MULTIPATH-AWARE JOINT SYMBOL TIMING AND CFO ESTIMATION IN MULTIUSER OFDM/OQAM SYSTEMS

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ABSTRACT

In this paper we consider the problem of data-aided synchronization in the up-link of a multiple access (MA) orthogonal frequency division multiplexing (OFDM) system based on offset quadrature amplitude modulation (OQAM). In particular, the joint maximum-likelihood (ML) phase offset, carrier-frequency offset (CFO) and symbol timing (ST) estimator, exploiting a short known preamble embedded in the burst received from each of U users in multipath channel, is considered. Under the assumption that the CFO of each user is sufficiently small, the considered approach leads to U different approximate ML (AML) estimators; moreover, delays, amplitudes and phase-offsets are jointly estimated for each path. Specifically, the phase, amplitude and CFO estimators are in closed form while the AML ST estimators require a one-dimensional maximization procedure. The performance of the proposed joint AML estimator is assessed via computer simulations both in AWGN and multipath channel.

1. INTRODUCTION

In the last years, the interest for filter-bank multicarrier (FBMC) systems is increased, since they provide high spectral containment. Therefore, they have been taken into account for high-data-rate transmissions over both wired and wireless frequency-selective channels. One of the most famous multicarrier modulation techniques is orthogonal frequency division multiplexing (OFDM), embedded in several standards such as digital audio and video broadcasting or Wi-Fi wireless LANs IEEE 802.11a/g. Other known types of FBMC systems are Filtered Multitone (FMT) systems, that have been proposed for very high-speed digital subscriber line standards [1] and are under investigation also for broadband wireless applications [2] and, moreover, OFDM based on offset QAM modulation (OQAM) [3].

Unlike OFDM, OFDM/OQAM systems do not require the presence of a cyclic prefix (CP) in order to combat the effects of frequency selective channels. The absence of the CP implies on the one hand the maximum spectral efficiency and, on the other hand, an increased computational complexity. However, since the subchannel filters are obtained by complex modulation of a single filter, efficient polyphase implementations are possible. Another fundamental difference between OFDM and OFDM/OQAM systems is the adoption in the latter case of pulse shaping filters very well localized in time and frequency [4].

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OFDM/OQAM systems are more sensitive to synchronization errors than single-carrier systems. In particular, the presence of carrier frequency-offset (CFO) and symbol timing (ST, i.e., the delay of the first multipath component) estimation errors can lead to a severe performance degradation. For this reason, it is very important to derive efficient synchronization schemes. In the last years several studies have been focused on blind or data-aided CFO and ST estimation for OFDM/OQAM systems. For example, in [5] the authors proposed a blind CFO estimator by exploiting the conjugate second-order cyclostationarity of the transmitted OFDM/OQAM signal. The derived estimator assures a satisfactory performance only when a large number of OFDM/OQAM symbols is considered. The second-order cyclostationarity property of the OFDM/OQAM signal is also exploited in [6] to obtain a blind joint CFO and ST estimator, while in [7] both the conjugate and the unconjugate second-order cyclostationarity of the OFDM/OQAM signal is used to derive the joint maximum likelihood (ML) CFO and carrier phase estimator. Finally, in [8] a synchronization scheme for data-aided ST and CFO estimation with robust acquisition properties in dispersive channels is developed. However, all cited estimators suffer from the drawback that they cannot be exploited in up-link FBMC communications. Moreover, the estimator derived in [9] is proposed under the assumption of additive white Gaussian noise (AWGN) channel.

In this paper we consider the problem of data-aided synchronization in the up-link of a multiple access (MA) OFDM/OQAM system. In particular, the joint ML phase offset, CFO and ST estimator exploiting a short known preamble embedded in the burst received from each of U users is considered. The paper extends the results in [9] since the estimation procedure is derived according to a multipath channel model. Specifically, under the assumption that the CFO of each user is sufficiently small, the considered approach leads to U different approximate ML (AML) estimators; moreover, delays, amplitudes and phase-offsets are jointly estimated for each path. In particular, the phase, amplitude and CFO estimators are in closed form while the AML ST estimators require a one-dimensional maximization procedure. The proposed joint estimator, derived with reference to the more hostile uplink scenario, admits as special case the joint estimator proposed in [10] for the downlink scenario. The performance of the proposed joint AML estimator is assessed via computer simulations and compared with that achieved by the joint AML estimator proposed in [9]. The paper is organized as follows. In Section II the OFDM/OQAM system model is described. In Section III the proposed data-aided estimator is described. In Section IV the performance analysis obtained by computer simulations is presented and discussed. Finally, conclusions are drawn in Section V.

