

Channel estimation methods for preamble-based OFDM/OQAM modulations

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Abstract—OFDM/OQAM is a special type of multi-carrier modulation that can be considered as an alternative to conventional OFDM with cyclic prefix (CP) for transmission over multi-path fading channels. Indeed, as it requires no guard interval, it has the advantage of a theoretically higher spectral efficiency. Furthermore if the pulse shape is well-localized in frequency, the resulting OFDM/OQAM signal satisfies stringent spectrum requirements. However, the classical channel estimation methods used for OFDM cannot be directly applied to OFDM/OQAM. In this paper we present an analysis of this problem and we describe a method for “pseudo” perfect channel estimation in OFDM/OQAM. Then we introduce two new realistic preamble-based channel estimation methods. The performance results are obtained by considering an IEEE 802.22 channel model with OFDM/OQAM and the proposed channel estimation methods are compared to the ones obtained with CP-OFDM.

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is an efficient Multi Carrier Modulation (MCM) to fight against multi-path fading channels. However its robustness to multi-path propagation effects comes from the insertion of a Guard Interval (GI) and is therefore obtained at the price of a reduced spectral efficiency. As furthermore the rectangular OFDM symbols lead to a $\sin(x)/x$ frequency spectrum, several alternatives have been researched to find better MCM schemes w.r.t. the frequency and/or time-frequency localization criteria.

As suggested in [1]–[3], OFDM/OQAM is a MCM modulation scheme, which may be the appropriate alternative. In OFDM/OQAM each subcarrier is modulated with a staggered Offset Quadrature Amplitude Modulation (OQAM). The basic concepts of OFDM/OQAM can be found in [4], [5]. But it is more recently [1] that OFDM/OQAM has been presented as a viable alternative to OFDM that could also be based on fast Fourier transform algorithms. Compared to conventional OFDM that transmits complex-valued symbols at a given symbol rate, OFDM/OQAM transmits real-valued symbols at twice this symbol rate and therefore has a similar spectral efficiency. Furthermore, in practice, it may provide a higher useful bit rate, since it operates without the addition of a guard interval, together with a pulse shaping that can be optimized according to given channel characteristics. However all the nice features of OFDM/OQAM come at the price of a relaxation of the orthogonality conditions that only hold in the real field. Consequently existing OFDM channel estimation methods cannot be directly applied in the case of OFDM/OQAM signals. Then, indeed a specific problem of imaginary intrinsic interference has to be solved.

Solutions have already been proposed for preamble-based [6] and scattered-based [7] channel estimation for OFDM/OQAM. Our aim here is to propose two new approaches leading to a reduced complexity. Furthermore, differently from [6], with one of them there is no need of an a priori knowledge of the pulse shaping function.

In section II, we give a short description of the continuous-time OFDM/OQAM modulation. Then, in section III, we provide an overview of the specific problem related to channel estimation for the OQAM modulation. Section IV is devoted to the presentation of the perfect channel estimation method for OFDM/OQAM and of the proposed preamble-based realistic estimation approaches. Finally, in section V, we compare in an IEEE802.22 context, these different channel OFDM/OQAM estimation algorithms to the ones obtained using a conventional OFDM modulation with Cyclic Prefix (CP-OFDM).

II. THE OFDM/OQAM MODULATION

We can write the baseband equivalent of a continuous-time OFDM/OQAM signal as follows [1]:

$$s(t) = \sum_{m=0}^{M-1} \sum_{n \in \mathbf{Z}} a_{m,n} \underbrace{g(t - n\tau_0) e^{j2\pi m F_0 t} e^{j\phi_{m,n}}}_{g_{m,n}(t)} \quad (1)$$

with M an even number of sub-carriers, $F_0 = 1/T_0 = 1/2\tau_0$ the subcarrier spacing, g the pulse shape and $\phi_{m,n}$ an additional phase term. The transmitted symbols $a_{m,n}$ are real-valued. They are obtained from a 2^{2K} -QAM constellation, taking the real and imaginary parts of these complex-valued symbols of duration $T_0 = 2\tau_0$, where τ_0 denotes the time offset between the two parts [1]–[3], [8].

Assuming a distortion-free channel, perfect reconstruction of real symbols is obtained owing to the following real orthogonality condition:

$$\Re \{ \langle g_{m,n} | g_{p,q} \rangle \} = \Re \left\{ \int g_{m,n}(t) g_{p,q}^*(t) dt \right\} = \delta_{m,p} \delta_{n,q}, \quad (2)$$

where, $\delta_{m,p} = 1$ if $m = p$ and $\delta_{m,p} = 0$ if $m \neq p$. For concision purpose we set $\langle g \rangle_{m,n}^{p,q} = -j \langle g_{m,n} | g_{p,q} \rangle$, with $\langle g_{m,n} | g_{p,q} \rangle$ a pure imaginary term for $(m,n) \neq (p,q)$.

However in practice for transmission over a realistic channel, the orthogonality property is lost, leading to ISI (Inter Symbol Interference) and ICI (Inter Carrier Interference). We will show that if the prototype filter g has good localization properties in time and frequency domains, a simple one tap

equalization process may be sufficient to restore the real orthogonality. However, this equalization requires channel estimates that are complex-valued. As the orthogonality is limited to the real field a specific estimation procedure has to be carried out. Recall that in practical OFDM systems the complex orthogonality is insured thanks to the introduction of a cyclic prefix (CP) of length Δ .

