A BLIND ITERATIVE CARRIER FREQUENCY OFFSET ESTIMATOR BASED ON A KALMAN APPROACH FOR AN INTERLEAVED OFDMA UPLINK SYSTEM

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ABSTRACT

When dealing with uplink transmissions of a coded orthogonal frequency division multiple access (OFDMA) system, frequency synchronization has to be addressed. To this end, we propose a blind carrier frequency offset (CFO) estimator based on an iterative architecture that combines the so-called minimum mean square error successive detector and a Kalman algorithm, i.e. the extended Kalman filter, the unscented Kalman filter and the central difference Kalman filter. Unlike the actual open literature, no training sequence in our CFO estimation is required. In addition, the computational cost of our approach is moderate. When considering an interleaved OFDMA uplink system over a doubly Rayleigh fading channel, simulations results clearly show the efficiency of the proposed algorithm in terms of CFO estimation and bit error rate performances.

1. INTRODUCTION

The common feature of many current wireless standards for high data rate transmission is the adoption of a multi-carrier system based on orthogonal frequency division multiplexing (OFDM). Currently, there is a strong interest in extending the OFDM concept to multiuser communication scenarios. Multiuser OFDM systems, also known as orthogonal frequency division multiple access (OFDMA), are becoming the preferred system for many new communication standards. This scheme was originally presented by Sari and Karam for cable TV network [1] and later adopted for digital terrestrial television in the DVB-RCT standard [2]. More recently, OFDMA has been used in WiMAX (802.16x) [3] and in 3GPP long-term evolution [4]. Unlike the conventional OFDM case where all sub-carriers are assigned to a single user, each sub-carrier is exclusively assigned to a particular user in an OFDMA network. The communication link between each user and the base station is modeled by a time-varying channel, whose response differs from one user to the other. For simultaneous transmission, OFDMA allocation algorithms [11] exploit this spectral diversity to allocate the communication resources to the different users, such as power, constellation size and necessary bandwidth to maximize the link efficiency. However, similarly to OFDM, OFDMA is sensitive to:

- timing errors between the incoming signal and the base station references used for reception and demodulation in the uplink case. They lead to inter-block interferences (IBI) and can be avoided by using a sufficiently long cyclic prefix between adjacent OFDMA bursts. However, to maintain an acceptable data throughput, the cyclic-prefix length is chosen greater than the length of the channel impulse response.
- carrier frequency offsets (CFOs): without CFO estimations/compensations, orthogonality between sub-carriers is no longer satisfied. It results in inter-channel interferences as well as multiple access interferences (MAI).

The CFO estimation problem for OFDMA uplink transmissions has been recently addressed in several papers, [5], [6] and [11]. More particularly in [8], the CFO is estimated at the base station, but the frequency adjustment is left to the mobile terminals. In [9], when the channel is time-varying, the frequency estimation and its correction are only done at the base station. In [7], Zhao *et al.* use an extended Kalman filter (EKF) to estimate the CFO, provided that a long training sequence is known. However, this assumption is not necessary satisfied in real cases when using the IEEE 802.16 standard, since CFO is subject to change between two successive frames. In [10], *Hou et al.* propose a minimum mean square error successive detector (MMSE-SD) to suppress the MAI, but the CFO is assumed to be known.
To our knowledge, the joint CFO estimation and MAI correction without training sequence have never been addressed yet in OFDMA systems. To obtain a maximum data rate transmission in an OFDMA uplink transmission, we propose an iterative receiver that combines a MMSE-SD with a CFO estimator based on a Kalman algorithm. Three methods are studied: the EKF and two different sigma points Kalman filters (SPKFs), namely the unscented Kalman filter (UKF) and the central difference Kalman filter (CDKF). One of our purposes is to compare their estimation performances. It should be noted that this new architecture has the advantage of avoiding the use of a preamble.
The paper is organized as follows. The OFDMA system and the signal models are presented in section 2. Section 3 shows how estimates of the synchronization parameters can be exploited to restore orthogonality among the received users' signals. Simulation results, which confirm our algorithm efficiency, are presented in section 4 and finally conclusions are given in section 5.

2. SYSTEM DESCRIPTION

Let us consider an OFDMA network consisting of a single base station and U simultaneously independent users (See figure 1). The available bandwidth B is divided among N sub-carriers, and we suppose a fair distribution of the bandwidth $B_u = B/U$ between each user.

