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JOINT COMPENSATION OF OFDM TRANSMITTER AND RECEIVER IQ IMBALANCE IN THE PRESENCE OF CARRIER FREQUENCY OFFSET

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ABSTRACT

Zero-IF based OFDM transmitters and receivers are gaining a lot of interest because of their potential to enable low-cost, low-power and less bulky terminals. However these systems suffer from In-phase/Quadrature-phase (IQ) imbalances in the front-end analog processing which may have a huge impact on the performance. We also consider the case where the local oscillator suffers from carrier frequency offset. As OFDM is very sensitive to the carrier frequency offset, this distortion needs to be taken into account in the derivation and analysis of any IQ imbalance estimation/compensation scheme. In this paper the effect of both transmitter and receiver IQ imbalance under carrier frequency offset in an OFDM system is studied and algorithms are developed to compensate for such distortions in the digital domain.

1. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is a popular, standardized technique for broadband wireless systems: it is used for Wireless LAN [1], Fixed Broadband Wireless Access [2], Digital Video & Audio Broadcasting [3], etc.

Recently, a lot of effort is spent in developing integrated, cost and power efficient OFDM receivers. The Zero-IF architecture (or Direct-Conversion architecture) is an attractive candidate. As the name suggests, the Zero-IF architecture converts the RF signal directly to baseband or vice-versa without any Intermediate Frequencies (IF). The Zero-IF architecture performs In-phase/Quadrature-phase (IQ) modulation and demodulation in the analog domain. As a result the matching between the analog I and Q paths and their components are imperfect. This leads to IQ imbalance distortion which significantly degrades the signal quality.

Rather than decreasing IQ imbalance by increasing the design time and the component cost, IQ imbalance can also be tolerated and then compensated digitally. Along with IQ imbalance, OFDM systems are also very sensitive to Carrier Frequency Offset (CFO). The performance degradation due to receiver IQ imbalance and CFO on OFDM systems have been investigated in [4] and [5]. Several compensation algorithms considering either only receiver IQ imbalance or transmitter IQ imbalance individually have been developed in [6],[7] and [8]. Recently, joint transmitter and receiver IQ imbalance compensation algorithms have been proposed in [9] and [10].

However, all the above mentioned algorithms ignore the problem of joint IQ imbalance at both transmitter and receiver along with CFO estimation and compensation. While the combination of CFO with receiver IQ imbalance has been studied in [11], to the best of our knowledge, no general solution has been proposed so far for the complete problem combining CFO with both transmitter and receiver IQ imbalance.

This report is organized as follows. In section II, we describe the impact of IQ imbalance and CFO in OFDM systems. In section III, the basics of a suitable transmitter and receiver IQ imbalance and CFO compensation scheme are explained. Section IV presents the various compensation algorithms. Our simulation results are shown in section V and finally the conclusion is given in section VI.

2. IQ IMBALANCE AND CFO MODEL

We analyze the effect of IQ imbalance in time and frequency domain. Frequency domain signals are underscored, while time domain signals are not. Signals are indicated in bold and scalar parameters in normal font. Superscripts $^*, ^T, ^H$ represent conjugate,transpose and Hermitian respectively.

In general, IQ imbalance (both transmitter and receiver imbalance) can be characterized by 2 parameters: an amplitude mismatch ε between the I and Q branch and a phase orthogonality mismatch $\Delta\phi$.

The complex baseband equation for IQ imbalance effect on an ideal time domain signal vector $\mathbf{r}$ is given by [12]:

$$\mathbf{r_{iq}} = (cos\Delta\phi + j\varepsilon sin\Delta\phi).\mathbf{r} + (\varepsilon cos\Delta\phi - jsin\Delta\phi).\mathbf{r}^* \quad (1)$$

$$\mathbf{r_{iq}} = \alpha.\mathbf{r} + \beta.\mathbf{r}^* \quad (2)$$

with $()^*$ the complex conjugate and

$$\alpha = cos\Delta\phi + j\varepsilon sin\Delta\phi \quad (3)$$

$$\beta = \varepsilon cos\Delta\phi - jsin\Delta\phi \quad (4)$$

If no IQ imbalance is present, then $\varepsilon = \Delta\phi = 0$ and thus $\alpha = 1$ and $\beta = 0$ and then (2) reduces to $\mathbf{r_{iq}} = \mathbf{r}$.