III. PROBLEM STATEMENT

Generally the CP-OFDM system is dimensioned in order to get a flat channel characteristic leading to the possibility of a simple zero forcing (ZF) equalization. Here we place ourselves in a similar context for OFDM/OQAM, we also call OQAM in short.

A. The channel model

The transmitted OFDM/OQAM signal is filtered by a channel with impulse response denoted $h(t, \tau)$ and we also consider the presence of an additive noise denoted by $\eta(t)$. Then the baseband version of the received signal can be written as follows:

$$y(t) = (h * s)(t) + \eta(t) \quad (3)$$

where $*$ denotes the convolution operation.

Denoting by $y_{wn}(t)$ the channel output without any noise component, and by Δ the maximum delay spread of the channel, we get:

$$y_{wn}(t) = \sum_{n=-\infty}^{\infty} \sum_{m=0}^{M-1} a_{m,n} \int_0^{\Delta} h(t, \tau) g_{m,n}(t - \tau) d\tau. \quad (4)$$

This expression can be expanded as follows:

$$y_{wn}(t) = \sum_{n=-\infty}^{\infty} \sum_{m=0}^{M-1} a_{m,n} e^{j\phi_{m,n}} e^{2j\pi m F_0 t} \times \int_0^{\Delta} h(t, \tau) g(t - \tau - n\tau_0) e^{-2j\pi m F_0 \tau} d\tau.$$

Let denote by T_g the length of the prototype filter with $T_g = kT_0$ and $k \geq 1$. As an example for the IOTA prototype [1], k is generally taken equal to 4. Then we assume that we have a flat fading channel at each subcarrier, which means that, $\frac{1}{T_g}$ is less than the coherence bandwidth of the channel $B_c \approx \frac{1}{2\Delta}$ [9]. Therefore the prototype function has relatively low variations in time over the interval $[0, \Delta]$ that is, $g(t - \tau - n\tau_0) \approx g(t - n\tau_0)$ for $\tau \in [0, \Delta]$. Then we get:

$$y_{wn}(t) = \sum_{n=-\infty}^{\infty} \sum_{m=0}^{2N-1} a_{m,n} g_{m,n}(t) H_m^{(c)}(t),$$

with

$$H_m^{(c)}(t) = \int_0^{\Delta} h(t, \tau) e^{-2j\pi m F_0 \tau} d\tau,$$

where $H_m^{(c)}(t)$ represents the complex response of the channel at instant t , if the channel is constant during the all duration

of the prototype then $H_m^{(c)}(t) = H_{m,n}^{(c)}$ and we get:

$$y(t) = \sum_{n=-\infty}^{\infty} \sum_{m=0}^{2N-1} a_{m,n} g_{m,n}(t) H_{m,n}^{(c)} + \eta(t)$$

If the channel is only constant on a short period of time, e.g. during $2\tau_0$, as assumed for CP-OFDM, or over $2\tau_0$, this last equality can be nevertheless approximately satisfied if the prototype function has most of its energy concentrated in a reduced time interval.

B. Demodulation and ZF equalisation

The demodulation of the received signal at the (m_0, n_0) position, noise taken apart, provides a complex symbol given by $y_{m_0, n_0}^{(c)} = \langle y | g_{m_0, n_0} \rangle$. After computations, we get the following expression:

$$y_{m_0, n_0}^{(c)} = H_{m_0, n_0}^{(c)} a_{m_0, n_0} + j \sum_{(p,q) \neq (0,0)} a_{m_0+p, n_0+q} H_{m_0+p, n_0+q}^{(c)} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \quad (5)$$

Let us first assume that the channel is ideal, that is $h(t, \tau) = \delta(t)$, i.e $H_{m,n}^{(c)} = 1$. In this particular case, (5) rewrite as:

$$y_{m_0, n_0}^{(c)} = a_{m_0, n_0} + j \sum_{(p,q) \neq (0,0)} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}$$

Recall from (2) that, $\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}$ is a pure real term, then the estimate of the transmitted symbol is obtained by:

$$\hat{a}_{m_0, n_0} = \Re \left\{ y_{m_0, n_0}^{(c)} \right\} = a_{m_0, n_0}. \quad (6)$$

Using similar principles for a realistic transmission conditions, the one tap ZF equalized signal can be written as:

$$\frac{y_{m_0, n_0}^{(c)}}{H_{m_0, n_0}^{(c)}} = a_{m_0, n_0} + I_{m_0, n_0} \quad (7)$$

where,

$$I_{m_0, n_0} = j \sum_{(p,q) \neq (0,0)} a_{m_0+p, n_0+q} \frac{H_{m_0+p, n_0+q}^{(c)}}{H_{m_0, n_0}^{(c)}} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}$$

is a complex-valued term. Then the estimated symbol is given by:

$$\begin{aligned} \hat{a}_{m_0, n_0} &= \Re \left\{ \frac{y_{m_0, n_0}^{(c)}}{H_{m_0, n_0}^{(c)}} \right\} = a_{m_0, n_0} \\ &- \sum_{(p,q) \neq (0,0)} a_{m_0+p, n_0+q} \Im \left\{ \frac{H_{m_0+p, n_0+q}^{(c)}}{H_{m_0, n_0}^{(c)}} \right\} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \\ &= a_{m_0, n_0} + \Re \{ I_{m_0, n_0} \}. \end{aligned}$$