In the following, we denote $(.)^H$ and $(.)^T$, the hermitian and transposition operations respectively. In addition $Re(.)$ is the real part of $(.)$, diag$(.)$ is a square zero matrix, the main diagonal of which is $(.)$ and $\mathbf{I}$ is the identity matrix.

2.1 Proposed OFDMA uplink model

The signal received by the base station is a superposition of the contributions from U active users. In the following, let $\mathbf{S}_u$ be the OFDMA symbol emitted by the uth user with $u \in \{1, \ldots U\}$:

$$\mathbf{S}_u = [S_u(0), S_u(1) \ldots S_u(N-1)]^T \tag{1}$$

According to the frequency allocation of each user [11], $S_u(k)$ can be non-zero if the kth carrier is allocated to the uth mobile terminal, for $k \in \{0, \ldots N-1\}$. Then let us introduce the column vector $\mathbf{S}$ that contains the information symbols on all the N sub-carriers:

$$\mathbf{S} = \mathbf{S}_1 + \mathbf{S}_2 + \ldots + \mathbf{S}_U \tag{2}$$

The corresponding transmitted signal from the uth user is given by:

$$X_u(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_u(k) e^{j2\pi nk/N} \tag{3}$$

where $-N_g \leq n \leq N-1$ and $N_g < N$ is the length of the cyclic prefix.

Moreover, let us assume that the channel impulse response of the uth user at time n is $\mathbf{h}_u(n) = [h_u(n,0), h_u(n,1), \ldots, h_u(n, L_u)]^T$ where L_u is the length of the maximum channel delay spread and $L_u \leq N_g$ so that the cyclic prefix discards the IBI. We suppose a multipath quasi-static doubly Rayleigh fading channel[1]. Hence, $\mathbf{h}_u$ does not vary during an OFDMA block transmission, even though it may vary from block to block.

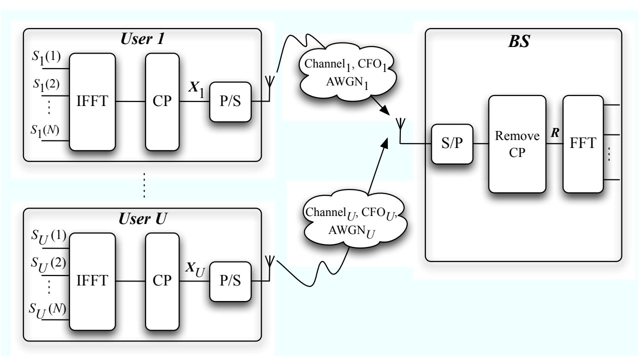

Figure 1: OFDMA system model

The U incoming waveforms are naturally combined by the received antenna. The resulting received signal at time n can be expressed as:

$$R(n) = \sum_{u=1}^{U} R_u(n) + B(n) \tag{4}$$

where $B(n)$ is a complex white gaussian noise with variance σ_b^2 while $R_u(n)$, the signal received from the uth user, can be expressed as:

$$R_u(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_u(k) \tilde{H}_u(n,k) e^{j2\pi nk/N} \tag{5}$$

where $\tilde{H}_u(n,k) = \sum_{l=0}^{L_u} h_u(n,l) exp(-j2\pi lk/N)$ is the channel frequency response associated with the kth sub-carrier of the uth user.

At the receiver, due to the propagation conditions, time offset and CFO are induced into the baseband signal. The received signal can be rewritten as follows:

$$R_u(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_u(k) \tilde{H}_u(n,k) e^{j2\pi(n-\tau_u)(k+\varepsilon_u)/N} \tag{6}$$

with ε_u and τ_u the normalized CFO to the sub-carrier spacing and timing error related to the uth user, respectively.

In order to restore orthogonality among each user sub-carrier, the synchronization error vector $\varepsilon = [\varepsilon_1, \varepsilon_2, \ldots, \varepsilon_U]$ must be estimated to compensate for the CFOs. By choosing an appropriate cyclic prefix length, namely $L_T = \max\{\tau_u + L_u\}$, the effects of the uplink timing errors are counteracted, i.e they are incorporated as a part of their channel responses[2]. After the cyclic prefix removal, the received signal can hence be expressed as:

$$R_u(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_u(k) H_u(n,k) e^{j2\pi n(k+\varepsilon_u)/N} \tag{7}$$

where $H_u(n,k) = \tilde{H}_u(n,k) e^{-j2\pi\tau_u(k+\varepsilon_u)/N}$.

