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Superimposed Training for Channel Estimation in FBMC-OQAM

(Invited Paper)

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***Abstract*—Wireless broadband communication systems are always requiring higher data rates. In order to achieve this goal, we should provide advanced schemes which are capable of improving the spectral efficiency and reusing, in a better way, the available radio spectrum resources. Filter Bank Multi-Carrier Offset Quadrature Amplitude Modulation (FBMC-OQAM) combined with Superimposed Training (ST) is a promising technique with a very high spectral efficiency. This improvement is due to the low out-of-band emissions of FBMC-OQAM, because it uses a well-designed prototype filter, and the lack of dedicated pilot tones owing to ST scheme. However, this combination is not straightforward due to the appearance of the self-interference at receiver side in FBMC-OQAM. In this paper, we provide a novel channel estimation which is capable of dealing with this self-interference in the context of combining these two techniques.**

I. Introduction

We are expecting a revolution in wireless technologies where the subscribers are demanding an unprecedented increase of data rates in the range of 1000 times [1]. In order to achieve this goal, we should use multi-carrier modulation techniques, which have a high spectral efficiency. The most widely-used is Orthogonal Frequency Division Multiplexing (OFDM). Despite its well-known advantages such as robustness against multi-path fading and ease of implementation, it also has severe drawbacks. One of the most important disadvantages is its high side lobes in the frequency domain, which are interfering other services placed at adjacent bands. In order to avoid this issue, it is necessary to leave several unused frequency resources (guard-bands) at both ends of the signal, reducing the spectral efficiency.

Currently, one of the potential waveform candidates to replace the well-known OFDM is Filter Bank Multi-Carrier Offset Quadrature Amplitude Modulation (FBMC-OQAM) [2]. Comparing these two techniques, FBMC-OQAM has a higher spectral efficiency than OFDM due to the suppression of the cyclic prefix and the low out-of-band emissions with the use of a well-designed prototype filter [3]. Therefore, FBMC-OQAM can make a better use of the available spectrum, reducing the existing guard-bands.

Multi-carrier modulation systems require channel equalization in order to perform a coherent detection. In OFDM, there already exist several methods for this purpose. However, those existing methods are not straightforward to be reused in FBMC-OQAM, because the intrinsic interference, caused by the prototype filters, must be canceled 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 [4] and scattered pilot-based [5]. All these techniques are using a considerable amount of time-frequency resources to allocate pilot information instead of data, thus, reducing the effective rate of the system.

Superimposed Training (ST) is a semi-blind channel estimation technique, which achieves a better spectral efficiency. It consists in arithmetically superimposing a pilot over the data signal [6], allowing us to use all available resources to transmit information and, thus, enhance the bandwidth efficiency. Indeed, this scheme has already been proposed for OFDM.

The combination of OFDM-OQAM and ST scheme has been addressed in [7]. It shows a great improvement in the spectral efficiency due to this combination. In this reference, the channel equalization is based on a correlated-based method in order to reduce the self-interference. However, it is difficult to evaluate the goodness of this proposal, since it does not show its effectiveness in terms of Mean Squared Error (MSE) of the channel estimation, or the Symbol Error Rate (SER) of the entire system.

The objective of this paper is to propose the combination of FBMC-OQAM with ST, where we can achieve even a better improvement in terms of spectral efficiency, compared to [5]. This extra improvement is due to the use of the well designed prototype filter considered in FBMC-OQAM, that we mentioned before. Moreover, we propose a novel channel estimation technique with self-interference cancellation. Finally, the performance of the system is evaluated through the MSE of channel estimation and SER, comparing it to the Auxiliary Pilot (AP) method [8].

The remainder of the paper is organized as follows. Section II provides the system model. Section III describes the proposed channel estimation method when FBMC-OQAM is combined with ST. Section IV presents the simulation results to provide a better understanding of the entire system. Finally, in section V some conclusions are pointed out.