Parameters (α, β) are used for analytical derivations, as they form a more tractable mathematical description of the IQ imbalance. For simulations, the more physically relevant parameters $(\varepsilon, \Delta\phi)$ will be used.

The IQ imbalance impact on the frequency domain signal vector $\underline{\mathbf{r}} = DFT\{\mathbf{r}\}$ is:

$$\begin{aligned}\underline{\mathbf{r}}_{\mathbf{iq}} &= DFT\{\mathbf{r_{iq}}\} \\ &= DFT\{\alpha.IDFT(\underline{\mathbf{r}}) + \beta.[IDFT(\underline{\mathbf{r}})]^*\} \\ \underline{\mathbf{r}}_{\mathbf{iq}} &= \alpha.\underline{\mathbf{r}} + \beta.\underline{\mathbf{r}}_{\mathbf{m}}{}^*\end{aligned} \quad (5)$$

Here $()_m$ denotes the mirroring operation in which the vector indices are reversed, such that $\underline{\mathbf{r}}_m[l] = \underline{\mathbf{r}}[l_m]$

$$\text{where} \begin{cases} l_m = 2 + N - l \ for \ l = 2..N \\ l_m = l \ for \ l = 1 \end{cases}$$

i.e., l_m is the mirror carrier of l and N is the number of carriers in an OFDM symbol.

When the impact of both transmitter and receiver IQ imbalance is considered, equations (2) and (5) change as follows:

$$\begin{aligned} \mathbf{r_{Tiq\text{-}Riq}} &= \alpha_r.(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*) + \beta_r(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)^* \\ &= (\alpha_r.\alpha_t + \beta_r.\beta_t^*).\mathbf{r} + (\alpha_r.\beta_t + \beta_r.\alpha_t^*).\mathbf{r}^* \\ \underline{\mathbf{r}}_{\mathbf{Tiq\text{-}Riq}} &= (\alpha_r.\alpha_t + \beta_r.\beta_t^*).\underline{\mathbf{r}} + (\alpha_r.\beta_t + \beta_r.\alpha_t^*).\underline{\mathbf{r}}_{\mathbf{m}}^* \end{aligned} \quad (6)$$

where $(\alpha_t, \beta_t), (\alpha_r, \beta_r)$ represent the transmitter and receiver IQ imbalance respectively.

If the transmission channel is frequency selective, it introduces a filtering that should be included in formulas (6). If the channel impulse response is shorter than the OFDM cyclic prefix, which is a standard assumption, formula (6) is changed to:

$$\begin{aligned} \mathbf{r_{Tiq\text{-}c\text{-}Riq}} =& \alpha_r.IDFT(\underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)) \\ &+ \beta_r(IDFT(\underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)))^* \\ \underline{\mathbf{r}}_{\mathbf{Tiq\text{-}c\text{-}Riq}} =& (\alpha_r.\alpha_t.\underline{\mathbf{c}} + \beta_r.\beta_t^*.\underline{\mathbf{c}}_{\mathbf{m}}^*) \star \underline{\mathbf{r}} \\ &+ (\alpha_r.\beta_t.\underline{\mathbf{c}} + \beta_r.\alpha_t^*.\underline{\mathbf{c}}_{\mathbf{m}}^*) \star \underline{\mathbf{r}}_{\mathbf{m}}^* \end{aligned} \quad (7)$$

where '$\star$' denotes component-wise vector multiplication and $\underline{\mathbf{c}}$ represents the channel's frequency response.

Equation (7) shows that due to transmitter and receiver IQ imbalance power leaks from the signal on the mirror carrier ($\underline{\mathbf{r}}_{\mathbf{m}}^*$) to the carrier under consideration ($\underline{\mathbf{r}}$) and thus causes Inter-Carrier-Interference (ICI). As OFDM is very sensitive to ICI, IQ imbalance causes severe performance degradation.

Since OFDM is very sensitive to CFO, this distortion also needs to be taken into account. CFO occurs when there is a frequency deviation between the sine wave produced by the receiver local oscillator and the transmitter local oscillator.