Thus the received signal contains no IBI and (4) can be rewritten as:

$$R(n) = \sum_{u=1}^{U} e^{j2\pi n\varepsilon_u/N} A_u(n) + B(n) \tag{8}$$

where $A_u(n)$ corresponds to the nth OFDMA symbol only affected by the propagation channel and expressed as:

$$A_u(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_u(k) H_u(n,k) e^{j2\pi nk/N}$$

By using (4) each received OFDMA block can be rewritten in a matrix form as:

$$\mathbf{R} = [R(0), R(1), \ldots, R(N-1)]^T = \mathbf{G}\mathbf{S} + \mathbf{B} \tag{9}$$

where $\mathbf{B}$ is a column vector that contains N samples of noise. In addition the transmission matrix $\mathbf{G}$ is defined by:

$$\mathbf{G} = \sum_{u=1}^{U} \mathbf{E}_u \mathbf{H}_u \mathbf{Q}_u \tag{10}$$

where $\mathbf{E}_u = \text{diag}\left[1, e^{j2\pi\varepsilon_u/N}, \ldots, e^{j2\pi(N-1)\varepsilon_u/N}\right]$; $\mathbf{Q}_u$ is a diagonal matrix where the kth coefficient of the main diagonal is given by $\mathbf{Q}_u(k,k) = 1$ if $S_u(k) \neq 0$, $\mathbf{Q}_u(k,k) = 0$ otherwise.The coefficient $\mathbf{H}_u(n,k)$ of the nth row and kth column of $\mathbf{H}_u$ satisfies $\mathbf{H}_u(n,k) = H_u(n,k) e^{j2\pi(n-1)(k-1)/N}$.

Due to the CFO, the received OFDMA block column vector $\mathbf{R}$ includes interferences both from the same user and from all the other users. In the next section we analyze the way to estimate and correct this CFO in order to obtain the decoded signals of each user.

3. FREQUENCY OFFSET ESTIMATION AND USER DETECTION

In uplink OFDMA systems, the receiver performance is affected by CFO and MAI. In this section, a new method is proposed to mitigate both impacts. Firstly we use the MMSE-SD to estimate the signal sent by each user, then the estimated signals are used by the Kalman algorithm to estimate the CFO (See figure 2). This method has the advantage of avoiding a preamble.

[1]Doubly Rayleigh fading channel means frequency and time selective Rayleigh fading channel.

[2]This paper is focused in the CFO estimation, so τ_u and L_u are supposed to be known.

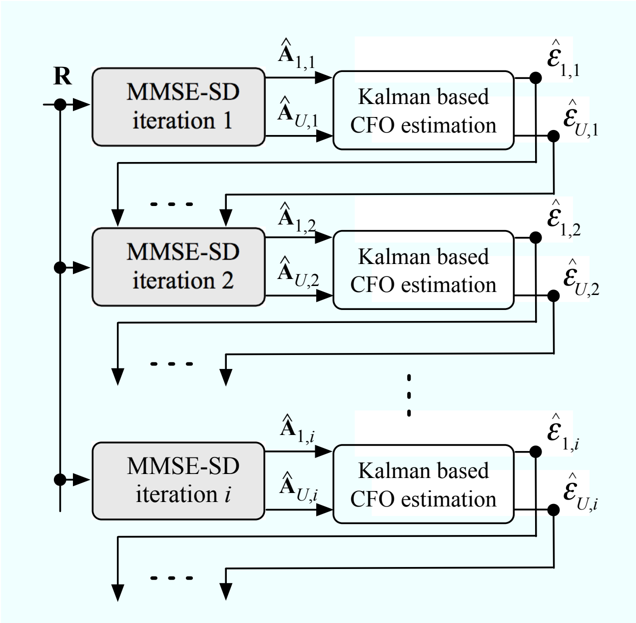

Figure 2: Proposed OFDMA receiver archietcture

3.1 MAI Suppression MMSE successive detection

According to [10], the MMSE-SD is robust against near-far effect, produced by the strong MAI. These phenomena are induced by the difference that may exist between two users in terms of the propagation loss. Instead of a joint multiuser decoding, we propose to combine a MMSE pre-detection scheme with a successive detection. The detection of transmitted interleaved OFDMA signal components operates in two major steps:

