EFFICIENT COMPENSATION OF FREQUENCY SELECTIVE TX AND RX IQ IMBALANCES IN OFDM SYSTEMS

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Abstract Radio frequency impairments such as in-phase/quadrature-phase (IQ) imbalances can result in a severe performance degradation in direct-conversion architectures. In this paper, a training based equalizer is developed to estimate and compensate the effects of these IQ imbalances along with channel distortions in an OFDM system. The proposed scheme provides a decoupling mechanism to estimate all the three frequency selective distortions in the communication system, namely the transmitter IQ imbalance, the receiver IQ imbalance and the channel dispersion. Once the IQ imbalance parameters are estimated, the proposed scheme then utilizes a standard one tap frequency domain equalizer to estimate and compensate the channel variations in the system. Simulation results show that the resulting calibrated equalizer requires a very small training overhead for an efficient, post-FFT equalization with performance very close to the ideal case.

1. Introduction

A direct-conversion based system is an attractive front-end radio architectural design for a communication engineer [1]. These systems are typically small in size and cheaper to implement. They also provide a very good flexibility in supporting growing number of wireless standards found in today's communication systems. However, the direct-conversion based architecture is generally very sensitive to any front-end component imperfections. These imperfections are unavoidable especially when cheaper components are used in the manufacturing process. The front-end imperfections can result in radio frequency (RF) impairments such as in-phase/quadrature-phase (IQ) imbalance, carrier

frequency offset (CFO), etc. The multi-carrier based communication systems such as OFDM [2] are found to be very sensitive to such RF impairments [3]- [4]. In this paper we study the effects of only IQ imbalance in an OFDM system. The RF impairments such as IQ imbalance can result in a severe performance degradation in an OFDM system, rendering the communication system useless, unless they are adequately compensated.

Recently several articles [3]- [9] have been published that address these effects of IQ imbalances in an OFDM system. In [5] and [6] efficient digital compensation schemes have been developed for the case of frequency independent IQ imbalance. [7] and [8] provide a joint compensation scheme in MIMO and SISO systems for three frequency selective distortions, namely the transmitter (Tx) IQ imbalance, the receiver (Rx) IQ imbalance and the channel dispersion. The joint compensation scheme provides a simple implementation alternative but the equalizer's training overhead is found to be large as the received signal has to be equalized at both the desired sub-carrier and its mirror component for any channel variation. Recently [9] proposed a compensation scheme based on special training symbols that decouples the frequency selective Rx IQ imbalance from the channel dispersion resulting in a reliable estimation with a small training overhead.

This paper is an extension of [8] and [9] as it targets to design a low training overhead equalizer for the general case of frequency selective Tx and Rx IQ imbalance along with channel dispersion for SISO systems. The estimation process is divided in two phases, the first phase is the calibration phase where the Tx and Rx IQ imbalance parameters are estimated. These parameters are only slowly time-varying components thus they do not have to be re-estimated frequently. In the second phase only the channel dispersion is estimated using a standard one tap equalizer. It is shown that the proposed equalizer provides an efficient and an effective compensation method for systems impaired with IQ imbalance distortions.

This paper is organized as follows: The input-output OFDM system model is presented in Section 2. Section 3 then explains the IQ imbalance compensation scheme. Computer simulations are shown in Section 4 and finally the conclusion is given in Section 5.

Notation: Vectors are indicated in bold and scalar parameters in normal font. Superscripts $\{\}^*, \{\}^T, \{\}^H$ represent complex conjugate, transpose and Hermitian respectively. $\mathbf{F}$ and $\mathbf{F}^{-1}$ represent the $N \times N$ discrete Fourier transform and its inverse. $\mathbf{I}_N$ is the $N \times N$ identity matrix and $0_{M\times N}$ is the $M \times N$ all zero matrix. Operators $\star$, . and

$\div$ denote linear convolution, component-wise vector multiplication and division respectively.

