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Efficient Compensation of Receiver IQ Imbalance in OFDM System

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ABSTRACT

Transceivers based on a direct conversion architecture introduces an unwanted limited image rejection and degrades the accuracy of carrier estimation due to presence of I/Q imbalance. In this paper, we proposed an efficient scheme to compensate IQ imbalance in OFDM receivers. Simulation results confirmed that the proposed scheme performance is very close to the ideal case.

Keywords- IQ imbalance, OFDM, Receiver, AWGN

1. Introduction

OFDM is a widely recognized and standardized modulation technique [1], [2]. Due to lower complexity and higher flexibility, direct conversion receivers are preferred to super heterodyne transceivers in OFDM systems. Unfortunately, OFDM transceivers are sensitive to non-idealities in the receiver front-end [3]. IQ imbalance has been identified as a key front-end effect for OFDM systems. In practice, due to manufacturing imperfections, there would be phase and amplitude mismatch in the phase (I) and quadrature (Q) branches of the IQ-receiver commonly known as IQ-imbalance [3]. The compensation of IQ-imbalance in MIMO-OFDM has gained a lot of attention recently [4], [5]. Phase noise is another major impairment caused by non-ideal oscillator which introduces random phase fluctuations at the oscillator output resulting in distortion of the signals [3]. The effect of phase noise is more severe in OFDM systems and is studied extensively in the literature [6], [7]. These works state that the effect of phase noise can be broken down into common phase error (CPE) and inter-carrier interference (ICI) with CPE is more prominent for systems with less phase noise. Analysis of performance and compensation techniques are proposed for OFDM and MIMO-OFDM systems in the literature and can be found in [7] -[9].

Joint IQ-imbalance and phase noise compensation techniques for single-input-single-output (SISO) systems have been previously discussed in [10] and [11]. However, joint treatment of these impairments for MIMO-OFDM systems have not been studied extensively. In [12], joint estimation and compensation scheme for MIMO-OFDM systems, assuming quasi-static channel is proposed. The authors first develop a compensation scheme for SISO systems and generalize it to the MIMO case. The proposed method requires more training symbols with increasing number of transmit-receive antennas and the complexity of the scheme is high due to the required maximum likelihood (ML) estimation. In this paper, we proposed an efficient scheme to compensate IQ imbalance in OFDM receivers and compared its performance with existing IQ imbalance compensation schemes. The paper is organized as follows, in section 2 system model is described. In section 2.1 proposed schemes for IQ imbalance compensation is explained. In section 3 simulation results are presented and in section 4 the paper is concluded.

2. System Model

At the transmitter serial data stream is converted to parallel data using serial to parallel converter, modulated and passed through N-point FFT block and cyclic prefix are added before transmitting the signal. Let $s_{n,k}$ be the k^{th} frequency domain OFDM symbol to be transmitted from the n^{th} transmit antenna whose time domain representation is given by $\overline{s_{n,k}}$ = $F^{-1}(s_{n,k})$ where F^{-1} is the inverse Fourier transform operation. The vector that contains all the N + N_{cp} + L − 1 discrete time domain received signal samples at the m^{th} receive antenna is then given by

$$\overline{r_{m,k}} = \sum_{n=1}^{M} h_{m,n} * \overline{s_{n,k}} + w_m$$

Where $h_{m,n}$ the channel impulse response of the multipath channel from the n^{th} transmit to the m^{th} receive antenna for the k^{th} OFDM symbol. $\overline{s_{n,k}}$ is the transmitted discrete time signal vector from the n^{th} antenna to m^{th} antenna. w_q is the added additive white Gaussian noise (AWGN) at the m^{th} receive antenna. At the receiver the cyclic prefix of the received signal is removed, the signal samples are then converted into parallel streams and fed into FFT block with size N. The equivalent frequency domain received signal model at the i^{th} subcarrier

$$r_k(i) = H_k(i)s_k(i) + \eta_k(i)$$

Where $r_k(i)$ is the received signal, $H_k(i)$ is the frequency response of the channel, $s_k(i)$ is the transmitted signal at i^{th} sub carrier of k^{th} OFDM symbol.