- interference cancelling (IC): during this step, the previous "detected" OFDMA signal components are subtracted out of the received signal. Indeed, let be $\hat{\mathbf{S}}_{\tilde{u},i}$ the estimation of the signal sent by the $\tilde{u}$th user at the ith iteration of the MMSE-SD, where $1 \le i \le I_{max}$ and I_{max} denotes the maximum iteration number. The user decoding order is denoted as $\tilde{u}$ where $\tilde{u} \in \{1,\ldots,U\}$; when $\tilde{u}=1$, it is associated with the maximum user signal interference noise ratio (SIR) whereas $\tilde{u}=U$ represents the user with the lowest SIR. In addition $\mathbf{Y}_{1,i} = \mathbf{R}\ \forall i$ and the so-called $(\tilde{u}+1)$th order MMSE-SD residual at the ith iteration $\mathbf{Y}_{\tilde{u}+1,i}\ \forall \tilde{u} \neq 1$ is the difference between the received signal $\mathbf{R}$ and the components transmitted by the detected user (namely those corresponding to the $\tilde{u}$ highest SIRs) (See figure 3).

$$\mathbf{Y}_{\tilde{u}+1,i} = \mathbf{Y}_{\tilde{u},i} - \mathbf{G}_{\tilde{u},i}\hat{\mathbf{S}}_{\tilde{u},i} \tag{11}$$

$$= \mathbf{R} - \sum_{l=1}^{\tilde{u}} \mathbf{G}_{l,i}\hat{\mathbf{S}}_{l,i} \tag{12}$$

$$\text{with } \mathbf{G}_{\tilde{u},i} = \begin{cases} \sum_{\tilde{u}=1}^{U} \hat{\mathbf{E}}_{\tilde{u},i}\mathbf{H}_{\tilde{u}}\mathbf{Q}_{\tilde{u}} & \text{if } \tilde{u}=1 \\ \mathbf{G}_{\tilde{u}-1,i}\left(\mathbf{I}-\mathbf{Q}_{\tilde{u}-1}\right) & \text{if } 2 \le \tilde{u} \le U \end{cases} \tag{13}$$

where $\hat{\mathbf{E}}_u = \operatorname{diag}\left[1, e^{j2\pi\hat{\varepsilon}_{u,i-1}/N}, \ldots, e^{j2\pi(N-1)\hat{\varepsilon}_{u,i-1}/N}\right]$ and $\hat{\varepsilon}_{\tilde{u},i}$ is the estimation of the CFO associated with the $\tilde{u}$th user at the ith iteration[3]. Since there is no *a priori* information of the offset related to each user, $\hat{\varepsilon}_{\tilde{u},0} = 0$ $\forall \tilde{u}$. A new iteration begins when all the users have been treated and when the estimation $\hat{\varepsilon}_{\tilde{u},i+1}$, using the Kalman estimator approach proposed in the next subsection, has been performed. For more details about the calculation of the SIR the reader is referred to [10].

- interference suppression (IS): this step aims at removing the interference stemming from the as-yet undecoded components, by using linear operations. As we suppose an ordered MMSE-SD, the output SIR must be available for each user. The purpose of this step is hence to filter the $\tilde{u}$th order MMSE-SD residual $\mathbf{Y}_{\tilde{u},i}$ by searching the suppression weight matrix $\mathbf{W}_{\tilde{u},i}$ minimizing the following criterion:

$$\underset{\mathbf{W}_{\tilde{u},i}}{\operatorname{argmin}} \left\| \hat{\mathbf{S}}_{\tilde{u},i} - \mathbf{S}_{\tilde{u}} \right\|^2$$

Then, by denoting σ_s^2 the signal power allocated on each of the sub-carriers, the suppression weight matrix for the selected $\tilde{u}$th user at the ith iteration satisfies:

$$\mathbf{W}_{\tilde{u},i} = \left(\frac{\sigma_b^2}{\sigma_s^2}\mathbf{I} + \mathbf{G}_{\tilde{u},i}^H\mathbf{G}_{\tilde{u},i}\right)^{-1}\mathbf{G}_{\tilde{u},i}^H \tag{14}$$