$\Re \{ I_{m_0, n_0} \}$ represents the Inter-Symbol Interference (ISI) between the successive real symbols. At this point, due to ISI, it seems difficult to get an accurate estimation of a_{m_0, n_0} . However, we are going to show under a few approximations that: $\Re \{ I_{m_0, n_0} \} \approx 0$, leading to reliable estimation of a_{m_0, n_0} .

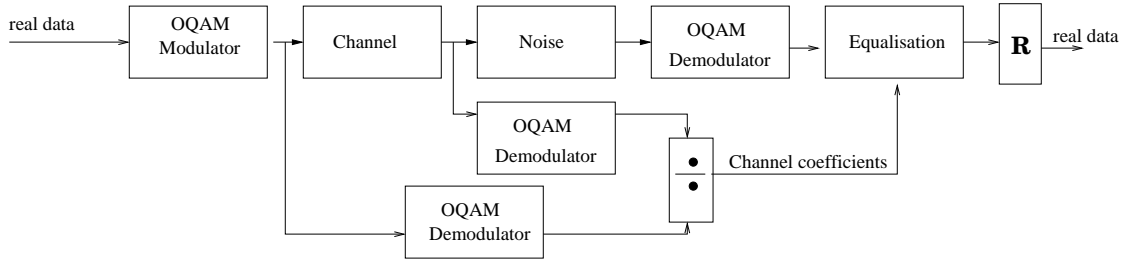


Fig. 1. Channel communication model for a perfect channel estimation with OQAM.

Let us define a neighborhood $\Omega_{\Delta m, \Delta n}$ of a given point (m_0, n_0) such that:

$$\Omega_{\Delta m, \Delta n} = \left\{ (p, q), |p| \leq \Delta m, |q| \leq \Delta n \mid H_{m_0+p, n_0+q}^{(c)} \approx H_{m_0, n_0}^{(c)} \right\},$$

and let us also set $\Omega_{\Delta m, \Delta n}^* = \Omega_{\Delta m, \Delta n} - (0, 0)$.

At this step we have to notice that Δn and Δm are chosen according to the coherence time (T_c) and bandwidth (B_c). It is worth mentioning that when B_c (resp. T_c) decreases, Δm (resp. Δn) also decreases. For well-dimensioned real systems, B_c encompasses a few sub-carriers ($\Delta m \geq 1$) and T_c is generally bigger than T_0 ($\Delta n \geq 1$). This allows us to write (7) as:

$$\begin{aligned} \frac{y_{m_0, n_0}^{(c)}}{H_{m_0, n_0}^{(c)}} &= a_{m_0, n_0} \\ &+ j \sum_{(p, q) \in \Omega_{\Delta m, \Delta n}^*} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \\ &+ j \sum_{(p, q) \notin \Omega_{\Delta m, \Delta n}^*} a_{m_0+p, n_0+q} \frac{H_{m_0+p, n_0+q}^{(c)}}{H_{m_0, n_0}^{(c)}} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \end{aligned} \quad (8)$$

Let us consider at first the case of a prototype function being well localized in time and frequency. Then, with increasing $|p|$ and $|q|$, $\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}$ becomes very close to zero. As an example with the IOTA function for $(p, q) \notin \Omega_{1,1}$ we have:

$$|\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}| < 0.04 \text{ and}$$

$$\frac{\sum_{(p, q) \notin \Omega_{1,1}} [\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}]^2}{\sum_{(p, q) \in \Omega_{1,1}^*} [\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}]^2} \approx 0.02.$$

Consequently, we also have:

$$\begin{aligned} &\left| \sum_{(p, q) \notin \Omega_{\Delta m, \Delta n}} a_{m_0+p, n_0+q} \frac{H_{m_0+p, n_0+q}^{(c)}}{H_{m_0, n_0}^{(c)}} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \right| \\ &\ll \left| \sum_{(p, q) \in \Omega_{\Delta m, \Delta n}^*} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \right|. \end{aligned}$$

Then (8) can be rewritten as:

$$\begin{aligned} \frac{y_{m_0, n_0}^{(c)}}{H_{m_0, n_0}^{(c)}} &\approx a_{m_0, n_0} \\ &+ j \sum_{(p, q) \in \Omega_{\Delta m, \Delta n}^*} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}. \end{aligned} \quad (9)$$

Let us denote by $ja_{m_0, n_0}^{(i)}$ the second term in (9), with

$$a_{m_0, n_0}^{(i)} = \sum_{(p, q) \in \Omega_{\Delta m, \Delta n}^*} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}. \quad (10)$$