2. System model

We consider an OFDM transmission over frequency selective fading channels. We assume a single-input single-output (SISO) system, but the results can be easily extended to multiple-input multiple-output (MIMO) systems. Let $\mathbf{S}$ be an uncoded frequency domain OFDM symbol of size $(N \times 1)$. This symbol is transformed to the time domain by an inverse discrete Fourier transform (IDFT) operation. A cyclic prefix (CP) of length ν is then added to the head of each symbol. The resulting time domain baseband signal $\mathbf{s}$ is then given as:

$$\mathbf{s} = \mathbf{P_{CI}}\mathbf{F}^{-1}\mathbf{S} \tag{1}$$

where $\mathbf{P_{CI}}$ is the cyclic insertion matrix given by:

$$\mathbf{P_{CI}} = \left[\begin{array}{c|c} 0_{(\nu \times N-\nu)} & \mathbf{I}_\nu \\ \hline & \mathbf{I}_N \end{array}\right]$$

We represent frequency selective (FS) IQ imbalance resulting from Tx front-end components by two mismatched filters with frequency responses given as $\mathbf{H}_{ti} = \mathbf{F}\{\mathbf{h}_{ti}\}$ and $\mathbf{H}_{tq} = \mathbf{F}\{\mathbf{h}_{tq}\}$. The frequency independent (FI) IQ imbalance is represented by amplitude and phase mismatch g_t and ϕ_t between the two front-end branches. Following [10], the baseband signal $\mathbf{p}$ after front-end distortions can be given as:

$$\mathbf{p} = \mathbf{g}_{ta} \star \mathbf{s} + \mathbf{g}_{tb} \star \mathbf{s}^* \tag{2}$$

where

$$\mathbf{g}_{ta} = \mathbf{F}^{-1}\mathbf{G}_{ta} = \mathbf{F}^{-1}\left\{\frac{\left[\mathbf{H}_{ti} + g_t e^{-\jmath\phi_t}\mathbf{H}_{tq}\right]}{2}\right\}$$

$$\mathbf{g}_{tb} = \mathbf{F}^{-1}\mathbf{G}_{tb} = \mathbf{F}^{-1}\left\{\frac{\left[\mathbf{H}_{ti} - g_t e^{\jmath\phi_t}\mathbf{H}_{tq}\right]}{2}\right\}$$

Here $\mathbf{g}_{ta}$ and $\mathbf{g}_{tb}$ are mostly truncated to length L_t and then padded with $N - L_t$ zero elements. They represent the combined FI-FS Tx IQ imbalance. $\mathbf{G}_{ta}$ and $\mathbf{G}_{tb}$ are the frequency domain representations of $\mathbf{g}_{ta}$ and $\mathbf{g}_{tb}$ respectively.

Finally, an expression similar to equation (2) can be used to model IQ imbalance at the receiver. Let $\mathbf{z}$ represent the down-converted baseband complex signal after being distorted by combined FI-FS Rx IQ imbalance $\mathbf{g}_{ra}$ and $\mathbf{g}_{rb}$ of length L_r. Then $\mathbf{z}$ will be given as:

$$\mathbf{z} = \mathbf{g}_{ra} \star \mathbf{r} + \mathbf{g}_{rb} \star \mathbf{r}^* \tag{3}$$

where

$$\mathbf{r} = \mathbf{c} \star \mathbf{p} + \mathbf{n}$$

Here $\mathbf{c}$ is the baseband representation of the multipath channel of length L and $\mathbf{n}$ is the additive white Gaussian noise (AWGN). Equation (2) can be substituted in equation (3) leading to

$$\begin{aligned}\mathbf{z} =&(\mathbf{g}_{ra} \star \mathbf{c} \star \mathbf{g}_{ta} + \mathbf{g}_{rb} \star \mathbf{c}^* \star \mathbf{g}^*_{tb}) \star \mathbf{s} + \mathbf{g}_{ra} \star \mathbf{n} + \\ &(\mathbf{g}_{ra} \star \mathbf{c} \star \mathbf{g}_{tb} + \mathbf{g}_{rb} \star \mathbf{c}^* \star \mathbf{g}^*_{ta}) \star \mathbf{s}^* + \mathbf{g}_{rb} \star \mathbf{n}^* \\ =&\mathbf{d}_1 \star \mathbf{s} + \mathbf{d}_2 \star \mathbf{s}^* + \mathbf{g}_{ra} \star \mathbf{n} + \mathbf{g}_{rb} \star \mathbf{n}^*\end{aligned} \tag{4}$$