Then (14) is used to decode the selected user and to obtain the estimated signal sent by uth user[4].

$$\hat{\mathbf{S}}_{\tilde{u},i} = \underset{\mathbf{S_m}\in\Omega}{\operatorname{argmin}}\left(\mathbf{Q}_{\tilde{u}}\mathbf{W}_{\tilde{u},i}\mathbf{Y}_{\tilde{u},i} - \mathbf{S_m}\right) \tag{15}$$

where $\Omega = \{S_1, S_2, \ldots S_m, \ldots S_M\}$ is the modulation constellation. At that stage, (15) is used to obtain the estimated signal of the $\tilde{u}$th user.

$$\hat{\mathbf{A}}_{\tilde{u},i} = \mathbf{G}_{\tilde{u},i}\hat{\mathbf{S}}_{\tilde{u},i} = \left[\hat{A}_{\tilde{u},i}(0), \hat{A}_{\tilde{u},i}(1) \ldots \hat{A}_{\tilde{u},i}(N-1)\right] \tag{16}$$

In the next section those estimated signals are used as preambles to estimate the CFO of each user.

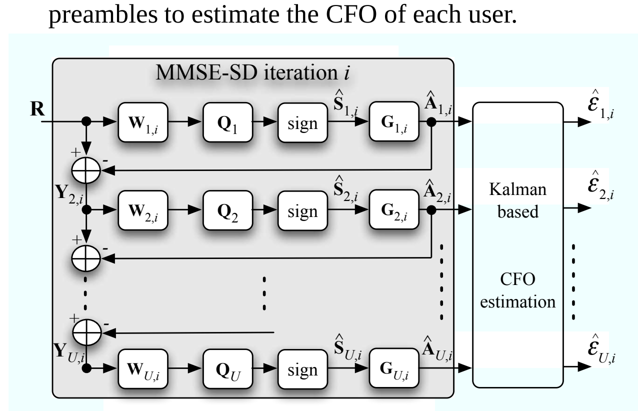

Figure 3: Iterative MMSE-SD and MAI suppression for BPSK modulation

3.2 Kalman approach based estimator: SPKF vs EKF

The CFO is estimated by a recursive method. The space-state model used to estimate the CFO is defined by:

State equation:
$$\varepsilon_{\tilde{u},i}(n) = \varepsilon_{\tilde{u},i}(n-1) \tag{17}$$

Measurement Equation:
$$\hat{R}_{\tilde{u},i}(n) = \hat{A}_{\tilde{u},i}(n)e^{j2\pi n\varepsilon_{\tilde{u},i}/N} + B_{\tilde{u}}(n) \tag{18}$$

where $B_{\tilde{u}}$ includes both the channel noise and the interferences from the other users. The variance of $B_{\tilde{u}}$ is assumed

[3] The matrix $\mathbf{H}_{\tilde{u}}$ is assumed to be known because in this paper a perfect channel estimation has been assumed.

[4] For example in BPSK modulation $\hat{\mathbf{S}}_{\tilde{u},i} = \operatorname{sign}\left(\mathbf{Q}_{\tilde{u}}\mathbf{W}_{\tilde{u},i}\mathbf{Y}_{\tilde{u},i}\right)$

to be $\sigma_{\tilde{u}}^2$. It should be noted that this space-state model is considered during one OFDMA block.
As the space-state model is non-linear, we suggest studying three kinds of Kalman algorithm:

- The EKF [12] consists in analytically propagating the Gaussian Random Vector (GRV) through the system dynamics, by means of a first-order linearization (Taylor expansion) of the non-linear equation (18) around the last available estimate of the state vector.
  However, due to the first order approximation, this approach may sometimes lead to large errors when evaluating the mean and the covariance matrix of the GRV that undergoes the non-linear transformation.
- When dealing with the SPKF [13], the state distribution is still approximated by a Gaussian distribution, but is now characterized by a set of points lying along the main eigenaxes of the GRV covariance matrix. Then, these so-called sigma-points propagate through the non-linear system (18). A weighted combination of the resulting values makes it possible to estimate the mean and the covariance matrix of the transformed random vector, i.e. the RV that undergoes the non-linear transformation. On one hand, the unscented Kalman filter is based on the unscented transformation. When the density is odd, the weights are chosen to provide the 2nd order Taylor expansion, around the mean of the RV. On the other hand, the CDKF is based on the 2nd order Sterling polynomial interpolation formula. The difference between CDKF and UKF stands in the way the mean and the covariance matrix of the transformed RV is calculated. For more details about SPKF algorithm description the reader is referred to [13].

