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Power allocation and capacity analysis for FBMC-OQAM with Superimposed Training

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ABSTRACT Superimposed Training (ST) is a semiblind channel estimation technique, proposed for orthogonal frequency division multiplexing (OFDM), where training sequences are added to data symbols, avoiding the use of dedicated pilot-subcarriers, and then increasing the available bandwidth compared to pilot symbol assisted modulation (PSAM). Filter bank multicarrier offset quadrature amplitude modulation (FBMC-OQAM) is a promising waveform technique considered to replace OFDM which takes advantage of well designed filters to avoid the use of cyclic prefix and reduce the out-band-emissions. In this paper, we provide the expressions of average channel capacity of FBMC-OQAM combined with either PSAM or ST schemes, taking into account imperfect channel estimation and the presence of the pilot sequences. In order to compute the capacity expression of our proposal, ST-FBMC-OQAM, we analyze the channel estimation error and its variance. The average channel capacity is deduced considering the noise, data interference from ST and the intrinsic self-interference of FBMC-OQAM. Additionally to maximize the average channel capacity, the optimal value of data power allocation is also obtained. Simulation results confirm the validity of the capacity analysis and demonstrate the superiority of ST-FBMC-OQAM over existing proposals.

INDEX TERMS Channel estimation, data interference, FBMC, superimposed training

I. INTRODUCTION

The wireless data traffic is expected to increase exponentially from 3 exabytes in 2010 to 500 exabytes in 2020, considering new concepts as internet of things (IoT), cloud computing, and new mobile data applications [1]. In order to support this huge demand, the industry and the academy have put their efforts in developing new generation technologies to improve the available bandwidth in wireless communications. Due to their advantages in spectral efficiency, new waveforms have appeared and can be considered as part of the evolution of mobile communications (5G and beyond) [2] [3].

Several wireless technologies such as Long Term Evolution (LTE) [4] and Wi-Fi [5] adopted orthogonal frequency division multiplexing (OFDM) [6] [7] as their multicarrier transmission waveform. While OFDM offers robustness

against multipath fading and convenience for a one-tap equalizer implementation, this is achieved at the expense of the efficiency of the entire system. OFDM requires adding a cyclic prefix (CP) whose length can represent up to 20% of the OFDM symbol in some scenarios. Additionally, OFDM exhibits high side lobes in the frequency domain, which also requires leaving guard-bands to prevent interference with the adjacent signals.

With the aim of addressing the aforementioned issues of OFDM, filter bank multicarrier offset quadrature amplitude modulation (FBMC-OQAM) [8] [9] is considered as a waveform candidate in order to substitute OFDM in some scenarios. FBMC-OQAM offers lower out-of-band emissions through the use of a well-designed prototype filter. This enables avoiding the use of undesirable guard-bands. Moreover, the symbols used in FBMC-OQAM are longer than those

used in OFDM to minimize the multipath effects of not using a CP. However, those symbols are overlapped to each other in order to maintain the data rate. According to [8], when the length of FBMC-OQAM symbols are long enough, the inter-symbol interference (ISI) caused by the channel can be neglected, as compared to the noise and channel estimation error.

Multicarrier modulation systems require channel equalization in order to perform a coherent detection. In OFDM, several methods exist for this purpose. However, these existing methods cannot be straightforwardly used in FBMC-OQAM. The reason is the presence of the intrinsic self-interference caused by the surrounding symbols, due to the use of the prototype filters, that must be taken into account before the equalization process. Two alternatives for placing the pilot information for channel estimation in FBMC-OQAM are proposed in the literature, namely preamble-based [10] and scattered pilot-based [11].

Focusing on scattered pilot-based techniques, several methods have been proposed [12]–[17]. The main idea consists in placing two or more pilot-symbols combined with a proper processing of them, either at the transmitter or receiver side, which is capable of compensating the intrinsic self-interference. By doing so, the received pilots become interference-free, and channel estimation can be performed in the same way as in OFDM systems. However, these techniques have several drawbacks and one of the most important issues is the increase in the number of pilot-symbols, as compared to OFDM, which seriously decreases the spectral efficiency; some auxiliary pilots have a power of 3.3dB higher than the power of data symbols, which means an additional waste of the valuable energy and increasing the peak-to-average power ratio (PAPR); all the proposed techniques require an additional complexity, either to compute or combine the auxiliary pilots at the receiver side in order to avoid the self-interference.

Superimposed Training combined with OFDM (ST-OFDM) consists in arithmetically superimposing pilot sequences over the data-signal, allowing to use of all time-frequency resources to transmit data information, thus, increasing the data rate. This technique is being proposed to be part of beyond-5G technologies to overcome boundaries such as pilot contamination in massive multiple input multiple output (MIMO) [18] or with new techniques like non orthogonal multiple access (NOMA) [19]. Indeed, [20] has fully characterized ST-OFDM providing the mean squared error (MSE) of the channel estimation and the capacity analysis of the entire system. Moreover, it has also provided a comparison between ST-OFDM and pilot symbol assisted modulation-OFDM (PSAM-OFDM) whose capacity was given in [21] where the former outperforms the latter in terms of capacity.

In [22], a first approach to the channel capacity of FBMC-OQAM, without considering ST, is provided, showing its superiority as compared to OFDM. However, it assumes ideal conditions, where the channel is perfectly known and pilot-symbols are not taken into account, which corre-

sponds to an unrealistic situation. Later, given the benefits of ST and FBMC-OQAM, we have proposed in [23] the combination of these two techniques where we provided some numerical results for the channel estimation error and symbol error rate (SER).

In this paper, we first study the average channel capacity for PSAM-FBMC-OQAM, considering imperfect channel estimation and the presence of pilots. Secondly, for our proposal ST-FBMC-OQAM, the analysis of the channel estimation error is presented and its variance is analytically derived. The average channel capacity is deduced considering the noise, data interference from ST and the intrinsic self-interference from FBMC-OQAM. Given the realistic channel capacity expressions of FBMC-OQAM and OFDM combined with both PSAM and ST, we can see that our proposal ST-FBMC-OQAM outperforms the other candidates. Finally, the optimal value of data power allocation to maximize the average channel capacity expression is obtained.