where $\mathbf{d}_1$ and $\mathbf{d}_2$ are the combined Tx IQ, channel and Rx IQ impulse responses of length $L_t + L + L_r - 2$. The down-converted received signal $\mathbf{z}$ is now separated from the CP at the receiver. This CP free received symbol is then transformed to the frequency domain by an discrete Fourier transform (DFT) operation. In this paper we consider the CP length ν to be sufficiently longer than both $\mathbf{d}_1$ and $\mathbf{d}_2$. The longer CP length results in a simple dot multiplication between the various frequency selective terms and the transmitted symbol in the frequency domain. This resulting signal $\mathbf{Z}$ can then be written as:

$$\begin{aligned}\mathbf{Z} =&\mathbf{F}\mathbf{P}_{\mathbf{CR}}\{\mathbf{z}\} \\ =&(\mathbf{G}_{ra}.\mathbf{G}_{ta}.\mathbf{C} + \mathbf{G}_{rb}.\mathbf{G}^*_{tb_m}.\mathbf{C}^*_m).\mathbf{S}^{(i)} + \mathbf{G}_{ra}.\widetilde{\mathbf{n}} \\ &+ (\mathbf{G}_{ra}.\mathbf{G}_{tb}.\mathbf{C} + \mathbf{G}_{rb}.\mathbf{G}^*_{ta_m}.\mathbf{C}^*_m).\mathbf{S}^{*(i)}_m + \mathbf{G}_{rb}.\widetilde{\mathbf{n}}^*_m\end{aligned} \tag{5}$$

where $\mathbf{P}_{\mathbf{CR}}$ is the cyclic removal matrix given as:

$$\mathbf{P}_{\mathbf{CR}} = \left[0_{(N\times\nu)} \;\middle|\; \mathbf{I}_N\right]$$

Here $\mathbf{G}_{ra}, \mathbf{G}_{rb}, \mathbf{C}$ and $\widetilde{\mathbf{n}}$ are the frequency domain representations of $\mathbf{g}_{ra}, \mathbf{g}_{rb}, \mathbf{c}$ and $\mathbf{n}$. The operator $()_m$ denotes the mirroring operation in which the vector indices are reversed, such that $\mathbf{S}_m[l] = \mathbf{S}[l_m]$ where $l_m = 2 + N - l$ for $l = 2 \ldots N$ and $l_m = l$ for $l = 1$. Equation (5) shows that due to IQ imbalance, the power leaks from the signal on the mirror carrier ($\mathbf{S}^*_m$) to the carrier under consideration ($\mathbf{S}$) and thus causes inter-carrier-interference (ICI). The ICI distortion due to Tx and Rx IQ imbalance results in a severe performance degradation and thus a compensation scheme is needed in the OFDM system. In the next section, we develop a digital compensation scheme for joint Tx and Rx IQ imbalance distortions in an OFDM system.

3. IQ imbalance compensation scheme

In the previous section, we obtained the equation for the received signal after being distorted by Tx IQ imbalance, channel and Rx IQ

imbalance respectively. In practice both Tx and Rx IQ imbalance distortions are relatively static and change very slowly when compared to the channel characteristics. Thus for a given channel dispersion characteristic, say $\mathbf{C}^{(1)}$, the received signal (5) can be rewritten as follows:

$$\begin{aligned}\mathbf{Z}^{(1)} =& \underbrace{(\mathbf{D}^{(1)} + \mathbf{N}_r.\mathbf{N}^*_{t_m}.\mathbf{D}^{*(1)}_m)}_{\mathbf{W}_1}.\mathbf{S} + \mathbf{G}_{ra}.\widetilde{\mathbf{n}} \\ &+ \underbrace{(\mathbf{N}_t.\mathbf{D}^{(1)} + \mathbf{N}_r.\mathbf{D}^{*(1)}_m)}_{\mathbf{W}_2}.\mathbf{S}^*_m + \mathbf{G}_{rb}.\widetilde{\mathbf{n}}^*_m\end{aligned} \tag{6}$$