2.1 IQ Imbalance Model

The time domain oscillator output with frequency independent IQ imbalance can be given as [23],[24]

$$y(t) = \cos(\omega_c t) + j\varepsilon \sin(\omega_c t + \phi)$$

Where ε and ϕ are amplitude and in-phase mismatches respectively. Y(t) can also be given as

$$\begin{aligned} y(t) &= (1+\varepsilon)\cos\Delta\varphi\Re\{x\} \\ &- j(1-\varepsilon)\sin\Delta\varphi\Re\{x\} \\ &+ j(1-\varepsilon)\cos\Delta\varphi\Im\{x\} \\ &- (1+\varepsilon)\sin\Delta\varphi\Im\{x\} \end{aligned}$$

$$y(t) = (\cos\Delta\phi - j\varepsilon\sin\Delta\phi).x - (\varepsilon\cos\Delta\phi - j\sin\Delta\phi).x^*$$

$$y(t) = \alpha.x + \beta.x^* \quad - \tag{1}$$

Where y(t) represents signal with imbalance, $\Re\{\cdot\}$ represents real part of signal and $\Im\{\cdot\}$ represents imaginary part of signal.

$$\alpha = \cos\Delta\phi + j\varepsilon\sin\Delta\phi$$

$$\beta = \varepsilon\cos\Delta\phi - j\sin\Delta\phi$$

If no IQ imbalance is present, then $\alpha = 1$ and $\beta = 0$ and then (1) reduces to $y = x$.

We analyze the effect of the IQ imbalance on OFDM transmission. We consider AWGN channel. If x is the transmitted OFDM symbol (in the frequency domain), then *IFFT*(x) is the incoming time domain signal in the receiver. Applying the IQ imbalance (1) and taking the FFT leads to

$$R = FFT\{\alpha \times IFFT(x) + \beta \times IFFT(x)^*\}$$

$$R = \alpha \times x + \beta \times x^*$$

2.2 Cost function for IQ imbalance

Y(t) is the received signal at the OFDM receiver, equation (1) can also be given as

$$y(t) = A\cos(\omega_c t) + j\alpha A\sin(\omega_c t + \phi + \varepsilon)$$

Where α amplitude is mismatch and ε is in-phase mismatch

At the OFDM receiver the received signal with IQ imbalance can be given as

$$\begin{aligned}R(m) &= \left(\frac{1}{N}\times\alpha\times\sum_{k=-N/2}^{N/2-1}X(k)e^{j2\pi mk/2N}\right)\\&+\left(\frac{1}{N}\times\beta\times\sum_{k=-N/2}^{N/2-1}X^{*}(k)e^{j2\pi mk/2N}\right)\end{aligned} \qquad (2)$$

where *X*(*k*) denote the frequency domain, transmitted data for the *k*-th subcarrier, N is the number of subcarriers. Equation (2) can be written as

$$\begin{aligned}&R(m) = \left(\frac{1}{N}\times(\cos\Delta\phi + j\varepsilon\sin\Delta\phi)\times\sum_{k=-N/2}^{N/2-1}X(k)e^{j2\pi mk/2N}\right)+\\&\left(\frac{1}{N}\times(\varepsilon\cos\Delta\phi - j\sin\Delta\phi)\times\sum_{k=-N/2}^{N/2-1}X^{*}(k)e^{j2\pi mk/2N}\right)\end{aligned}$$

$$\begin{aligned}R(m) &= \left(\frac{1}{N}\times\left(\frac{e^{\Delta\phi}+e^{-\Delta\phi}}{2}+j\varepsilon\frac{e^{\Delta\phi}-e^{-\Delta\phi}}{2j}\right)\times\sum_{k=-N/2}^{N/2-1}X(k)e^{j2\pi mk/2N}\right)\\&+\left(\frac{1}{N}\times\left(\varepsilon\frac{e^{\Delta\phi}+e^{-\Delta\phi}}{2}-j\frac{e^{\Delta\phi}-e^{-\Delta\phi}}{2j}\right)\times\sum_{k=-N/2}^{N/2-1}X^{*}(k)e^{j2\pi mk/2N}\right)\end{aligned}$$

