Power Efficient Scattered Pilot Channel Estimation for FBMC/OQAM

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***Abstract*—Filter bank multicarrier transmission with offset-QAM (FBMC/OQAM) is a promising candidate waveform for the next mobile communication systems as it is well suited for many new scenarios and challenges like improved spectral efficiency, spectrum sharing approaches or high mobility scenarios. It does not require a cyclic prefix (CP) leading to a higher spectral efficiency than orthogonal frequency division multiplexing with CP (CP-OFDM) and the flexibility of the transmit and receive filters enables higher throughput in spectral sharing and high mobility scenarios. One aspect to be considered is efficient channel estimation that is needed in order to realize these gains. As the classical channel estimation used for CP-OFDM cannot be applied directly to FBMC/OQAM, new competitive solutions are needed. One promising solution for the pilot design suited for FBMC introduces an auxiliary pilot (precoding symbol) that nullifies the intrinsic interference at the pilot position, but this leads to increased power on these auxiliary pilots. In this paper, a new non-linear scattered pilot design and channel estimation technique is proposed that addresses the problem of increased power of the auxiliary pilot. In addition to the reduced transmit power, we also observe that the receive power at the pilot position is increased with this new proposal which leads to an improved channel estimation performance.**

I. INTRODUCTION

Multicarrier schemes are promising candidates to achieve high data rate transmission. Orthogonal frequency division multiplexing (OFDM) has been adopted for different wireless standards such as WiFi, WiMax and Long Term Evolution (LTE) [1]. With the use of cyclic-prefix (CP), OFDM provides numerous advantages such as the efficient implementation through fast Fourier transforms (FFT) to combat severe multipath fading for broadband signals and its good affinity with multiple-input multiple-output (MIMO) systems. On the other hand, it is also widely recognized that CP-OFDM has high out-of-band radiation and a loss in spectral efficiency due to the CP and the performance suffers from high mobility, carrier frequency offset (CFO), and synchronization errors.

To overcome those drawbacks of CP-OFDM, a number of research activities on multicarrier transmission targeting the future 5G systems are ongoing. That includes generalized frequency division multiplexing (GFDM) [2], universal-filtered multicarrier (UFMC) [3] and filter bank multicarrier with offset-QAM (FBMC/OQAM) [4]. Furthermore, recently some new FBMC schemes have been proposed using circular convolution to combat the problem of the filter tails which decreases the efficiency for short bursts [5], [6]. In this paper, we study classical FBMC/OQAM scheme which has desired properties such as high spectral efficiency, low out-of-band radiation, and high robustness against high mobility, CFO, and synchronization errors. The high spectral efficiency comes from the fact that no CP is used for FBMC/OQAM. Instead, prototype filters are applied to achieve better localization in both frequency and time for obtaining the desired properties [7], [8].

All those desired features of FBMC/OQAM, however, come at the price of the relaxation of the orthogonality condition from the complex field to the real field. This means that several important techniques that have been developed for CP-OFDM such as transmit diversity technique from the orthogonal design and scattered pilot-based channel estimation cannot be directly applied to FBMC/OQAM. The main reason behind this is the presence of the so-called intrinsic interference, which we will elaborate later in detail. In this paper we focus on the pilot-based channel estimation problem for FBMC/OQAM, in particular with scattered pilot design suitable also in highly mobile scenarios.

The main challenge of coherent channel estimation based on scattered pilot design for FBMC/OQAM is the random nature of the intrinsic interference observed at the pilot position at the receiver since the intrinsic interference is caused by the data symbols surrounding the pilot. In [9] and [10] the authors propose two different approaches to combat the problem. One approach introduces a spreading method that is applied to data symbols located immediately next to a pilot position. Despreading has to be performed at the receiver. This approach can increase the complexity and the intrinsic interference caused by data symbols other than the immediate neighbors of the pilot position may not be handled well. Another approach introduces a precoding symbol to nullify the intrinsic interference at the pilot position at the receiver side. This approach can be applied to cancel the intrinsic interference caused by any number of data symbols around the pilot position. However, it can be power inefficient since the resulting precoding symbol may lead to high transmit power. We propose a new power efficient solution that extends the idea in [10].

The remaining part of this paper is organized as follows. In Section II we present the FBMC/OQAM system model. Then, in Section III we describe the channel estimation problem for FBMC/OQAM and two of the available solutions. Section IV introduces our new proposed method. In Section V we compare the performance obtained by CP-OFDM and FBMC/OQAM with the different scattered pilot based channel estimation solutions. Conclusions and future work are given in Section VI.