Notation: $j \triangleq \sqrt{-1}$, superscript $(\cdot)^*$ denotes the complex conjugation, $\Re[\cdot]$ real part, $|\cdot|$ absolute value, $(\cdot)^T$ transpose and $\angle[\cdot]$ the argument of a complex number in $[-\pi, \pi)$. Finally, lower case boldface symbols indicate column vectors.

2. SYSTEM MODEL

Let us consider an MA OFDM/OQAM system with U users and N subcarriers. The received signal in the up-link of a multipath channel with N_c paths, when the information-bearing signal of the $m-$th user $s_m(t)$ presents a timing offset $\tau_{m,i}$ over the i-th path (of gain $\gamma_{m,i}$), a CFO normalized to subcarrier spacing $\epsilon_m = \Delta f_m T$ and a carrier phase offset ϕ_m, can be written as

$$r(t) = \sum_{m=1}^{U} e^{j2\pi\Delta f_m t} \sum_{i=1}^{N_c} \gamma_{m,i} e^{j\phi_{m,i}} s_m(t-\tau_{m,i}) + n(t) \quad (1)$$

where $n(t)$ is a zero-mean complex-valued white Gaussian noise process with independent real and imaginary part, each with two-sided power spectral density N_0. The received signal $r(t)$ is filtered with an ideal low-pass filter with a bandwidth of $1/T_s$, where T_s denotes the sampling period. The sampled signal $s_m(kT_s)$ is equal to

$$s_m(kT_s) = \sqrt{\frac{N_u}{2N_m}} \left[s_m^R(kT_s) + j s_m^I(kT_s) \right] \quad (2)$$

with

$$s_m^R(kT_s) = \sum_{p=0}^{S-1} \sum_{\ell \in \mathscr{A}_m} a_{p,\ell}^R e^{j\ell\left(\frac{2\pi}{N}k + \frac{\pi}{2}\right)} g(kT_s - pT) \quad (3)$$

$$s_m^I(kT_s) = \sum_{p=0}^{S-1} \sum_{\ell \in \mathscr{A}_m} a_{p,\ell}^I e^{j\ell\left(\frac{2\pi}{N}k + \frac{\pi}{2}\right)} g(kT_s - T/2 - pT) \quad (4)$$

where $T = NT_s$ is the OFDM/OQAM symbol interval and S denotes the number of information-bearing symbols in the burst. In (3) and (4) $\mathscr{A}_m$ is the set of subcarriers of size N_m allocated to the $m-$th user, N_u is the number of active subcarriers, the sequences $a_{p,\ell}^R$ and $a_{p,\ell}^I$ indicate the real and imaginary part of the complex data symbols transmitted on the ℓth subcarrier during the pth OFDM/OQAM symbol, while $g(t)$ is the real transmitted pulse-shaping filter. Finally, we assume that the different delays of the paths of each multipath channel $\tau_{m,i}$ are sufficiently separated ($|\tau_{m,i_1} - \tau_{m,i_2}| > \sigma_\tau$) so that they can be resolved by means of the available samples.

3. JOINT SYMBOL TIMING AND CFO ESTIMATOR

In this section we consider the problem of data-aided synchronization in the up-link of an MA OFDM-OQAM system transmitting over multipath channel. In particular, we derive the joint ML ST, CFO and phase-offset estimator exploiting a short known preamble embedded in the burst received from each of the U users. Specifically, the known preamble of the mth user, is given by

$$z_m(kT_s) = \sqrt{\frac{N_u}{2N_m}} \sum_{p=0}^{L-1} \sum_{\ell \in \mathscr{A}_m} e^{j\ell\left[\frac{2\pi}{T}kT_s + \frac{\pi}{2}\right]} \times \left[a_{p,\ell}^R g(kT_s - pT) + j a_{p,\ell}^I g\left(kT_s - pT - T/2\right) \right] \quad (5)$$