where $\mathbf{D}^{(1)} = \mathbf{G}_{ra}.\mathbf{G}_{ta}.\mathbf{C}^{(1)}$, $\mathbf{N}_t = \frac{\mathbf{G}_{tb}}{\mathbf{G}_{ta}}$ and $\mathbf{N}_r = \frac{\mathbf{G}_{rb}}{\mathbf{G}^*_{ra_m}}$. Here $\mathbf{W}_1$ represents the scaling factor of the desired sub-carrier and $\mathbf{W}_2$ represents the amount of interference from the mirror sub-carrier. In equation (6), both $\mathbf{W}_1$ and $\mathbf{W}_2$ can be estimated by designing a training based two tap frequency domain equalizer (FEQ). The first input of the FEQ is applied to the OFDM training sequence $\mathbf{S}$ and and the second input to its mirror complex conjugate $\mathbf{S}^*_m$. Similarly a two tap FEQ can also be applied to the received signal $\mathbf{Z}^{(1)}$ and its mirror conjugate $\mathbf{Z}^{*(1)}_m$ in order to estimate the transmitted symbol $\mathbf{S}$. The FEQ can be designed based on a maximum-likelihood (ML), least-square (LS) or minimum mean-square-error (MMSE) criteria [8]. Once the FEQ coefficients are trained, the equalizer can then be applied on all the forthcoming OFDM data symbols as long as the channel characteristics remain the same $\mathbf{C}^{(1)}$. This is the principle behind the joint compensation scheme in [7] and [8].

In the case of a variation in the channel characteristics, say to $\mathbf{C}^{(2)}$, the received signal $\mathbf{Z}^{(2)}$ can now be written as:

$$\begin{aligned}\mathbf{Z}^{(2)} =& \underbrace{(\mathbf{D}^{(2)} + \mathbf{N}_r.\mathbf{N}^*_{t_m}.\mathbf{D}^{*(2)}_m)}_{\mathbf{W}_3}.\mathbf{S} + \mathbf{G}_{ra}.\widetilde{\mathbf{n}} \\ &+ \underbrace{(\mathbf{N}_t.\mathbf{D}^{(2)} + \mathbf{N}_r.\mathbf{D}^{*(2)}_m)}_{\mathbf{W}_4}.\mathbf{S}^*_m + \mathbf{G}_{rb}.\widetilde{\mathbf{n}}^*_m\end{aligned} \tag{7}$$

Here once again, the coefficients $\mathbf{W}_3$ and $\mathbf{W}_4$ can be estimated by a training based two tap FEQ. Is is to be noted that both Tx and Rx IQ imbalance parameters $\mathbf{N}_t$ and $\mathbf{N}_r$ are small values as $\mathbf{G}_{ta} >> \mathbf{G}_{tb}$ and $\mathbf{G}_{ra} >> \mathbf{G}_{rb}$. This is true even for large IQ imbalance values as seen in practise. Thus for the moment we can ignore the dot multiplication of $\mathbf{N}_r$ and $\mathbf{N}^*_{t_m}$ in equations (6) and (7) as their contribution to the two equations are relatively small. We can now write $\widetilde{\mathbf{D}}^{(1)} = \mathbf{W}_1$ and

$\overset{\sim}{\mathbf{D}}^{(2)} = \mathbf{W}_3$ where $\overset{\sim}{\mathbf{D}}^{(q)}$ is the estimate of $\mathbf{D}^{(q)}$ (for q=1,2). Now the estimates $\overset{\sim}{\mathbf{N}}_t$ and $\overset{\sim}{\mathbf{N}}_r$ of $\mathbf{N}_t$ and $\mathbf{N}_r$ can be obtained by the following equation:

$$\begin{bmatrix} \overset{\sim}{\mathbf{N}}_t \\ \overset{\sim}{\mathbf{N}}_r \end{bmatrix} \Leftarrow \begin{bmatrix} \overset{\sim}{\mathbf{D}}^{(1)} & \overset{\sim}{\mathbf{D}}_m^{*(1)} \\ \overset{\sim}{\mathbf{D}}^{(2)} & \overset{\sim}{\mathbf{D}}_m^{*(2)} \end{bmatrix}^{-1} \begin{bmatrix} \mathbf{W}_2 \\ \mathbf{W}_4 \end{bmatrix} \tag{8}$$