The remainder of this paper is organized as follows. Section II describes the system model, the effects of the channel in FBMC-OQAM and the ST proposal. In Section III, the ST-FBMC-OQAM channel estimation mechanism is studied. Section IV presents the analysis of the average channel capacity. In section V, the analysis of the power allocation is provided. In section VI, the simulation results are exposed and, finally in section VII the conclusions are presented.

Notation: \mathbf{x} represents a vector where $x[k]$ denotes its k -th element. \mathbf{X} denotes a matrix where $X[k, m]$ is the element of the k -th row and m -th column. $\Re(x)$ and $\Im(x)$ are the real and imaginary part of x , respectively. The superscripts $(\cdot)^T, (\cdot)^H$ and $(\cdot)^*$ denote transpose, Hermitian transpose and complex conjugate operations, respectively. $*$ denotes the convolution operation. $\mathbb{E}\{\mathbf{x}\}$ represents the expected value of \mathbf{x} . $\mathbb{E}_M\{\mathbf{x}\} = (1/M) \sum_{m=0}^{M-1} x[m]$ denotes the average over M elements of \mathbf{x} . $\mathcal{CN}(0, \sigma^2)$ denotes the circularly-symmetric and zero-mean complex normal distribution with the variance of σ^2 . $\mathbf{1}_{1 \times M}$ denotes a matrix of ones of size $1 \times M$. \mathbf{I}_K denotes an identity matrix of size $K \times K$. \otimes denotes the Kronecker product, $\text{vec}\{\cdot\}$ represents the vectorization of an array and $\lceil \cdot \rceil$ represents the rounding up to the next integer operation.

II. SYSTEM MODEL

The system model is focused on a point-to-point link, where the FBMC-OQAM waveform is transmitted. Firstly, an analysis of FBMC-OQAM without considering the channel effects is presented. Then, the channel is taken into account and the corresponding system model is exposed.

A. FBMC-OQAM

Let us denote \mathbf{S} as a matrix containing the set of $K \times M/2$ complex data symbols to be transmitted, where K represents the number of parallel subcarriers and M is the number of symbols. The complex symbols belong to a QAM constellation with an average power denoted as P_S/K where P_S represents the total multicarrier symbol power. Given \mathbf{S} , it

is transformed into OQAM real symbols, where the real and imaginary parts of \mathbf{S} are separated to build \mathbf{S}_o

$$S_o[k, m = 2m'] = \Re(S[k, m']), \quad (1)$$

$$S_o[k, m = 2m' + 1] = \Im(S[k, m']), \quad (2)$$

with $k = 0 \cdots K - 1$ and $m' = 0 \cdots M/2 - 1$. Note that the size of \mathbf{S}_o is $K \times M$, so m ranges here up to $M - 1$.

We consider a critically sampled discrete-time system model, where sampling period is given by $T_S = \frac{T}{K}$, where T corresponds to the transmission time of K symbols. Then, the transmitted signal can be represented as

$$x[n] = \sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} g \left[n - m \frac{K}{2} \right] e^{j \frac{2\pi}{K} \Delta f k \left(n - m \frac{K}{2} \right)}, \quad (3)$$

where $\Delta f = \frac{1}{T}$ is the subcarrier spacing and the phase $\phi[k, m]$ is given by

$$\phi[k, m] = \begin{cases} 0, & k + m \text{ even} \\ \frac{\pi}{2}, & k + m \text{ odd} \end{cases}. \quad (4)$$

We define $g_k[n] = g[n] e^{j \frac{2\pi}{K} \Delta f k n}$ as a time-frequency shifted version of $g[n]$. Then, $x[n]$ can be written as

$$x[n] = \sum_{m=-\infty}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} g_k \left[n - m \frac{K}{2} \right]. \quad (5)$$

This operation is also known as the synthesis filter bank (SFB). Note that $g[n]$ and $\phi[k, m]$ are designed so that $g_k[n]$ is orthogonal in the real field satisfying

$$\Re \{ \langle g_k[n], g_{k'}[n] \rangle \} = \begin{cases} 1, & k = k' \text{ and } n = n' \\ 0 & \text{otherwise} \end{cases}. \quad (6)$$

In a realistic environment, the transmitted signal goes through a frequency-selective fading channel

$$y[n] = x[n] \otimes h[\tau, n] + w[n]. \quad (7)$$

where \otimes corresponds to the convolution operation. The channel impulse response can be written as

$$h[\tau, n] = \sum_{l=0}^{L-1} h[\tau_l, n] \delta[\tau - \tau_l] \quad (8)$$

where L indicates the total number of paths and τ_l represents the l -th tapped delay where $l = \{0, 1, \dots, L - 1\}$. Furthermore, considering channels with frequency-flat subcarriers and constant in the duration of the filter as in [10], the received signal can be represented as

$$R_o[k_0, m_0] = e^{-j\phi[k_0, m_0]} \left(y[n] \otimes g_{k_0}^* \left[m_0 \frac{K}{2} - n \right] \right) \downarrow_{\frac{K}{2}} \quad (9)$$

where $(\cdot) \downarrow_{\frac{K}{2}}$ indicates that the signal within the expression is sampled at $K/2$ symbol spacing for OQAM modulation. Then, recovering the information from the filtered signal, we

obtain that

$$y[n] \otimes g_{k_0}^* \left[m_0 \frac{K}{2} - n \right] = \left(\sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} \sum_{l=0}^{L-1} h[\tau_l, n] g_k \left[n - \tau_l - m \frac{K}{2} \right] \right) \otimes g_{k_0}^* \left[m_0 \frac{K}{2} - n \right], \quad (10)$$

then,

$$\left(y[n] \otimes g_{k_0}^* \left[m_0 \frac{K}{2} - n \right] \right) \downarrow_{\frac{K}{2}} = \sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} H_{k_0, m_0}[k, m], \quad (11)$$

where the channel frequency response $H_{k_0, m_0}[k, m]$ can be written as

$$H_{k_0, m_0}[k, m] = \left(\sum_{l=0}^{L-1} h[\tau_l, n] \sum_{\eta=-\infty}^{\infty} g_k \left[\eta - \tau_l - m \frac{K}{2} \right] \cdot g_{k_0}^* \left[\eta + n - m_0 \frac{K}{2} \right] \right) \downarrow_{\frac{K}{2}}. \quad (12)$$