where $a_{p,\ell}^R, a_{p,\ell}^I, 0 \le p \le L-1, \ell \in \mathscr{A}_m$, denote the known pilot symbols of the m-th user. By considering an observations window of total length $N\eta$ containing the non-zero support of the preamble received from each user, the log-likelihood function in multipath channel for the unknown parameters ϵ_m, $\tau_{m,i}$, $\gamma_{m,i}$ and $\phi_{m,i}$, $m = 1, \ldots, U$, and $i = 1, \ldots, N_c$ is given by (up to an irrelevant multiplicative factor)

$$\log\Lambda\left(\widetilde{\boldsymbol{\gamma}}, \widetilde{\boldsymbol{\phi}}, \widetilde{\boldsymbol{\tau}}, \widetilde{\boldsymbol{\varepsilon}}\right) = -\sum_{k=0}^{N\eta-1} \left| r(kT_s) - \sum_{m=1}^{U} e^{j\frac{2\pi}{N}\widetilde{\varepsilon}_m k} \sum_{i=1}^{N_c} \widetilde{\gamma}_{m,i} e^{j\widetilde{\phi}_{m,i}} z_m^{\widetilde{\tau}_{m,i}}(kT_s) \right|^2 \quad (6)$$

where $\widetilde{\boldsymbol{\gamma}} \triangleq [\widetilde{\boldsymbol{\gamma}}_1^T \ldots \widetilde{\boldsymbol{\gamma}}_U^T]^T$, $\widetilde{\boldsymbol{\gamma}}_m \triangleq [\widetilde{\gamma}_{m,1} \ldots \widetilde{\gamma}_{m,N_c}]^T$, $\widetilde{\boldsymbol{\phi}} \triangleq [\widetilde{\boldsymbol{\phi}}_1^T \ldots \widetilde{\boldsymbol{\phi}}_U^T]^T$, $\widetilde{\boldsymbol{\phi}}_m \triangleq [\widetilde{\phi}_{m,1} \ldots \widetilde{\phi}_{m,N_c}]^T$, $\widetilde{\boldsymbol{\tau}} \triangleq [\widetilde{\boldsymbol{\tau}}_1^T \ldots \widetilde{\boldsymbol{\tau}}_U^T]^T$, $\widetilde{\boldsymbol{\tau}}_m \triangleq [\widetilde{\tau}_{m,1} \ldots \widetilde{\tau}_{m,N_c}]^T$, $\widetilde{\boldsymbol{\varepsilon}} \triangleq [\widetilde{\varepsilon}_1 \ldots \widetilde{\varepsilon}_U]^T$,

$$\begin{aligned} z_m^\tau(kT_s) &\triangleq z_m(kT_s - \tau) \\ &= \sqrt{\frac{N_u}{2N_m}} \sum_{p=0}^{L-1} \sum_{\ell \in \mathscr{A}_m} e^{j\ell\left[\frac{2\pi}{T}(kT_s - \tau) + \frac{\pi}{2}\right]} \\ &\times \Big[a_{p,\ell}^R g(kT_s - pT - \tau) \\ &+ j a_{p,\ell}^I g\left(kT_s - pT - T/2 - \tau\right) \Big] \end{aligned} \quad (7)$$

and the notation of the type $\tilde{x}$ indicates trial value of x.

By replacing (7) in (6) and dropping irrelevant multiplicative and additive factors the log-likelihood function results to be equivalent to

$$\lambda\left(\widetilde{\boldsymbol{\gamma}}, \widetilde{\boldsymbol{\phi}}, \widetilde{\boldsymbol{\tau}}, \widetilde{\boldsymbol{\varepsilon}}\right) = \left\{ \underbrace{2\sum_{k=0}^{\eta N-1} \Re\{r(kT_s) r_I^*(k)\}}_{\triangleq A} - \underbrace{\sum_{k=0}^{\eta N-1} |r_I(k)|^2}_{\triangleq B} \right\} \quad (8)$$

where

$$r_I(k) \triangleq \sum_{m=1}^{U} e^{j\frac{2\pi}{N}\widetilde{\varepsilon}_m k} \sum_{i=1}^{N_c} \widetilde{\gamma}_{m,i} e^{j\widetilde{\phi}_{m,i}} z_m(kT_s - \widetilde{\tau}_{m,i}). \quad (9)$$

