# OFDM RECEIVER DESIGN IN THE PRESENCE OF FREQUENCY SELECTIVE IQ IMBALANCE AND CFO

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**ABSTRACT**

Direct conversion receivers (DCR) are the preferred choice for the RF front-end in modern communication devices. These devices are simple and cheap to implement but for the case of sensitive multicarrier systems these devices may not maintain the required level of performance as regards image rejection (IR), carrier frequency offset (CFO) and direct current offset. In the presence of this non-ideal behaviour it is not possible to achieve a high signal to interferer ratio (SIR). This situation necessitates the use of digital signal processing (DSP) schemes to efficiently mitigate these effects and also to relax the stringent requirements on receiver design. In this paper we study the design of a receiver architecture which can jointly estimate the channel impulse response (CIR), IQ imbalance and CFO using long training sequences (LTS) which are already a part of WLAN standards. The proposed schemes provide an excellent performance/complexity trade-off.

## 1. INTRODUCTION

OFDM systems are becoming widely accepted in all modern communication standards. They find application in WiMax (IEEE 802.16), WLAN (IEEE 802.11n) and LTE schemes. Their biggest advantage is in being more spectrally efficient and maximizing the system throughput. The simple and low cost direct conversion receiver (DCR) is a preferred choice because it keeps the hardware cost to minimum but this receiver suffers from problems like IQ imbalance, carrier frequency offset (CFO), direct current offset (DCO) and phase noise (PN). Therefore it is essential that the receiver can handle the effects of these practical limitations in the digital domain. The traditional approach is to transmit certain training sequences and then do joint estimation of both the channel and the non-ideal behaviour parameters. From the literature we know that these schemes require multiple long training symbols (LTS) to achieve acceptable channel estimates.

Several methods have been proposed to estimate transmit and receive IQ imbalance in OFDM systems and [1] proposed an adaptive least mean square (LMS) scheme which estimates frequency selective (FS) and frequency independent (FI) IQ imbalance. However this scheme is slow to converge. The estimate of FI IQ imbalance and channel impulse response (CIR) for OFDM systems has been considered using pre- and post-FFT processing [2]. The authors have used the frequency domain model of channel and IQ imbalance to create a set of linear equations with the help of special training symbols. This scheme exploits the fact that in the presence of IQ imbalance in an OFDM system each subcarrier effects its (frequency domain) mirror image. The main drawback of this scheme is that it requires several OFDM symbols to adaptively estimate both IQ imbalance and the channel coefficients. In addition, these symbols are non-standard and so may not be useful in estimating other system parameters such as timing offset, phase noise or CFO. Another adaptive scheme has been proposed by [3] for joint estimation of CFO, TX/RX FI IQ imbalance and CIR. A joint channel IQ imbalance and PN estimation scheme has been proposed in [4] using a multidimensional grid search to minimize the overall cost function. However, all of these schemes are computationally complex and require several training sequences to converge to a reliable estimate.

The idea of joint IQ imbalance and channel estimation for an OFDM system has been considered in [5] where the authors have designed an optimal training sequence to jointly estimate FS IQ imbalance and the channel (as a set of two independent channels related to the useful and the interfering signal). A nonlinear least square (NLLS) estimation of CFO and FS IQ imbalance has been proposed by [6]. More recently [7], some authors have proposed a simple and robust joint channel, IQ imbalance and CFO estimation for MIMO-OFDM systems. This scheme also uses the LTS sequences available in WLAN standards. But this scheme too is limited to FI IQ imbalance and does not consider the FS imbalance model which is essential in broad bandwidth systems such as WLAN and WiMax.

In this paper we study a low complexity joint CIR, CFO, and IQ imbalance estimation for an OFDM system. In contrast to other works proposed in the literature which require many training symbols, our scheme can estimate the channel and the IQ imbalance using only the training sequences which are already part of WLAN standards. The proposed joint estimation scheme is very simple and robust. Another advantage of the proposed system is that it can perform channel estimation for FS and FI IQ imbalance in the presence of frequency selective CIR regardless of the delay spread of the LPF impulse responses (see Fig. 1) as long as we ensure a sufficiently long guard interval and an appropriate channel impulse response length is considered for estimation.

