Channel Estimation in Filter Bank-Based Multicarrier Systems: Challenges and Solutions

Eleftherios Kofidis

Dept. of Statistics and Insurance Science
University of Piraeus, Greece,
and
Computer Technology Institute and Press “Diophantus” (CTI), Greece.
Email: kofidis@unipi.gr

ABSTRACT
Filter bank-based multicarrier (FBMC) techniques have attracted a lot of interest in the last decade as a competitive alternative to the long established OFDM, particularly in wireless applications. Their potential stems from their increased ability to carrying a flexible spectrum shaping together with a major increase in spectral efficiency and robustness to synchronization requirements, features of fundamental importance in future mobile networks. A particular type of FBMC, the so-called FBMC/OQAM system, consisting of pulse shaped OFDM carrying offset QAM (OQAM) symbols, has received increasing attention due to, among other features, its potential for maximum spectral efficiency. It suffers, however, from an inherent inter-carrier/inter-symbol interference that complicates signal processing tasks such as channel estimation. The goal of this paper is to concisely present the state-of-the-art in channel estimation for FBMC/OQAM, putting emphasis on the problems that are still open and outlining related on-going research. Preamble-based channel estimation is given special attention in the realistic scenario of highly frequency/time selective channels and some new results are reported in this context.

Index Terms— Channel estimation, FBMC, intrinsic interference, OFDM, OQAM, pilot, preamble

1. INTRODUCTION
In its corresponding article, Wikipedia stresses the following as the first of the key research topics for the attainment of 5G goals: “New data coding and modulation techniques, including filter bank multicarrier or non-orthogonal multiple access schemes” [1]. Indeed, filter bank-based multicarrier (FBMC) techniques have been recently shown to offer an attractive alternative to OFDM as the modulation scheme of choice in future mobile networks. Their potential stems from (among others) their increased ability to carrying a flexible spectrum shaping together with a major increase in spectral efficiency and robustness to synchronization requirements. Notable examples of modern wireless applications adopting FBMC include cognitive radio [2, 3], beyond-LTE [4], and the accommodation of broadband data services between narrowband (e.g., professional mobile radio (PMR)) channels [5].

A particular type of FBMC, the so-called FBMC/OQAM (or OFDM/OQAM) system, consisting of pulse shaped OFDM carrying offset QAM (OQAM) symbols, has received increasing attention due to, among other features, its potential for maximum spectral efficiency [6]. It suffers, however, from an inherent (intrinsic) inter-carrier/inter-symbol interference that complicates signal processing tasks such as channel estimation [7]. Most of the research on FBMC/OQAM channel estimation so far has relied on the assumption of channels that are still open and outlining related on-going research. Preamble-based channel estimation is given special attention in the realistic scenario of highly frequency/time selective channels and some new results are reported in this context.

The goal of this paper is to concisely present the state-of-the-art in channel estimation for FBMC/OQAM, putting emphasis on the problems that are still open and outlining related on-going research. Preamble-based channel estimation is given special attention in the challenging scenario of highly frequency/time-selective channels and some new results are reported in this context.
2. ON CHANNEL ESTIMATION IN FBMC/OQAM

The (baseband) output of an FBMC/OQAM synthesis filter bank (SFB) can be written as

\[ s(l) = \sum_{m=0}^{M-1} \sum_{n} d_{m,n}g_{m,n}(l), \]

where \((m, n)\) refers to the \(m\)th subcarrier and the \(n\)th FBMC symbol, \(d_{m,n}\) are real OQAM symbols, \(M\) is the number of subcarriers, \(g_{m,n}(l) = g(l-n\frac{M}{2}) e^{j\frac{2\pi}{M}(l-\frac{m}{M})} \times e^{j\pi m},\) with \(g\) being the employed prototype filter impulse response (assumed of unit energy) with length \(L_g\), \(j = \sqrt{-1},\) and \(\varphi_{m,n} = (m+n)\frac{\pi}{2}+mn\pi\). Moreover, usually \(L_g = KM\), with \(K\) being the overlapping factor. The corresponding output of a channel with impulse response \(h\) of length \(L_h\) will be