For the sake of space, we do not detail the manipulations given in [11] that can be used in order to simplify (12) as

$$H_{k_0, m_0}[k, m] = \sum_{l=0}^{L-1} h[\tau_l, n] D_{k_0, m_0}[k, m, \tau_l] e^{-j \frac{2\pi}{K} \Delta f k \tau_l}, \quad (13)$$

where the ambiguity function $D_{k_0, m_0}[k, m, \tau_l]$ can be expressed as

$$D_{k_0, m_0}[k, m, \tau_l] = \sum_{\eta=-\infty}^{\infty} g \left[\eta - \tau_l - \Delta_m \frac{K}{2} \right] g[\eta] \cdot e^{j \frac{2\pi}{K} \Delta f \Delta_k \eta} e^{-j \frac{2\pi}{K} \Delta f k \Delta_m \frac{K}{2}} \quad (14)$$

and $\Delta_k = k - k_0$ and $\Delta_m = m - m_0$. The received signal (9) after the synthesis can be seen as follows

$$R_o[k_0, m_0] = e^{-j\phi[k_0, m_0]} \cdot \left(\sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} H_{k_0, m_0}[k, m] \right). \quad (15)$$

Finally, simplifying (15) the expression can be reduced to

$$R_o[k_0, m_0] = \underbrace{H_{k_0, m_0}[k_0, m_0] S_o[k_0, m_0]}_{\text{desired signal}} + W[k_0, m_0] + e^{-j\phi[k_0, m_0]} \underbrace{\left(\sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} H_{k_0, m_0}[k, m] \right)}_{\substack{m \neq m_0 \text{ and } k \neq k_0 \\ \text{interference}}}, \quad (16)$$

where $W[k, m]$ is a random variable with a distribution $W \sim$

$\mathcal{CN}(0, \sigma_w^2)$ defined in [8]. It is assumed that the prototype filters have a unitary gain. Note that \mathbf{H} is a matrix of size $K \times M$.

B. SUPERIMPOSED TRAINING (ST) FOR FBMC-OQAM

For the FBMC-OQAM system, let us define a known pilot sequence vector \mathbf{q} of size $K \times 1$, whose real part must be the same as imaginary part. We apply (1) and (2) in order to obtain \mathbf{q}_o of size $K \times 2$. Then, we build the pilot sequence matrix \mathbf{Q}_o as follows

$$\mathbf{Q}_o = \mathbf{q}_o \otimes \mathbf{1}_{1,M/2}. \quad (17)$$

Note that all columns of \mathbf{Q}_o have the same value for a given row. This property is very important in order to estimate the channel aided by an averaging process. Additionally, we assume that the transmitted signal is composed by data and pilots where P is the total power, $P = P_S + P_P$. The parameter $P_S = \alpha P$ corresponds to the power assigned to the data symbols, $P_P = (1 - \alpha)P$ represents the power assigned to the pilots and α , which varies between $0 < \alpha < 1$, is the data power coefficient. When we add \mathbf{S}_o with \mathbf{Q}_o , we must satisfy the available power constraint given by the following expression

$$\frac{P}{2K} = \alpha \mathbb{E} \{ \|\mathbf{S}_o\|^2 \} + (1 - \alpha) \mathbb{E} \{ \|\mathbf{Q}_o\|^2 \}. \quad (18)$$

where $\|\cdot\|$ represents the Euclidean norm. Therefore, the new data-pilot matrix \mathbf{T}_o is given by

$$\mathbf{T}_o = \sqrt{\alpha} \mathbf{S}_o + \sqrt{1 - \alpha} \mathbf{Q}_o. \quad (19)$$

Once we have the new data-pilot sequence given by (19), we can transmit it through the FBMC-OQAM baseband model in (15), where \mathbf{S}_o is replaced by \mathbf{T}_o .

III. ANALYSIS OF THE CHANNEL ESTIMATION ERROR FOR ST-FBMC-OQAM

Given the system model detailed in the previous section, any element of the received signal \mathbf{R}_o can be derived as

$$R_{o k_0, m_0} = \sqrt{\alpha} R_o^S[k_0, m_0] + \sqrt{1 - \alpha} R_o^Q[k_0, m_0] + W[k_0, m_0], \quad (20)$$

where $R_o^S[k_0, m_0]$ and $R_o^Q[k_0, m_0]$ correspond to the received data and pilot signal, respectively, affected by the channel path and filters as in (15),

$$\begin{aligned} R_o^S[k_0, m_0] &= H_{k_0, m_0}[k_0, m_0] S_o[k_0, m_0] + W[k_0, m_0] + \\ &+ e^{-j\phi[k_0, m_0]} \underbrace{\left(\sum_{m=0}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]} H_{k_0, m_0}[k, m] \right)}_{k \neq k_0 \text{ and } m \neq m_0}, \\ R_o^Q[k_0, m_0] &= H_{k_0, m_0}[k_0, m_0] Q_o[k_0, m_0] + W[k_0, m_0] + \\ &+ e^{-j\phi[k_0, m_0]} \underbrace{\left(\sum_{m=0}^{M-1} \sum_{k=0}^{K-1} Q_o[k, m] e^{j\phi[k, m]} H_{k_0, m_0}[k, m] \right)}_{k \neq k_0 \text{ and } m \neq m_0}, \end{aligned} \quad (21)$$