The rest of the paper is now organized as follows. Section 2 presents the model of a typical DCR in the presence of non-ideal receiver behaviour. The proposed optimal NLLS and reduced complexity solutions for joint FS IQ imbalance, CIR and CFO estimation are presented in section 3. Section 4 presents the scheme for joint CFO, IQ imbalance mitigation, channel equalization and data detection. Section 5 presents the MSE and BER performance of the proposed scheme and section 6 concludes this paper.

Notation: Boldface small letters represent vectors, while boldface capital letters represent matrices. The superscripts $*$, $T$, and $H$ represent respectively conjugate, transpose and Hermitian operators; $\Re\{\cdot\}$ and $\Im\{\cdot\}$ respectively represent the real and imaginary components of a complex number; $\otimes$ is the continuous time convolution operator; diag $(\cdot)$ represents the construction of a diagonal matrix from a vector; $\mathbf{I}$ is the identity matrix; $\mathbf{F}$ is a (unitary) DFT matrix of size $N \times N$; $\mathbf{F}$ is then partitioned into sub-blocks, i.e. $\mathbf{F}{=}[\mathbf{W}|\mathbf{V}]$ where $\mathbf{W}$ is the $N{\times}L'$ portion of $\mathbf{F}$.

## 2. SIGNAL MODEL

The typical diagram of a DCR device is illustrated in Fig. 1. We can see that in the case of lack of orthogonality in the inphase and quadrature-phase arms of the IQ receiver the system will suffer from the image problem. This can severely degrade the performance of the receiver and effectively limit the achievable SNR. A low cost local oscillator can also introduce a linear frequency offset to the received data sequence and so it will further degrade the performance of the receiver. These problems motivate us to design a simple yet robust receiver architecture which can handle the effects of this non-ideal behaviour and improve the operating SNR even in presence of severe IQ imbalance, CFO and DCO. The imbalance occurs due to non-uniform amplitude and phase gains in the I and Q branch of the DCR.

There are several ways of representing IQ imbalance in the literature and we will use the model employed in [2, 8]. The received signal at the RF front-end is defined as $r_{RF}(t){=}\Re\{r(t)e^{j2\pi f_c t}\}$ where $f_c$ is the carrier frequency and $r(t)$ is the equivalent complex baseband signal defined as $r(t){:=}c(t){\otimes}x(t){+}n(t)$, where $x(t)$ is the baseband equivalent of transmitted RF signal, $c(t)$ is the equivalent transmission channel and $n(t)$ is the equivalent zero mean white noise. As illustrated in Fig. 1 the low pass filter (LPF) gains of the I and Q branches are not matched exactly. The received signals in the I and Q branch in the presence of FS IQ imbalance are given as

$$\begin{aligned} y^I(t) ={}&(1+\eta)[\cos(2\pi\Delta ft-\theta)\cdot\Re\{r(t)\} \\ &-\sin(2\pi\Delta ft-\theta)\cdot\Im\{r(t)\}]{\otimes}g^I(t) \qquad (1) \end{aligned}$$

$$\begin{aligned} y^Q(t) ={}&(1-\eta)[\sin(2\pi\Delta ft+\theta)\cdot\Re\{r(t)\} \\ &+\cos(2\pi\Delta ft+\theta)\cdot\Im\{r(t)\}]{\otimes}g^Q(t) \qquad (2) \end{aligned}$$