The term B in (8) can be easily re-written as

$$B = \sum_{m_1, m_2 = 1}^{U} \sum_{i_1, i_2 = 1}^{N_c} \widetilde{\gamma}_{m_1,i_1} \widetilde{\gamma}_{m_2,i_2} e^{j(\widetilde{\phi}_{m_1,i_1} - \widetilde{\phi}_{m_2,i_2})} C \quad (10)$$

with

$$C \stackrel{\triangle}{=} \sum_{k=0}^{\eta N-1} e^{j\frac{2\pi}{N}(\widetilde{\varepsilon}_{m_1}-\widetilde{\varepsilon}_{m_2})k} z_{m_1}(kT_s-\widetilde{\tau}_{m_1,i_1}) z_{m_2}^*(kT_s-\widetilde{\tau}_{m_2,i_2}). \tag{11}$$

Two approximations are considered now. The first approximation consists in noticing that $C \simeq 0$ for $m_1 \neq m_2$: it derives from the fact that the scalar product among the signals of two different users can be neglected since they are practically separated in frequency (the small offsets Δf_{m_1} and Δf_{m_2} are not able to introduce significant frequency overlapping). The second approximation consists in noticing that $C \simeq 0$ for $i_1 \neq i_2$, i.e., also the different delays of the same user are considered able to produce orthogonal signals since they are assumed to be sufficiently wideband. Here the assumption $|\tau_{m,i_1} - \tau_{m,i_2}| > \sigma_\tau$ for $i_1 \neq i_2$ is crucial. By using these two approximations, the term B is independent of the frequency offsets $\widetilde{\varepsilon}_m$ and of the phases $\widetilde{\phi}_{m,i}$:

$$B \simeq \sum_{m=1}^{U}\sum_{i=1}^{N_c} \widetilde{\gamma}_{m,i}^2 \sum_{k=0}^{\eta N-1} |z_m(kT_s-\widetilde{\tau}_{m,i})|^2. \tag{12}$$

Moreover, the term A in (8) can be easily re-written as

$$A = \sum_{m=1}^{U}\sum_{i=1}^{N_c} \Re\left\{ Z_m(\widetilde{\gamma}_{m,i}, \widetilde{\phi}_{m,i}, \widetilde{\tau}_{m,i}, \widetilde{\varepsilon}_m)\right\} \tag{13}$$

with

$$Z_m(\gamma,\phi,\tau,\varepsilon) \stackrel{\triangle}{=} \sqrt{\frac{N_u}{2N_m}} \gamma e^{-j\phi} c_m(\varepsilon,\tau) \tag{14}$$

where

$$c_m(\varepsilon,\tau) \stackrel{\triangle}{=} \sum_{\ell\in\mathscr{A}_m} e^{-j\ell\frac{\pi}{2}} e^{j\frac{2\pi}{T}\ell\tau} \times \sum_{p=0}^{L-1}\left[a_{p,\ell}^R w_p^{(\ell)}(\varepsilon,\tau) - j a_{p,\ell}^I w_p^{(\ell)}(\varepsilon,\tau+T/2)\right] \tag{15}$$

with

$$w_p^{(\ell)}(\varepsilon,\tau) \stackrel{\triangle}{=} \sum_{k=0}^{N\eta-1} r(kT_s) e^{-j\frac{2\pi}{N}\varepsilon k} e^{-j\frac{2\pi}{N}k\ell} g(kT_s-\tau-pT). \tag{16}$$

In order to approximate the term A in (13), let us assume that

$$e^{-j\frac{2\pi}{N}\varepsilon k} g(kT_s) \simeq g(kT_s) \tag{17}$$

which holds provided that the CFO of each user is sufficiently small:

$$e^{-j\frac{2\pi}{N}\varepsilon k} \simeq 1 \qquad \forall k \in \{0,1,\ldots,\rho N\} \tag{18}$$

where $\rho N T_s = \rho T$ is the length of the prototype filter $g(t)$. From (16) and (17) it follows that

$$w_p^{(\ell)}(\varepsilon,\tau) \simeq e^{-j2\pi\varepsilon p} e^{-j\frac{2\pi}{N}\varepsilon\theta} w_p^{(\ell)}(0,\tau). \tag{19}$$

