation with minimal increase in modulator complexity.**

1. INTRODUCTION

In today's world of miniaturization, there is ever-increasing pressure to reduce the cost of communication chips. This is the driving force to develop transceivers with higher levels of integration. However, many receivers used today 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 need for higher levels of integration hence low cost and power has resulted in new transceiver architectures like the low-IF and the zero-IF receivers that reduce the off-chip components required. The zero-IF receiver provides high level of integration but suffers from the DC-offsets and analog I and Q channel mismatches [4] which has prevented it from being used in high-end applications. The low-IF receiver on the other hand alleviates all the above problems enabling high performance and high levels of integration at the same time. However, due to its two-path structure, analog I and Q processing is extremely sensitive to phase and gain mismatches. With practical analog components, mismatches are unavoidable resulting in imperfect image rejection. The precision with which the two analog paths can be matched determines to what extend the image signal can be suppressed. In order to achieve 50dB image signal rejection, phase and gain errors must be constraint to 0.2° and 0.05dB. In this paper we analyse the effects of the analog front-end phase and gain mismatches on the low-IF receivers performance. We then propose a feasible and simple adaptive (blind) digital correction scheme at the baseband to compensate the analog front-end impairments without a substantial increase in modulator complexity.

The paper is organised as follows: Section 2 analysis of nonlinear analog front-end is given. In Section 3, the blind mismatch correction scheme is derived. Simulation results and concluding remarks are given in sections 4 and 5 respectively.

2. ANALYSIS OF NONLINEAR ANALOG FRONT-END

In the low-IF receiver, the incoming signal is directly quadrature downconverted to a low-IF, not much greater than the signal bandwidth. In comparison with zero-IF receivers, the receiver is insensitive to parasitic baseband signals like dc-offset voltages and self-mixing products that would otherwise be generated as in the case of the zero-IF receiver [5]. With this receiver 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. However, quadrature mixing, which in theory provides infinite image signal suppression, provides finite image rejection due to mismatch between the mixers and quadrature generator for the local oscillator (LO). This may fall short of some standards' requirements. Figure 1 depicts a low IF architecture with all the phase and gain mismatches of the analog I and Q paths incorporated into the quadrature mixer as φ_ε and α_ε.

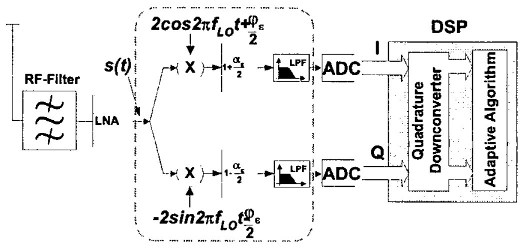

Figure 1 The Low IF receiver

The following is an outline of an analysis to determine the effects of these errors on the quadrature receivers. These imbalances cause the generation of an image signal, which limit the useful dynamic range of the receiver. Evaluate the performance of the receiver in the presence of adjacent channel interferer and analog front-end impairments. Incoming signal s(t) consists of the desired signal u(t) at f_{RF} and the interferer i(t) at f_{INT} where $f_{INT}=f_{RF}-2f_{IF}$. Received signal s(t) can be expressed as:

$$s(t)=\tfrac{1}{2}[u(t)e^{j(2\pi f_{RF}t)}+u^*(t)e^{-j(2\pi f_{RF}t)}]+[i(t)e^{j(2\pi f_{IMG}t)}+i^*(t)e^{-j(2\pi f_{IMG}t)}] \quad (1)$$

where u(t) *and* i(t) are the complex envelopes of the bandpass signals at f_{RF} and f_{IMG} respectively and * is the complex conjugate. The non-ideal LO signal can be expressed as:

$$x_{LO}=2(1+\tfrac{\alpha_\varepsilon}{2})\cos(2\pi f_{LO}t+\tfrac{\varphi_\varepsilon}{2})-j2(1-\tfrac{\alpha_\varepsilon}{2})\sin(2\pi f_{LO}t-\tfrac{\varphi_\varepsilon}{2}) \quad (2)$$

This can be re-written as:

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$$x_{LO} = e^{j(2\pi f_{LO}t)}(g_1 e^{j\frac{\varphi_\varepsilon}{2}} - g_2 e^{-j\frac{\varphi_\varepsilon}{2}}) + e^{-j(2\pi f_{LO}t)}(g_1 e^{-j\frac{\varphi_\varepsilon}{2}} + g_2 e^{j\frac{\varphi_\varepsilon}{2}}) \quad (3)$$

where $g_1=(1+0.5\alpha_\varepsilon)$, $g_2=(1-0.5\alpha_\varepsilon)$. LO signal can be shown as:

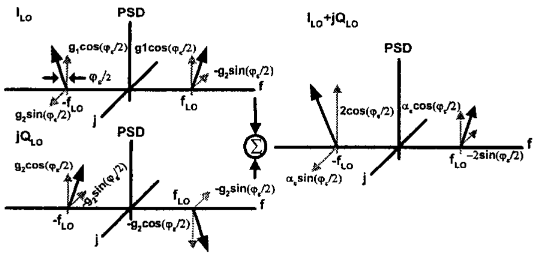


Figure 2 Frequency response of mismatched LO signals

As can be seen from Figure 2, instead of having a single tone at $-f_{LO}$ we have, due to the analog mismatches, an image at $+f_{LO}$ as well. Quadrature mixing s(t) with x_{LO} and Low-Pass Filtering (LPF) results in an IF signal:

$$s_{IF}(t) = \tfrac{1}{2}\left[u(t)e^{j2\pi f_{IF}t}(1+\alpha_\varepsilon e^{-j\varphi_\varepsilon}) + u^*(t)e^{-j2\pi f_{IF}t}(1-\alpha_\varepsilon e^{j\varphi_\varepsilon})\right] + \tfrac{1}{2}\left[i^*(t)e^{j2\pi f_{IF}t}(1-\alpha_\varepsilon e^{-j\varphi_\varepsilon}) + i(t)e^{-j2\pi f_{IF}t}(1+\alpha_\varepsilon e^{j\varphi_\varepsilon})\right] \quad (4)$$

where the desired signal u(t) is corrupted by the interferer i*(t) leaked in-band due to phase and gain mismatches. There is also a leakage from the desired signal into image channel. This is depicted in Figure 3.

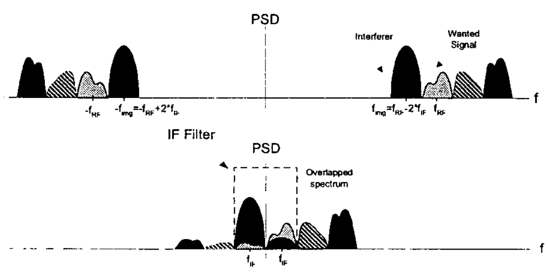


Figure 3 RF to IF down-conversion

If we let A_{sig} to denote the amplitude of the desired and A_{img} the amplitude of the image signal; as a function of gain and phase errors, the image rejection ratio (IRR) in decibels can be written as in equation (5) and shown graphically in Figure 4.

$$10\log\left(\frac{A_{img}}{A_{sig}}\right)^2 = 10\log\left(\frac{2-2\cos\varphi_\varepsilon+0.5\alpha_\varepsilon^{\,2}(1+\cos\varphi_\varepsilon)}{2+2\cos\varphi_\varepsilon+0.5\alpha_\varepsilon^{\,2}(1-\cos\varphi_\varepsilon)}\right) \quad (5)$$

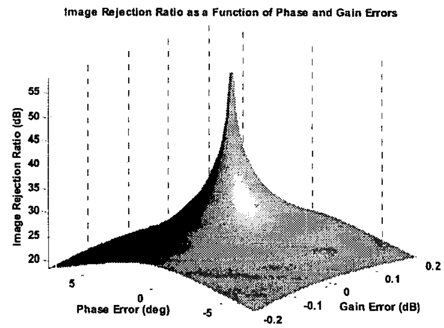


Figure 4 Image Reject ratio as a function of α_ε and φ_ε

In order to achieve 50dB image signal rejection, phase and gain errors must be constraint to 0.2° and 0.05 dB. Designing low-power analog components with such accuracy is not a trivial task.

3. BLIND CORRECTION SCHEME

In the ideal case the in-phase $s_I(t)$ and the quadrature $s_Q(t)$ signals are orthogonal and there is no crosstalk between them e.g. they are not correlated. Hence:

$$E\left[s_I(n)\times s_Q(n-k)\right]=0, \qquad \forall k, \qquad (6)$$

where $E[\bullet]$ denotes expectation. In the case where there are no mismatch errors, the transmitted in-phase and quadrature components relate to the received ones at the baseband as:

$$\begin{bmatrix} r_I(t) \\ r_Q(t) \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} s_I(t) \\ s_Q(t) \end{bmatrix} \qquad (7)$$

From (7) we can see that there is no crosstalk between the quadrature components hence they are still not correlated with each other and orthogonal. In the presence of analog front-end impairments however, this relationship no longer holds. Corrupted baseband signal can be expressed as:

$$\begin{bmatrix} r_I(t) \\ r_Q(t) \end{bmatrix} = \underbrace{\begin{bmatrix} (1+0.5\alpha_\varepsilon)\cos\frac{\varphi_\varepsilon}{2} & (1+0.5\alpha_\varepsilon)\sin\frac{\varphi_\varepsilon}{2} \\ (1-0.5\alpha_\varepsilon)\sin\frac{\varphi_\varepsilon}{2} & (1-0.5\alpha_\varepsilon)\cos\frac{\varphi_\varepsilon}{2} \end{bmatrix}}_{M} \begin{bmatrix} s_I(t) \\ s_Q(t) \end{bmatrix} \qquad (8)$$

where M is the mixing matrix which transforms the received $s_I(t)$ and $s_Q(t)$ signals to new $r_I(t)$ and $r_Q(t)$ signals. Effect of this on the signal constellations of QPSK and GMSK modulation schemes are shown in Figures 5 and 6.