As $a_{m_0, n_0}^{(i)}$ is a pure real term, taking the real part in (9) we get:

$$\hat{a}_{m_0, n_0} = \Re \left\{ \frac{y_{m_0, n_0}^{(c)}}{H_{m_0, n_0}^{(c)}} \right\} \approx a_{m_0, n_0}. \quad (11)$$

So the ISI can be neglected in practice for well-localized prototype functions. Therefore we can have an accurate detection of a_{m_0, n_0} as long as we know the channel coefficient $H_{m_0, n_0}^{(c)}$ at the receiver side. In practice we have to estimate $H_{m_0, n_0}^{(c)}$. From (9) and (10) we can write:

$$y_{m_0, n_0}^{(c)} \approx H_{m_0, n_0}^{(c)} (a_{m_0, n_0} + ja_{m_0, n_0}^{(i)}). \quad (12)$$

Let us assume that a pilot symbol a_{m_0, n_0} is transmitted at a position (m_0, n_0) that is known by the receiver. The value of $a_{m_0, n_0}^{(i)}$ defined in (10) depends on the unknown data that are transmitted around the pilot position, then the channel value $H_{m_0, n_0}^{(c)}$ cannot be directly derived from (12).

IV. THE PROPOSED CHANNEL ESTIMATION METHODS

A. Pseudo perfect estimation

Assuming now for the rest of this paper that the channel is locally time and frequency invariant and that the prototype function is well localized in time and frequency. We can deduce from our previous analysis, that at any time, n , and for any subcarrier, m , the received signal can be accurately approximated by:

$$y_{m, n}^{(c)} \approx H_{m, n}^{(c)} (a_{m, n} + ja_{m, n}^{(i)}). \quad (13)$$

For an ideal distortion-free channel (13) rewrites as

$$y_{m, n}^{(I)} \approx a_{m, n} + ja_{m, n}^{(i)} \quad (14)$$

Therefore the channel coefficient $H_{m, n}^{(c)}$ is the ratio between (13) and (14). So differently from the OFDM system, here the channel coefficients are only approximated. Thus, we only provide a ‘‘pseudo’’ perfect channel estimation. Figure 1 gives the block diagram to perform ‘‘pseudo’’ perfect channel estimation in OQAM. This block diagram is only for simulation purpose and cannot be used in a realistic transmission context. However we can derive from this ideal case a realistic channel estimation method. It is a preamble-based method that we have called Interference Approximation Method (IAM).

B. Interference approximation method (IAM)

The preamble is assumed to be perfectly known at the receiver and for the OQAM modulation its length is taken equal at least to $3\tau_0$. As for OFDM the preamble can be limited to one complex symbol duration $2\tau_0$, it seems at first that the OQAM involves more overhead than OFDM. Indeed to get the channel coefficients for all frequencies $3M$ real data are required for OQAM and only $2M$ for OFDM, i.e. there is a loss of $M/2$ complex data for OQAM. But in the case of a classical frame-by-frame transmission mode the situation is different if we compare with CP-OFDM. Indeed for a frame of length Q , there is a loss of ΔQ complex data due to the CP. Situations where $\Delta Q \geq (M/2)$, giving the advantage to OQAM, may be often encountered in practice. But let us now examine more in details the IAM method, for a_{m_0, n_0} symbols located in the middle of the preamble, a_{m_0, n_0-1} and a_{m_0, n_0+1} are also part of the preamble so the imaginary term can be approximated by:

$$a_{m_0, n_0}^{(i)} \approx \sum_{(p,q) \in \Omega_{\Delta^1, \Delta^1}^*} a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}. \quad (15)$$

Therefore, noise taken apart, the channel is estimated by:

$$H_{wn, m_0, n_0}^{(c)} = \frac{y_{m_0, n_0}^{(c)}}{(a_{m_0, n_0} + ja_{m_0, n_0}^{(i)})} \quad (16)$$

In the presence of a noise $n^{(c)}$ the channel estimation becomes:

$$H_{m_0, n_0}^{(c)} = H_{wn, m_0, n_0}^{(c)} + \frac{\eta^{(c)}}{(a_{m_0, n_0} + ja_{m_0, n_0}^{(i)})} \quad (17)$$

Then the greater the power of $(a_{m_0, n_0} + ja_{m_0, n_0}^{(i)})$ the better the estimation will be. We assume at first, for a variant named IAM1, that the real-valued coefficients, $a_{m, n}$, of the OQAM are i.i.d. with a variance of σ_a^2 . That means the power of the complex OFDM pilot symbol has to be equal to $2\sigma_a^2$ without boosting. In fact in OQAM/IAM1 we consider that we have a ‘‘pseudo pilot’’ $b_{m, n} = a_{m, n} + ja_{m, n}^{(i)}$ such that:

$$\begin{aligned} E(b_{m, n}) &= 0 \\ E(|b_{m, n}|^2) &= \sigma_a^2 \left(1 + \sum_{(p,q) \in \Omega_{\Delta^m, \Delta^n}^*} |\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}|^2 \right) \end{aligned}$$

In appendix it is shown that for a real-valued discrete-time orthogonal prototype, we have:

$$\sum_{(p,q) \neq (0,0)} [\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}]^2 = 1 \quad (18)$$