II. FBMC/OQAM SYSTEM MODEL

Fig. 1 shows the system block diagram of FBMC/OQAM. The transmit and receive filter operation can be implemented

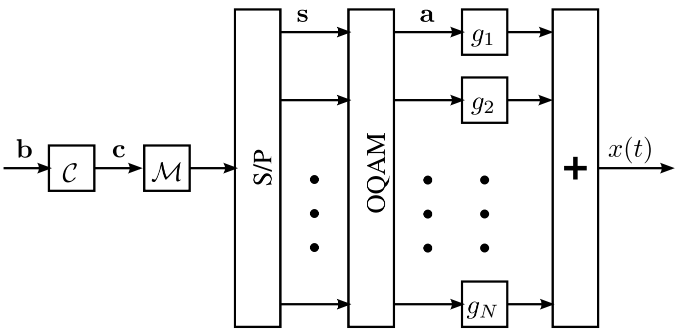

Fig. 1. System model for a FBMC transmitter.

in different ways. The polyphase structure described in [11] is usually assumed for the computationally efficient implementation, but here we show the block diagram that is analytically equivalent to the polyphase structure for the ease of exposition.

The information bit vector $\mathbf{b}$ is first channel encoded with the code $\mathcal{C}$ to get the code bit vector $\mathbf{c}$ and then mapped to complex QAM symbol vector $\mathbf{s}$ by the mapping function $\mathcal{M}$. These complex symbols are then split and rearranged into real valued OQAM symbols from the symbol alphabet $\mathcal{A}$ denoted as $a_{m,n}$ where m denotes the subcarrier index and n the time instance, respectively. The baseband equivalent system model of the FBMC/OQAM transmit signal can then be written as

$$x(t) = \sum_{m=0}^{M-1} \sum_{n \in \mathbb{Z}} a_{m,n} g(t - n\tau_0) e^{j2\pi m F_0 t} e^{j\frac{\pi}{2}(n+m)} \;, \quad (1)$$

where M is an even number of subcarriers and $F_0 = 1/(2\tau_0)$ is the subcarrier spacing.

The prototype filter g is chosen such that it is orthogonal in the real domain

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

where δ is Kronecker delta. According to this real orthogonality, $\langle g_{m,n} | g_{p,q} \rangle$ is purely imaginary and therefore we define $\langle g \rangle^{p,q}_{m,n} = -j\langle g_{m,n} | g_{p,q} \rangle$ which is then purely real valued.

The signal at the receiver after the analysis filter can be approximated as (cf. (5)) [9]

$$y_{m,n} = H_{m,n}(a_{m,n} + j\, I_{m,n}) + \eta_{m,n}, \quad (3)$$

and contains the real valued data symbols $a_{m,n}$ as well as so-called intrinsic interference $I_{m,n}$ due to the lacking orthogonality in the complex domain. $H_{m,n}$ denotes the complex valued channel coefficient at subcarrier index m and symbol index n and the additive white Gaussian noise after the receive filtering is denoted as $\eta_{m,n}$.

The intrinsic interference, $j\, I_{m,n}$, is purely imaginary in case of real valued data and depends on the data symbols in the neighborhood and the shape of the prototype filter. This neighborhood can be defined as a set of frequency and time index pairs (p, q) depending on the time and frequency localization characteristics of the prototype filter according to

$$\Omega_{\Delta m, \Delta n} = \{(p,q), |p| \leq \Delta m, |q| \leq \Delta n | H_{m+p,n+q} \approx H_{m,n}\} - \{(0,0)\}, \quad (4)$$

where $H_{m,n}$ is assumed to be constant in the neighborhood $\Omega_{\Delta m, \Delta n}$. Furthermore, we define

$$I_{m,n} = \sum_{(p,q) \in \Omega_{\Delta m, \Delta n}} a_{m+p,n+q} \langle g \rangle^{p,q}_{m,n}, \quad (5)$$

and assume any interference from symbols outside the neighborhood $\Omega_{\Delta m, \Delta n}$ to be negligible due to the good localization of the prototype filter.