In [11], it was shown that when CFO is present together with receiver IQ imbalance, the resulting baseband signal can be written as:

$$\mathbf{r_{cfo\text{-}Riq}} = \alpha_r.\mathbf{r} \star e^{j2\pi\Delta f.\mathbf{t}} + \beta_r.\mathbf{r}^* \star e^{-j2\pi\Delta f.\mathbf{t}} \quad (8)$$

where Δf is CFO in the system and $\mathbf{t}$ is the time vector.

For the case involving CFO as well as transmitter and receiver IQ imbalance, the above equation changes to:

$$\begin{aligned} \mathbf{r_{Tiq\text{-}cfo\text{-}Riq}} =& \alpha_r.(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*) \star e^{j2\pi\Delta f.\mathbf{t}} \\ &+ \beta_r.(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)^* \star e^{-j2\pi\Delta f.\mathbf{t}} \\ =& (\alpha_r.\alpha_t.\mathbf{r} + \alpha_r.\beta_t.\mathbf{r}^*) \star e^{j2\pi\Delta f.\mathbf{t}} \\ &+ (\beta_r.\alpha_t^*.\mathbf{r}^* + \beta_r.\beta_t^*.\mathbf{r}) \star e^{-j2\pi\Delta f.\mathbf{t}} \end{aligned} \quad (9)$$

Finally, if the channel is frequency selective, with frequency response $\underline{\mathbf{c}}$, formula (9) can be generalized to:

$$\begin{aligned} \mathbf{r_{Tiq\text{-}c\text{-}cfo\text{-}Riq}} =& \alpha_r.(IDFT(\underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*))) \star e^{j2\pi\Delta f.\mathbf{t}} \\ &+ \beta_r.(IDFT(\underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)))^* \star e^{-j2\pi\Delta f.\mathbf{t}} \end{aligned} \quad (10)$$

The joint effect of both transmitter and receiver IQ imbalance along with CFO results in a severe ICI. Thus a digital compensation is required which limits the achievable operating SNR at the receiver and the achievable data rates.

3. IQ IMBALANCE AND CFO COMPENSATION

Following the joint model from the previous section, formula (10), we know that the received OFDM symbol in time domain is given as:

$$\mathbf{r_{Tiq\text{-}c\text{-}cfo\text{-}Riq}} = \alpha_r.\mathbf{p} \star e^{j2\pi\Delta f.\mathbf{t}} + \beta_r.\mathbf{p}^* \star e^{-j2\pi\Delta f.\mathbf{t}} \quad (11)$$

where $\mathbf{p} = IDFT(\underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*))$. Equation (11) explicitly shows only the CFO and the receiver IQ imbalance. The channel filtering and the distortion due to transmitter IQ imbalance are hidden in the definition of $\mathbf{p}$ and so not considered for the time being.

We now assume that the CFO can be estimated accurately in the OFDM system. In practice, several CFO estimation schemes indeed exist that are found to be sufficiently robust against the IQ imbalance [11] and [13]. Thus, given a good estimate of Δf, we first perform an element wise multiplication of the received distorted symbol in equation (11) with the estimated negative frequency offset $e^{-j2\pi\Delta f.\mathbf{t}}$. This results in a vector $\mathbf{r_1}$ as follows:

$$\begin{aligned} \mathbf{r_1} &= \mathbf{r_{Tiq\text{-}c\text{-}cfo\text{-}Riq}} \star e^{-j2\pi\Delta f.\mathbf{t}} \\ &= \alpha_r.\mathbf{p} + \beta_r.\mathbf{p}^* \star e^{-2j2\pi\Delta f.\mathbf{t}} \\ &= \alpha_r.\mathbf{p} + \beta_r.\mathbf{q} \end{aligned} \quad (12)$$