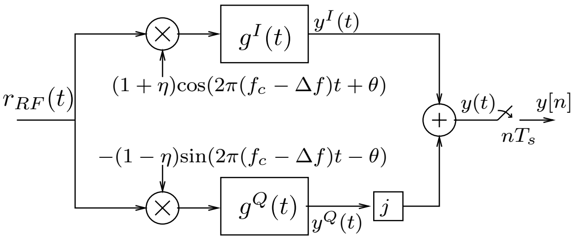


**Fig. 1**: The equivalent model of a DCR in the presence of FS IQ mismatch and CFO.

where $\eta$ represents the amplitude imbalance, $\theta$ is the phase imbalance, $\Delta f$ is the CFO introduced by the LO, and $g^I(t)$ and $g^Q(t)$ are the impulse responses of the LPF's in the inphase and quadrature-phase branches respectively. In the case of non FS IQ imbalance then $g^I(t){=}g^Q(t) = g(t)$. So from (1) and (2), the output of the DCR in Fig. 1 is:

$$\begin{aligned} y(t) ={}&y^I(t) + jy^Q(t) \\ ={}&\Big[(\cos(\theta) - j\eta\sin(\theta))\Big(\frac{g^I(t)+g^Q(t)}{2}\Big) \\ &+ (\eta\cos(\theta) - j\sin(\theta))\Big(\frac{g^I(t)-g^Q(t)}{2}\Big)\Big]{\otimes}r(t)e^{j2\pi\Delta ft} \\ &+ \Big[(\eta\cos(\theta) + j\sin(\theta))\Big(\frac{g^I(t)+g^Q(t)}{2}\Big) \\ &+ (\cos(\theta) + j\eta\sin(\theta))\Big(\frac{g^I(t)-g^Q(t)}{2}\Big)\Big]{\otimes}r^*(t)e^{-j2\pi\Delta ft}. \end{aligned} \qquad (3)$$

Re-writing the filter mismatch terms in (3) we define

$$\begin{aligned} k^1(t) &:= \frac{g^I(t)+g^Q(t)}{2} \\ k^2(t) &:= \frac{g^I(t)-g^Q(t)}{2}. \end{aligned} \qquad (4)$$

Similarly, the complex amplitude and phase mismatches in (3) are defined as follows

$$\begin{aligned} \mu &:= \cos(\theta) - j\eta\sin(\theta) \\ \nu &:= \eta\cos(\theta) + j\sin(\theta). \end{aligned} \qquad (5)$$

In the case of no FI imbalance (i.e. $\theta{=}\eta{=}0$), then $\mu = 1$ and $\nu = 0$. Now, from (3),(4) and (5) we have

$$\begin{aligned} y(t) ={}&h^D(t){\otimes}\Big(c(t){\otimes}x(t)\Big)e^{j2\pi\Delta ft} \\ &+ h^I(t){\otimes}\Big(c^*(t){\otimes}x^*(t)\Big)e^{-j2\pi\Delta ft}{+}\tilde{n}(t) \end{aligned} \qquad (6)$$

where

$$\begin{aligned} h^D(t) &= \mu k^1(t) + \nu^* k^2(t) \\ h^I(t) &= \nu k^1(t) + \mu^* k^2(t) \end{aligned} \qquad (7)$$

and $\tilde{n}(t){=}\Big(h^D(t)e^{j2\pi\Delta ft}{+}h^I(t)e^{-j2\pi\Delta ft}\Big){\otimes}n(t)$. We define $h_\mu(t){:=}h^D(t){\otimes}c(t)$ and $h_\nu(t){:=}h^I(t){\otimes}c^*(t)$ as equivalent overall channel models pertaining to the desired and image signals respectively [6]. Then after sampling at the baud rate ($1/T_s$) and removing the cyclic prefix, we get

$$\mathbf{y}^{(i)} = \mathbf{E}^{(i)}\mathbf{X}_c\mathbf{h}_\mu + \mathbf{E}^{(i)^*}\mathbf{X}_c^*\mathbf{h}_\nu + \mathbf{n}^{(i)} \qquad (8)$$