Given the designed pilot-sequence in the previous subsection the received pilot-signal can be simplified as

$$\begin{aligned} R_o^Q[k_0, m_0] &= \tilde{H}[k_0, m_0] q_o + W[k_0, m_0] + \\ &+ q_o e^{-j\phi[k_0, m_0]} \sum_k \sum_m e^{j\phi[m, n]} H_{k_0, m_0}[k, m], \end{aligned} \quad (22)$$

where, assuming that the channel frequency response satisfies the quasi-stationarity condition and no Doppler effect is present, the channel $h[\tau_l, n] = h[\tau_l]$ and $\tilde{H}[k_0] = H_{k_0, m_0}[k_0, m_0]$. Furthermore, taking into account that in multi-carrier modulations it is satisfied that $K \gg \tau_l$ by design, $D_{k_0, m_0}[k, m, \tau_0] \simeq D_{k_0, m_0}[k, m, \tau_{L-1}]$, and $\tilde{H}[k] \simeq \tilde{H}[k \pm \frac{1}{2}]$. Thus,

$$\begin{aligned} R_o^Q[k_0, m_0] &\simeq \tilde{H}[k_0] q_o + W[k_0, m_0] + \\ &+ q_o e^{-j\phi[k_0, m_0]} \sum_k \tilde{H}[k] \sum_m e^{j\phi[m, n]} D_{k_0, m_0}[k, m] \approx \\ &\approx \tilde{H}[k_0] (q_o + j B_q[k_0, m_0]), \end{aligned} \quad (23)$$

where $j B_q$ is the total amount of self-interference produced by adjacent symbols.

In order to estimate the channel, we average \mathbf{R}_o over M consecutive symbols, $\{m, \dots, m + M\}$, for each subcarrier k . The expression is given by

$$\begin{aligned} \mathbb{E}_M \{ R_o[k, m] \} &= \mathbb{E}_M \{ W[k, m] \} + h[k] \times \\ &\times \left(\sqrt{\alpha} (\mathbb{E}_M \{ S_o[k, m] \} + j \mathbb{E}_M \{ B_S[k, m] \}) + \right. \\ &\left. + \sqrt{1 - \alpha} (\mathbb{E}_M \{ Q_o[k, m] \} + j \mathbb{E}_M \{ B_Q[k, m] \}) \right). \end{aligned} \quad (24)$$

Taking into account that each element of \mathbf{q}_o have the same value, we can say that $q_o[k] = \mathbb{E}_M \{ Q_o[k, m] \}$ and $b_q[k] = \mathbb{E}_M \{ B_Q[k, m] \}$, and defining the self-interference and noise value $u[k]$ as

$$\begin{aligned} u[k] &= \mathbb{E}_M \{ W[k, m] \} + h[k] \times \\ &\times (\sqrt{\alpha} (\mathbb{E}_M \{ S_o[k, m] \} + j \mathbb{E}_M \{ B_S[k, m] \})). \end{aligned} \quad (25)$$

Thus, (24) can be simplified by

$$\mathbb{E}_M \{ R_o[k, m] \} = h[k] \sqrt{1 - \alpha} (q_o[k] + j b_q[k]) + u[k]. \quad (26)$$

Note that the sequence $q_o[k] + jb_q[k]$ is known as pseudo-pilot and it is deterministic, where according to [17], it will help to improve the quality of the channel estimation. The term $jb_q[k]$ and its power can be computed following the Appendix formulation using the pilot sequence instead of the data sequence. Its vectorized form is denoted as \mathbf{b}_q and its power as P_{b_q} .

The variance of the interference and noise signal σ_u^2 can be obtained through $\mathbb{E}\{\mathbf{u}\mathbf{u}^H\} = \sigma_u^2 \mathbf{I}_K$ where

$$\mathbb{E}\{\mathbf{u}\mathbf{u}^H\} = \mathbb{E}\left\{ \left(\frac{1}{M} \overline{\mathbf{W}} + \frac{\sqrt{\alpha}}{M} \mathbf{h} (\overline{\mathbf{S}}_o + j\overline{\mathbf{B}}_s) \right) \times \left(\frac{1}{M} \overline{\mathbf{W}} + \frac{\sqrt{\alpha}}{M} \mathbf{h} (\overline{\mathbf{S}}_o + j\overline{\mathbf{B}}_s) \right)^H \right\}, \quad (27)$$

where \mathbf{S}_o , \mathbf{B}_s and \mathbf{W} represent the vectorized expression of each element of the self-interference and noise values. $\overline{\mathbf{W}}$ represents $\overline{\mathbf{W}} = \sum_{i=0}^M \mathbf{W}^i$. The overlined variables $\overline{\mathbf{S}}_o$ and $\overline{\mathbf{B}}_s$ follow the same definition.

Note that the power of the intrinsic interference comes from OQAM symbols, and then, the corresponding power is one half of the power of QAM symbols. Thus, the corresponding variances can be represented as follows

$$\sigma_u^2 = \frac{\sigma_w^2}{M} + \frac{\sigma_H^2 \alpha P}{2KM} + \frac{\sigma_H^2 \alpha P}{2KM} \frac{\sigma_v^2}{M}, \quad (28)$$

and, after reordering, the variance can be expressed as

$$\sigma_u^2 = \frac{1}{M} \left(\sigma_w^2 + \frac{\sigma_H^2 \alpha P}{2K} \left(1 + \frac{\sigma_v^2}{M} \right) \right). \quad (29)$$

The first term of (29) corresponds to the AWGN noise. The second term represents the data interference. The third term corresponds to the intrinsic self-interference in the proposed system, which is a linear combination of the interference produced by the surrounding elements of the affected symbol and also by the channel and data power. The resultant power of the intrinsic self-interference produced by the surrounding symbols (σ_v^2) is characterized in the Appendix. The number of FBMC symbols that can be transmitted in a coherence time, where the channel does not change, is represented by the previously defined parameter M . Thus, if this period of time increases then the stationarity of the channel would allow a greater averaging and thus a reduced noise and interference signal σ_u^2 would be obtained, as it can be observed in (29).