Similarly, we also perform an element wise multiplication of the complex conjugate of the received signal with the negative frequency offset $e^{-j2\pi\Delta f.\mathbf{t}}$ resulting in a vector $\mathbf{r_2}$ as follows:

$$\begin{aligned} \mathbf{r_2} &= (\mathbf{r_{Tiq\text{-}c\text{-}cfo\text{-}Riq}})^* \star e^{-j2\pi\Delta f.\mathbf{t}} \\ &= (\alpha_r^*.\mathbf{p}^* \star e^{-j2\pi\Delta f.\mathbf{t}} + \beta_r^*.\mathbf{p} \star e^{j2\pi\Delta f.\mathbf{t}}) \star e^{-j2\pi\Delta f.\mathbf{t}} \\ &= \beta_r^*.\mathbf{p} + \alpha_r^*.\mathbf{p}^* \star e^{-2j2\pi\Delta f.\mathbf{t}} \\ &= \beta_r^*.\mathbf{p} + \alpha_r^*.\mathbf{q} \end{aligned} \quad (13)$$

Both the signals $\mathbf{r_1}$ and $\mathbf{r_2}$ consist of two contributions '$\mathbf{p}$' and '$\mathbf{q}$' scaled by different weighing factors. '$\mathbf{p}$' is called the desired signal and '$\mathbf{q}$' the undesired signal. This is because in frequency domain, the former gives rise to the desired signal, while the latter yields a mirror image and causes ICI (because of the complex conjugate), subject to leakage caused by the exponential term ($e^{-2j2\pi\Delta f.\mathbf{t}}$).

Transforming equations (12) and (13) to the frequency domain, we obtain:

$$\begin{bmatrix}\underline{\mathbf{r_1}} & \underline{\mathbf{r_2}}\end{bmatrix} = \begin{bmatrix}\underline{\mathbf{p}} & \underline{\mathbf{q}}\end{bmatrix} \cdot \begin{bmatrix}\alpha_r & {\beta_r}^* \\ \beta_r & {\alpha_r}^*\end{bmatrix} \tag{14}$$

Equation (14) can be written more explicitly for each component (frequency bin) 'l' of the OFDM symbol:

$$\begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l]\end{bmatrix} = \begin{bmatrix}\underline{\mathbf{p}}[l] & \underline{\mathbf{q}}[l]\end{bmatrix} \cdot \begin{bmatrix}\alpha_r & {\beta_r}^* \\ \beta_r & {\alpha_r}^*\end{bmatrix} \tag{15}$$

From this it follows that the desired signal $\underline{\mathbf{p}}[l]$ can be obtained by taking an appropriate linear combination of $\underline{\mathbf{r_1}}[l]$ and $\underline{\mathbf{r_2}}[l]$, i.e.

$$\begin{aligned}
\begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l]\end{bmatrix} \cdot \begin{bmatrix}\kappa \\ \lambda\end{bmatrix} &= \begin{bmatrix}\underline{\mathbf{p}}[l] & \underline{\mathbf{q}}[l]\end{bmatrix} \cdot \begin{bmatrix}\alpha_r & {\beta_r}^* \\ \beta_r & {\alpha_r}^*\end{bmatrix} \cdot \begin{bmatrix}\kappa \\ \lambda\end{bmatrix} \\
\begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l]\end{bmatrix} \cdot \begin{bmatrix}\kappa \\ \lambda\end{bmatrix} &= \begin{bmatrix}\underline{\mathbf{p}}[l] & \underline{\mathbf{q}}[l]\end{bmatrix} \cdot \begin{bmatrix}1 \\ 0\end{bmatrix} \\
\Rightarrow \underline{\mathbf{p}}[l] &= \begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l]\end{bmatrix} \cdot \begin{bmatrix}\kappa \\ \lambda\end{bmatrix}
\end{aligned} \tag{16}$$

From the definition of $\mathbf{p}$ it follows that $\underline{\mathbf{p}} = \underline{\mathbf{c}} \star DFT(\alpha_t.\mathbf{r} + \beta_t.\mathbf{r}^*)$, hence

$$\begin{aligned}
\underline{\mathbf{p}}[l] &= \underline{\mathbf{c}}[l].(\alpha_t.\underline{\mathbf{r}}[l] + \beta_t.\underline{\mathbf{r_m}}^*[l]) \\
\Rightarrow \underline{\mathbf{p}}[l] &= \underline{\mathbf{c}}[l].\underline{\widetilde{\mathbf{p}}}[l]
\end{aligned} \tag{17}$$

where $\underline{\widetilde{\mathbf{p}}}[l] = (\alpha_t.\underline{\mathbf{r}}[l] + \beta_t.\underline{\mathbf{r_m}}^*[l])$