Analogously, under the same assumption (17), it follows that

$$w_p^{(\ell)}\left(\varepsilon,\tau+\frac{T}{2}\right) \simeq e^{-j2\pi\varepsilon(p+\frac{1}{2})} e^{-j\frac{2\pi}{N}\varepsilon\theta} w_p^{(\ell)}\left(0,\tau+\frac{T}{2}\right). \tag{20}$$

From (15), (19), and (20), it follows that, for $L=1$,

$$c_m(\varepsilon,\tau) \simeq e^{-j\frac{2\pi}{N}\varepsilon\theta}\left\{P_m(\tau) + e^{-j\pi\varepsilon} Q_m(\tau)\right\} \tag{21}$$

with

$$P_m(\tau) \stackrel{\triangle}{=} \sum_{\ell\in\mathscr{A}_m} e^{-j\frac{\pi}{2}\ell} e^{j\frac{2\pi}{T}\ell\tau} a_{0,\ell}^R w_0^{(\ell)}(0,\tau) \tag{22}$$

and

$$Q_m(\tau) \stackrel{\triangle}{=} \sum_{\ell\in\mathscr{A}_m} e^{-j\frac{\pi}{2}(\ell+1)} e^{j\frac{2\pi}{T}\ell\tau} a_{0,\ell}^I w_0^{(\ell)}(0,\tau+T/2). \tag{23}$$

Using the approximation (12) for B, (8) can be re-written as

$$\sum_{m=1}^{U}\sum_{i=1}^{N_c}\left\{ b\Re\left\{\widetilde{\gamma}_{m,i} e^{-j\widetilde{\phi}_{m,i}} c_m(\widetilde{\varepsilon}_m,\widetilde{\tau}_{m,i})\right\} - \widetilde{\gamma}_{m,i}^2 d_m(\widetilde{\tau}_{m,i})\right\} \tag{24}$$

with $b \stackrel{\triangle}{=} 2\sqrt{\frac{N_u}{2N_m}}$, $d_m(\tau) \stackrel{\triangle}{=} \sum_{k=0}^{\eta N-1}|z_m(kT_s-\tau)|^2$. It follows that the approximate maximum-likelihood estimates (denoted with the superscript (AML)) are

$$\phi_{m,i}^{(AML)} = \angle c_m(\varepsilon_m^{(AML)}, \tau_{m,i}^{(AML)}) \tag{25}$$

$$\gamma_{m,i}^{(AML)} = \sqrt{\frac{N_u}{2N_m}} \frac{|c_m(\varepsilon_m^{(AML)}, \tau_{m,i}^{(AML)})|}{d_m(\tau_{m,i}^{(AML)})} \tag{26}$$

and

$$(\boldsymbol{\tau}_m^{(AML)}, \varepsilon_m^{(AML)}) = \arg\max_{\widetilde{\boldsymbol{\tau}}_m, \widetilde{\varepsilon}_m} \sum_{i=1}^{N_c} \frac{|c_m(\widetilde{\varepsilon}_m, \widetilde{\tau}_{m,i})|^2}{d_m(\widetilde{\tau}_{m,i})}. \tag{27}$$

Using the approximation (21) in (27) it follows that

$$\varepsilon_m^{(AML)} = \frac{1}{\pi}\angle\sum_{i=1}^{N_c} P_m^{(d)*}(\tau_{m,i}^{(AML)}) Q_m^{(d)}(\tau_{m,i}^{(AML)}) \tag{28}$$

and

$$\boldsymbol{\tau}_m^{(AML)} = \arg\max_{\widetilde{\boldsymbol{\tau}}_m} \left\{ \sum_{i=1}^{N_c}\left[|P_m^{(d)}(\widetilde{\tau}_{m,i})|^2 + |Q_m^{(d)}(\widetilde{\tau}_{m,i})|^2\right] + 2\left|\sum_{i=1}^{N_c} P_m^{(d)}(\widetilde{\tau}_{m,i}) Q_m^{(d)*}(\widetilde{\tau}_{m,i})\right|\right\} \tag{29}$$

with

$$P_m^{(d)}(\tau) \stackrel{\triangle}{=} \frac{P_m(\tau)}{d_m(\tau)} \qquad Q_m^{(d)}(\tau) \stackrel{\triangle}{=} \frac{Q_m(\tau)}{d_m(\tau)}. \tag{30}$$