$$\begin{aligned}R(m) &= \left(\frac{1}{N}\times\frac{((1+\varepsilon)e^{\Delta\phi}+(1-\varepsilon)e^{-\Delta\phi})}{2}\times\sum_{k=-N/2}^{N/2-1}X(k)e^{j2\pi mk/2N}\right.\\&+\frac{1}{N}\times\frac{((\varepsilon-1)e^{\Delta\phi}+(\varepsilon+1)e^{-\Delta\phi})}{2}\times\sum_{k=-N/2}^{N/2-1}X^{*}(k)e^{j2\pi mk/2N}\end{aligned}$$

$$\begin{aligned}R(m) &= \left(\frac{1}{N}\times\frac{((1+\varepsilon)e^{\Delta\phi}+(1-\varepsilon)e^{-\Delta\phi})}{2}\times\sum_{k=-N/2}^{N/2-1}X(k)e^{j2\pi mk/2N}\right.\\&+\frac{1}{N}\times\frac{((\varepsilon-1)e^{\Delta\phi}+(\varepsilon+1)e^{-\Delta\phi})}{2}\times\sum_{k=-N/2}^{N/2-1}X^{*}(k)e^{j2\pi mk/2N}\end{aligned} \qquad (3)$$

Equation (3) can be rearranged as

$$
\begin{aligned}
R(m) = (&\frac{(1+\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} X(k) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{(1-\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} X(k) e^{j2\pi mk/2N} e^{-\Delta\phi} \\
&+ \frac{(\varepsilon-1)}{2N} \times \sum_{k=-N/2}^{N/2-1} X^{*}(k) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{(\varepsilon+1)}{2N} \times \sum_{k=-N/2}^{N/2-1} X^{*}(k) e^{j2\pi mk/2N} e^{-\Delta\phi}
\end{aligned}
$$

$$
\begin{aligned}
R(m) = (&\frac{(1+\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{(1+\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} + \\
&\frac{(1-\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi} \\
&+ \frac{(1-\varepsilon)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi} \\
&+ \frac{(\varepsilon-1)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&- \frac{(\varepsilon-1)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{(\varepsilon+1)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi} \\
&- \frac{(\varepsilon+1)}{2N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}
\end{aligned}
$$

$$
\begin{aligned}
R(m) = (&\frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi} \\
&+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi} \\
&- \frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi})
\end{aligned}
$$

$$R(m) = (\frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi}$$
$$+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}$$
$$- \frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}$$
$$+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi})$$

Let

$$R_a(m) = \frac{1}{N}(abs(\varepsilon \times IFFT(\Re(x(k)) \times e^{\Delta\phi}$$
$$+ IFFT(\Re(x(k)) \times e^{-\Delta\phi})$$
$$+ abs(-\varepsilon \times IFFT(\Im(x(k)) \times e^{-\Delta\phi}$$
$$+ IFFT(\Im(x(k)) \times e^{\Delta\phi})$$

Where abs refers to absolute value.

$$R_a(m) = (\frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi}$$
$$+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Re(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}$$
$$+ \frac{\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}$$
$$+ \frac{1}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{\Delta\phi})$$

Then $R_y(m) = R(m) - R_a(m)$ will result

$$R_y(m) = (R(m) - R_a(m)) = -\frac{2\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi mk/2N} e^{-\Delta\phi}$$

At the receiver let us consider time shift version of $R_y(m)$ i.e. R(m-t)

$$R_y(m-t) = -\frac{2\varepsilon}{N} \times \sum_{k=-N/2}^{N/2-1} \Im(X(k)) e^{j2\pi(m-t)k/2N} e^{-\Delta\phi}$$

If in-phase mismatch doesn't exist i.e. $\phi = 0$

Then $FFT(R_y(m)) = FFT(R_y(m-t)) \times e^{j2\pi tk/2N}$ (4)