Once $\overset{\sim}{\mathbf{N}}_t$ and $\overset{\sim}{\mathbf{N}}_r$ are known, we can then substitute $\overset{\sim}{\mathbf{N}}_{t_m}^*$ and $\overset{\sim}{\mathbf{N}}_r$ in $\mathbf{W}_1$ and $\mathbf{W}_3$. This results in a more precise estimate of $\overset{\sim}{\mathbf{D}}^{(q)}$, given by the following equation:

$$\begin{bmatrix} \overset{\sim}{\mathbf{D}}^{(1)} \\ \overset{\sim}{\mathbf{D}}^{(2)} \end{bmatrix} \Leftarrow \begin{bmatrix} \mathbf{W}_1 & \mathbf{W}_{1_m}^* \\ \mathbf{W}_3 & \mathbf{W}_{3_m}^* \end{bmatrix} \begin{bmatrix} 1 \\ -\overset{\sim}{\mathbf{N}}_r.\overset{\sim}{\mathbf{N}}_{t_m}^* \end{bmatrix} . \frac{1}{1 - \overset{\sim}{\mathbf{N}}_r.\overset{\sim}{\mathbf{N}}_{r_m}^*.\overset{\sim}{\mathbf{N}}_t.\overset{\sim}{\mathbf{N}}_{t_m}^*} \tag{9}$$

The new estimates of $\overset{\sim}{\mathbf{D}}^{(q)}$ and $\overset{\sim}{\mathbf{D}}_m^{*(q)}$ are then re-substituted in equation (8) to obtain a more precise estimate of IQ imbalance parameters. Thus equations (8) and (9) can be repeated a number of times until sufficiently good estimates of $\overset{\sim}{\mathbf{N}}_t$ and $\overset{\sim}{\mathbf{N}}_r$ are obtained. It is observed that in the noiseless case, 2-3 iterations already lead to an accurate estimation. We will call the estimation of $\overset{\sim}{\mathbf{N}}_t$ and $\overset{\sim}{\mathbf{N}}_r$ as the calibration phase from now on.

Once the calibration phase is over, we can now compensate the Rx IQ imbalance by the following equation:

$$\begin{aligned} \mathbf{Z}_t &= \mathbf{Z} - \overset{\sim}{\mathbf{N}}_r.\mathbf{Z}_m^* \\ &= \underbrace{\left(\mathbf{G}_{ra} - \frac{\mathbf{G}_{rb}.\mathbf{G}_{rb_m}^*}{\mathbf{G}_{ra_m}^*}\right)}_{\mathbf{G}_{rx}}.\mathbf{C}.(\mathbf{G}_{ta}.\mathbf{S} + \mathbf{G}_{tb}.\mathbf{S}_m^*) + \overset{\sim}{\mathbf{n}}_1 \\ &= \mathbf{D}_c.(\mathbf{S} + \mathbf{N}_t.\mathbf{S}_m^*) + \overset{\sim}{\mathbf{n}}_1 \end{aligned} \tag{10}$$

where $\mathbf{D}_c = \mathbf{G}_{rx}.\mathbf{C}.\mathbf{G}_{ta}$ is the composite channel and $\overset{\sim}{\mathbf{n}}_1$ is the noise term. The superscript from the received symbol $\mathbf{Z}$ and $\mathbf{C}$ has been dropped here to make the equation concise. Equation (10) shows that the composite channel estimate $\overset{\sim}{\mathbf{D}}_c$ can now be obtained based on a transmitted training symbol $\mathbf{S}$ and $\mathbf{S}_m^*$ as follows:

$$\overset{\sim}{\mathbf{D}}_c = \frac{\mathbf{Z}_t}{\mathbf{S} + \overset{\sim}{\mathbf{N}}_t.\mathbf{S}_m^*} \tag{11}$$