where '$i$' is the symbol index and the CFO process affecting the $i$-th OFDM symbol is defined as $\mathbf{E}^{(i)}{=}\mathrm{diag}(\{e^{j\frac{2\pi}{N}\epsilon[(i-1)(N+N_g)+N_g+n]}\}_{n=0}^{N-1})$; $N$ and $N_g$ are respectively the size of the OFDM symbol and the cyclic prefix; $\epsilon$ is the CFO coefficient normalized to the subcarrier spacing, i.e. $\epsilon{=}T\Delta f$ $(T{=}NT_s)$; $\mathbf{n}^{(i)}{\sim}\mathcal{CN}(\mathbf{0},\sigma_n^2\mathbf{I})$ is the circularly symmetric complex additive white Gaussian noise; $\mathbf{X}_c$ is the $N{\times}L'$ circulant matrix constructed from the transmitted OFDM pilot symbols, i.e., $(\mathbf{x}{=}\mathbf{F}^H\mathbf{s})$, where $\mathbf{s}$ is the known pilot symbol vector; $\mathbf{h}_\mu$ and $\mathbf{h}_\nu$ are respectively the equivalent overall channel impulse responses pertaining to the useful and the interfering signals in (8). For simplicity we assume that these channels have same length, i.e., $L'{\geq}L_h{+}L_g{+}1$, where $L_h$ and $L_g$ are the respective lengths of the equivalent discrete-time CIR $(\mathbf{h})$ and the two LPF's in Fig. 1. In practice the normalized CFO can take any value of offset, which can then be decomposed into integer and fractional parts, namely $\epsilon{=}\kappa{+}\phi$, where $\kappa$ and $\phi$ are respectively integer and fractional, i.e. $-\pi{\leq}\phi{\leq}\pi$. However in practice the maximum detectable range of the CFO using a LTS is defined as $|\phi|{<}N\pi/2(N{+}N_g)$ [9]. In the literature, several schemes are available which can estimate the integer and fractional part of the CFO process. However within the scope of this work we limit ourselves to the fractional part of the CFO only.

In this next section we propose an optimal joint IQ imbalance, CIR and the CFO estimation scheme. Subsequently, a reduced complexity scheme (RCS) is also proposed which provides a good complexity/performance trade-off.

## 3. PARAMETER ESTIMATION

In this section we propose two estimation schemes. The first scheme is the NLLS scheme which is the optimal scheme but it is computationally intensive. We then propose a suboptimal joint estimation scheme which provides acceptable estimates of system parameters with much lower implementation complexity.

### 3.1. Nonlinear Least Squares (NLLS) Scheme

Let us assign (as defined in the IEEE 802.11 WLAN standard) two LTS training sequences in the first two symbols of an OFDM block (i.e. $i{=}1,2$). So (8) becomes

$$\underbrace{\begin{bmatrix}\mathbf{y}^{(1)}\\ \mathbf{y}^{(2)}\end{bmatrix}}_{\bar{\mathbf{y}}} = \underbrace{\begin{bmatrix}\hat{\mathbf{E}}^{(1)}\mathbf{X}_c & (\hat{\mathbf{E}}^{(1)}\mathbf{X}_c)^*\\ \hat{\mathbf{E}}^{(2)}\mathbf{X}_c & (\hat{\mathbf{E}}^{(2)}\mathbf{X}_c)^*\end{bmatrix}}_{\mathbf{A}_\phi}\begin{bmatrix}\mathbf{h}_\mu\\ \mathbf{h}_\nu\end{bmatrix} + \underbrace{\begin{bmatrix}\mathbf{n}^{(1)}\\ \mathbf{n}^{(2)}\end{bmatrix}}_{\bar{\mathbf{n}}} \quad (9)$$

where $\hat{\mathbf{E}}^{(i)}$ is the estimate of $\mathbf{E}^{(i)}$ in (8) via (11) and (16) . The LS channel estimates (given the CFO estimates) are:

$$\begin{bmatrix}\hat{\mathbf{h}}_\mu\\ \hat{\mathbf{h}}_\nu\end{bmatrix} = \left(\mathbf{A}_{\hat{\phi}}^H\mathbf{A}_{\hat{\phi}}\right)^{-1}\mathbf{A}_{\hat{\phi}}^H\bar{\mathbf{y}}. \quad (10)$$

The NLLS estimate of the CFO process can be found from

$$\hat{\phi}_{\text{opt}} = \arg\max_{\phi} \quad \bar{\mathbf{y}}^H\mathbf{A}_\phi(\mathbf{A}_\phi^H\mathbf{A}_\phi)^{-1}\mathbf{A}_\phi^H\bar{\mathbf{y}} \quad (11)$$

where maximization is performed via grid search. Using the estimates of $\phi$ from (16) as an initial search point in (11) the search convergence time can be minimized. The NLLS scheme yields optimal CFO and joint channel estimates at the expense of additional complexity.

