A Simplified Scattered Pilot for FBMC/OQAM in Highly Frequency Selective Channels

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***Abstract*—Filterbank multi-carrier with offset quadrature amplitude modulation (FBMC/OQAM) avoids the usage of the cyclic prefix (CP) by using more localized prototype filters. However, channel estimation in highly frequency selective channels is very challenging, especially for the scattered pilots due to the intrinsic interference leaked from the neighboring symbols. This paper presents a novel pilot design called Composite Pilot Pair (CPP), which exploits the localization property of the filters to enable a simple estimation procedure at the receiver. The proposed CPP design and corresponding receiving algorithm have been evaluated on an FBMC-based, Long Term Evolution (LTE) like mobile system. The evaluation shows that the proposed algorithm outperforms the conventional auxiliary pilot algorithms substantially, both in terms of power overhead and pilot mean square error (MSE), whereas the overall receive complexity is low. Moreover, the proposed method respects the frame design principles for the mobile downlink, and is applicable to both common and dedicated pilot symbols in the next generation mobile systems.**

I. Introduction

Filterbank multi-carrier (FBMC) transmission scheme [1] is first proposed in 1960s, approximately at the same time as cyclic prefix orthogonal frequency division multiplexing (CP-OFDM). The most important difference between FBMC and CP-OFDM is that FBMC uses waveforms which are better localized in the frequency domain. Recently, the FBMC related modulation schemes were put in the research focus because this waveform difference could possibly be translated into several advantages for the next generation mobile wireless systems [2], [3], such as synchronization robustness and lower out-of-band interference leakage.

In this paper, FBMC with offset quadrature amplitude modulation (OQAM) modulation is studied. Current research works have revealed that, by using frequency localized prototype filters, such as the PHYDYAS filters [4], [5], with well designed linear multi-tap (pre-)equalizers and decoders, the bit error rate (BER) and block error rate (BLER) performances under highly frequency selective channels are better for FBMC compared with CP-OFDM [6], [7], [8], [9]. Meanwhile, FBMC is attaining a higher spectral efficiency gain by removing the CP. However, these works are based on perfect channel knowledge assumptions for the receivers and transmitters, which are in reality very challenging to achieve.

The scattered pilot based channel estimation becomes difficult because of the real field orthogonality for FBMC. This orthogonality cannot be established before the channel knowledge is acquired at the receiver. To solve the problem, [10] proposed an auxiliary pilot (AUP) scheme, where the intrinsic image interference can be perfectly pre-canceled under the assumption of the flat channel in the pilot neighborhood.

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The highly frequency selective channel case is, however, more challenging. There, the inter-carrier interference (ICI) and inter-symbol interference (ISI) to the pilot are present after the analysis filterbank (AFB) at the receiver due to a strong channel distortion [11]. A scrambling based AUP algorithm is proposed in [12] to solve the power overhead problems for certain prototype filters at certain AUP positions. Iterative channel estimation algorithms based on successive interference cancellation (SIC) were proposed in [13], but they require joint channel estimation and symbol detection (even decoding) of all surrounding payload symbols. This is not only of significant complexity, but also difficult to apply to practical systems, where scattered common reference symbols (known as CRS in LTE) are used not only for demodulation purpose to the assigned user in the resource block, but also for common usage such as synchronization and channel quality indicator (CQI) measurements.

The method in this paper attempts to resolve the above-explained issues in the AUP and SIC based algorithms for FBMC/OQAM channel estimation. The main contributions can be summarized as follows

- The detection for interference cancellation is based on a new composite pilot pair (CPP) scheme, which scrambles the AUP over the two most critical interfering positions.
- The proposed pilot interference cancellation is not optimal but very effective, as it is realized by a simple SIC receiver based on a small neighborhood of pilot position.
- The interference cancellation is performed using a more accurate channel model compared to the reference works, which boosts the performance.

The method works similarly to the standard AUP algorithm in the low SNR regime, where noise is the limiting factor for channel estimation, and outperforms AUP algorithm in higher SNR regime, where the distorted intrinsic interference is the limiting factor.