If the channel coefficient is known, the detection of the OQAM symbols can be done by simply taking the real value after equalization (cf. (3)):

$$\hat{a}_{m,n} = Re\left\{\frac{y_{m,n}}{H_{m,n}}\right\} \approx a_{m,n} + \eta'_{m,n}. \quad (6)$$

We observe that the pure imaginary intrinsic interference is gone and thus, the channel knowledge is essential for restoring the orthogonality.

In the next section we discuss pilot design and channel estimation techniques for obtaining this channel knowledge.

III. Channel Estimation problem for FBMC/OQAM & prior arts solution

Assuming that a_{m_0,n_0} is a real pilot symbol transmitted at a position (m_0, n_0) that is known by the receiver. The received pilot after the analysis filter can be written as

$$y_{m_0,n_0} = H_{m_0,n_0}(a_{m_0,n_0} + j\, I_{m_0,n_0}) + \eta_{m_0,n_0}. \quad (7)$$

The value of I_{m_0,n_0} defined in (5) is dependent on the data that are transmitted around the pilot position and are unknown at the receiver, hence a coherent channel estimation:

$$\hat{H}_{m_0,n_0} = \frac{y_{m_0,n_0}}{(a_{m_0,n_0} + j\, I_{m_0,n_0})}, \quad (8)$$

cannot be performed at the receiver.

One idea to combat this problem is to ensure that the intrinsic interference at the pilot position becomes zero:

$$I_{m_0,n_0} \stackrel{!}{=} 0. \quad (9)$$

In [9] the authors propose satisfying this condition by introducing spreading codes for the data around the pilot position. At the transmitter side, the data around the pilot position are spread by a spreading matrix. The detailed mathematical algorithm to get a spreading matrix ensuring (9) is available in [9]. At the receiver side despreading the demodulated data around the pilot is necessary. The major drawback of this solution is the increased complexity. Furthermore, the intrinsic interference caused by data symbols other than the immediate neighbors of the pilot position is ignored.

The authors of [10] propose a computationally simpler solution by precoding only at one position (μ, ν) around the pilot position. For convenience, we define another subset of the neighborhood using (4)

$$\Omega^*_{\Delta m,\Delta n} = \Omega_{\Delta m,\Delta n} - \{(\mu,\nu)\}, \tag{10}$$

and the partial interference from this subset can be written as

$$I^*_{m_0,n_0} = \sum_{(p,q)\in\Omega^*_{\Delta m,\Delta n}} a_{m_0+p,n_0+q}\langle g\rangle^{p,q}_{m_0,n_0}. \tag{11}$$

In order to cancel the intrinsic interference at the pilot position, the precoding symbol is proposed to be selected as

$$a_{\mu,\nu} = -\frac{I^*_{m_0,n_0}}{\langle g\rangle^{\mu,\nu}_{m_0,n_0}}. \tag{12}$$

Using this approach and by referring to (7) and (8), the term I_{m_0,n_0} becomes zero and the coherent channel estimation in (8) can be performed at the receiver.

It is noted that the precoding symbol in (12) needs to be transmitted using the position (μ,ν) in addition to each of the pilot position. The spectral efficiency of this new scheme is, however, similar to that of CP-OFDM which uses one complex-valued pilot symbol to recover one complex channel coefficient H_{m_0,n_0} whereas here we use two real-valued OQAM symbols, i.e., the pilot and precoding symbols.

However, the major problem of this approach is the possibly high transmit power for the precoding symbol (cf. (11),(12)). For a well localized filter the maximum amplitude of the precoding symbol is

$$\sum_{(p,q)\in\Omega^*_{\Delta m,\Delta n}} \frac{\max\limits_{a_{m,n}\in\mathcal{A}}(|a_{m,n}|)\langle g\rangle^{p,q}_{m_0,n_0}}{\langle g\rangle^{\mu,\nu}_{m_0,n_0}}. \tag{13}$$

One possible solution to lower the transmit power of the precoding symbol in (13) is to choose the position of the precoding symbol to have the maximum impact on the interference, i.e., choose the position (μ,ν) with maximum value of $\langle g\rangle^{\mu,\nu}_{m_0,n_0}$. But still the amplitude can be multiples of the maximum amplitude of the modulation alphabet used for the data symbols leading to higher transmit power (e.g., for OQPSK with $1/\sqrt{2}$ maximum amplitude of the modulation alphabet the average precoding symbol amplitude is 1.03). Our new proposal to combat the increased power of the precoding symbols will be presented in the next section.