\[ y(l) = \sum_{k=0}^{L_h-1} h(k)s(l-k) + w(l), \]

with \(w(l)\) assumed to be zero mean Gaussian noise with variance \(\sigma^2\). For pulses \(g\) that are well localized in both time and frequency, the interference from frequency-time (FT) points outside a neighborhood \(\Omega_{p,q}\) of \((p, q)\) is negligible. If, moreover, the time/frequency dispersion of the channel is low enough that its response can be viewed as almost constant over this neighborhood, one can write the analysis filter bank (AFB) output at the FT point \((p, q)\) as

\[ y_{p,q} \approx H_{p,q}c_{p,q} + \eta_{p,q}, \]

where \(H_{p,q}\) is the channel frequency response (CFR),

\[ c_{p,q} = d_{p,q} + jn_{p,q} \]

is the virtual transmitted symbol at \((p, q)\), with \(jn_{p,q}\) being the interference from the neighboring FT points, and \(\eta_{p,q}\) is the corresponding noise component. When known pilots are transmitted at that FT point and its neighborhood \(\Omega_{p,q}\), the quantity in (4) can be approximated and serve as a pseudo-pilot to compute an estimate of the CFR at the corresponding FT point, just like in OFDM. This observation underlies the so called Interference Approximation Methods (IAM) [7]. A review of these and other preamble-based channel estimation methods that rely on the model (3) is given in [7]. Analogous methods have also been proposed for multiple antenna (MIMO) systems (see [7] for a complete review). MIMO-FBMC/OQAM is more challenging as it entails the need to address both filter bank- and antenna-induced interferences [10]. This paper will only consider the single antenna case.

When scattered pilots are employed, as it is the case for estimating and tracking time varying channels, the neighbors of \((p, q)\) carry unknown (data) symbols, and hence one cannot (directly) approximate the interference in (4). However, by properly choosing one of the neighboring symbols, say at the point \((r, t),\) this interference can be forced to zero. The symbol at \((r, t)\) is then known as help or auxiliary pilot [2]. This is the idea that lies behind most of the scattered pilots based methods. Others cancel the interference by applying an appropriate transmission pattern around the pilot and/or transmitting zeros at the FT points that contribute most to the interference. Iterative joint estimation/detection schemes that overcome the need for a help pilot have been also reported. In all of these methods (see [7] for a list of references), (3) is assumed to be valid. To the best of the author’s knowledge, the more realistic scenario of a highly time/frequency selective channel has been only considered in [11], where a “help” pilot is computed on the basis of the channel FT correlation function. A Kalman filter-based pilot-assisted approach for time-varying (yet of low frequency selectivity) channels was recently presented in [12].

3. PREAMBLE-BASED CHANNEL ESTIMATION

3.1. State of the art

For preamble-based channel estimation, the progress beyond the state of the art recorded in [7] has been more significant. Extensions to relay-based networks recently appeared [13]. Frequency smoothing (or averaging) was considered in [14, 15] as a means of improving upon IAM. The frequency selectivity of the subchannels in an FBMC/OQAM system was explicitly taken into account in [16], where the estimation of their impulse responses was addressed via a structured approach, i.e., using a formulation that explicitly separates their unknown (physical) and known (filter bank-based) parts. An iterative procedure for optimizing the corresponding training sequences was proposed in [17]. In [18], a more spectrally efficient training scheme was developed and tested, where only a subset of the (active) subcarriers are employed as pilot tones, with the rest carrying information symbols. In the vein of joint estimation/detection, the subchannel response estimates are computed with the aid of an iterative (EM-based) algorithm. Subchannel estimates were also computed in [19] based on frequency sampling. The channel impulse response (CIR) itself was first estimated in [20]. The main idea was to model the subchannel CIRs as Taylor polynomials (see also [8]) whose order reflects the degree of their frequency selectivity. Together with the adoption of a path-delay model for the channel, this lead to an estimation procedure that treats the problem as one of array processing. A more general, model-independent formulation of the CIR estimation problem was recently developed in [21, 22]. Optimal preamble design in this context was addressed in [21, 23]. A glimpse of this work is given below, and some new results on optimal comb-type preamble design are reported.