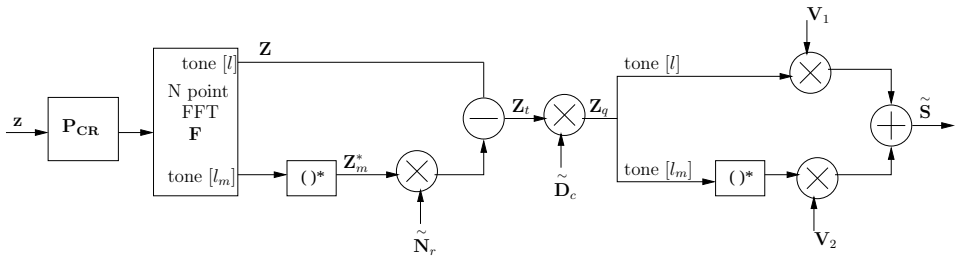


Figure 1. Compensation scheme for Tx Rx IQ imbalance and the channel in an OFDM system

Once the composite channel has been estimated and then compensated from the received signal $\mathbf{Z}_t$, the final step then involves the compensation of only Tx IQ imbalance term from the received signal. The estimate of the transmitted signal $\widetilde{\mathbf{S}}$ can then be obtained by the following equation:

$$\widetilde{\mathbf{S}} = \begin{bmatrix} \mathbf{V}_1 & \mathbf{V}_2 \end{bmatrix} \begin{bmatrix} \mathbf{Z}_q \\ \mathbf{Z}^*_{qm} \end{bmatrix} \tag{12}$$

where $\mathbf{Z}_q = \frac{\mathbf{Z}_t}{\widetilde{\mathbf{D}}_c}$, $\mathbf{V}_1 = \frac{1}{1-\widetilde{\mathbf{N}}_t.\widetilde{\mathbf{N}}^*_{tm}}$ and $\mathbf{V}_2 = \frac{\widetilde{\mathbf{N}}_t}{\widetilde{\mathbf{N}}_t.\widetilde{\mathbf{N}}^*_{tm}-1}$. We call this second phase of the equalization process as channel estimation and Tx Rx IQ imbalance compensation phase. It should be noted that any channel variation requires the re-estimation of only the composite channel $\widetilde{\mathbf{D}}_c$. The remaining equalization process including the calibration phase and the IQ imbalance compensation scheme does not change. Thus once the calibration phase is over we then require only one training symbol to estimate the composite channel. The two phase scheme finally results in a lower training overhead as will be verified in the simulation section. The proposed equalization scheme when applied to the OFDM data symbol is shown in **Figure 1**.

The estimates $\mathbf{W}_1 \ldots \mathbf{W}_4$ and $\widetilde{\mathbf{D}}_c$ can be further improved by utilizing a frequency smoothing operation. In this scheme we first transform $\mathbf{W}_1 \ldots \mathbf{W}_4$ and $\widetilde{\mathbf{D}}_c$ to their time domain representation by an IDFT operation. Once in time domain, the first and last L' terms at the top and bottom of their time domain sequence are left unchanged while the remaining $N - 2L'$ terms are forced to zero. This is because the impulse response length of these estimates are considered to be of finite length but they may be split at both the top and the bottom of their time domain sequence. The safe length for $2L'$ is then always the CP

length. Finally the time domain sequences are again transformed to the frequency domain by a DFT operation. This simple scheme results in an overall frequency smoothing operation [9]. The smoothing operation helps in accelerating the convergence of the equalizer coefficients resulting in an overall improved performance as will be shown in the simulation section.