IV. Proposed Power efficient pilot design

In this section we firstly provides a short description of the basic principle of our proposed technique and then describe the concrete procedure that is derived from the principle. Furthermore, we discuss certain optimization as well as some other practical aspects.

A. Basic principle

Instead of forcing the intrinsic interference at the pilot position to be zero (cf. (9)) as discussed in Section III, also other values may be chosen for the precoding symbol depending on the actual data symbols influencing the intrinsinc interference on the pilot position (cf. (11)). This idea is inspired by the well-known Tomlinson-Harashima precoding [13], [14] where the originally limited set of discrete points is repeated periodically and all periodically repeated points in the set correspond to the same information. Then out of this set of equivalent points referring to the same information one point is chosen that minimizes the transmit power, which in the end leads to a coding gain. This idea of a set of equivalent points leading to different transmit powers is transferred to the problem of channel estimation for FBMC with OQAM in the following.

B. Procedure

We introduce a finite discrete set $\mathcal{X}$ of possible interference terms that are assumed *a-priori* known both at the transmitter and receiver. And we define the following optimization problem that is to choose the value of the interference out of $\mathcal{X}$ such that the corresponding precoding symbol minimizes the transmit power of the precoding symbol:

$$\min_{X\in\mathcal{X}} |a_{\mu,\nu}|^2, \tag{14}$$

with

$$a_{\mu,\nu} = -\frac{I^*_{m_0,n_0} - X}{\langle g\rangle^{\mu,\nu}_{m_0,n_0}}. \tag{15}$$

This minimization can be described as

$$\min_{X\in\mathcal{X}} \left|\frac{I^*_{m_0,n_0} - X}{\langle g\rangle^{\mu,\nu}_{m_0,n_0}}\right|^2, \tag{16}$$

that can be equivalently simplified to:

$$\min_{X\in\mathcal{X}} \left|I^*_{m_0,n_0} - X\right|. \tag{17}$$

This optimization represents the closest neighbor search between $I^*_{m_0,n_0}$ and the possible values of $X\in\mathcal{X}$. This leads to a *non-zero* intrinsic interference of $I_{m_0,n_0} = X$ (cf. (9)).

At the receiver side the pilot symbol including the real valued pilot *plus* the intrinsic interference $I_{m_0,n_0} = X$ have to be known to perform the simple zero forcing channel estimation according to (8). The actual value of X is not known a-priori to the receiver as it depends on the random data symbols around the pilot as shown in (5) and therefore has to be estimated before the actual channel estimation. Unfortunately, only non-coherent amplitude detection can be done to estimate the value of X because the channel knowledge required for coherent estimation is not available yet.

To this end we define the absolute amplitude as

$$S = |1 + jX| \tag{18}$$

assuming the real valued pilot a_{m_0,n_0} is always set equal to 1 without loss of generality. The required steps in order to perform the channel estimation on the pilot positions are summarized below.

- Measure the absolute value at the pilot position $|y_{m_0,n_0}|$
- Compare this measurement with all possible absolute values S according to the set $\mathcal{X}$ and choose the most likely value of X denoted as $\hat{X}$
- Perform zero forcing channel estimation with the estimated pilot $1 + j\hat{X}$ according to $\hat{H}_{m_0,n_0} = y_{m_0,n_0}/(1 + j\hat{X})$

TABLE I. Optimization of set $\mathcal{X}$.

1 :	$L := 3$	Define set size of $\mathcal{X}$, 3 for example
	$S_{\max} = \max \sqrt{1 + \left\|I^*_{m_0,n_0}\right\|^2}$	Find meaningfull range for S
	$\Delta S = (S_{\max} - 1)/(L-1)$	Equal distance between values of S
	$\mathcal{S} = \{1, \Delta S, \dots, L\Delta S\}$	Calculate elements of $\mathcal{S}$
5 :	$\mathcal{X} = \{(-1)^l \sqrt{S-1} \mid S \in \mathcal{S}\}$	Calculate values of $\mathcal{X}$

After the channel estimation on the scattered pilot positions as described above, any interpolation can be performed to obtain the channel estimation for the whole time-frequency grid.

C. Optimization of set $\mathcal{X}$

As the detection of the transmitted value of X is essential for the whole channel estimation performance, the set $\mathcal{S}$ of possible values of $S = |1 + jX|$ should be chosen properly. Table I summarizes the procedure used in this paper to find the set $\mathcal{X}$ using the pseudo codes.