Notation: j denotes $\sqrt{-1}$. x denotes a scalar value, $\mathbf{x}$ denotes a vector whose first element is $x[0]$ and $\mathbf{X}$ denotes a matrix whose first element is $X[0,0]$. $\Re(x)$ and $\Im(x)$ are the real and imaginary part of x respectively. $*$ denotes the convolution operation. $\otimes$ denotes the Kronecker product of two matrices. $\mathbf{1}_{M,N}$ denotes a matrix of ones of size $M \times N$. $E[\mathbf{x}]$ denotes the expectation of $\mathbf{x}$. $E_N[\mathbf{x}] = (1/N)\sum_{n=0}^{N-1} x[n]$ denotes the average over N elements of $\mathbf{x}$.

II. System Model

The system model is focused on a point-to-point link, where a FBMC-OQAM waveform is transmitted.

A. FBMC-OQAM

Let $\mathbf{S}$ denote a matrix containing the set of $M \times N/2$ complex information symbols to be transmitted, which belong to a QAM constellation. Given $\mathbf{S}$, we transform it into OQAM real symbols, where the real and imaginary parts of $\mathbf{S}$ are separated to build $\mathbf{S_o}$

$$S_o[m,2n] = \left\{ \begin{array}{ll} \Re(S[m,n]), & m \quad even \\ \Im(S[m,n]), & m \quad odd \end{array} \right\} \tag{1}$$

and

$$S_o[m,2n+1] = \left\{ \begin{array}{ll} \Im(S[m,n]), & m \quad even \\ \Re(S[m,n]), & m \quad odd \end{array} \right\}. \tag{2}$$

with $m = 0 \cdots M-1$ and $n = 0 \cdots N/2-1$. Note that the size of $\mathbf{S_o}$ is $M \times N$, where M represents the number of parallel subcarriers and N represents the number of FBMC-OQAM symbols transmitted over the time.

The baseband model of the FBMC-OQAM transmitted signal is given by

$$x(t) = \sum_{n=0}^{N-1}\sum_{m=0}^{M-1} S_o[m,n]e^{j\phi[m,n]} g\left(t - n\frac{T_s}{2}\right) e^{j2m\pi f_s t}, \tag{3}$$

where $f_s = 1/T_s$ is the subcarrier spacing, $g[t]$ is the prototype filter impulse response and the phase $\phi[m,n]$ is given by

$$\phi[m,n] = \left\{ \begin{array}{ll} 0, & m+n \ \ even \\ \frac{\pi}{2}, & m+n \ \ odd \end{array} \right\}. \tag{4}$$

We can define $g_{m,n}[t]$ as a time-frequency shifted version of $g[t]$, which can expressed by

$$g_{m,n}(t) = g\left(t - n\frac{T_s}{2}\right) e^{j2m\pi f_s t}. \tag{5}$$

Therefore, we can rewrite (3) as

$$x(t) = \sum_{n=0}^{N-1}\sum_{m=0}^{M-1} S_o[m,n]e^{j\phi[m,n]} g_{m,n}(t). \tag{6}$$

Note that this operation is also known as the Synthesis Filter Bank (SFB).

In the absence of channel effects (distortion-free and noiseless), the demodulated signal at subcarrier m_0 and symbol n_0 can be expressed by

$$\begin{aligned} R_o[m_0,n_0] &= \langle x(t), g_{m_0,n_0}(t)\rangle = \\ &= \sum_n \sum_m \int_{-\infty}^{\infty} S_o[m,n]e^{j\phi[m,n]} g_{m,n}(t) g^*_{m_0,n_0}(t)dt = \\ &= \underbrace{S_o[m_0,n_0]e^{j\phi[m_0,n_0]}}_{\text{desired signal}} + \underbrace{\sum_{(m,n)\neq}\sum_{(m_0,n_0)} V_{m_0,n_0}[m,n]}_{\text{self-interference}}, \end{aligned} \tag{7}$$

where $V_{m_0,n_0}[m,n]$ is the interference contribution due to the residual components of the prototype filter, caused by the adjacent symbols. Its expression is given by

$$V_{m_0,n_0}[m,n] = \int_{-\infty}^{\infty} S_o[m,n]e^{j\phi[m,n]} g_{m,n}(t) g^*_{m_0,n_0}(t)dt. \tag{8}$$

Note that the operation in (7) is also known as the Analysis Filter Bank (AFB).