By combining equations (16) and (17) we obtain:

$$\underline{\widetilde{\mathbf{p}}}[l] = \begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l]\end{bmatrix} \cdot \begin{bmatrix}\kappa[l] \\ \lambda[l]\end{bmatrix} \tag{18}$$

which means that by taking an appropriate (now frequency bin dependent) linear combination of $\underline{\mathbf{r_1}}[l]$ and $\underline{\mathbf{r_2}}[l]$, signal $\underline{\widetilde{\mathbf{p}}}[l]$ is obtained which is only distorted by transmitter IQ imbalance, i.e

$$\underline{\widetilde{\mathbf{p}}}[l] = \alpha_t.\underline{\mathbf{r}}[l] + \beta_t.\underline{\mathbf{r_m}}^*[l] = \alpha_t.\underline{\mathbf{r}}[l] + \beta_t.\underline{\mathbf{r}}^*[l_m] \tag{19}$$

where l_m is the mirror index of l.

In a similar fashion, an appropriate linear combination of $\underline{\mathbf{r_1}}[l_m]$ and $\underline{\mathbf{r_2}}[l_m]$ can then be found, i.e

$$\underline{\widetilde{\mathbf{p}}}[l_m] = \begin{bmatrix}\underline{\mathbf{r_1}}[l_m] & \underline{\mathbf{r_2}}[l_m]\end{bmatrix} \cdot \begin{bmatrix}\tau[l_m] \\ \upsilon[l_m]\end{bmatrix} \tag{20}$$

leading to a signal $\underline{\widetilde{\mathbf{p}}}[l_m]$ equal to

$$\begin{aligned}
\underline{\widetilde{\mathbf{p}}}[l_m] &= \alpha_t.\underline{\mathbf{r}}[l_m] + \beta_t.\underline{\mathbf{r_m}}^*[l_m] \\
&= \alpha_t.\underline{\mathbf{r}}[l_m] + \beta_t.\underline{\mathbf{r}}^*[l] \\
\underline{\widetilde{\mathbf{p}}}^*[l_m] &= \beta_t^*.\underline{\mathbf{r}}[l] + \alpha_t^*.\underline{\mathbf{r}}^*[l_m]
\end{aligned} \tag{21}$$

Formula (19) and (21) can be combined into

$$\begin{bmatrix}\underline{\widetilde{\mathbf{p}}}[l] & \underline{\widetilde{\mathbf{p}}}^*[l_m]\end{bmatrix} = \begin{bmatrix}\underline{\mathbf{r}}[l] & \underline{\mathbf{r}}^*[l_m]\end{bmatrix} \cdot \begin{bmatrix}\alpha_t & \beta_t^* \\ \beta_t & \alpha_t^*\end{bmatrix} \tag{22}$$

From this it follows that the transmitted signal $\underline{\mathbf{r}}[l]$ can be obtained by taking an appropriate linear combination of $\underline{\widetilde{\mathbf{p}}}[l]$ and $\underline{\widetilde{\mathbf{p}}}^*[l_m]$. By substituting the formula for $\underline{\widetilde{\mathbf{p}}}[l]$ and $\underline{\widetilde{\mathbf{p}}}^*[l_m]$ in the formulas (18) and (20), $\underline{\mathbf{r}}[l]$ can be obtained directly as:

$$\underline{\mathbf{r}}[l] = \begin{bmatrix}\underline{\mathbf{r_1}}[l] & \underline{\mathbf{r_2}}[l] & \underline{\mathbf{r_1}}^*[l_m] & \underline{\mathbf{r_2}}^*[l_m]\end{bmatrix} \cdot \begin{bmatrix}\chi[l] \\ \psi[l] \\ \gamma[l] \\ \rho[l]\end{bmatrix} \tag{23}$$

where the weights $\chi[l], \psi[l], \gamma[l], \rho[l]$ are derived from all the previously used weights. This formula demonstrates that a receiver structure can be designed that exactly compensates for the transmitter and the receiver IQ imbalances, the CFO and the channel effect. The coefficients $\chi[l], \psi[l], \gamma[l], \rho[l]$ can be computed from the $\alpha_t, \beta_t, \alpha_r, \beta_r, \Delta f$ and $\underline{\mathbf{c}}$, if these are available. In the next section various algorithms are explained which compensate the joint transmitter and receiver IQ imbalance along with CFO by the initialization of $\chi[l], \psi[l], \gamma[l], \rho[l]$.