The initialization parameters of the algorithm are: $\hat{\varepsilon}_{u,1}(-1) = 0$, and for $i \geq 2$ $\hat{\varepsilon}_{\tilde{u},i}(-1) = \hat{\varepsilon}_{\tilde{u},i-1}$.
In order to improve the Kalman estimation we consider a simple MAI cancellation strategy as in [7]. The results from the $(n-1)$th recursion are used to estimate and to eliminate different users' signals in the nth recursion.

$$\text{MAI estimation: } \hat{R}_{\tilde{u},i}^{(est)}(n) = \hat{A}_{\tilde{u},i}(n)e^{j2\pi n\hat{\varepsilon}_{\tilde{u},i}(n-1)k/N} \quad (19)$$

$$\text{MAI correction: } \hat{R}_{\tilde{u},i}(n) = R(n) - \sum_{j=1, j\neq\tilde{u}}^{U} \hat{R}_{\tilde{u},i}^{(est)}(n) \quad (20)$$

After some recursions the algorithm can estimate the value of the uth user CFO, which is denoted as $\hat{\varepsilon}_{\tilde{u},i}$.

4. SIMULATION RESULTS

We consider an OFDMA interleaved uplink system, which is composed of 4 users sharing 512 sub-carriers and with cyclic prefix $N_g = 128$. We suppose a transmission over a Rayleigh quasi-static frequency selective channel composed of 3 multipaths. QPSK is used to modulate the information bits. All the simulations are performed under the assumption of perfect knowledge of channel impulse response.
We propose to carry out two kinds of tests.
Test 1: we test our CFO estimation algorithm for 4 users with different CFO. The receiver is assumed to use a sufficient training period to perform the estimation. The users' CFO estimation errors are considered fixed during an OFDMA symbol, but they are modeled as independent zero-mean Gaussian random variables with a certain variance equal to the mean square error (MSE) that varies between the different

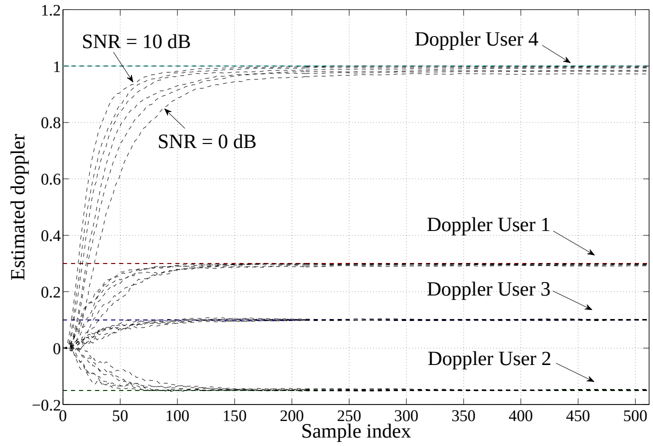

Figure 4: Kalman approach for CFO estimation with a known preamble

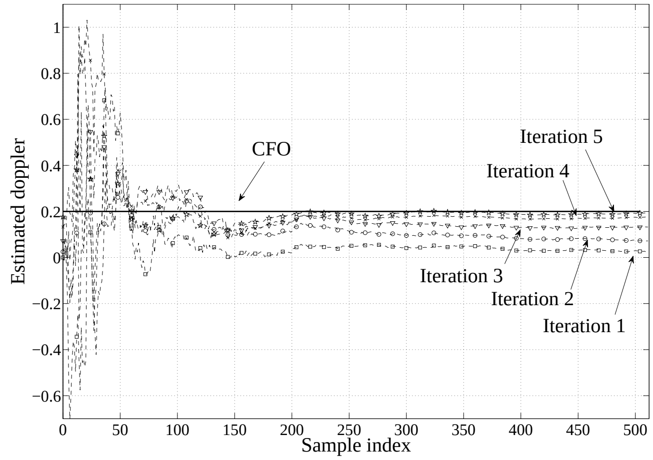

Figure 5: MMSE-SD combined with Kalman approach for CFO estimation, without preamble