We can see the overall estimation using (28) as an averaging over the different estimations according to the strength of each path. Consequently, if the considered paths are sufficiently strong the phase term $\angle P_m^{(d)*}(\widetilde{\tau}_{m,i}) Q_m^{(d)}(\widetilde{\tau}_{m,i})$ is practically independent of i and, then,

$$\left|\sum_{i=1}^{N_c} P_m^{(d)}(\widetilde{\tau}_{m,i}) Q_m^{(d)*}(\widetilde{\tau}_{m,i})\right| \simeq \sum_{i=1}^{N_c}\left|P_m^{(d)}(\widetilde{\tau}_{m,i}) Q_m^{(d)*}(\widetilde{\tau}_{m,i})\right|. \tag{31}$$

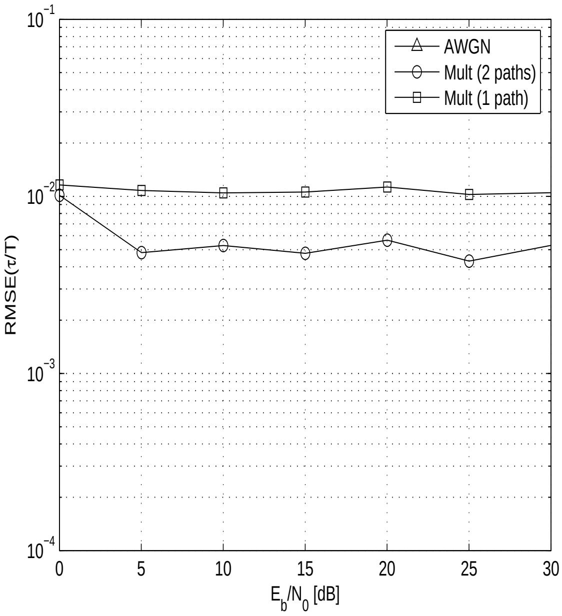

Figure 1: Performance of the proposed AML symbol timing estimator in AWGN and multipath channel.

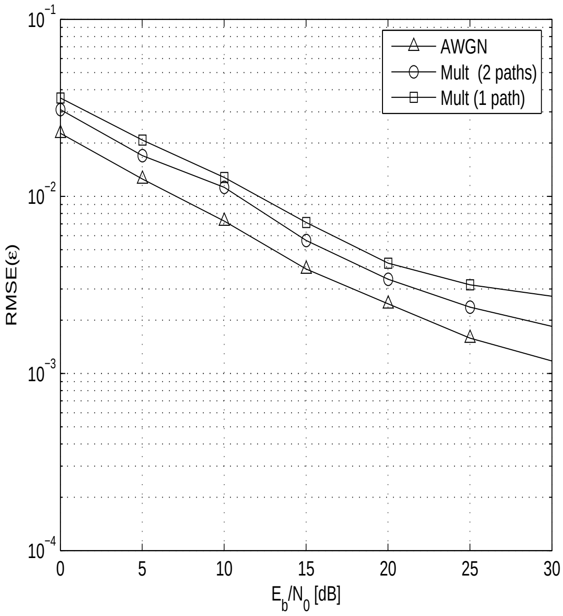

Figure 2: Performance of the proposed AML CFO estimator in AWGN and multipath channel.

Note that if in addition to the considered strong paths, also weaker paths are accounted for, the approximation (31) is still valid due to the contained contribution to the sum in the left hand side of (31). Hence, (29) becomes

$$\boldsymbol{\tau}_m^{(AML)} = \arg\max_{\widetilde{\boldsymbol{\tau}}_m} \sum_{i=1}^{N_c} \left\{ |P_m^{(d)}(\widetilde{\tau}_{m,i})| + |Q_m^{(d)}(\widetilde{\tau}_{m,i})| \right\}^2 . \tag{32}$$

Under the assumption $|\tau_{m,i_1} - \tau_{m,i_2}| > \sigma_\tau$, where σ_τ is related to the behavior of the functions $|P_m^{(d)}(\cdot)|$ and $|Q_m^{(d)}(\cdot)|$, the values of the delays $\tau_{m,i}^{(AML)}$ are the local maxima of $|P_m^{(d)}(\cdot)| + |Q_m^{(d)}(\cdot)|$, and, finally, the ST is the smallest among the estimated delays.
Note that the computational complexity of the proposed method is linear with N_c in (28) while it is practically independent of N_c in delay estimation.