4. COMPENSATION ALGORITHMS

4.1 Least-Square Compensation

The Least Square (LS) estimate of the coefficient vector $\chi[l], \psi[l], \gamma[l], \rho[l]$ can be computed as [14]:

$$\begin{bmatrix}\chi[l] \\ \psi[l] \\ \gamma[l] \\ \rho[l]\end{bmatrix} = (\mathbf{A}[l]^H \mathbf{A}[l])^{-1} \mathbf{A}[l]^H \underline{\mathbf{d}}[l] \tag{24}$$

where $\mathbf{A}[l]$ is the data matrix given for $i = K$ training symbols

$$\mathbf{A}[l] = \begin{bmatrix}
\underline{\mathbf{r_1}}[l]^{(1)} & \underline{\mathbf{r_2}}[l]^{(1)} & \underline{\mathbf{r_1}}^*[l_m]^{(1)} & \underline{\mathbf{r_2}}^*[l_m]^{(1)} \\
\underline{\mathbf{r_1}}[l]^{(2)} & \underline{\mathbf{r_2}}[l]^{(2)} & \underline{\mathbf{r_1}}^*[l_m]^{(2)} & \underline{\mathbf{r_2}}^*[l_m]^{(2)} \\
\vdots & \vdots & \vdots & \vdots \\
\underline{\mathbf{r_1}}[l]^{(K)} & \underline{\mathbf{r_2}}[l]^{(K)} & \underline{\mathbf{r_1}}^*[l_m]^{(K)} & \underline{\mathbf{r_2}}^*[l_m]^{(K)}
\end{bmatrix}$$

and $\mathbf{d}[l]$ is the desired data vector containing K transmitted training symbols

$$\mathbf{d}[l] = \begin{bmatrix}\underline{\mathbf{d}}[l]^{(1)} & \underline{\mathbf{d}}[l]^{(2)} & \ldots & \underline{\mathbf{d}}[l]^{(K)}\end{bmatrix}^T$$

Regularization can be used when it is required to combat ill-conditioning in the data $\mathbf{A}[l]$. Thus depending on the number of training symbols, K realizations of the above equation can be collected to perform the LS estimation of the coefficient matrix. The resulting coefficient vector can then be substituted in formula (23) to obtain the transmitted OFDM symbol $\underline{\mathbf{r}}$.