OFDMA symbols. We have noticed that the EKF, UKF and CDKF provide very similar results. Therefore in figure 4 we only show the results obtained when using a CDKF. According to figure 4 we can notice that our approach makes it possible to accurately estimate the CFO in a recursive way.
Test 2: the users' CFO estimation errors are modeled as independent zero-mean Gaussian random variables with a variance of 0.35, that varies between the different OFDMA symbols, but they are considered fixed during an OFDMA symbol. Figure 5 shows the results. The proposed algorithm provides a "good" estimation of the CFO in an iterative way without the knowledge of a preamble. As expected, the first iteration leads to poor performance, but iterating especially up to 5 times our approach makes it possible to obtain good performance.
In figure 6 we show the performance of our algorithm in terms of BER, we clearly see that if the iteration increases, a better performance is obtained. One can notice a gain of approximately 12dB between the first and the fifth iteration.
To demonstrate the feasibility of the proposed scheme in a more realistic scenario, we model the users' CFOs as independent random variables uniformly distributed in the range (-0.05, 0.05) around an increasing linear CFO from 0.2 to 0.7. Figure 7 shows our estimation algorithm robustness for tracking a variable CFO. The multiple frequency offsets also have been assumed varying between OFDMA symbols. The proposed iterative scheme has the ability to track the fre-

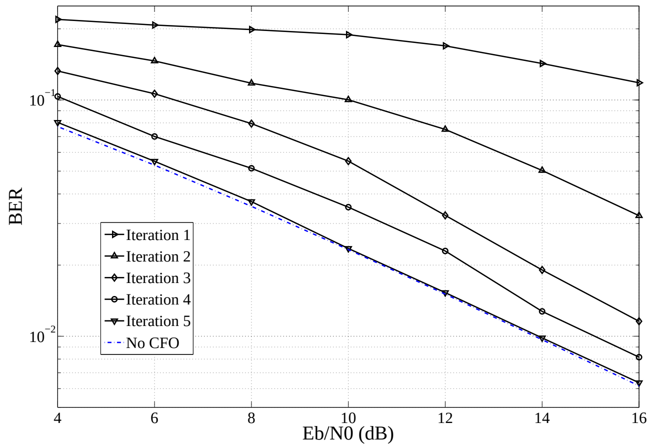

Figure 6: BER performance, random uniformly distributed CFO, 5 iterations

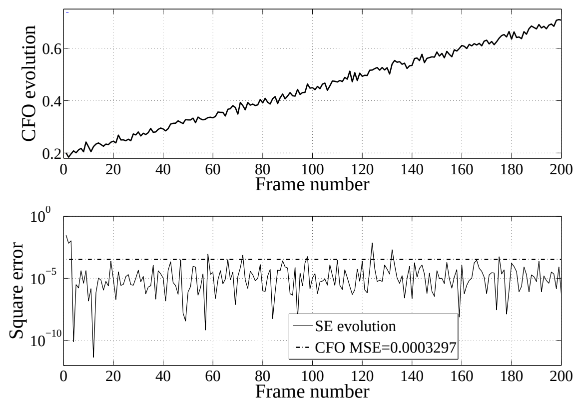

Figure 7: Carrier frequency offset tracking, SNR=4dB

quency offsets as depicted in figure 7.
The main difference with the existing CFO estimation systems is that our receiver is able to estimate the different user CFO without preambles.

5. CONCLUSIONS

The blind MAI suppression scheme for an interleaved OFDMA uplink transmission followed by the CFO estimation algorithm based on Kalman filter can jointly estimate and detect respectively each user CFO and frame, with no need of long training signal. This enables to maintain a maximum transmission data rate, particularly in the IEEE 802.16 standard context, where CFO is subject to change between two successive frames.
In addition, simulation results demonstrate that the proposed scheme can effectively suppress the MAI caused by a relatively large CFO, with sufficient robustness to CFO variations. The decoding of interleaved OFDMA is an ordered serial processing that combines interference suppression and interference cancelation techniques. The iterative decoding applies simple hard interference cancelation techniques, resulting moderate complexity.
The CFO estimation performance of three different Kalman approaches give quite similar results. When using the SPKF, there is no need to calculate Jacobians or Hessians. The computational complexity is moderate.

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