The values of $\mathcal{S}$ should be distinguishable as much as possible at the receiver in terms of amplitudes at the pilot position. Consequently, the possible values of S should have maximum distance leading to a uniform distribution of the values of S over the meaningful range. The meaningful range depends on the selection rule of a specific X. If the transmit power should be minimized (as it is the case in this paper), the meaningful range for X is determined by twice the maximum amplitude of $I^*_{m_0,n_0}$. A value for X that is larger than the maximum value of $I^*_{m_0,n_0}$ would not be optimal according to the rule described in (17) as there is always a better value for X being closer to $I^*_{m_0,n_0}$.

A uniform distribution of the values of S excludes values for X that lead to the same S, e.g., ± 1 would have the same S and would therefore not be distinguishable (cf. (18)). The approach chosen here is to first choose a certain number of equally distant values for S within the meaningful range of amplitudes as indicated in lines 2 to 4 of Table I, calculate the corresponding values for $|X|$ and use them with alternating signs to finally get the values for the set $\mathcal{X}$ as shown in line 5.

The size of the set $\mathcal{X}$ defined in line 1 directly influences the distance between the values of $\mathcal{S}$ and therefore the probability of detection errors at the receiver. A larger set size leads to lower transmit power of the precoding symbol as the closest neighbor according to (17) will be much closer on average, but the detection errors at the receiver and the complexity can become higher. Therefore, the set size is a parameter and may be chosen to find a compromise between minimizing the detection error probability and minimizing the transmit power as well as the complexity.

D. Increased pilot power at the receiver

Another advantage of the proposed scheme compared to the zero forcing approach is that any value of X unequal to 0 in fact leads to a higher receive power and therefore to a higher SNR at the pilot position (cf. (3),(9)). This side effect comes at no additional cost or optimization but leads to better channel estimation performance. Therefore, with the proposed solution we get a lower transmit power and a higher receive power at the same time and both effects improve the channel estimation. The bottleneck of the scheme is the estimation of X and this limits the overall performance as shown in the next section.

TABLE II. Simulation parameters.

FFT size	1024
Channel model	6-taps exponentially decaying: power profile (dB) = 0, -2, -4, -6, -8, -10 delay profile (μs) = 0, 0.5, 1, 1.5, 2, 2.5 $f_D = 0$
Channel coding	Convolutional code: $K = 7, g_1 = (133)_o, g_2 = (171)_o, r = \frac{1}{2}$
Frame length	12 and 40 OFDM symbols (24 and 80 FBMC symbols)
Channel estimation	Perfect and scattered pilot based estimations
Modulation	QPSK / OQPSK
OFDM cyclic prefix	64 samples (32μs)
FBMC prototype filter	PHYDYAS

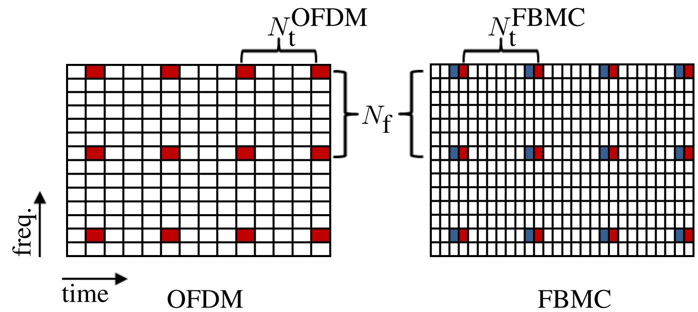

Fig. 2. Scattered pilots.

V. Performance evaluation

In this section, we examine the performance of the new proposed FBMC/OQAM channel estimation method and compare the results with that of the proposed solution in [10] and with that of CP-OFDM. Our simulations have been carried out with a channel model and parameters that are shown in Table II where we refer to the method in [10] and the new proposed methods of this paper as "SoA" and "New", respectively. The neighborhood used for the intrinsic interference calculation is $\Omega_{1,3}$. The power of one FBMC pilot symbol is set to be equal to the power of one OFDM pilot symbol for a fair channel estimation comparison. After performing the channel estimation at the pilot positions, a two-cascaded one dimensional FIR filtering based interpolation [12]; first filtering in the time direction and followed by filtering in frequency direction, is performed to get the channel coefficients at all frequency and time locations. Fig. 2 shows the time frequency grid where pilot symbols are inserted among data symbols for both CP-OFDM and FBMC/OQAM systems. The red symbols represents the complex-valued and real-valued pilots for the cases of CP-OFDM and FBMC/OQAM, respectively, and the blue symbols represent the precoding symbols for the case of FBMC/OQAM. The distance between two pilot symbols in frequency direction is $N_f = 6$ for both CP-OFDM and FBMC/OQAM and in time direction, $N_t^{\text{OFDM}} = 4$ for CP-OFDM and $N_t^{\text{FBMC}} = 2 \cdot N_t^{\text{OFDM}} = 8$ for FBMC/OQAM.