Given (7) and (8), it is straightforward to show that the desired signal is always orthogonal to the self-interference thanks to both $\phi[m,n]$ and the prototype filter design. Therefore, the expression in (7) can be simplified by

$$\begin{aligned} &R_o[m_0,n_0] = \\ &= \left\{ \begin{array}{ll} S_o[m_0,n_0] + jB[m_0,n_0], & \phi[m_0,n_0] = 0 \\ jS_o[m_0,n_0] + B[m_0,n_0], & \phi[m_0,n_0] = \frac{\pi}{2} \end{array} \right\}, \end{aligned} \tag{9}$$

where $B[m_0,n_0]$ is a real quantity that results from the self-interference described in (7). Then, the total interference is either real or pure imaginary depending on $[m_0,n_0]$ values.

Finally, all the real symbols $\mathbf{S_o}$ are recovered taking the real or imaginary part of $R_o[m_0,n_0]$ depending on the index values $[m_0,n_0]$.

B. Effects of the Channel in FBMC-OQAM

In the previous subsection, we have provided the analytical model for FBMC-OQAM under a scenario of absence of channel effects. However, in a realistic environment, the signal $x(t)$ goes through a frequency-selective fading channel. The received signal can be expressed as

$$y(t) = h(t,\tau) * x(t) + w(t), \tag{10}$$

where $h(t,\tau)$ represents the channel impulse response of the multipath channel with maximum delay spread τ_{max} and $w(t)$ is the Additive White Gaussian Noise (AWGN) with distribution $w \sim \mathcal{CN}(0,\sigma_w^2)$.

According to (10), (9) will be also modified by the channel effects. Without loss of generality and for the sake of saving space, we focus on the case where $\phi[m_0,n_0] = 0$. Note that for the other case, the analysis is analogous. Now, the received signal is given by

$$\begin{aligned} R_o[m_0,n_0] = H[m_0,n_0]\left(S_o[m_0,n_0] + jB[m_0,n_0]\right) + \\ + W[m_0,n_0], \end{aligned} \tag{11}$$

where $H[m_0,n_0]$ and $W[m_0,n_0]$ are the frequency response of the channel and noise respectively at subcarrier m_0 and symbol n_0. Note that $\mathbf{H}$ is a matrix of size of $M \times N$, where each element is defined by

$$H[m,n] = H(f,t)\Big|_{f=mf_s, \quad t=n\frac{T_s}{2}}, \tag{12}$$

where $H(f,t)$ is defined in [9] by

$$H(f,t) = \int_0^{\tau_{max}} h(t,\tau)e^{-j2\pi mf\tau}d\tau. \tag{13}$$

Additionally, $W[m_0,n_0]$ is also an AWGN random variable with distribution $W \sim \mathcal{CN}(0, M\sigma_w^2)$.

Inspecting (11), it is obvious that if we want to apply any equalization method developed for OFDM, we will have to remove the self-interference part of the equation previously. Otherwise, the channel estimation would not be as accurate as it should, and the performance of entire system would be compromised.

C. Auxiliary Pilot (AP)

Channel estimation may be obtained with the insertion of pilot information in the time-frequency grid. For the case of scattered-pilot symbols in FMBC-OQAM, the AP scheme is a very simple and effective way canceling the interference at the transmitter side. For each Traditional Pilot (TP), we have to place another AP close to it in order to compensate its interference term. Then, we can perform the channel estimation.

According to [8], if we have a TP placed at $[m_0,n_0]$, the AP will be placed at $[m_0,n_0+1]$. The interference caused by the AP at $[m_0,n_0]$ should cancel the self-interference of TP described in (11). Thus, the AP value should be computed as

$$S_o[m_0,n_0+1] = -j\frac{\sum\limits_{\substack{(m,n)\neq \\ (m,n)\neq}} \sum\limits_{\substack{(m_0,n_0) \\ (m_0,n_0+1)}} V_{m_0,n_0}[m,n]}{V_{m_0,n_0}[m_0,n_0+1]}. \tag{14}$$

At the receiver side, the self-interference part of (11) has been mitigated. Now, we can perform any traditional channel estimation scheme.