4. NUMERICAL RESULTS AND COMPARISONS

In this section the performance of the proposed joint AML estimator in (28) and (32) is assessed via computer simulations. A number of 500 Monte Carlo trials has been performed under the following conditions (unless otherwise stated)

1. The considered MA OFDM/OQAM system has a bandwidth $B = 1/T_s = 44.8$ MHz and a total number of subcarriers $N = 1024$.
2. The data symbols $a_{p,l}^R$ and $a_{p,l}^I$ are the real and imaginary part of QPSK symbols.
3. The number of users is $U = 4$ each with $N_m = N_u/4 = 228$ active subcarriers.
4. The length of the considered prototype filter [11], designed with the frequency sampling technique, is ρN where the overlap parameter $\rho = 4$.
5. The value of E_b/N_0 for users 2,3, and 4 has been fixed at $E_b/N_0 = 16\,\text{dB}$. Moreover, their carrier phase, CFO and ST (assumed to be an integer multiple of the sampling period T_s) are uniformly distributed in $[-\pi, \pi)$, $[-0.5, 0.5)$ and $\{-T_s N/2, T_s N/2 - 1\}$, respectively.
6. The length of the training sequence is $L = 1$ OFDM/OQAM symbol; moreover, to reduce the interference due to the data symbols, the useful data in the whole burst is delayed with respect to the preamble of the burst by one OFDM/OQAM symbol interval.
7. The considered multipath channel model is the ITU Vehicular A [12], which has six multipaths with differential delays 0, 0.31, 0.71, 1.09, 1.73 and 2.51 μs and relative power 0, -1, -9, -10, -15, and $-20\,\text{dB}$. Moreover, the channel is fixed in each run but it is independent from one run to another.
8. Blockwise allocation scheme is adopted.

Note that although the proposed estimator has been derived by considering only the preamble of each user, in the simulations the burst of each user contains the exploited preamble and the information bearing data.

In the simulations we have tested the performance of the derived joint estimators with reference to the two choices: $N_c = 1$ and $N_c = 2$. Note that the choice $N_c = 1$ is equivalent to the joint synchronization algorithm proposed in [9] and, therefore, the two values of N_c are chosen in order to appreciate the advantages of the proposed extension.

Figure 1 displays the root mean square error (RMSE) (normalized to the OFDM/OQAM interval T) of the AML ST estimator for the first user as a function of E_b/N_0. Moreover, Fig. 2 shows the RMSE (normalized to the intercarrier spacing $1/T$) of the AML CFO estimator for the first user. The results show that the proposed estimators exploiting $N_c = 2$ paths outperforms the previously derived estimators almost in the whole range of E_b/N_0 values. Note that in

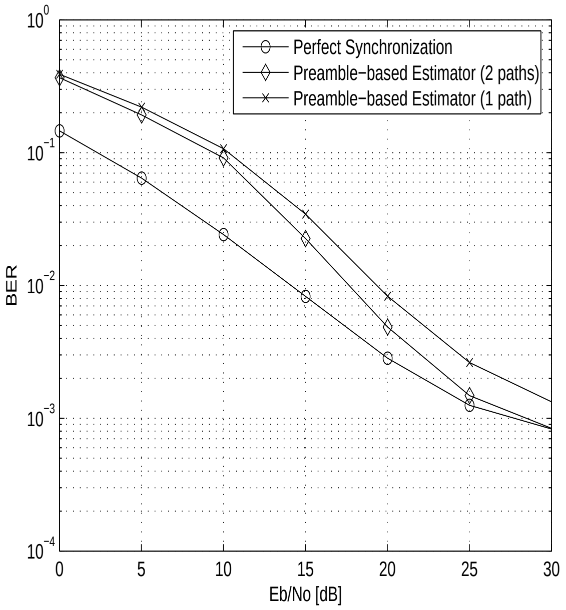

Figure 3: BER of the proposed joint estimator in multipath channel.

Fig. 1 is not reported the curve for AWGN since perfect ST estimates were obtained in all the performed simulation runs.