The saving in the transmit power of the precoding symbol for FBMC/OQAM is shown in Fig. 3. The upper and lower histograms show the precoding symbol power distribution for the SoA and New methods, respectively. It can be seen that the precoding symbol power range of the new proposed method

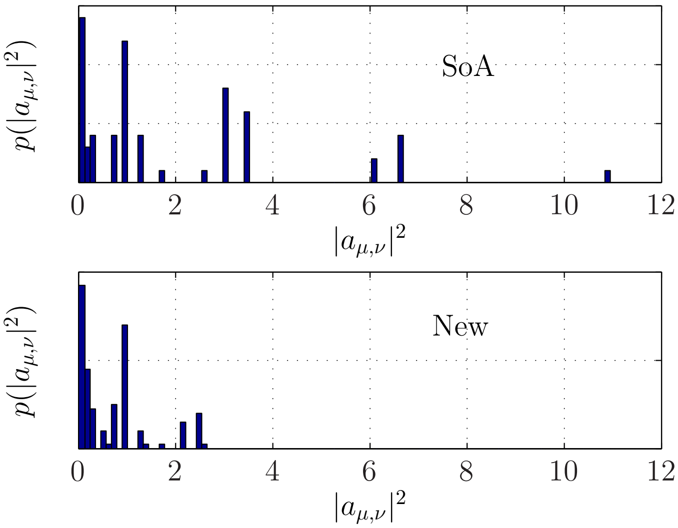

Fig. 3. Precoding Symbol Power Distribution. The upper histogram refers to SoA, the lower one to the new proposal.

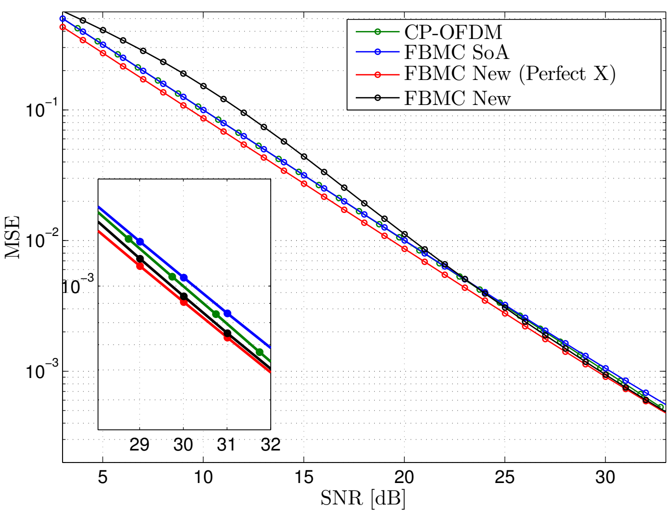

Fig. 4. MSE of the channel estimation at pilot positions versus SNR.

in much smaller then that of the SoA method. In addition, the simulations showed the average precoding symbol transmit power of the new proposed method is around 2.4 dB less compared to SoA.

The performance of the different channel estimation methods is evaluated firstly by a comparison of the mean squared error (MSE) at the pilot positions as a function of the SNR. It can be observed in Fig. 4 that the SoA method has the similar performance compared to CP-OFDM channel estimation and the new proposed method with perfect knowledge of X gives a 0.6 dB better performance compared to the SoA method. This is due to the fact that the received pilot symbol power is higher. The errors in the estimation of X degrade the performance at low SNR, however at high SNR the performance converges to the case of perfect knowledge of X. It should be noted, however, that the SoA method spends more transmit power for the precoding symbols as compared to the new proposal, as we observed in Fig. 3, which is not captured in Fig. 4.