### 3.2. Reduced Complexity Scheme (RCS)

In this section we propose a simple solution for joint CFO, FS IQ imbalance and CIR estimation in the frequency domain. The model of (8) can now be equivalently rewritten in the frequency domain notation as

$$\mathbf{y}_f^{(i)} = \mathbf{F}\mathbf{y}^{(i)} = \mathbf{C}^{(i)}\mathbf{S}\mathbf{W}\mathbf{h}_\mu + \mathbf{C}^{(i)^H}\mathbf{Q}\mathbf{S}^H\mathbf{W}^*\mathbf{h}_\nu + \mathbf{z}^{(i)} \quad (12)$$

where $\mathbf{S}{=}\mathrm{diag}\{\mathbf{s}\}$ is the known pilot sequence; $\mathbf{Q}$ is the permutation matrix defined as $\mathbf{Q}{=}\mathbf{F}\mathbf{F}^T$; $\mathbf{z}^{(i)}{=}\mathbf{F}\mathbf{n}^{(i)}$; $\mathbf{C}^{(i)}$ is the circulant matrix of the CFO process which is defined as $\mathbf{C}^{(i)}{=}\mathbf{F}\mathbf{E}^{(i)}\mathbf{F}^H$ [4]. Decomposing this expression further we get

$$\begin{aligned}\mathbf{y}_f^{(i)} &= c_0^{(i)}\mathbf{S}\mathbf{W}\mathbf{h}_\mu{+}c_0^{(i)^*}\mathbf{Q}\mathbf{S}^H\mathbf{W}^*\mathbf{h}_\nu\\ &\quad{+}\underline{\mathbf{C}}^{(i)}\mathbf{S}\mathbf{W}\mathbf{h}_\mu{+}\underline{\mathbf{C}}^{(i)^*}\mathbf{Q}\mathbf{S}^H\mathbf{W}^*\mathbf{h}_\nu{+}\mathbf{z}^{(i)} \quad (13)\end{aligned}$$

where $\mathbf{C}^{(i)}{=}c_0^{(i)}\mathbf{I}{+}\underline{\mathbf{C}}^{(i)}$ and $c_0^{(i)}{=}e^{j\frac{2\pi}{N}\epsilon[(i-1)(N+N_g)+N_g+\frac{N}{2}]}$ is the mean value of the CFO process commonly known as common phase rotation (CPR), while $\underline{\mathbf{C}}^{(i)}$ is the intercarrier interference (ICI) component of the CFO which is caused by the spectral leakage of adjacent subcarriers. Taking the data dependent interference and additive noise into account we can re-write (13) as

$$\mathbf{y}_f^{(i)} \quad = \mathbf{S}\mathbf{W}\underline{\mathbf{h}}_\mu^{(i)}{+}\mathbf{Q}\mathbf{S}^H\mathbf{W}^*\underline{\mathbf{h}}_\nu^{(i)}{+}\tilde{\mathbf{z}}^{(i)} \quad (14)$$

where $\underline{\mathbf{h}}_\mu^{(i)} = c_0^{(i)}\mathbf{h}_\mu$ and $\underline{\mathbf{h}}_\nu = c_0^{(i)^*}\mathbf{h}_\nu$ and $\tilde{\mathbf{z}}^{(i)}$ is the sum of the remaining interfering terms in (13). So

$$\mathbf{y}_f^{(i)} = \underbrace{\begin{bmatrix}\mathbf{S}\mathbf{W} & \mathbf{Q}\mathbf{S}^H\mathbf{W}^*\end{bmatrix}}_{\mathbf{A}}\begin{bmatrix}\underline{\mathbf{h}}_\mu^{(i)}\\ \underline{\mathbf{h}}_\nu^{(i)}\end{bmatrix}+\tilde{\mathbf{z}}^{(i)}.$$