3.2. Some new results

As it is common in preamble-based channel estimation, the channel is assumed time invariant over the training sequence
duration. Thus, in order to be able to cope with fast fading environments, the latter should be as small as possible. In contrast to what is assumed in other works ([19, 16, 17, 18]), the preamble here is constructed so as to consist of only one pilot FBMC symbol, followed by one (or more!) symbol(s) of all zeros. This is to protect the pilots from being interfered by the unknown data (or control) samples of the current frame. Then, $d_{m,n}$ is nonzero only for $n = 0$ and the AFB output is considered at time $q = 0$. One can then write [21, 22]

$$y = \Gamma h + \eta,$$

(5)

with $y = \left[y_{m,0}\right]_{m=0}^{M-1}$ and similarly for $\eta$, while $(\Gamma)_{p+1,k+1}$ can be seen to be the response of the transmultiplexer to this input for a channel equal to a $k$-samples delay and can be easily computed. The noise $\eta$ is known to be zero mean Gaussian with the same variance as $w$, while its covariance matrix is known to be given by $C_\eta = \sigma^2 B$, with $B$ being a priori computable based on the knowledge of $g$ [21]. The Gauss-Markov estimate of $h$ is then given by

$$\hat{h} = (\Gamma^H B^{-1} \Gamma)^{-1} \Gamma^H B^{-1} y$$

(6)

with mean squared error (MSE) equal to $\sigma^2 \text{tr}\{(\Gamma^H B^{-1} \Gamma)^{-1}\}$. The problem of choosing the pilot symbol $d = [d_{m,0}]_{m=0}^{M-1}$ so that this MSE is minimized with a constraint on the transmit energy at the SFB output, namely $d^H B d \leq \mathcal{E}$, was addressed and solved in [21, 23]. Write $\Gamma$ in the form

$$\Gamma = \mathcal{G} D,$$

(7)

where $D = I_{L_h} \otimes d$. One can then show that

$$\mathcal{G}(kM + 1 : (k+1)M) = W^k G_k, \quad k = 0, 1, 2, \ldots, L_h - 1$$

where $W = \text{diag}\{(w_{kM}^{0})_{m=0}^{M-1}\}$, $w_M = e^{-j\pi/2}$, and $G_k$ is a Hermitian (almost) tridiagonal circulant matrix, which is DFT-diagonalizable with (real) eigenvalues $\{\lambda_k\}_{k=1}^{M-1}$. In fact, $G_0 = B$. Then a (generally complex valued) optimal $d$ can be built as a scaled column of the DFT matrix, whose index depends on the eigenvalues of the $G_k$ matrices [23]. Constraining $d$ to be real-valued can be seen to correspond to all equal $d_{m,0}$’s or equal with alternating signs. The former choice turns out to be the best: $d = \pm \sqrt{\mathcal{E}/M} 1_M$ [23]. Such a choice leads to the following simple computational procedure for implementing (6):

1) Take the first $L_h$ terms$^2$ of IDFT($y$)

2) Divide by $\sqrt{\mathcal{E}} \text{diag}(\lambda_{0}, \lambda_{1}, \ldots, \lambda_{L_h-1})$