Then, as already explained, for a prototype g being well localized in time and frequency $E(|b_{m, n}|^2) \approx 2\sigma_a^2$ which shows that the pilot power is nearly the same for OQAM and OFDM. On the other hand, the preamble can be chosen in order to get a maximum power of the interference component $a_{m, n}^{(i)}$. Indeed if in (15) all the $a_{m_0+p, n_0+q} \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}$ terms

have all the same sign, then

$$\begin{aligned} E(|a_{m, n}^{(i)}|^2) &= \sigma_a^2 \left(\sum_{(p,q) \in \Omega_{\Delta^m, \Delta^n}^*} |\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}| \right)^2 \\ &\geq \sigma_a^2 \sum_{(p,q) \in \Omega_{\Delta^m, \Delta^n}^*} |\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0}|^2 \\ &\geq \sigma_a^2 \end{aligned}$$

Therefore, as the ‘‘pseudo’’ pilot power is higher, the channel estimation will be of course improved. This second variant of the interference approximation method is denoted IAM2.

C. Estimation using pairs of real pilots (POP)

This second method is based on the insertion of a pair of real-valued pilots (POP) at positions known by the receiver. In practice they can be placed at two consecutive time positions having the same subcarrier index and where the channel is quasi-invariant. Let us denote by $(m_1, n_1), (m_2, n_2)$ the two reference symbol positions. We have:

$$\begin{cases} y_{m_1, n_1}^{(c)} = H_{m_1, n_1}^{(c)} (a_{m_1, n_1} + ja_{m_1, n_1}^{(i)}) \\ y_{m_2, n_2}^{(c)} = H_{m_2, n_2}^{(c)} (a_{m_2, n_2} + ja_{m_2, n_2}^{(i)}) \end{cases} \quad (19)$$

We introduce a parameter C that is the ratio between the imaginary and real part of $H_{m_1, n_1}^{(c)}$, i.e. $H_{m_1, n_1}^{(c)} = H_{m_1, n_1}^{(r)} + jH_{m_1, n_1}^{(i)}$ and $C = H_{m_1, n_1}^{(i)} / H_{m_1, n_1}^{(r)}$. Then taking the real and imaginary part of each equation in (19), we get:

$$\begin{cases} y_{m_1, n_1}^{(r)} = H_{m_1, n_1}^{(r)} a_{m_1, n_1} - CH_{m_1, n_1}^{(r)} a_{m_1, n_1}^{(i)} \\ y_{m_1, n_1}^{(i)} = CH_{m_1, n_1}^{(r)} a_{m_1, n_1} + H_{m_1, n_1}^{(r)} a_{m_1, n_1}^{(i)} \\ y_{m_2, n_2}^{(r)} = H_{m_2, n_2}^{(r)} a_{m_2, n_2} - CH_{m_2, n_2}^{(r)} a_{m_2, n_2}^{(i)} \\ y_{m_2, n_2}^{(i)} = CH_{m_2, n_2}^{(r)} a_{m_2, n_2} + H_{m_2, n_2}^{(r)} a_{m_2, n_2}^{(i)} \end{cases} \quad (20)$$

As the channel is invariant we set $H_{m_2, n_2}^{(c)} = H_{m_1, n_1}^{(c)}$ and by a combination of the above equations we get:

$$\begin{cases} X_1 = a_{m_2, n_2} y_{m_1, n_1}^{(r)} - a_{m_1, n_1} y_{m_2, n_2}^{(r)} \\ = CH_{m_1, n_1}^{(r)} (-a_{m_2, n_2} a_{m_1, n_1} + a_{m_1, n_1} a_{m_2, n_2}^{(i)}) \\ X_2 = a_{m_2, n_2} y_{m_1, n_1}^{(i)} - a_{m_1, n_1} y_{m_2, n_2}^{(i)} \\ = H_{m_1, n_1}^{(r)} (a_{m_2, n_2} a_{m_1, n_1}^{(i)} - a_{m_1, n_1} a_{m_2, n_2}^{(i)}) \end{cases} \quad (21)$$

Then the X_1/X_2 ratio gives:

$$C = \frac{a_{m_2, n_2} y_{m_1, n_1}^{(r)} - a_{m_1, n_1} y_{m_2, n_2}^{(r)}}{a_{m_1, n_1} y_{m_2, n_2}^{(i)} - a_{m_2, n_2} y_{m_1, n_1}^{(i)}} \quad (22)$$

Then the first two equations in (20) rewrites as:

$$\begin{cases} y_{m_1, n_1}^{(r)} = H_{m_1, n_1}^{(r)} a_{m_1, n_1} - CH_{m_1, n_1}^{(r)} a_{m_1, n_1}^{(i)} \\ Cy_{m_1, n_1}^{(i)} = C^2 H_{m_1, n_1}^{(r)} a_{m_1, n_1} + CH_{m_1, n_1}^{(r)} a_{m_1, n_1}^{(i)} \end{cases}$$

The sum of the two above equations gives:

$$H_{m_1, n_1}^{(r)} a_{m_1, n_1} (1 + C^2) = y_{m_1, n_1}^{(r)} + Cy_{m_1, n_1}^{(i)}$$

Therefore the real component of the channel is:

$$H_{m_1, n_1}^{(r)} = \frac{y_{m_1, n_1}^{(r)} + Cy_{m_1, n_1}^{(i)}}{a_{m_1, n_1} (1 + C^2)}$$

and its imaginary part is given by:

$$H_{m_1, n_1}^{(i)} = CH_{m_1, n_1}^{(r)}$$

So, we have estimated the channel coefficients. This method, that does not require any knowledge of the prototype function, can be used as preamble-based method and also as a scattered-based channel estimation method.