Regarding the structure of this paper, Section II introduces the system model, while Section III explains the proposed pilot

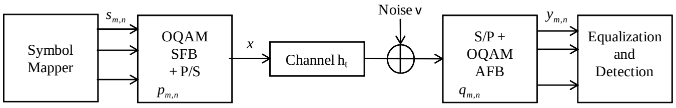

Fig. 1. A block diagram of FBMC/OQAM system

TABLE I. $A_{m,n}[m', n']$ RESPONSE FOR PHYDYAS FILTER, CASE n EVEN

$[m', n']$	$n-2$	$n-1$	n	$n+1$	$n+2$
$m-1$	$0.1250j$	$0.2058j$	$0.2393j$	$0.2058j$	$0.1250j$
m	0.0002	$0.5644j$	1	$-0.5644j$	0.0002
$m+1$	$-0.1250j$	$0.2058j$	$-0.2393j$	$0.2058j$	$-0.1250j$

structure CPP and the corresponding estimation algorithms. Section IV presents the performance evaluation, and Section V concludes the paper. The following mathematical notations are used: The sets of integer and real numbers are denoted as $\mathbb{Z}$ and $\mathbb{R}$, respectively. $\Re(c)$ returns the real part of complex number c, $(\cdot)^*$ denotes complex conjugation, $(\star)$ denotes discrete convolution, $\lfloor . \rfloor_{\downarrow N}$ denotes signal decimation by factor N, $j = \sqrt{-1}$, and $\delta_{m,n}$ is the discrete Dirac function.

II. System Model

For an FBMC/OQAM system, the real pulse amplitude modulated (PAM) or OQAM symbol is modulated on the m-th subcarrier of the n-th symbol (denote as $s_{m,n}$). Assuming that the Fast Fourier Transform (FFT) size is M with the subcarrier spacing $\frac{1}{T}$, the discrete transmit signal $x[k]$ is written as

$$x[k] = \sum_{n\in\mathbb{Z}} \sum_{m=1}^{M} s_{m,n} \underbrace{p(k - \frac{nT}{2}) e^{j\frac{2\pi m}{M}(k-D)} d_{m,n}}_{\triangleq p_{m,n}[k]} \quad (1)$$

where $p[k]$ is the prototype filter in discrete form defined in $(k = 0, ..., L-1)$ with filter length $L = KM$, K is the overlapping factor, $D = \frac{L-1}{2}$ is the causal delay factor, and $d_{m,n} = j^{m+n}$ is the phase term. We denote the equivalent transmit filter bank as $p_{m,n}[k]$.

At the receiver, the matched filter bank with $q_{m,n}[k] = p^*_{m,n}[k]$ is adopted. Therefore, the demodulated symbol after receiving filter bank at position (m, n) can be derived as:

$$\begin{aligned} y_{m,n} &= \left\lfloor \sum_k x[k] \sum_{m,n} q_{m,n}[k] \right\rfloor_{\downarrow \frac{M}{2}} \\ &= s_{m,n} A_{m,n}[m, n] + \sum_{(m',n')\in\Omega^0_{m,n}} s_{m',n'} A_{m,n}[m', n'] \\ &= s_{m,n} A_{m,n}[m, n] + \quad y^{(i)}_{m,n} \end{aligned} \quad (2)$$

where set $\Omega^0_{m,n} \triangleq \{(m', n') | (m', n') \neq (m, n)\}$ is defined as the local neighborhood of the symbol (m, n) due to the localization property of filter $p[k]$ and $y^{(i)}_{m,n}$ is pure imaginary, since the down-sampled ambiguity function $A_{m,n}[m', n']$ has the following property for the perfect reconstruction (PR) filter [11]:

$$\begin{aligned} A_{m,n}[m', n'] &\triangleq \lfloor A_{m,n,m',n'}[k] \rfloor_{\downarrow \frac{M}{2}} \\ &= \begin{cases} 1 & \text{for } (m,n) \\ \text{pure imaginary} & \text{otherwise} \end{cases} \end{aligned} \quad (3)$$

where

$$A_{m,n,m',n'}[k] \triangleq \sum_k p_{m',n'}[k] q_{m,n}[k] \quad (4)$$

is the ambiguity function for the transmultiplexer (TMUX) filter banks (cf. Table I).

In this paper, we assume the prototype filter $p[k]$ as the PHYDYAS project filter ($K = 4$) [4], [5], which is non PR (NPR) in the sense that (3) holds approximately.