The main drawback of this method is that it is not only reducing the amount of the valuable resources for data transmission, but also increasing the complexity at transmitter side in order to compute all the APs for each symbol frame.

D. Superimposed Training (ST) for FBMC-OQAM

ST is a well-known method already applied to OFDM, which consists in adding the TP over the data symbols before transmission. Its main advantage is the improvement of the available rate.

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

$$\mathbf{Q_o} = \mathbf{q_o} \otimes \mathbf{1_{1,N/2}}. \tag{15}$$

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 suppose that our system has an available power per time-frequency resource P_a. When we add $\mathbf{S_o}$ with $\mathbf{Q_o}$, we must satisfy the available power constraint given by the following expression

$$P_a = \alpha E\left[|\mathbf{S_o}|^2\right] + (1-\alpha)E\left[|\mathbf{Q_o}|^2\right], \;\; 0 \leq \alpha \leq 1, \tag{16}$$

where α is a power coefficient. Therefore, the new data-pilot matrix $\mathbf{T_o}$ is given by

$$\mathbf{T_o} = \sqrt{\alpha}\mathbf{S_o} + \sqrt{(1-\alpha)}\mathbf{Q_o}. \tag{17}$$

Once we have the new data-pilot sequence given by (17), we can transmit it through the FBMC-OQAM system using (3), where $\mathbf{S_o}$ is replaced by $\mathbf{T_o}$.

III. Channel Equalization for ST-FBMC-OQAM

Given the system model detailed in the previous section, the new received signal can be expressed as

$$\begin{aligned} &R_o[m_0,n_0] = H[m_0,n_0]\cdot \\ &\cdot\Big(\sqrt{\alpha}\left(S_o[m_0,n_0] + jB_S[m_0,n_0]\right) + \; + W[m_0,n_0] + \\ &+\sqrt{(1-\alpha)}\left(Q_o[m_0,n_0] + jB_Q[m_0,n_0]\right)\Big), \end{aligned} \tag{18}$$

where $jB_S[m_0,n_0]$ is the interference caused by $S_o[m_0,n_0]$ and $jB_Q[m_0,n_0]$ by $Q_o[m_0,n_0]$.

Given (18), first we need to estimate the channel based on the pilot sequence, and then, we detect the data after performing the equalization.

In order to estimate the channel, we average $\mathbf{R_o}$ over the variable n for each subcarrier m. The expression is given by

$$\begin{aligned} &E_N\left[R_o[m,n]\right] = E_N\left[H[m,n]\right]\cdot \\ &\cdot\Big(E_N\left[\sqrt{\alpha}\left(S_o[m,n] + jB_S[m,n]\right)\right] + E_N\left[W[m,n]\right] + \\ &+E_N\left[\sqrt{(1-\alpha)}\left(Q_o[m,n] + jB_Q[m,n]\right)\right]\Big). \end{aligned} \tag{19}$$

The channel frequency response can be considered invariant during N symbols. Therefore

$$H[m,n] = h[m], \quad n = 0\cdots N-1. \tag{20}$$

Looking at (19), if N is large enough, then the following statements hold:

- The mean value of the noise tends to be zero. Using the Central Limit Theorem and some normal distribution properties, it is straightforward to obtain its distribution, which is given by

$$E_N\left[W[m,n]\right] \sim \mathcal{CN}\left(0, \frac{M\sigma_w^2}{N}\right). \tag{21}$$

- The mean value of the OQAM symbols tend to be zero, and its distribution can be expressed by

$$E_N\left[\sqrt{\alpha}S_o[m,n]\right] \sim \mathcal{CN}\left(0, \frac{\alpha}{N}\right). \tag{22}$$

- Additionally, the interference caused by neighbour OQAM symbols also tends to be zero, and its distribution can be expressed by

$$E_N\left[\sqrt{\alpha} j B_S[m,n]\right] \sim \mathcal{CN}\left(0, \frac{\alpha \sum_n \sum_m \sigma^2_{v[m,n]}}{N}\right), \tag{23}$$

where $\sigma^2_{v[m,n]}$ denotes the variance of the self-interference.
- The pilot sequence for the m-th subcarrier always contains the same value at every time instant, therefore it is a deterministic value. Moreover, the surrounding neighbour pilot sequences are also deterministic values.