In (26), we apply a least squares (LS) technique in order to estimate the channel so that

$$\begin{aligned} \hat{h}[k] &= (\sqrt{1-\alpha}(q_o[k] + jb_q[k]))^{-1} \mathbb{E}_M \{\mathbf{R}_o\}[k] \\ &= h[k] + (\sqrt{1-\alpha}(q_o[k] + jb_q[k]))^{-1} u[k]. \end{aligned} \quad (30)$$

The channel estimation error can be written as

$$\begin{aligned} \Delta h[k] &= \hat{h}[k] - h[k] \\ &= (\sqrt{1-\alpha}(q_o[k] + jb_q[k]))^{-1} u[k]. \end{aligned} \quad (31)$$

The variance of the channel estimation error is defined as

$$\begin{aligned} \sigma_{\Delta h}^2 &= \mathbb{E} \left((\sqrt{1-\alpha}(q_o[k] + jb_q[k]))^{-1} u[k] \right. \\ &\quad \left. \left((\sqrt{1-\alpha}(q_o[k] + jb_q[k]))^{-1} u[k] \right)^H \right) = \\ &= K \frac{\sigma_u^2}{(1-\alpha)P} \mathbf{I}_K. \end{aligned} \quad (32)$$

Then, the received QAM symbols can be recovered from the received OQAM symbols after the known pilot sequences are removed, and joining the OQAM parts as follows

$$\begin{aligned} \tilde{S}[k_0, m'_0] &= \Re(R_o[k_0, m_0 = 2m'_0]) + \\ &\quad + j\Re(R_o[k_0, m_0 = 2m'_0 + 1]). \end{aligned} \quad (33)$$

Hence, the received complex symbols for m -th subcarrier are given by

$$\begin{aligned} \tilde{S}[k, m'] &= \Re \{ \sqrt{\alpha} \cdot H[k, 2m'] (S_o[k, 2m'] + jB_s[k, 2m']) \\ &\quad + \sqrt{1-\alpha} \cdot \Delta H[k, m] (Q_o[k, 2m'] + jB_Q[k, 2m']) + W[k, 2m'] \} \\ &\quad + j\Re \{ \sqrt{\alpha} \cdot H[k, m] (S_o[k, 2m'+1] + jB_s[k, 2m'+1]) \\ &\quad + \sqrt{1-\alpha} \cdot \{ \Delta H[k, m] (Q_o[k, 2m'+1] + jB_Q[k, 2m'+1]) \} \\ &\quad + W[k, 2m'+1] \}. \end{aligned} \quad (34)$$

The signal-to-noise ratio (SNR) of the received complex symbol for each subcarrier is defined as

$$\gamma_{\text{SNR}} = \frac{P\sigma_H^2}{2K\sigma_w^2}, \quad (35)$$

where $\sigma_H^2 = \sum_{l=0}^{L-1} \sigma_{h_l}^2$ is the power gain of the channel.

IV. ANALYSIS OF THE AVERAGE CHANNEL CAPACITY

For this analysis, the average channel capacity is defined as [20] [21]

$$C = A \cdot \mathbb{E} \{ \log_2(1 + \bar{\gamma}) \} \approx A \cdot \log_2(1 + \bar{\gamma}), \quad (36)$$

where $\bar{\gamma}$ corresponds to the signal-to-interference and noise ratio (SINR) of the system analyzed where A corresponds to $\frac{K-M_P}{K}$ in PSAM and to a unitary value in ST.

In PSAM schemes, N_P is defined as the number of pilot-symbols devoted to channel estimation where the channel can be perfectly recovered in the absence of noise and interference, when pilots are equipowered and equispaced. To guarantee an estimation with a minimum interpolation error the condition $N_P \geq L$ has to be accomplished as in [20] and [21]. We assume that the minimum number of pilots that are required in a multicarrier symbol is lower bounded by the number of channel taps $N_P = L$ to ensure channel acquisition. The maximum number of data resources can be obtained using the minimum number of pilots, i.e. $(K - L)$. Therefore, the maximum ratio of data symbols over the total number of subcarriers is $\frac{K-L}{K}$. Thus, the maximum α is determined as $\alpha_{\max} = 1 - L/K$. For comparison purposes, N_P is calculated as $N_P = \lceil (1 - \alpha)K \rceil$, i. e. the integer resulting from rounding up the product of the pilot power allocation by the number of subcarriers. Note that for ST schemes,

where it is assumed that every subcarrier transmits a pilot, the condition ($L \leq N_p$) is always satisfied.

Note that the CP length affects the average channel capacity, since it influences the signal-to-noise ratio applied over a multicarrier symbol. To take account for the CP in the following subsections, γ_{SNR} is penalized in each case.

A. PREVIOUS WORK IN OFDM

PSAM-OFDM

In this case, the channel estimation is not affected by either intrinsic self-interference or data interference of superimposition. Then, we have that $\sigma_u^2 = \sigma_w^2/M$. Thus, the average channel capacity obtained, as in [21], will be

$$C_{\text{PSAM}}^{\text{OFDM}} = \frac{K - N_p}{K} \mathbb{E} \{ \log_2 \{ 1 + \bar{\gamma}_{\text{PSAM}}^{\text{OFDM}} \} \}, \quad (37)$$

where the $\bar{\gamma}_{\text{PSAM}}^{\text{OFDM}}$ is defined as

$$\bar{\gamma}_{\text{PSAM}}^{\text{OFDM}} = \frac{\alpha(1-\alpha)KM}{\frac{1}{\gamma_{\text{SNR}}^{\text{OFDM}}} (K + KM(1-\alpha))}. \quad (38)$$

and $\gamma_{\text{SNR}}^{\text{OFDM}} = (K) \gamma_{\text{SNR}} / (K + \text{CP})$ where the SNR is penalized due to the presence of CP.