If in-phase mismatch exists and $\phi \neq 0$, then time shift invariant property of DFT does not hold any more and

$$FFT(R_y(m)) \neq FFT(R_y(m-t)) \times e^{j2\pi tk/2N}$$

Thus, the in-phase mismatch should be compensated exactly before operation FFT on R(m) and R(m-t). we introduce a trial in-phase mismatch estimation value $\widehat{\phi}$ and compensate the effects of in-phase mismatch in *R*$_y$*(m)* and *R*$_y$*(m-t)*, by forming the compensated signals $R_y(m)e^{-\Delta\phi}$ and $R_y(m-t)e^{j2\pi tk/2N}e^{-\Delta\phi}$ In the frequency domain, the compensated signals can be expressed as follows:

$$r_y(l,\hat{\phi}) = \sum_{m=0}^{N-1}\{R_y(m)e^{-\Delta\phi}\}e^{-j2\pi ml/N}$$

$$r_{iy}(l,\hat{\phi}) = \sum_{m=0}^{N-1}\{R_y(m-t)e^{-\Delta\phi}e^{j2\pi tk/2N}\}e^{-j2\pi ml/N}$$

As given in equation (4) if there is no in-phase mismatch i.e. $\phi = \hat{\phi}$, then $r_y(l,\hat{\phi}) - r_{iy}(l,\hat{\phi}) = 0$, if in-phase mismatch exists i.e. $\phi \neq \hat{\phi}$ then $r_y(l,\hat{\phi}) - r_{iy}(l,\hat{\phi}) = 0$, using $r_y(l,\hat{\phi})$ & $r_{iy}(l,\hat{\phi})e^{j2\pi ml/N}$ we define least squares cost function

$$\varsigma(\hat{\phi}) = \sum_{m=0}^{N-1}| r_y(l,\hat{\phi}) - r_{iy}(l,\hat{\phi})e^{j2\pi ml/N} |^2$$

Therefore in-phase mismatch is estimated and compensated searching for $\hat{\phi}$ that minimizes the value of cost function given in equation (5). Thus, the in-phase mismatch estimate is given as

$$\hat{\phi} = \arg\min_{\hat{\phi}} \varsigma(\hat{\phi}) = \arg\min_{\hat{\phi}} \sum_{m=0}^{N-1}| r_y(l,\hat{\phi}) - r_{iy}(l,\hat{\phi})e^{j2\pi ml/N} |^2 \qquad (5)$$

If quadrature-phase mismatch (ε)=0 then $\alpha = \Re(\alpha)$, if quadrature-phase mismatch exists i.e. $(\varepsilon) \neq 0$ then $\alpha \neq \Re(\alpha)$ using the relation, we define least squares cost function

$$\xi(\hat{\varepsilon}) = |\,\alpha - \Re(\alpha)\,|^2$$

Therefore quadrature-phase mismatch is estimated and compensated searching for $\hat{\varepsilon}$ that minimizes the value of cost function given in equation (6). Thus, the quadrature-phase mismatch estimate is given as

$$\hat{\varepsilon} = \arg\min_{\hat{\varepsilon}} \xi(\hat{\varepsilon}) = \arg\min_{\hat{\varepsilon}} |\,\alpha - \Re(\alpha)\,|^2 \qquad - \qquad (6)$$

3. Simulation Results

A typical OFDM system is simulated to evaluate the performance of the compensation scheme for receiver I/Q imbalance in AWGN channel. The performance comparison is made with an ideal system with no front-end distortion and with a system with no compensation algorithm included. The parameters employed in the simulation are as follows: OFDM symbol length of N=64 , cyclic prefix of CP=16 . The step size of the adaptive equalizer is kept at 0.2.