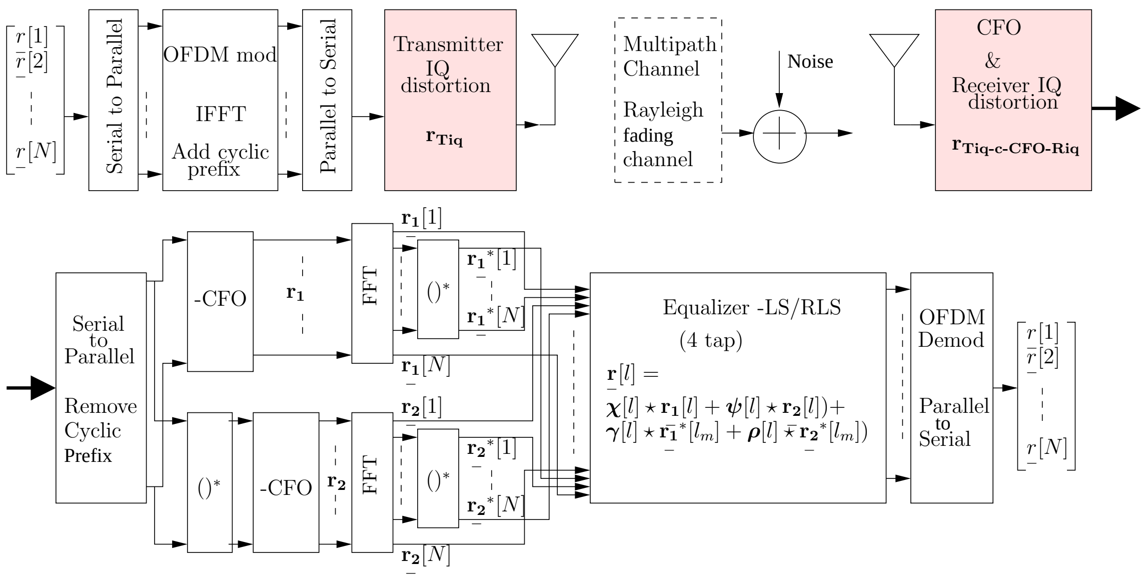


Figure 1: OFDM system with post-FFT compensation of transmitter and receiver IQ imbalance with CFO.

4.2 Adaptive Equalization

We propose a training based system with 4 tap adaptive equalizer to compensate transmitter and receiver IQ imbalance with CFO. The equalizer taps $\mathbf{w}[l]$ can be updated adaptively to the coefficient vector $\chi[l], \psi[l], \gamma[l], \rho[l]$ by any of the standard adaptive algorithms. For our simulations we have used the RLS scheme for its superior convergence properties.

The RLS algorithm to compute the coefficient vector is listed in Algorithm 1 [14]. To better illustrate the update equations, we use the time (or iteration) index i. As a result let $\mathbf{w}[l]^{(i)}$ represent the equalization vectors at time instant i. δ is the regularization factor which is a small positive constant.

Algorithm 1 RLS direct equalization with CFO
For all the carriers in OFDM symbol l=1..N compute
Initialize the algorithm by setting

$$\mathbf{w}[l]^{(i=0)} = 0_{4\times 1}$$
$$\mathbf{P}^{(i=0)} = \delta^{-1}\mathbf{I}_{4\times 4}$$

For each iteration $i = 1\ldots K$ compute

$$\mathbf{u}[l]^{(i)} = \begin{bmatrix} \underline{\mathbf{r_1}}[l]^{(i)} & \underline{\mathbf{r_2}}[l]^{(i)} & \underline{\mathbf{r_1}}^*[l_m]^{(i)} & \underline{\mathbf{r_2}}^*[l_m]^{(i)} \end{bmatrix}^T$$
$$\xi^{(i)} = \mathbf{d}[l]^{(i)} - \mathbf{w}[l]^{H(i-1)}\mathbf{u}[l]^{(i)}$$
$$\mathbf{w}[l]^{(i)} = \mathbf{w}[l]^{(i-1)} + \frac{\mathbf{P}^{(i-1)}\mathbf{u}[l]^{(i)}}{1+\mathbf{u}[l]^{H(i)}\mathbf{P}^{(i-1)}\mathbf{u}[l]^{(i)}}\xi^{*(i)}$$
$$\mathbf{P}^{(i)} = \mathbf{P}^{(i-1)} - \frac{\mathbf{P}^{(i-1)}\mathbf{u}[l]^{(i)}}{1+\mathbf{u}[l]^{H(i)}\mathbf{P}^{(i-1)}\mathbf{u}[l]^{(i)}}\mathbf{u}^{H(i)}\mathbf{P}^{(i-1)}$$

At the end of the training the weights $\mathbf{w}[l]$ correspond to the coefficient vector $\chi[l], \psi[l], \gamma[l], \rho[l]$. The weights $\mathbf{w}[l]$ are then substituted in formula (23) to obtain the transmitted symbol $\underline{\mathbf{r}}[l]$