But let us now examine its behavior in presence of a noise term. Then (20) becomes:

$$\begin{cases} Y_1 = CH_{m_1, n_1}^{(r)} (-a_{m_2, n_2} a_{m_1, n_1}^{(i)} + a_{m_1, n_1} a_{m_2, n_2}^{(i)}) + \eta_1 \\ Y_2 = H_{m_1, n_1}^{(r)} (a_{m_2, n_2} a_{m_1, n_1}^{(i)} - a_{m_1, n_1} a_{m_2, n_2}^{(i)}) + \eta_2 \end{cases}$$

where η_1 and η_2 are the noise terms at each pilot position. The quality of this channel estimation is dependent upon how close $\frac{Y_1}{Y_2}$ is to C .

$$\frac{Y_1}{Y_2} = C \frac{1 + \frac{\eta_1}{v H_{m_1, n_1}^{(r)}}}{1 - \frac{\eta_2}{v H_{m_1, n_1}^{(r)}}}$$

where: $v = -a_{m_2, n_2} a_{m_1, n_1}^{(i)} + a_{m_1, n_1} a_{m_2, n_2}^{(i)}$.

The quality of the channel estimation is directly linked to the power of v . Indeed if v power is high, the estimation will be good enough but if it close to zero then the noise is enhanced and we will have a poor channel estimation. The problem here is that v depends on $a_{m_1, n_1}^{(i)}$ and on $a_{m_2, n_2}^{(i)}$, which also are depending on the real data around the pilot positions. v is then a random variable that can be sometimes close to zero. This randomness will impact the performance as we will see in the simulation results.

V. SIMULATION RESULTS

Our simulations have been carried out with a channel model and modulation parameters that are borrowed from the IEEE802.22 standard. IEEE802.22 is a new 802 LAN/MAN standard that is still in progress. It aims at constructing Wireless Regional Area Network (WRAN) utilizing free TV bands. In this context one of the main interest for pulse-shaped multicarrier modulation systems comes from the possibility to more easily satisfy the transmission frequency mask. As furthermore IEEE802.22 may take advantage of some of the principles of cognitive radio, the possibility with OFDM/OQAM of generating agile waveforms adds another interest for this modulation scheme. The channel profile and the main parameters of the system used are given below:

- Sampling frequency: 9.14 MHz
- Number of paths: 6
- Power profile (in dB) : -6.0, 0.0, -7.0, -22.0, -16.0, -20.0.
- Delay profile (μs) : -3, 0, 2, 4, 7, 11
- FFT size: 2048
- Guard interval composed of 130 samples (14.22 μs)
- QPSK modulation and Convolutional channel coding ($K = 7$ with $g_1 = (133)_o$, $g_2 = (171)_o$ and code rate= $\frac{1}{2}$)
- Frame length: 41 OFDM symbols

In order to accurately satisfy the approximations that are presented in the previous sections, we only use here prototype filters being well-localized in time and frequency. This criterion also allows us to guarantee a frequency behavior that is better than the OFDM one. The simulations are carried out with a discrete-time signal model (cf. the appendix) and

prototype filters of finite length, denoted by L . So the IOTA prototype filter we use results from the truncation of a IOTA pulse of length $4T_0$ and contains $L = 4M = 8192$ taps. It will be designated as IOTA4. We also use another prototype filter, that results from a direct optimization, with $L = M = 2048$ coefficients, of the time-frequency localization (TFL) criterion [3]. We designate it by TFL1. As explained in section IV the preamble is perfectly known at the receiver side and its content depends on the channel estimation method under consideration. Therefore different frame configurations have to be examined. They are depicted in Fig. 2. For OFDM/OQAM we have compared the two variants of the interference approximation method: IAM1 and IAM2, with the channel estimation method using Pairs of Pilots, denoted POP. All these methods are also compared with a CP-OFDM modulation having the frame configuration depicted in Fig. 2A. In Fig. 2 we can see that, as indicated previously, preambles have different lengths: $2\tau_0$ for OFDM (see Fig. 2A) and OQAM/POP (see Fig. 2D) and $3\tau_0$ for OQAM/IAM (see Figs. 2B and C). Note also that for OQAM/IAM1, the $a_{m,n}$ coefficients of the preamble result from a random selection of real-valued extracted from a QPSK constellation.

As usual, the performance of the different estimation methods are evaluated by a comparison of the Bit Error Rate (BER) as a function of the E_b/N_0 ratio, with E_b the useful bit energy and N_0 the monolateral noise density. First of all, it can be seen in Fig. 3 that in case of a perfect channel estimation, OQAM performs better than OFDM. For a BER of 10^{-3} , the performance gain is around 0.3 dB. This difference is mainly due to the no use of a guard interval (GI) in OQAM ($10 \log((2048 + 130)/(2048)) \approx 0.3$ dB). The gain could have been still more important if we had transformed the no use of a GI into a more powerful coding rate, i.e. using a coding rate of 0.5 for OFDM and one of 0.47 for OQAM.