ST-OFDM

For this case considering that there is no intrinsic self-interference, but degradation caused by the data superimposition, the interference and noise variance will be $\sigma_u^2 = \sigma_w^2/M + \alpha P \sigma_H^2 / KM$. Thus, the average channel capacity can be written as in [20]

$$C_{\text{ST}}^{\text{OFDM}} = \mathbb{E} \{ \log_2 \{ 1 + \bar{\gamma}_{\text{ST}}^{\text{OFDM}} \} \}, \quad (39)$$

where the SINR is

$$\bar{\gamma}_{\text{ST}}^{\text{OFDM}} = \frac{\alpha(1-\alpha)KM}{\alpha K + \frac{1}{\gamma_{\text{SNR}}^{\text{OFDM}}} (K + KM(1-\alpha))}. \quad (40)$$

and $\gamma_{\text{SNR}}^{\text{OFDM}} = (K) \gamma_{\text{SNR}} / (K + \text{CP})$, where the SNR is penalized due to the presence of CP.

Note that for comparison purposes, (38) and (40) were adapted from [21] and [20], respectively, but now considering the channel estimation error in frequency domain.

B. AVERAGE CHANNEL CAPACITY FOR PSAM-FBMC

The average channel capacity in this case can be represented as

$$C_{\text{PSAM}}^{\text{FBMC}} = \frac{K - N_p}{K} \mathbb{E} \{ \log_2 \{ 1 + \bar{\gamma}_{\text{PSAM}}^{\text{FBMC}} \} \} \approx \frac{K - N_p}{K} \log_2 \{ 1 + \bar{\gamma}_{\text{PSAM}}^{\text{FBMC}} \}, \quad (41)$$

where $\bar{\gamma}_{\text{PSAM}}^{\text{FBMC}}$ corresponds to the SINR in PSAM-FBMC. Assuming that the interference is fully canceled the variance of the channel estimation error can be approximated using

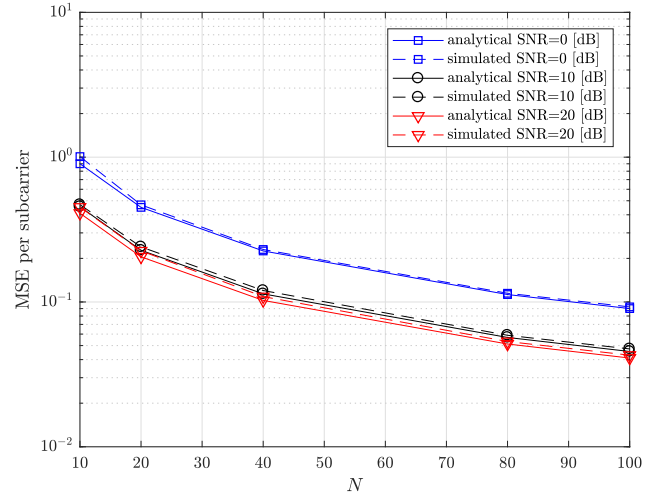


FIGURE 1: Average mean squared error versus M symbols for different SNR $\gamma_{\text{SNR}} = \{0, 10, 20\}$ [dB].

(32)

$$\begin{aligned} \sigma_{\Delta h}^2 &= K \frac{1}{(1-\alpha)P} \frac{1}{M} \left(\sigma_w^2 + \frac{\alpha P \sigma_H^2}{2K} \sigma_v^2 \right) \\ &\approx K \frac{1}{(1-\alpha)P} \frac{1}{M} (\sigma_w^2). \end{aligned} \quad (42)$$

Taking into account (42) similarly as in [21], the $\bar{\gamma}_{\text{PSAM}}^{\text{FBMC}}$ can be represented as

$$\bar{\gamma}_{\text{PSAM}}^{\text{FBMC}} = \frac{\sigma_H^2 \frac{P_s}{M}}{\sigma_{\Delta h}^2 \frac{P_s}{M} + \sigma_w^2}, \quad (43)$$

where simplifying

$$\bar{\gamma}_{\text{PSAM}}^{\text{FBMC}} = \frac{\alpha}{\alpha \frac{\sigma_{\Delta h}^2}{\sigma_H^2} + \frac{M}{P} \frac{\sigma_w^2}{\sigma_H^2}}, \quad (44)$$

and $\sigma_{\Delta h}^2 / \sigma_H^2$ is obtained from (42)

$$\frac{\sigma_{\Delta h}^2}{\sigma_H^2} = K \frac{1}{(1-\alpha)P} \frac{1}{M} \left(\frac{\sigma_w^2}{\sigma_H^2} \right). \quad (45)$$

Then, the SINR is obtained similarly to PSAM-OFDM which differs from PSAM-FBMC due to the presence of a CP. Thus,

$$\bar{\gamma}_{\text{PSAM}}^{\text{FBMC}} = \frac{\alpha(1-\alpha)KM}{\frac{1}{\gamma_{\text{SNR}}^{\text{FBMC}}} (K + KM(1-\alpha))}, \quad (46)$$

where $\gamma_{\text{SNR}}^{\text{FBMC}} = \gamma_{\text{SNR}}$ because no CP is present in this case.