These results concern the block-type (full) preamble. In a comb-type (sparse) preamble, only a set of isolated (surrounded by nulls) subcarriers carry pilots. For economy and to simplify the presentation, it will be assumed that there are no more than $L_h$ pilot tones, indexed as $\mathcal{P} = \{p_1, p_2, \ldots, p_{L_h}\} \subseteq \{0, 1, \ldots, M - 1\}$. Moreover, $L_h$ is assumed to divide $M$, and $M/L_h \geq 2$. The preamble design problem then consists of (a) finding the right places for the pilot tones, and (b) choosing the values for the pilot symbols. One can see that regardless of the choice of $\mathcal{P}$, the corresponding submatrix of $G_k$ equals $\alpha_k I_{L_h}$, where

$$\alpha_k = \sum_{l=k}^{L_h-1} g(l-k)g(l), \quad \text{for } k = 0, 1, \ldots, L_h - 1$$

(8)

is the lag-$k$ autocorrelation of $g$ (decreasing with $k$), with $\alpha_0 = 1$. The optimal solution then turns out to be given by equipowered and equispaced pilot tones, with

$$|d_{p_i,0}| = \sqrt{\mathcal{E}/L_h}, \quad i = 1, 2, \ldots, L_h$$

(9)

and $p_i = p_0 + (i-1)(M/L_h)$, $i = 1, 2, \ldots, L_h$

(10)

with $p_0$ freely chosen. The resulting MSE is given by

$$\text{MSE} = \sigma^2 \sum_{k=0}^{L_h-1} \frac{1}{\alpha_k^2} \mathcal{E} = \frac{\mathcal{E}/L_h}{\text{SNR}_{\text{subc}}},$$

(11)

where $\text{SNR}_{\text{subc}} = \mathcal{E}/L_h$ denotes the per-subcarrier SNR. It is of interest to note that this generalizes previous corresponding results, valid for channels of relatively (to $M$) low dispersion, where $\alpha_k \approx 1$ for all $k$ [7].

With this preamble, the estimator is greatly simplified, reducing to a simple procedure which resembles the one usually employed in OFDM. Let $p_0 = 0$ for the sake of simplicity and without loss of generality. Define $H_\alpha$ as the $L_h$-point DFT of $h$ weighted (or windowed) by $\alpha = [\alpha_0 \alpha_1 \ldots \alpha_{L_h-1}]$. Then one can show that the estimator can be implemented as follows:

1) Compute the “windowed” CFR vector first: $(\hat{H}_\alpha)_i = y_{p_i,0}/d_{p_i,0}$, $i = 1, 2, \ldots, L_h$

2) Compute the “windowed” impulse response via IFFT and divide by the weights $\alpha_k$’s to arrive at the CIR estimate: $\hat{h} = \text{IDFT}(\hat{H}_\alpha) \odot \alpha$

An example of the performance of the above estimator is shown in Fig. 1, where the normalized MSE (NMSE) is plotted versus the SNR, for a realistic scenario involving non-negligible interference from the data. The results of using one, two, and three guard (null) FBMC symbols between the pilot tones and the payload are depicted. Note how the proposed estimator manages to mitigate the well known error floor effect.

$^1$One guard FBMC symbol is not always sufficient and non-negligible interference may still exist. Efficiently addressing the problem of the data interfering with the preamble is a crucial question that can be considered to be still open. Important related contributions include the memory preloading technique for transmitting/receiving a periodic preamble [6] and channel estimation methods that are based on some iterative joint estimation/detection procedure for handling the unknown interference (cf. [7] for a review and some simulation results). The problem of how to truncate the tails (due to a long $g$) of the transmitted burst is also relevant in this context [6].

$^2$Observe that IDFT($y$) may also be directly found from the (polyphase) AFB structure at no extra cost.

$^3$All $M$ subcarriers are considered as candidate pilot tones. The question of optimally placing pilots in the presence of inactive subcarriers (null edges) is still open!
Fig. 1. Comparison of estimation performances for Veh-A channels of length $L_h = 11$ using $N_p = 16$ pilot tones. Filter banks designed as in [24] with $M = 64$ and $K = 3$ were employed.

4. REFERENCES