Finally, Fig. 3 reports the bit error rate (BER) obtained in multipath channel by exploiting the proposed joint AML algorithm and a one-tap equalizer with perfect knowledge of the channel and of the residual timing offset. Moreover, in order to appreciate the advantages of the proposed algorithms, we have not compensated at all the residual frequency offset error; however, to avoid to underestimate the performance of the actual receiver, we have evaluated by computer simulation the bit-error rate only on the sixth transmitted multicarrier symbol of the burst of the first user. Note that since 500 Monte Carlo trials have been performed, the BER has been estimated using $2 \times N_m \times 500 = 228 \times 10^3$ bits. The results show that in the considered scenario the proposed algorithm with the choice $N_c = 2$ can assure a reduced BER in comparison to the case $N_c = 1$ and, moreover, the obtained BER is nearly equal to that achieved in the case of perfect synchronization for moderate and high values of $\frac{E_b}{N_0}$.

5. CONCLUSIONS

In this paper we have dealt with the problem of data-aided synchronization for MA OFDM/OQAM systems. In particular, we have extended to the case of multipath channel the joint ML estimator for the phase offset, the CFO and the ST of each of U users derived in [9]. The extension has preserved the nice properties of the estimators in [9]; in particular, when the CFO of each user is sufficiently small, the derived approach leads to U different AML joint phase offset, CFO and ST estimators. More specifically, for each user the phase estimate and the CFO estimate are in closed form while ST estimate requires a one-dimensional maximization procedure. The performance of the proposed AML joint estimator has been assessed via computer simulations. The numerical results have shown that with only one training symbol (and the data burst delayed by one OFDM/OQAM symbol with respect to the preamble), the adoption of the proposed extension assures a contained performance degradation also when the frequency-domain processing devoted to the residual CFO compensation is limited.

REFERENCES

[1] G. Cherubini, E. Eleftheriou, S. Ölccer, and J.M. Cioffi, "Filter bank modulation techniques for very high-speed digital subscriber lines," *IEEE Commun. Mag.*, vol. 38, pp. 98-104, May 2000.

[2] T. Ihalainen, T. Hidalgo Stitz, M. Rinne, and M. Renfors, "Channel equalization in filter bank based multicarrier modulation for wireless communications," *EURASIP Journal on Advances in Signal Processing*, vol. 2007 (2007), Article ID 49389.

[3] D. Lacroix, N. Goudard, and M. Alard, "OFDM with Guard Interval Versus OFDM/OffsetQAM for High Data Rate UMTS Downlink Transmission," Proc. of *VTC'01* Fall, Atlantic City, NJ, USA, October 2001.

[4] P. Siohan, C. Siclet and N. Lacaille, "Analysis and design of OFDM/OQAM systems based on filterbank theory," in *IEEE Trans. Signal Processing*, vol. 50, no. 5, pp. 1170-1174, May 2002.

[5] P. Ciblat and E. Serpedin, "A fine blind frequency offset estimator for OFDM/OQAM systems," *IEEE Trans. Signal Processing*, vol. 52, pp. 291-296, January 2004.

[6] H. Bölcskei, "Blind estimation of symbol timing and carrier frequency offset in wireless OFDM systems," *IEEE Trans. Commun.*, vol. 49, pp. 988-999, June 2001.

[7] T. Fusco and M. Tanda, "Blind frequency-offset estimation for OFDM/OQAM systems," *IEEE Trans. Signal Processing*, vol. 55, pp. 1828-1838, May 2007.

[8] T. Fusco, A. Petrella and M. Tanda, "Data-aided symbol timing and CFO synchronization for filter-bank multicarrier systems," *IEEE Transactions on Wireless Communications*, vol. 8, no. 5, pp. 2705-2715, May 2009.

[9] T. Fusco, A. Petrella and M. Tanda, "Joint symbol timing and CFO estimation in multiuser OFDM/OQAM systems," Proc. of *SPAWC 2009*, Perugia, Italy, June 21-24, 2009.

[10] T. Fusco, A. Petrella and M. Tanda, "Joint symbol timing and CFO estimation for OFDM/OQAM systems in multipath channels," *EURASIP Journal on Advances in Signal Processing*, vol. 2010 (2010), Article ID 897607.

[11] M. Bellanger, "Specification and design of a prototype filter for a filter bank based multicarrier transmission," Proc. of *ICASSP'01*, vol. 4, 2001.

[12] Recommendation ITU-R M. 1225: "Guidelines for evaluation of radio transmission technologies for IMT-2000," 1997.