Thus, (19) can be simplified by

$$\begin{aligned} E_N\left[R_o[m,n]\right] = h[m]\cdot \\ \sqrt{(1-\alpha)}\left(q_o[m] + jb_Q[m]\right) + u[m], \end{aligned} \tag{24}$$

where $jb_Q[m]$ is the deterministic interference vector which can be computed using (7) and (8); and $u[m]$ is a residual component. This last term appears because the averaging process is not able to completely suppress the noise and interferences. Its distribution can be expressed by

$$u \sim \mathcal{CN}\left(0, \frac{M\sigma_w^2 + \alpha\left(1 + \sum_n \sum_m \sigma^2_{v[m,n]}\right)|h[m]|^2}{N}\right). \tag{25}$$

Given the expression in (24), we can apply a Least Squares (LS) technique in order to estimate the channel, which is given by

$$\widehat{\mathbf{h}} = \frac{E_N\left[\mathbf{R_o}\right]}{\sqrt{(1-\alpha)}\left(\mathbf{q_o} + j\mathbf{b_Q}\right)}. \tag{26}$$

Note that the sequence $\sqrt{(1-\alpha)}\left(\mathbf{q_o} + j\mathbf{b_Q}\right)$ is a known sequence at the receiver side. Therefore, we can store it in a look-up-table saving extra amount of computations and reducing the complexity and signal processing delay.

Finally, in order to obtain the equalized symbols with a Zero-Forcing (ZF) technique, we just need to apply the following expression

$$\widehat{\mathbf{S_o}} = \widehat{\mathbf{H}}^{-1}\mathbf{R_o} - \sqrt{(1-\alpha)}\mathbf{Q_o}, \tag{27}$$

where $\widehat{\mathbf{H}}$ is defined as

$$\widehat{\mathbf{H}} = \widehat{\mathbf{h}} \otimes \mathbf{1_{1,N}}. \tag{28}$$

IV. Numerical Results

In this section we show some numerical results comparing the performance of AP and the proposed ST scheme. First, we define the simulation parameters.

A. Simulation Parameters

In table I, we can see the numeric values for the variables that we defined in the previous sections. In order to set the value of N, we have computed the coherence time of the channel given the maximum Doppler frequency. In this case, we have chosen a well-known LTE EVA channel model with an intermediate value of Doppler frequency [10] [11]. The coherence time [12] is slightly larger than 6ms, which corresponds to about 12 slots [13]. This means that the channel remains almost invariant up to 12 slots or 168 FBMC-OQAM symbols.

TABLE I
Simulation Parameters

M	64	**Overlap. Factor**	4
N	up to 168	**Prototype Filter**	PHYDYAS
Subc. Spacing	15 KHz	**Constellation**	QPSK
Channel Model	LTE EVA 70Hz	**Slot time**	0.5ms
1 Slot = 7 OFDM-QAM symbols = 14 FBMC-OQAM symbols			

In the particular case of AP scheme, we are placing a total of 32 pilots tones (16 TP and 16 AP) for each slot, equally distributed over the frequency axis. Once we have performed the channel estimation at pilot positions, we perform an interpolation. Finally, we equalize the symbols with a ZF technique.

B. Definition of Signal-to-Noise (SNR)

The SNR is defined by

$$\eta = E\left[|x(t)|^2\right] / \sigma_w^2. \tag{29}$$

However, in the case of AP scheme, the value η should be penalized due to the use of dedicated pilot tones. These pilots tones mean a reduction of the system rate. Therefore, the effective η_e can be defined as

$$\eta_e = \eta\left(1 - \frac{\text{\#pilots}}{\text{\#resources in one slot}}\right) = \eta\left(1 - \frac{32}{14M}\right). \tag{30}$$

In the case of ST scheme, the effective η_e is directly η, because all resources are available for data transmission.