A simple least-squares estimator can determine the modified CIR as

$$\begin{bmatrix}\hat{\underline{\mathbf{h}}}_\mu^{(i)}\\ \hat{\underline{\mathbf{h}}}_\nu^{(i)}\end{bmatrix} = \left(\mathbf{A}^H\mathbf{A}\right)^{-1}\mathbf{A}^H\mathbf{y}_f^{(i)}. \quad (15)$$

Note that this estimator does not require any matrix inversion during run-time, as the pilot sequence is known a-priori at the receiver. The LS estimates of $\underline{\mathbf{h}}_\mu^{(i)}$ and $\underline{\mathbf{h}}_\nu^{(i)}$ are obtained using the two LTS sequences which are defined in the IEEE 802.11 standards. Assuming that the channel does not change within a transmission frame and all OFDM symbols are affected by same CFO coefficient, it can be easily shown that we can estimate the CFO as

$$\hat{\phi} = \frac{1}{\alpha}\tan^{-1}\left(\frac{\Im\{\underline{\mathbf{h}}_\mu^{(1)^H}\underline{\mathbf{h}}_\mu^{(2)}\}}{\Re\{\underline{\mathbf{h}}_\mu^{(1)^H}\underline{\mathbf{h}}_\mu^{(2)}\}}\right) \quad -\frac{\pi}{2} \leq \phi \leq \frac{\pi}{2} \quad (16)$$

where $\alpha{=}2\pi(N{+}N_g)/N$. The performance of this scheme suffers from residual interference in the high SNR region. Once the CFO is known then the channel can be estimated as in (10).

In the next section we present a joint single step channel and impairment mitigation scheme.

## 4. SINGLE STAGE CFO, IQ IMBALANCE AND CHANNEL EQUALIZATION

In contrast to previously proposed schemes which compensate IQ imbalance, CFO and CFR in multiple steps, we propose to compensate IQ imbalance, CFO and equalize the channel in the frequency domain using a single stage equalizer. The conventional FI IQ imbalance can be compensated from the received signal in (8) (as [7])

$$\mathbf{y}^{\text{I/Q}} = \mu^{-1}\left[\frac{\mathbf{y} - \frac{\nu}{\mu^*}\mathbf{y}^*}{1-\left|\frac{\nu}{\mu}\right|^2}\right] \tag{17}$$

where $\mathbf{y}^{\text{I/Q}}$ is the time domain sequence after IQ imbalance compensation where we have dropped the symbol index '$i$' in (8). This scheme is limited to the case when IQ imbalance is FI and the equivalent base-band channel will be estimated after both IQ imbalance and CFO are compensated. However in the case of FS IQ imbalance, we can estimate the CIR and FS IQ imbalance jointly using (10) and estimate the CFO using (16). Therefore channel equalization, CFO and IQ imbalance compensation can be performed in a single step. Re-writing (8) in the frequency domain

$$\mathbf{y}_f = \mathbf{C}\mathbf{\Lambda}_\mu\mathbf{s} + \mathbf{C}^H\mathbf{\Lambda}_\nu\mathbf{Q}\mathbf{s}^* + \mathbf{z}. \tag{18}$$

Using (17) with (18) it is not difficult to show that a joint channel, CFO and IQ imbalance equalizer can be devised as

$$\hat{\mathbf{s}}{=}\left(\mathbf{I}{-}\mathbf{\Lambda}_\nu\mathbf{Q}\mathbf{\Lambda}_\mu^{-*}\mathbf{\Lambda}_\nu^{*}\mathbf{Q}\mathbf{\Lambda}_\mu^{-1}\right)^{-1}\left[\mathbf{\Lambda}_\mu^{-1}\hat{\mathbf{C}}^H\left(\mathbf{y}_f{-}\mathbf{\Lambda}_\nu\mathbf{Q}\mathbf{\Lambda}_\mu^{-*}\mathbf{y}_f^*\right)\right] \tag{19}$$

where $\mathbf{\Lambda}_\mu$ and $\mathbf{\Lambda}_\nu$ are diagonal matrices of the equivalent CFR responses obtained from (10) and $\hat{\mathbf{C}}$ is the circulant matrix constructed from the estimated CFO process obtained from (16).