In the case when only Tx IQ imbalance or Rx IQ imbalance is present in the system (but not both at the same time), then the entire equalization structure can be further simplified. In this case both $\mathbf{N}_t$ and $\mathbf{N}_r$ can be directly derived from $\mathbf{W}_1$ and $\mathbf{W}_2$. The estimation of $\mathbf{W}_3$ and $\mathbf{W}_4$ is not required. Thus in the presence of only Tx IQ imbalance in the system (i.e. $\mathbf{N}_r = 0$), the estimate $\widetilde{\mathbf{N}}_t = \frac{\mathbf{W}_2}{\mathbf{W}_1}$. Also the equation (10) is now simplified to $\mathbf{Z}_t = \mathbf{Z}$. While in the case of only Rx IQ imbalance in the system (i.e. $\mathbf{N}_t = 0$), the estimate $\widetilde{\mathbf{N}}_r = \frac{\mathbf{W}_2}{\mathbf{W}_{1m}^*}$. The equation (11) is now modified as $\widetilde{\mathbf{D}}_c = \frac{\mathbf{Z}_t}{\mathbf{S}}$ and equation (12) as $\widetilde{\mathbf{S}} = \frac{\mathbf{Z}_q}{\widetilde{\mathbf{D}}_c}$. The equalization is now similar to the scheme proposed in [9].

4. Simulation Results

We have simulated a SISO based OFDM system to evaluate the performance of the proposed compensation scheme. The performance comparison is made with an ideal system with no front-end distortion and a system with joint compensation algorithm [8] included. The performance curves are also drawn for a system with no compensation algorithm in place. The parameters used in the simulation are as follows: we consider OFDM symbol of length $N = 128$, CP length $\nu = 16$ and the constellation size= 64QAM. The mismatched filter impulse responses at Tx and Rx are [0.01, 0.5 0.09] for the I branch and [0.09 0.5, 0.01] for the Q branch. The frequency independent amplitude and phase imbalances at both Tx and Rx are 10% and 10° respectively. It should be noted that we have considered quite large values of IQ imbalances in order to observe the robustness of the compensation scheme. The multipath channel length equals 4 taps. The taps of the multipath channel are chosen independently with complex Gaussian distribution. During the frequency smoothing operation, we have left the top $L' = 8$ and bottom $L' = 8$ elements in the time domain sequence of $\mathbf{W}_1 \ldots \mathbf{W}_4$ and $\widetilde{\mathbf{D}}_c$ as unchanged while the remaining $N - 2L'$ elements are forced to zero.

We have initially used 8 training symbols to estimate $\widetilde{\mathbf{N}}_t$ and $\widetilde{\mathbf{N}}_r$. The **Figure 2a** shows the number of iterations required by the equalizer to estimate $\widetilde{\mathbf{N}}_t$ and $\widetilde{\mathbf{N}}_r$ accurately. The curves depict these estimates as

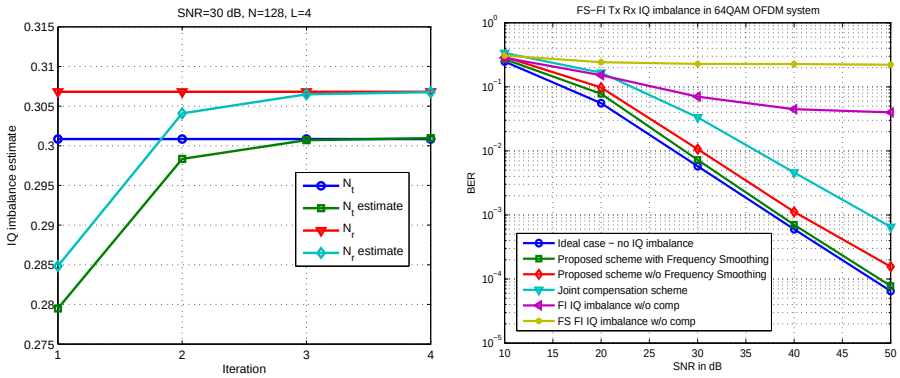


Figure 2. (a) IQ imbalance estimation with iterative method (b) BER vs SNR of OFDM system with FS-FI Tx Rx IQ imbalance