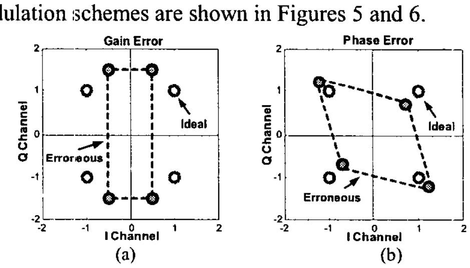
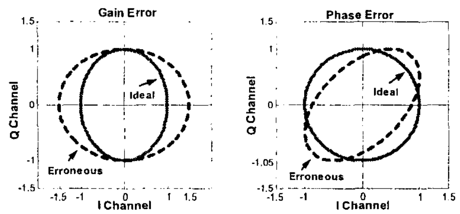


Figure 5 QPSK constellation with (a) Gain, (b) Phase imbalance (gain error of 6dB and phase error of 15°)



(a) (b)

Figure 6 GMSK constellation with (a) Gain, (b) Phase imbalance (phase error of 15° and gain error of 6dB)

As it can be seen from equation (8) there is now crosstalk between the quadrature components. This results in loss of orthogonality and correlation between the quadrature components. The crosstalk matrix H(z) can be modelled as:

$$H(z)=\begin{bmatrix} 1 & H_Q(z) \\ H_I(z) & 1 \end{bmatrix} \tag{9}$$

where $H_Q(z)$ determines the crosstalk from the $s_Q(t)$ to $s_I(t)$ and $H_I(z)$ determines the crosstalk from the $s_I(t)$ to $s_Q(t)$ channels. Both $H_I(z)$ and $H_Q(z)$ depend on the amount of analog front-end phase and gain mismatch. Figure 7 depicts the cross-correlation as a function of phase and gain mismatches. As the phase and gain errors increase the percentage cross-correlation increases as well. It is zero in the ideal case i.e. fully matched system, as expected.

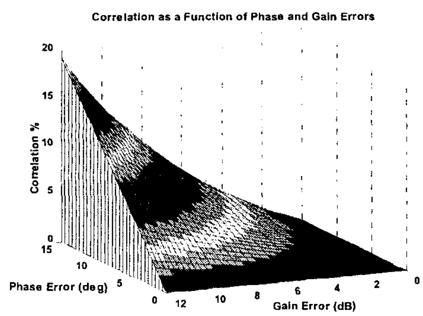


Figure 7 Cross-correlation as function of phase and gain errors

Crosstalk model and our proposed solution (outlined block) are depicted in Figure 8.

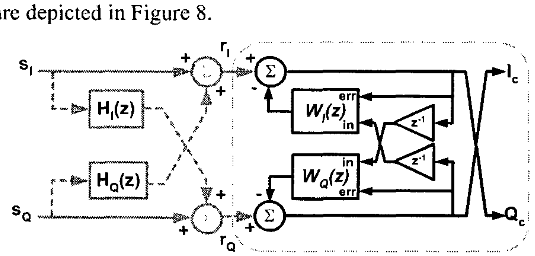


Figure 8 Proposed LMS based algorithm

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. Our approach differs from the other estimation and correction methods that have been reported in the literature [6]-[8] in that, no pilot tone is required i.e. blind algorithm. Hence the performance degradation due the errors in the test tone generation for wide range of frequencies to cover wide bandwidths is eliminated.

The proposed algorithm in steady state can be mathematically expressed as [9]:

$$\begin{bmatrix} 1 & W_I(z)z^{-1} \\ W_Q(z)z^{-1} & 1 \end{bmatrix} \times \begin{bmatrix} I_C \\ Q_C \end{bmatrix} = H(z) \times \begin{bmatrix} r_I \\ r_Q \end{bmatrix} \tag{10}$$

Solving this for I_C and Q_C:

$$\begin{bmatrix} I_C \\ Q_C \end{bmatrix} = \frac{1}{1-W_Q(z)W_I(z)z^{-2}} \times \begin{bmatrix} 1 & -W_I(z)z^{-1} \\ -W_Q(z)z^{-1} & 1 \end{bmatrix} \times H(z) \times \begin{bmatrix} r_I \\ r_Q \end{bmatrix} \tag{11}$$