The different channel estimation methods are also evaluated by a comparison of the bit error rate (BER) performance

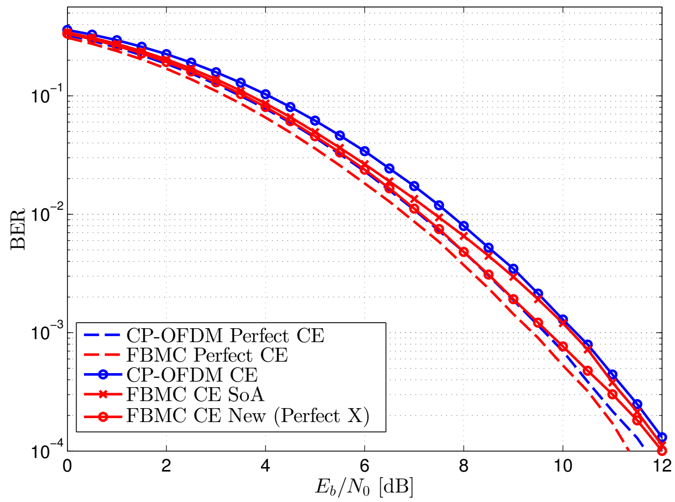

Fig. 5. BER performance for the different channel estimation methods.

as a function of the E_b/N_0 ratio, with E_b the energy per information bit and N_0 the noise power spectral density. First of all, it can be seen in Fig. 5 that in the case of perfect channel knowledge, FBMC/OQAM performs better than CP-OFDM. The performance gain is around 0.26 dB for a BER of 10^{-3}. This gain is mainly due to the no use of cyclic-prefix in FBMC/OQAM ($10\log((1024+64)/1024) \approx 0.26$ dB). It can be also seen that the new proposed method with perfect knowledge of X performs better than the SoA method with a gain around 0.5 dB for a BER of 10^{-3}. This is due to the higher received pilot power and the better MSE performance.

The impact of the estimation errors of X on the BER performance is shown in Fig. 6. It can be seen that the larger number of FBMC/OQAM symbols per frame the closer the performance to the case of perfect knowledge of X. This is due to the fact that with a high number of symbols per frame, the impact of the error in estimating X is lower which is in turn due to the time direction filtering interpolation with time invariant channels. For 80 FBMC/OQAM symbols per frame, the performance is comparable to that of CP-OFDM. Further investigations are ongoing to evaluate the performance of the new proposed method with time varying channels.

VI. Conclusions

In this paper, we proposed a new scattered pilot based channel estimation method for FBMC/OQAM systems in order to overcome the problem of high precoding symbol transmit power of the SoA method. In addition, the higher received pilot symbol provides a MSE gain around 0.6 dB at the pilot positions and a E_b/N_0 gain around 0.5 dB at a BER of 10^{-3} when the value of X is perfectly known at the receiver side. The BER performance of the new proposed method is highly affected by the accuracy of the estimation of X but it is still comparable with that of CP-OFDM for high number of symbols per frame and with lower precoding symbol transmit power compared to the SoA method. Future work will be to investigate more accurate techniques for estimating X and to apply the new proposed method over time varying channels. We will also tackle the issue of MIMO channel estimation schemes with FBMC/OQAM.

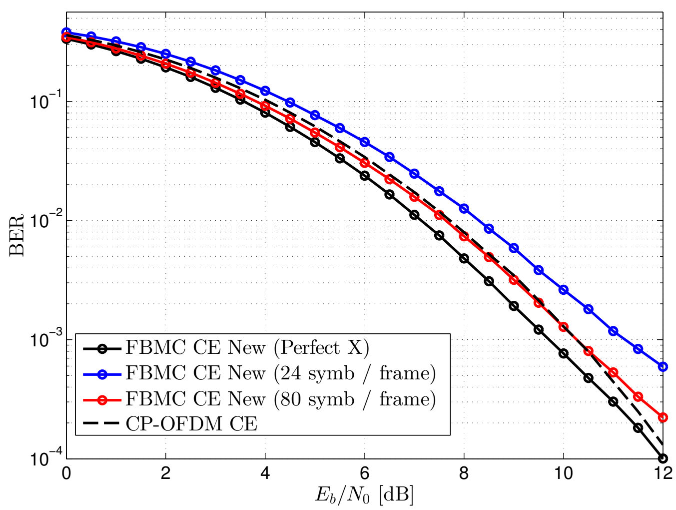

Fig. 6. BER performance with X estimation for different frame lengths.

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