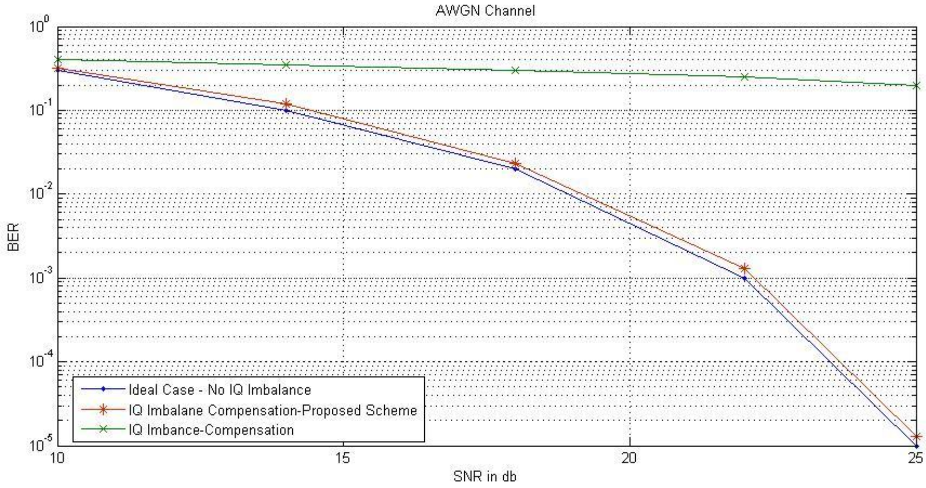

Figure 1 BER Vs SNR comparison for OFDM system with no IQ imbalance, IQ imbalance with compensation using proposed scheme and without compensation in AWGN channel (non fading).

We consider IQ amplitude imbalance of $\varepsilon = 5\%$ and phase imbalance of $\phi = 5^\circ$, at the receiver. Fig.1shows the performance curves obtained for BER versus SNR for an uncoded 64QAM OFDM system. With no compensation scheme in place, the OFDM system is unusable. Even for the case when there is only transmitter and receiver IQ imbalance and no CFO, the BER is very high. For the case with the proposed compensation scheme employed, the curves are very close to the ideal situation with no front-end distortion.

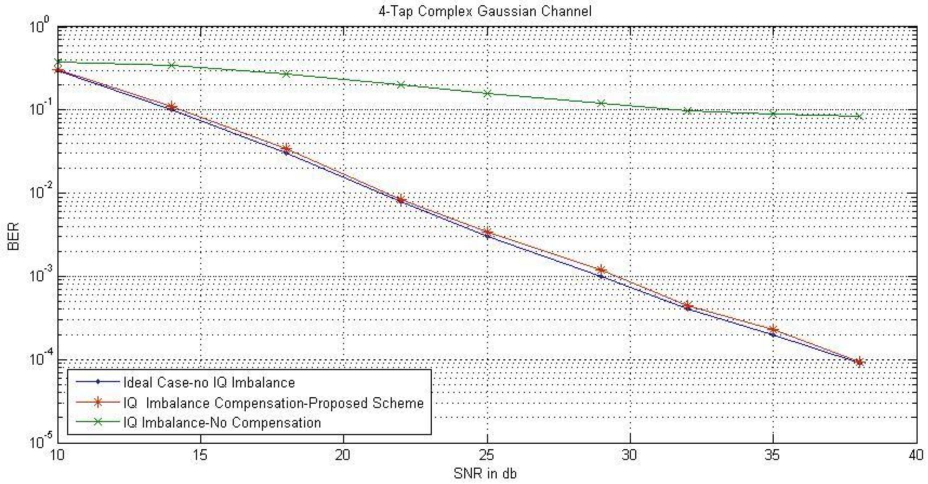

Figure 2 BER Vs SNR comparison for OFDM system with no IQ imbalance, IQ imbalance with compensation using proposed scheme and without compensation in 4-Tap Complex Gaussian channel (fading).

The compensation performance depends on how accurately the adaptive equalizer coefficients can converge to the ideal values. The design of zero-IF receivers typically yields an IQ imbalance on the order of [19]. The performance curves clearly demonstrate that for such IQ imbalance values compensation is necessary to enable a high data rate communication. Moreover, very large IQ imbalance values can be corrected just as easily. Thus, the presented IQ mitigation allows to greatly relax the zero-IF design specifications.

Conclusion

The I/Q imbalance in radio frequency (RF) impairments in direct conversion architecture based transceivers results in severe performance degradation. In this paper, we proposed an efficient scheme to compensate IQ imbalance in OFDM receivers. Simulation results confirmed that the proposed scheme performance is very near to ideal case (no IQ imbalance).

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