C. Results

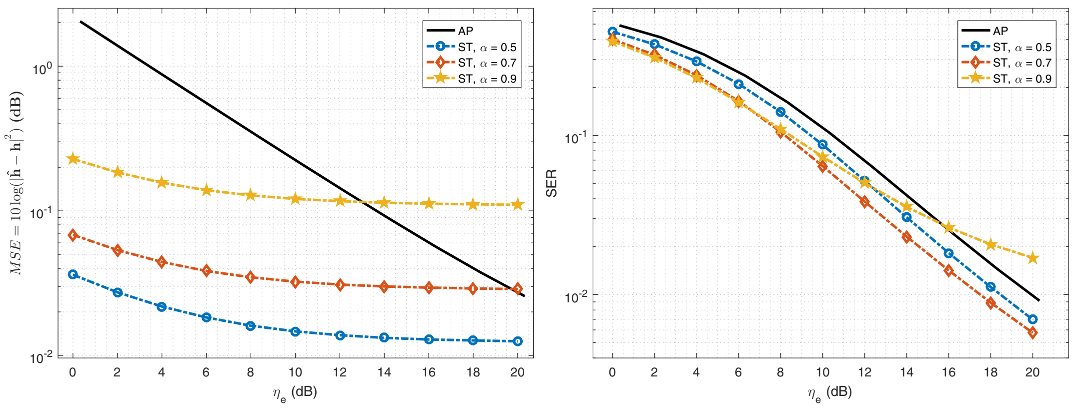
In Fig. 1 we compare the performance between AP and the proposed ST scheme, where ST has different values of α. In Fig. 1a, we can see that the results for $\alpha \geq 0.7$ are much better than AP, up to one order of magnitude. Thus, we must be careful setting the optimal value of α. Additionally, in Fig. 1b we can also see that we have around 2.5dB of SNR improvement compared to AP in the best case.

In Fig. 2 we compare again the performance between AP and ST schemes, when the channel estimation for ST has been averaged over a different length of N FBMC-OQAM symbols. We can also see that, at the low SNR region, ST outperforms AP in all cases. When $N = 6 \times 14$, ST even outperforms AP for the high SNR scenario.

V. Conclusions

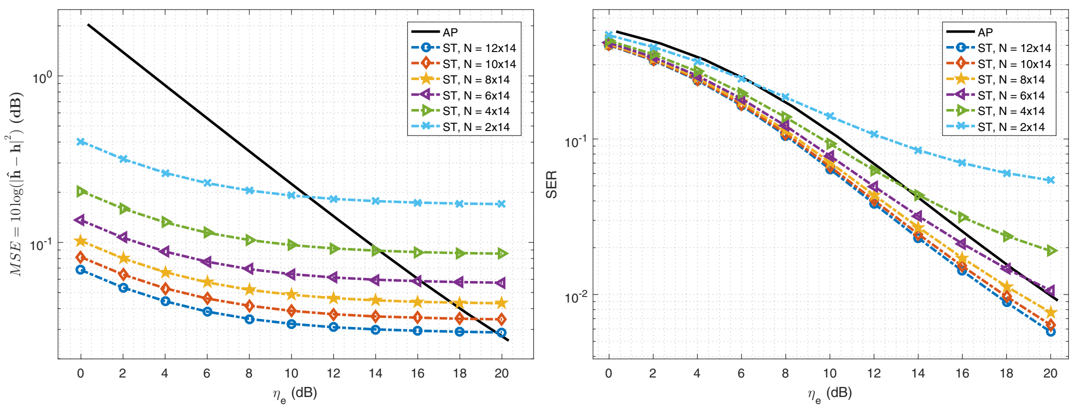
In this paper we have proposed the combination of ST with FBMC-OQAM in order to improve the spectral efficiency. We have also introduced an effective technique to estimate the channel under this scenario.

Our proposal outperforms previous works, such as AP, without increasing power. Additionally, ST technique provides full availability of all time-frequency resources for data transmission.



(a) MSE. (b) SER.

Fig. 1. Comparative results of AP and ST for $N = 168$ and different values of α.



(a) MSE. (b) SER.

Fig. 2. Comparative results of AP and ST for $\alpha = 0.7$ and different values of N.

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