The results obtained for the realistic, preamble-based, channel estimation methods are reported in Fig. 4. For OQAM systems, to designate the results obtained with a given estimation method and a given prototype, both acronyms are combined. For instance IOTA4/IAM1 corresponds to the result when IAM1 is run with the IOTA4 prototype.

The performance results are compared at BER= 10^{-3} . It can be noted that IOTA4/IAM1 and TFL1/IAM1 give performance that are approximately 1 dB worst than the OFDM one. This is due to the fact that $a_{m,n}^{(i)}$ is a random variable that can be close to zero. Then the power of the pilot symbol is reduced to the power of $a_{m,n}$, i.e. is equal to σ_a^2 , whereas for OFDM the power of the pilot is always $2\sigma_a^2$.

With the IAM2 variant, the preamble is chosen in order to maximize the power of $a_{m,n}^{(i)}$. Then, with the IOTA4 prototype, we get $a_{m,n}^{(i)} = \pm 0.882$ while the TFL1 prototype gives $a_{m,n}^{(i)} = \pm 1.076$. So, IOTA4/IAM2 is 1 dB better than CP-OFDM and the gain with TFL1/IAM2 is around 1.3 dB w.r.t. CP-OFDM. TFL1/IAM2 outperforms IOTA4/IAM2 because the power of its pseudo-pilot is higher, as the power value of $a_{m,n}^{(i)}$ is higher for TFL1 than for IOTA4.

IOTA4/POP gives results that are similar to the ones of IOTA4/IAM1. The small gain at low E_b/N_0 is mainly due

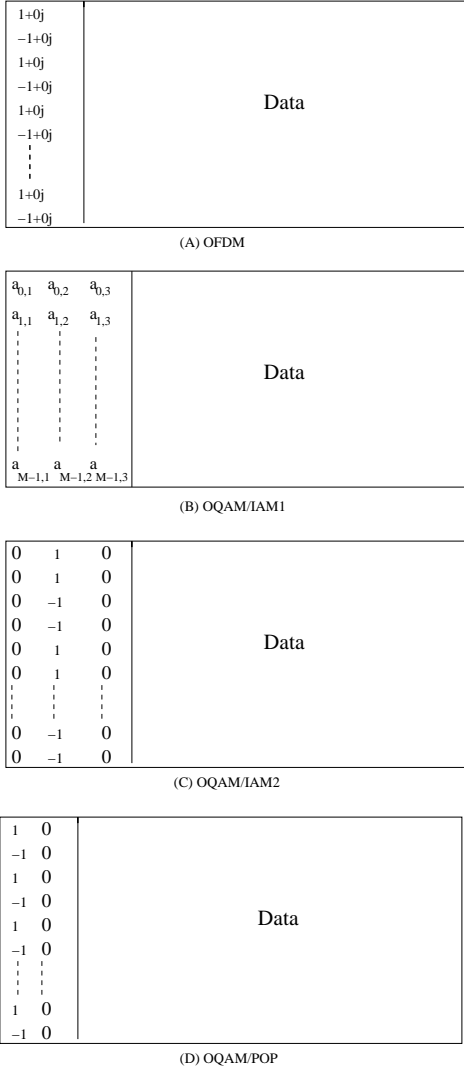


Fig. 2. Frame configurations for the different channel estimation methods.

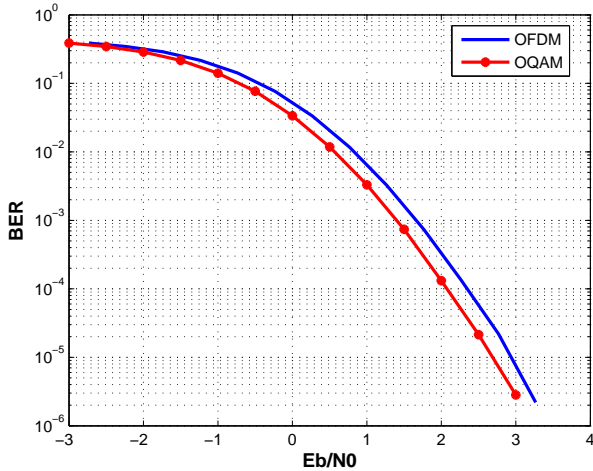


Fig. 3. Results for a perfect channel estimation.