A. Prior Arts for FBMC/OQAM Channel Estimation

If we assume the channel impulse response $h_t[l]$ with $(l = 1, ..., L_{ch})$ in the time domain, the receiving signal after the matched filter bank can be derived as

$$\begin{aligned} y'_{m,n} &= \left\lfloor \left(\sum_k (x[k] \star h_t[k]) \sum_{m,n} q_{m,n}[k] \right) \right\rfloor_{\downarrow \frac{M}{2}} + w_{m,n} \\ &= \sum_{m,n} s_{m,n} \lfloor A_{m,n,m',n'}[k] \star h_t[k] \rfloor_{\downarrow \frac{M}{2}} + w_{m,n} \end{aligned} \quad (5)$$

where $w_{m,n}$ is the additive white Gaussian noise v after AFB.

In the case of AWGN or short delay spread channel ($L_{ch} \ll L$), the convolution with a well localized filter can be simplified as scalar multiplication given as:

$$\lfloor A_{m,n,m',n'}[k] \star h_t[k] \rfloor_{\downarrow \frac{M}{2}} \approx A_{m,n}[m', n'] \cdot h_{m,n} \quad (6)$$

Therefore, in this case, the model (5) renders

$$\begin{aligned} y'_{m,n} &\approx h_{m,n} s_{m,n} + \\ &\quad \sum_{\Omega^0_{m,n}} h_{m,n} s_{m',n'} A_{m,n}[m', n'] + w_{m,n} \\ &= h_{m,n} \left(s_{m,n} + y^i_{m,n} \right) + w_{m,n} \end{aligned} \quad (7)$$

Assume that the scattered main pilot (MP) $P_{m,n}$ is allocated on the position of (m, n). The receive pilot symbol $\tilde{P}_{m,n}$ is written as

$$\tilde{P}_{m,n} = h_{m,n} \left(P_{m,n} + P^i_{m,n} \right) + w_{m,n} \quad (8)$$

Javaudin et al.'s [10] method is to insert an AUP $U_{m,n+1}$, usually at the nearby position of the main pilot $\Omega^0_{m,n}$, to manipulate the intrinsic response at the main pilot position so that $P^i_{m,n} = 0$, in other words,

$$U_{m,n+1} = -\frac{\sum_{(m',n')\in\Omega^0_{m,n}\backslash\{(m,n+1)\}} s_{m',n'} A_{m,n}[m', n']}{A_{m,n}[m, n+1]} \quad (9)$$

so that at the receiver, the least square estimator (LSE) is depicted as:

$$\hat{h}_{m,n} = \frac{\tilde{P}_{m,n}}{P_{m,n}} \quad (10)$$

For highly selective channel case, the simple model of (6) cannot hold as the receive signal is strongly distorted. Instead of using auxiliary pilots to precancel the interference, [13]

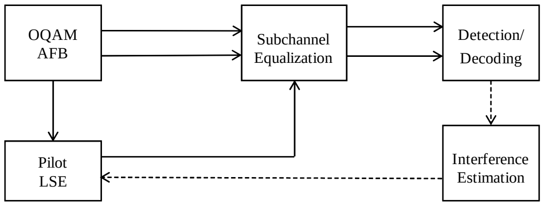

Fig. 2. Lélé et al. SIC based channel estimation [13]

sustains the model of (6) and proposes an iterative channel estimation and detection algorithm to predict the pilot symbol response. Simply put, the $i+1$-th channel LSE iteration is based on the i-th symbol detection:

$$\hat{h}_{m,n}^{(i+1)} = \frac{\tilde{P}_{m,n}}{P_{m,n} + \sum_{\Omega_{m,n}^0} s_{m',n'}^{(i)} A_{m,n}[m',n']} \tag{11}$$

However, the proposed algorithm has the following limitations:

- Without the support of image interference pre-cancellation, the receiving algorithm must always be conducted in iterative manner, even in the low SNR regime
- The algorithm usually needs low iteration number (such as three) to reach a satisfying performance. However, as it is based on the inaccurate channel model of (7), it is very slowly converging to the bound if channel is highly selective.
- In practical systems such as the LTE downlink, when CRS is embedded in a precoded block, it is difficult for the other users to detect the payload symbols.