C. AVERAGE CHANNEL CAPACITY FOR ST-FBMC

The average channel capacity in this case can be represented as

$$C_{\text{ST}}^{\text{FBMC}} = \mathbb{E} \{ \log_2 \{ 1 + \bar{\gamma}_{\text{ST}}^{\text{FBMC}} \} \} \approx \log_2 \{ 1 + \bar{\gamma}_{\text{ST}}^{\text{FBMC}} \}. \quad (47)$$

Applying (23)- [20] on (34) and taking into account the power constraints of the OQAM transmission system, $\bar{\gamma}_{\text{ST}}^{\text{FBMC}}$

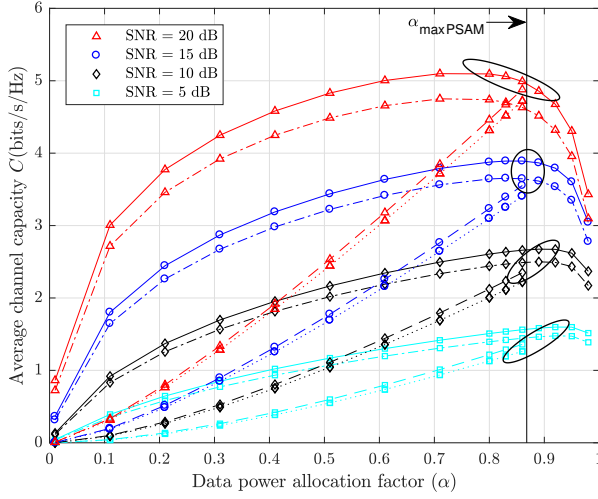


FIGURE 2: Average channel capacity of OFDM-PSAM (dashed), OFDM-ST (dot-dashed), FBMC-AuP (dotted) and FBMC-ST (solid-line) techniques versus α , for different SNR.

can be represented as

$$\bar{\gamma}_{ST}^{FBMC} = \frac{\sigma_H^2 \frac{P_S}{M}}{\sigma_{\Delta_h}^2 \frac{P_S}{M} + \sigma_{\Delta_h}^2 \frac{P_P}{M} + \sigma_W^2}, \quad (48)$$

where, simplifying the expression, we can find that

$$\bar{\gamma}_{ST}^{FBMC} = \frac{\alpha}{\frac{\sigma_{\Delta_h}^2}{\sigma_H^2} + \frac{M}{P} \frac{\sigma_W^2}{\sigma_H^2}}, \quad (49)$$

where $\sigma_{\Delta_h}^2/\sigma_H^2$ is obtained replacing (29) in (32) as

$$\frac{\sigma_{\Delta_h}^2}{\sigma_H^2} = K \frac{1}{(1-\alpha)P} \frac{1}{M} \left(\frac{\sigma_W^2}{\sigma_H^2} + \frac{\alpha P}{2K} (1+\sigma_v^2) \right), \quad (50)$$

and σ_W^2/σ_H^2 can be obtained from (35). Therefore, (49) can be simplified as

$$\bar{\gamma}_{ST}^{FBMC} = \frac{\alpha (1-\alpha) K M}{\alpha \frac{K}{2} (1 + \sigma_v^2) + \frac{1}{\bar{\gamma}_{ST}^{FBMC}} (K + K M (1-\alpha))}. \quad (51)$$

where $\bar{\gamma}_{ST}^{FBMC} = \gamma_{SNR}$, because no CP was used.

V. ANALYSIS OF THE POWER ALLOCATION

The data power allocation factor α represents the ratio of data-signal power with respect to the total power. As it can be observed in previous expressions, the average channel capacity depends on this factor. Moreover, it is easy to show that the capacity expressions are convex thus, the optimal solutions can be derived.

A. PSAM-FBMC

The average channel capacity in PSAM-FBMC can be obtained replacing (46) in (41). The value of α that maximizes

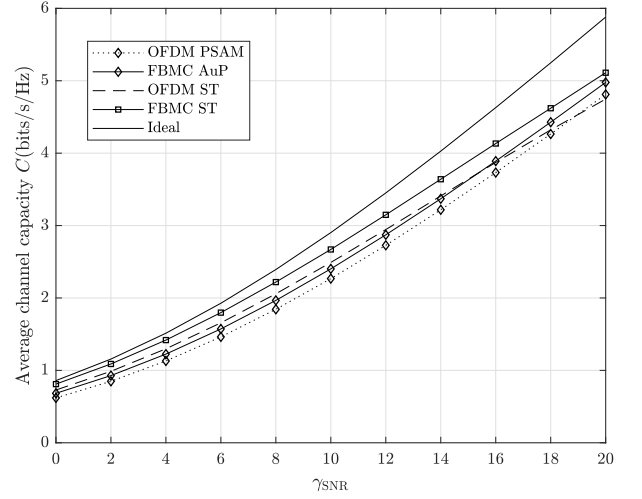


FIGURE 3: Average channel capacity of OFDM and FBMC techniques with ST and PSAM cases versus SNR.

the capacity can be found deriving $\partial C_{PSAM}/\partial \alpha = 0$. Thus,

$$\begin{aligned} \alpha_{opt}^{PSAM-FBMC} &= \alpha : -\log_2 (1 + \bar{\gamma}_{AuP}^{FBMC}) \\ &= \frac{1}{\ln(2)} \frac{(1-2\alpha)(1+M(1-\alpha)) + M\alpha(1-\alpha)}{(1+M(1-\alpha))(1-\alpha)} \end{aligned} \quad (52)$$

The value of α has to be selected between the maximum and optimal value. If the $\alpha_{opt} \leq \alpha_{max}$, α_{opt} can be used. If $\alpha_{opt} \geq \alpha_{max}$, α_{max} has to be used because a greater value of α would not guarantee the channel identifiability conditions.

B. ST-FBMC

The average channel capacity in ST-FBMC can be obtained replacing (51) in (47). The value of α that maximizes the capacity can be computed through $\partial C_{ST}^{FBMC}/\partial \alpha = 0$. Thus, $\alpha^2(KM - \frac{K}{2}(1+\sigma_v^2)\gamma_{SNR}) - 2\alpha K(1+M) + K(1+M) = 0$, and

$$\alpha_{opt}^{ST-FBMC} = \frac{2K(1+M) - \sqrt{4K^2(1+M)^2 - 4A(K+KM)}}{2A}, \quad (53)$$

where $A = KM - \frac{K}{2}(1+\sigma_v^2)\gamma_{SNR}$.