5. SIMULATION RESULTS

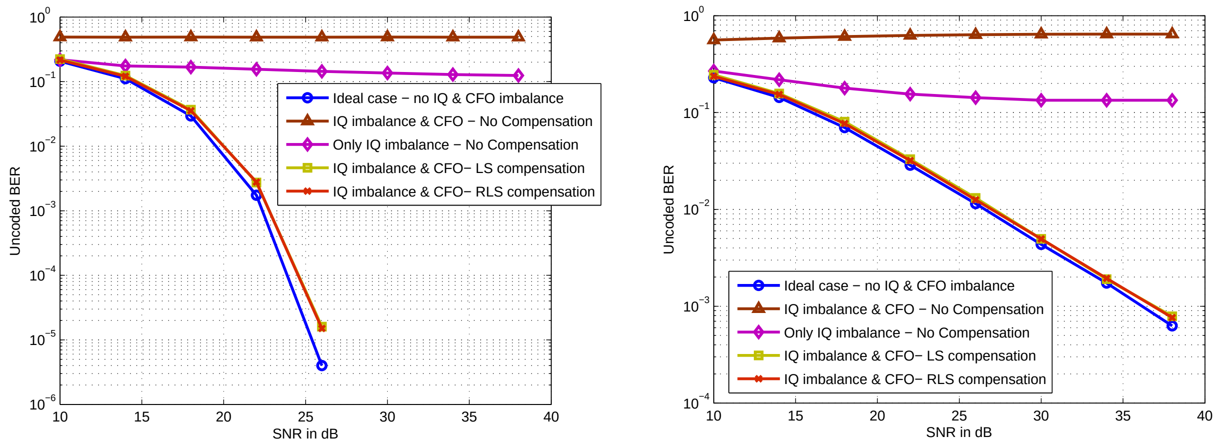
A typical OFDM system (similar to IEEE 802.11a) is simulated to evaluate the performance of the compensation scheme for transmitter and receiver IQ imbalance under CFO. The performance comparison is made with an ideal system with no front-end distortion and with a system with no compensation algorithm included. An end-to-end OFDM system with the compensation scheme is shown in **Figure 1**. The equalizer has four taps and the tap values are calculated using one of the algorithms proposed in the previous section.

It is noted that the receiver structure shown in Figure 1 generalizes earlier structures for specific subproblems. Reference [8] and [11] for instance apply to the compensation of CFO with either transmitter or with receiver IQ imbalance but not both. In this case the output of any one FFT branch can be taken (instead of both outputs) as the compensation can then be obtained with a two tap adaptive equalizer.

The parameters used in the simulation are: OFDM symbol length of $N = 64$, cyclic prefix of $CP = 16$. There are two different channel profiles: 1) an additive white Gaussian noise (AWGN) channel with a single tap unity gain and 2) a multipath channel with 4 taps where the taps are chosen independently with complex Gaussian distribution. Every channel realization is independent of the previous one and the BER results depicted are from averaging the BER curves over several independent channels.

We consider IQ amplitude imbalance $\varepsilon = 5\%$ and phase imbalance $\Delta\phi = 5^\circ$ at both transmitter and receiver. The CFO ($\zeta = \Delta f NT$) is assumed to be 0.32 where T is the sampling period. The IQ imbalance distortion taken are typical values achievable in practical integrated circuit implementations.

Figure 2 shows the performance curves (BER vs SNR) for uncoded 64QAM OFDM system. With no compensation scheme in place, the OFDM system is completely 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 compensation scheme employed, the curves are very close to the ideal situation with no front-end distortion.



(a) AWGN flat channel (non-fading)

(b) 4-tap complex Gaussian channel (fading)

Figure 2: BER vs SNR simulated for 64QAM constellation with training length of 15 OFDM symbols in LS and RLS solutions, transmitter phase imbalance of $\Delta\phi_t = 5°$, transmitter amplitude imbalance of $\varepsilon_t = 5\%$, receiver phase imbalance of $\Delta\phi_r = 5°$, receiver amplitude imbalance of $\varepsilon_t = 5\%$ and CFO of $\zeta = 0.32$ $(\zeta = \Delta f NT)$

The design of Zero-IF receivers typically yields an IQ imbalance on the order of $(\varepsilon, \Delta\phi) = (2-3\%, 2-3°)$ [15]. When CFO is also present, the system certainly cannot work without a suitable compensation technique. Moreover, very large IQ imbalance values can be corrected just as easily with this compensation scheme. Thus the presented IQ-CFO mitigation allows to greatly relax the Zero-IF design specifications.

6. CONCLUSION

In this report the joint effect of OFDM transmitter and receiver IQ imbalance under CFO is studied and algorithms are developed to compensate for such distortions in the digital domain. The algorithms provide a very efficient post-FFT equalization which leads to near ideal compensation.

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