Implementation of the above equalizer is low complexity as it does not require matrix inversion (the $(\cdot)^{-1}$ term in (19) is a diagonal matrix). We will compare our work with well-cited joint channel, CFO and FS IQ imbalance estimation schemes. The adaptive estimators proposed in the references [2, 3] require several short and long pilot symbols to converge to a reliable estimate. The complexity of these schemes is of the order $K\mathcal{O}(N^2)$, where $K$ is the number of training symbols. The complexity of the joint CFO, FI IQ imbalance estimation scheme proposed in [7] is of the order of $\mathcal{O}(N)$. Our scheme provides a closed-form expression for joint estimation of CIR, CFO and (FS/FI) IQ imbalance process with complexity order of $2\mathcal{O}(NL'^2)$, where $L'$ is the length of the equivalent CIRs as defined for (8). If we can say that $L' \ll N$, then the complexity of the proposed scheme is far less than other adaptive channel estimation schemes.

## 5. SIMULATION RESULTS

In this paper we consider a typical SISO OFDM system like a WLAN/WiMAX transmission system. The number of subcarriers in each OFDM symbol is $N{=}64$. The system bandwidth is assumed to be 20MHz and the sampling rate is defined as ($T_s{=}0.05\mu s$). The subcarrier spacing is assumed to be $\Delta F{=}312.5$ KHz. We consider an $L_h{=}6$ tap Rayleigh fading process with exponential power delay profile $e^{-\gamma l}$ with $\gamma{=}0.2$ and $l{=}0, 1, \ldots, L_h{-}1$. To mitigate the effects of inter symbol interference (ISI), the guard interval is assumed to be longer than CIR, i.e. $N_g{=}10$. The transmitted data is assumed to be taken from a 16 QAM constellation and no channel coding is used in these simulations. Each OFDM block contains 10 OFDM symbols. The first two symbols ($i{=}1, 2$)of each block are known training sequences chosen from a BPSK constellation according to the criterion presented in [5]. The IQ imbalance equation of (5) is assumed to be a random variable with uniformly distributed amplitude imbalance, $\eta{\sim}\mathcal{U}$[-$0.1, 0.1$], and phase imbalance, $\theta{\sim}\mathcal{U}$[-$10°, 10°$]. The normalized CFO is also assumed to be uniformly distributed, $\epsilon{\sim}\mathcal{U}$[-$0.43, 0.43$]. The LPF gains of the inphase and quadrature arms of the DCR used in our implementation are $\mathbf{k}^1{=}[0.01, 0.95, 0.1]^T$ and $\mathbf{k}^2{=}[0.01, 0.05, 0.01]^T$, where $k^j[n]{=}\{k^j(nT_s)\}|_{n=-1}^{1}$ in (7). These values present a plausible model for a frequency selective IQ imbalance which will be estimated in conjunction with the true CIR. Similar models have been used in [5] and [6]. The IQ imbalance process is assumed to be varying much more slowly than the CIR.

The simulations are performed using 5,000 Monte-Carlo channel realizations for each SNR. Fig. 2 illustrates the MSE performance of the proposed channel estimation scheme. It can be seen that the proposed estimator has a performance close to the CRLB [4]. The effects of CFO estimation error in RCS cause a slight degradation in channel estimation at high SNR.

The MSE performance of the CFO estimates is illustrated in Fig. 3. The estimate of the CFO obtained through RCS is very accurate in the low SNR region. The CRLB analysis for CFO estimation (in the absence of IQ imbalance) available in [10] is presented as a reference.