the mean of the absolute values for all the N tones of an OFDM symbol taken together (i.e $\Xi\{|\widetilde{\mathbf{N}}_t|\}$ and $\Xi\{|\widetilde{\mathbf{N}}_r|\}$). It is observed that even at low SNR=30dB, 3-4 iterations lead to accurate estimation. Once the equalizer is calibrated we can then obtain the performance curves with only one training symbol. In the simulation we have utilized two training symbols as this is the minimal requirement for the joint compensation scheme in [7]- [8]. The **Figure 2b** shows the performance curves (BER vs SNR) obtained for such a system. The BER results depicted are obtained by taking the average of the BER curves over 10^4 independent channels. It can be observed from the figure that with no compensation scheme in place, the system is completely unusable. The BER is also very high for the case of only FI Tx Rx IQ imbalance in the system. The proposed compensation scheme (with and without frequency smoothing operation) provides a very good performance results. The results obtained are very close to the ideal case even when only two training symbols are used in the system. The difference between the proposed scheme and the joint compensation scheme [8] is almost 9 dB at BER of 10^{-3}. Thus the proposed compensation scheme requires very low training overhead for an effective compensation.

5. Conclusion

In this paper the joint effect of Tx-Rx IQ imbalance along with multi-path channel distortions has been studied. A generally applicable compensation scheme has been developed that can decouple all the three frequency selective distortions, namely the Tx IQ imbalance, Rx IQ imbalance and the channel. The proposed equalizer works in two phases.

The first phase is the calibration phase where both the Tx and Rx IQ imbalance parameters are estimated. In the second phase, the equalizer estimates and compensates the channel along with IQ imbalance parameters. Once the calibration phase is over, the equalizer then has a very low training overhead requirement for effective compensation. Simulation results verify the effectiveness of the proposed scheme.

References

[1] A.A. Abidi,"Direct-conversion radio transceivers for digital communications", *IEEE Journal on Solid-State Circuits*, vol. 30, pp. 1399-1410, Dec. 1995.

[2] "IEEE standard 802.11a-1999: wireless LAN medium access control (MAC) & physical layer (PHY) specifications, high-speed physical layer in the 5 GHz band," 1999.

[3] C.L.Liu, "Impact of I/Q imbalance on QPSK-OFDM-QAM detection", *IEEE Transactions on Consumer Electronics*, vol. 44, no. 8, pp. 984-989, Aug. 1998.

[4] T.Pollet, M. Van Bladel and M. Moeneclaey, "BER sensitivity of OFDM systems to carrier frequency offset and wiener phase noise," *IEEE Transactions on Communications*, vol. 43, no. (2/3/4), pp. 191-193, 1995.

[5] J.Tubbax, B.Come, L.Van der Perre, M.Engels, M.Moonen and H.De Man, "Joint compensation of IQ imbalance and carrier frequency offset in OFDM systems", in *Proceedings Radio and Wireless Conference, Boston, MA*, Aug. 2003, pp. 39-42.

[6] A.Tarighat and A.H.Sayed. "OFDM systems with both transmitter & receiver IQ imbalances", in *Proceedings IEEE 6th SPAWC Workshop, New York, NY*, June 2005, pp. 735-739.

[7] T.Schenck, P Smulders and E.Fledderus, "Estimation and compensation of frequency selective Tx/Rx IQ imbalance in MIMO OFDM systems", in *Proceedings IEEE ICC, Istanbul, Turkey*, June 2006, pp. 251-256.

[8] D.Tandur and M.Moonen, "Joint compensation of OFDM frequency selective transmitter and receiver IQ imbalance", *Eurasip Journal on Wireless Communications and Networking (JWCN)*, Volume 2007 (2007).

[9] L.Anttila, M. Valkama and M.Renfors, "Efficient mitigation of frequency-selective I/Q imbalances in OFDM receivers", in *Proceedings IEEE 68th VTC conference, Calgary, Canada*, Sept. 2008.

[10] M.Valkama,M.Renfors and V.Koivunen, "Compensation of frequency-selective IQ imbalances in wideband receivers: models and algorithms", in *Proceedings IEEE 3rd Workshop on Signap Processing Advances in Wireless Communications (SPAWC)*, Taoyuan, Taiwan, Mar. 2001, pp. 42-45.