VI. SIMULATION RESULTS

A comparison between OFDM and FBMC technologies has been carried out. Also, two channel estimation techniques, ST and PSAM, have been applied independently, using PHYDYAS filter for FBMC-OQAM implementation. For a fair comparison, the data and pilot power are the same for every analyzed system, where $M=64$ is considered.

The channel is assumed to be Rayleigh distributed with $L=8$ and a power delay profile $\mathbb{E}\{|h_l|^2\} = e^{-\frac{l}{10}}$. The CP length is assumed to be equal to L in the OFDM cases. The obtained results are described in the following paragraphs.

In Fig. 1, the channel estimation error is depicted for $\gamma_{SNR} = \{0, 10, 20\}$ [dB] and $\alpha=0.8$ where the pilots are su-

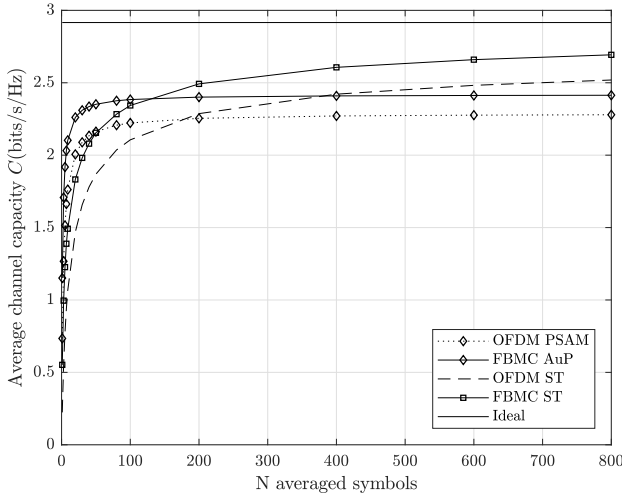


FIGURE 4: Average channel capacity of OFDM and FBMC techniques with ST and PSAM cases versus N for SNR=10 dB.

perimposed to QPSK data. Note that this constellation is specified for demonstration purposes. Similar results can be obtained, for MSE performance, using a 16QAM constellation, not shown for the sake of space. The analytical expression deduced in (32) matches with the MSE performance obtained in simulations.

Fig. 2 depicts the average channel capacity with respect to the data power allocation factor. The optimal α for different SNR can be obtained using (53) and this value coincides with the optimal α obtained through exhaustive search performed by simulations. It can be observed (rounded in circles) the superiority of ST schemes in comparison with the average channel capacity of the PSAM strategies in OFDM and FBMC systems. In this case the averaging is set to 100 slots and using 7 symbols per slot.

Fig. 3 illustrates the comparison of ST and PSAM for different multicarrier transmission systems as a function of the SNR, respectively. We can see that for specially low and medium SNR, ST-FBMC system is superior to dedicated pilot techniques. This is because the interference, produced by the superimposed information and by intrinsic self-interference becomes comparatively greater than the noise with high SNR. The scenario is tested using an averaging of 100 slots and 7 symbols per slot.

In Fig. 4, the average channel capacity is depicted as a function of the number of averaged symbols using a fixed SNR [dB]. We can see that ST overcomes dedicated pilot schemes in OFDM and FBMC. Moreover, ST-FBMC outperforms ST-OFDM, particularly in channels with a certain stationarity, what makes ST-FBMC an outstanding option for physical layer in future wireless communication systems.

VII. CONCLUSIONS

ST-FBMC-OQAM provides a better performance than PSAM-FBMC-OQAM and any OFDM schemes in terms of

average channel capacity, due to the fact that either CP or dedicated pilots sequences are not required.

We provided an analytical expression of the channel estimation error of ST-FBMC-OQAM in order to compute the channel capacity expression. This MSE is based on three terms: the noise, the intrinsic self-interference produced by the prototype filter, and the superimposed data-signal given by ST systems. However, with the averaging technique, we can effectively mitigate these undesirable terms. Additionally, we also detailed the optimum power allocation for the pilots in order to obtain the highest capacity.

APPENDIX

Let us consider an $K_f \times M_f$ self-interference intrinsic matrix $\mathcal{V} = \sum_{m=-\infty}^{M-1} \sum_{k=0}^{K-1} S_o[k, m] e^{j\phi[k, m]}$, which was previously characterized using an analogous method in [24]. The parameters K_f and M_f correspond to the number of subcarriers and time samples analyzed affected by the filters, respectively. The total self-interference term is computed, in an extended matrix form, at the k -th subcarrier for M consecutive symbols as

$$\mathcal{V}_{e[M \times (K_f \times L_T)]} = [\mathcal{V}_1[M \times L_T], \mathcal{V}_2[M \times L_T], \dots, \mathcal{V}_{K_f}[M \times L_T]] \quad (54)$$

where $L_T = M_f + M - 1$.

$$\mathcal{V}_{S[K_f \times L_T] \times 1} = \mathcal{V}_e^T 1_{[M \times 1]} \quad (55)$$

thus, the accumulated self-interference intrinsic power over M consecutive symbols is

$$\sigma_v^2 = \mathbb{E} \{ \mathcal{V}_S \mathcal{V}_S^H \}. \quad (56)$$

Note that in Table 1 we can see the corresponding $\mathcal{V}[m, n]$ coefficients for a PHYDYAS filter.

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TABLE 1: Intrinsic interference caused to data placed at $[k_0, m_0]$ using the PHYDYAS prototype filter.

	$m_0 - 3$	$m_0 - 2$	$m_0 - 1$	m_0	$m_0 + 1$	$m_0 + 2$	$m_0 + 3$
$k_0 - 1$	$-j0.0429$	0.125	$j0.2058$	0.2393	$-j0.2058$	-0.125	$j0.0429$
k_0	-0.0668	0	0.5644	1	0.5644	0	-0.0668
$k_0 + 1$	$j0.0429$	-0.125	$-j0.2058$	0.2393	$j0.2058$	0.125	$-j0.0429$

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