It should be mentioned here that the scheme of [7] can not be used for estimation of FS IQ imbalance whereas [3] can estimate FS IQ imbalance but will require many training sequences to obtain reliable estimates. In contrast to [2] and [3] the proposed schemes (using LTS sequences (which are part of the IEEE 802.11 standard)) can estimate the FS IQ imbalance as effectively as FI IQ imbalance as long as a sufficient cyclic prefix is available. The BER performance for the uncoded 16 QAM modulation scheme using NLLS and RCS

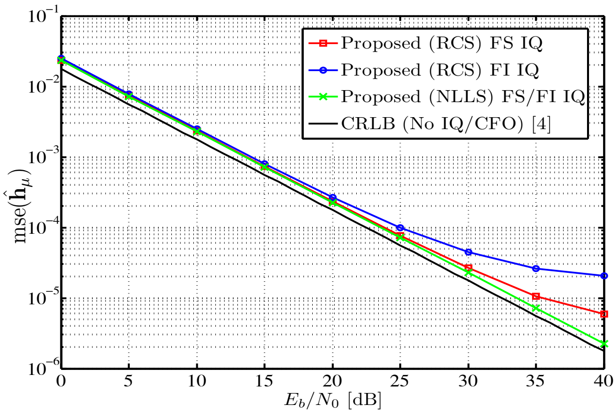


**Fig. 2**: MSE of channel estimation using (10), (16) and (11), in the presence of IQ imbalance and normalized CFO with $N$=64, Rayleigh channel with $L_h$=6 taps.

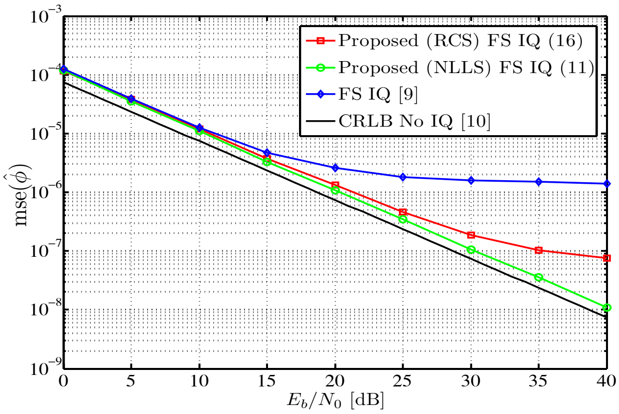


**Fig. 3**: MSE performance of CFO estimation with (11) and (16) with $N$=64, Rayleigh channel with $L_h$=6 taps.

estimates is presented in Fig. 4. The BER performance of the proposed low complexity scheme is within a 1 dB range of the ideal case for the system under consideration.

## 6. CONCLUSIONS

In this paper we have studied joint estimation of channel and FS/FI IQ imbalance in the presence of CFO for OFDM systems. We have devised a simple closed form expression for estimation of CFO without any a priori information about CIR and FS IQ imbalance. The CIR and FS IQ imbalance is estimated as a set of two independent channels corresponding to the desired and interfering signals. The proposed schemes use ordinary training sequences already defined in the WLAN standards.

We have also proposed a low complexity single stage CFO, FS IQ imbalance compensation and channel equalization scheme which can effectively deal with FS/FI IQ imbalance in the presence of a frequency selective channel impulse response. The simulation results show that the proposed schemes provide an excellent performance/complexity trade-off.

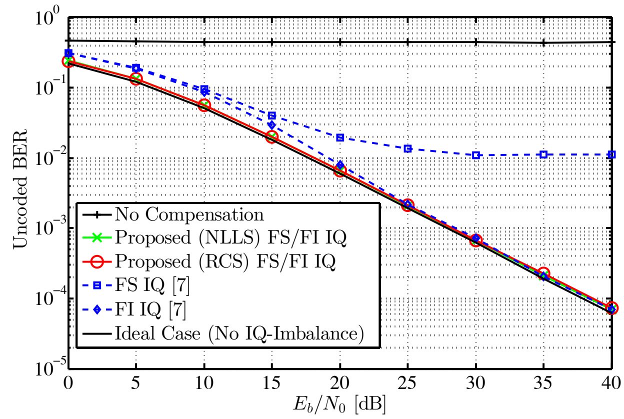


**Fig. 4**: BER performance of proposed schemes with $N$=64, Rayleigh channel with $L_h$=6 taps.

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