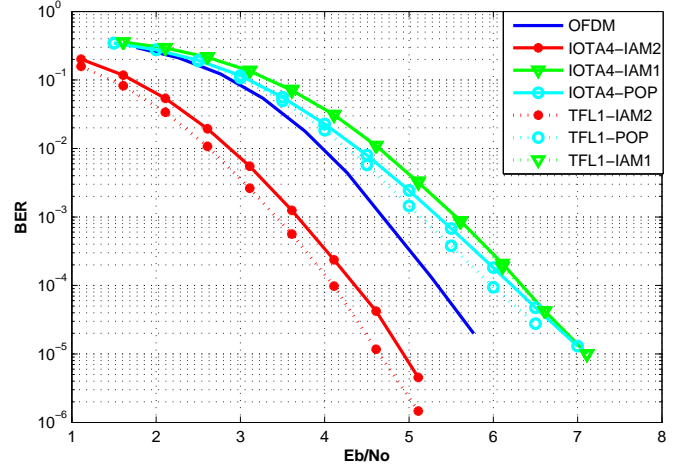


Fig. 4. Results for the different preamble-based methods.

to the gain brought by a reduction of the preamble duration which is $2\tau_0$ instead of $3\tau_0$. The randomness of the parameter v , as we said in the previous section, explains the 1 dB loss compared to CP-OFDM. TFL1/POP provides a 0.2 dB gain compared to IOTA4/POP. This can be explained by the fact that a short length prototype may be more robust to delay spread than a longer one. This slight difference can be explained by the impact of the prototype's length in the presence of a delay spread. Indeed, for a shorter prototype the resulting interference is more limited in the time domain. This effect cannot be observed with IAM1. Then, indeed, the power of the $a_{m,n}^{(i)}$ term is the most predominant factor.

VI. CONCLUSION

In this paper, we have presented two preamble-based channel estimation methods for OFDM/OQAM modulation systems. Our experimental setup, i.e. the channel model and the system parameters, is inspired from the IEEE802.22 standard and it also includes the simulation of CP-OFDM. The first channel estimation method for OFDM/OQAM (IAM) requires a preamble of $3\tau_0$ duration whereas the second one (POP method) only needs a $2\tau_0$ duration, i.e. the equivalent of one complex OFDM symbol. However, with one of the variant of the interference approximation method (IAM2), which includes an efficient design of the IAM preamble, the power of the pseudo-pilot symbol is maximized. So that OQAM/IAM2 provides the best results with a 2 dB gain compared to OQAM/POP and 1 dB gain when comparing with CP-OFDM. It has been also shown that we always get better or, at least, equal performance with the shortest OQAM prototype, the TFL1 prototype being better than the IOTA4 one. In future work the performance of these systems will be analyzed in the presence of Doppler spread and synchronisation errors.

The OFDM/OQAM discrete-time expression derived from (1) is given by:

$$s[k] = \sum_{m=0}^{M-1} \sum_{n \in \mathbf{Z}} a_{m,n} \underbrace{g[k - nN] e^{j \frac{2\pi}{M} m(k - \frac{L-1}{2})}}_{g_{m,n}[k]} e^{j \phi_{m,n}} \quad (23)$$

with $N = M/2$ and L the length of the prototype filter. For discrete-time prototypes, the real orthogonality condition can be written for instance as shown in [10]:

$$\sum_{n \in \mathbf{Z}} \sum_{m \in \mathbf{Z}} g[k + nN] g[k + mN] z^{m-n} = \frac{1}{N} \quad (24)$$

In the following we show that any real-valued prototype being orthogonal and with unit energy satisfies equation (18). So let show that:

$$S = \sum_{p=0}^{M-1} \sum_{q \in \mathbf{Z}} \left| \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \right|^2 = 2 \quad (25)$$

First we have:

$$\left| \langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \right|^2 = \sum_{k \in \mathbf{Z}} \sum_{l \in \mathbf{Z}} g[l] g[k] g[k+qN] g[l+qN] e^{j \frac{2\pi}{M} p(k-l)} \quad (26)$$

which can be rewritten as follows

$$S = \sum_{k \in \mathbf{Z}} \sum_{l \in \mathbf{Z}} g[l] g[k] \sum_{n \in \mathbf{Z}} g[k+nN] g[l+nN] \underbrace{\sum_{m=0}^{M-1} \left(e^{j \frac{2\pi}{M} (k-l)} \right)^m}_{A_{k,l}}.$$

There are two possible cases

- $k - l \neq aM$ with $a \in \mathbf{Z}$. then $A_{k,l} = 0$
- $k - l = aM = 2aN$ with $a \in \mathbf{Z}$. then $A_{k,l} = M$

which lead to

$$S = M \sum_{k \in \mathbf{Z}} \sum_{a \in \mathbf{Z}} g[k+2aN] g[k] \sum_{n \in \mathbf{Z}} g[k+nN] g[k+(n+2a)N] \quad (27)$$

Using the orthogonality condition (24), for $a \neq 0$, we can write:

$$\forall k, \sum_{n \in \mathbf{Z}} g[k+nN] g[k+(n+2a)N] = 0 \quad (28)$$

while for $a = 0$ we get:

$$\forall k, \sum_{n \in \mathbf{Z}} g[k+nN]^2 = \frac{1}{N} \quad (29)$$

Then

$$S = \frac{M}{N} \underbrace{\sum_{k \in \mathbf{Z}} g[k]^2}_{=1 \text{ normalized prototype}} = 2$$

and since $|\langle g \rangle_{0,0}^{0,0}| = 1$, we have:

$$\sum_{(p,q) \neq (0,0)} \left[\langle g \rangle_{m_0+p, n_0+q}^{m_0, n_0} \right]^2 = 1 \quad (30)$$

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