Based on the orthogonal principle, the filter $W_I(z)$ decorrelates its error signal Q_C with the input signal $I_C(z)z^{-1}$ while the filter $W_Q(z)$ decorrelates its input signal $Q_C(z)z^{-1}$ with the error signal I_C. Thus, the output signals I_C and Q_C are statistically uncorrelated. That is:

$$E[I_C(n) \times Q_C(n-k)] = 0, \qquad \forall k, \tag{12}$$

It is clear that the choice of the following solution:

$$\begin{bmatrix} W_I(z)z^{-1} \\ W_Q(z)z^{-1} \end{bmatrix} = \begin{bmatrix} H_I(z) \\ H_Q(z) \end{bmatrix} \tag{13}$$

yields:

$$\begin{bmatrix} I_C(n) \\ Q_C(n) \end{bmatrix} = \begin{bmatrix} s_I(n) \\ s_Q(n) \end{bmatrix} \tag{14}$$

which satisfies (12). Therefore, the filter $W_I(z)$ identifies the crosstalk between the I and Q whereas $W_Q(z)$ identifies the crosstalk between the Q and I channels. Hence the crosstalk effect is eliminated and quadrature channels are orthogonal again.

4. SIMULATION RESULTS

Simulations were carried out using a GMSK modulated signal in the presence of Additive White Gaussian Noise (AWGN) having phase and gain errors of 15° and 6dB respectively. Figure 9 depicts the response of the system to a constant phase input. The left hand plot of Figure 9 depicts the error, I^2+Q^2, plots, for the ideal, erroneous and corrected systems. The ideal is 1; the erroneous varies between 0.74 and 1.26 giving a maximum deviation of 26% from the ideal. In the case of the corrected system output values vary between 0.9524 and 1.003 hence the maximum deviation from the ideal is 0.04 i.e. 4%, giving an improvement of 84%.

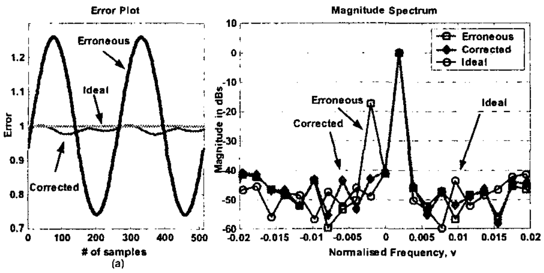


Figure 9 Simulation results for constant phase input

The right hand plot depicts the magnitude response of the resulting complex tone at f_{LO}. As it can be seen from this graph the image response generated, at $-f_{LO}$ due to the analog impairments is pulled down by almost 26 dB, approaching the ideal case. 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 10, 11 and 12 are the results of the application of the adaptive system.

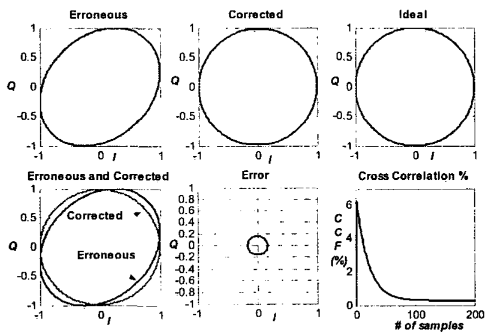


Figure 10 Signal constellation diagrams of the system

As it can be seen from the simulation results of Figure 10, the I and Q channels were correlated by 6% this was reduced all the way down to 0.12%. An 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. Error plot of Figure 11 depicts I^2+Q^2 for ideal, erroneous and corrected systems.

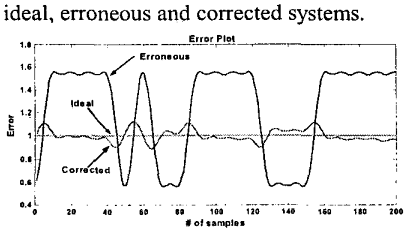
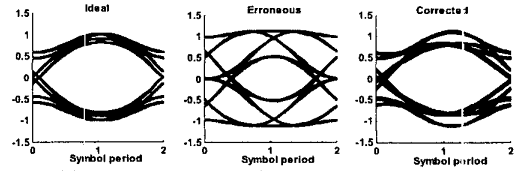


Figure 11 Error plot for GMSK signal

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.



(a) (b) (c)

Figure 12 Eye Diagrams (a) Original, (b) erroneous and (c) Corrected

As it can be seen from Figure 12, the eye diagram of the erroneous system, (b), is closed. The output of our adaptive system, (c), substantially opens the eye converting it back to the original state, (a). Hence, resulting in reduced sensitivity to timing errors and noise margin. Also, zero crossing distortion due to analog impairments is removed.

5. CONCLUSIONS

In this paper we have presented the design of a blind adaptive system to combat the phase and gain mismatches in the I and Q analog signal processing paths of the quadrature 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 with minimal increase in modulator complexity.

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