B. Equalizer and Detector

For a 1-tap ZF equalizer, it is apparent that the demodulation process for the n-th symbol on the m-th subcarrier can be given as

$$\Re\left(\frac{h_{m,n}^* y'_{m,n}}{|h_{m,n}|^2}\right) = s_{m,n} + \Re\left(\frac{h_{m,n}^* w_{m,n}}{|h_{m,n}|^2}\right) \tag{12}$$

In long delay spread channels, a multi-tap equalizer needs to be adopted to mitigate the ISI and ICI interference due to the strong distortion by the channel, which will not be discussed in this paper.

III. Proposed Schemes and Estimation Algorithm

In this paper, we propose a pilot design and channel estimation approach based on the following ideas:

- Instead of using a single axillary pilot defined by (9), we propose a CPP to capture the major part of the channel-induced distortion.
- Instead of an iterative channel estimation based on all the neighboring symbols, the proposed algorithm bases the iterations on only two symbols from the local neighborhood.
- Instead of using the simplified model (6), we use the original complex model (5). We remark that when taking channel selectivity into account, the symmetry of the ambiguity function (4) does not result in a symmetric influence of the interference from the symbols in the pilot neighborhood. The complexity of channel convolution can be reduced by the pruned FFT.

A. Composite Pilot Pair

For frequency localized filters such as PHYDYAS, it is important to note that, the most significant intrinsic interference to the pilot $P_{m,n}$ due to channel distortion under long delay spread channels, originates from the neighboring pair of symbols in time domain. In other words, the critical positions for the interferences are

$$(m, n-1) \qquad \text{and} \qquad (m, n+1).$$

For this reason, we devise the CPP together with the pilot $P_{m,n}$ in the following way

$$\begin{array}{lll} \text{for} & (m,n): & P_{m,n} \\ & (m,n-1): & C_{m,n}^1 \\ & (m,n+1): & C_{m,n}^2 \\ & \text{other } (m',n') \in \Omega_{m,n}^0: & s_{m',n'} \end{array} \tag{13}$$

where $C_{m,n}^1$ and $C_{m,n}^2$ is defined by:

$$\begin{aligned} C_{m,n}^1 &= \frac{1}{\sqrt{2}} s_{m,n}^* \gamma - \frac{1}{2} \sum_{\Omega_{m,n}^0 \setminus \{m,n\pm1\}} \frac{s_{m',n'} A_{m,n}[m',n']}{A_{m,n}[m,n-1]} \\ C_{m,n}^2 &= -\frac{1}{\sqrt{2}} s_{m,n}^* \gamma - \frac{1}{2} \sum_{\Omega_{m,n}^0 \setminus \{m,n\pm1\}} \frac{s_{m',n'} A_{m,n}[m',n']}{A_{m,n}[m,n+1]} \end{aligned} \tag{14}$$

where γ is the scalar factor for adjusting the payload symbol power in the CPP algorithm. For a fair comparison with the AUP scheme [10], we keep it at the level to make the effective power constant for both schemes. A detailed explanation is given in the power overhead subsection. We remark that scrambling of pilots by (14) would in the idealized case precancel almost all (depending on the size of the considered neighborhood) interference to the pilot, similarly to [10].

For highly selective channels, there will be residual interference left, which is mostly leaked from the CPP symbols. We remove them by devising an SIC based iterative algorithm.

The proposed iterative procedure can be described as follows (the iteration step number is denoted by (i)):

1) Channel estimation $\tilde{h}_{m,n}^{(i)}$ based on $\tilde{P}_{m,n}^{(i)}$ (This is also the initialization step.).
2) Detect $s_{m,n}^{*(i)}$ by combining $C_{m,n}^1$ and $C_{m,n}^2$ based on (14):
$$s_{m,n}^* = \frac{\sqrt{2}}{\gamma}(C_{m,n}^1 - C_{m,n}^2) \tag{15}$$
3) Calculate $\tilde{A}_{m,n,m,n\pm1}^{(i)}$ by:
$$\left\lfloor A_{m,n,m',n'}[k] \star h_t[k]^{(i)} \right\rfloor_{\downarrow \frac{M}{2}} \tag{16}$$

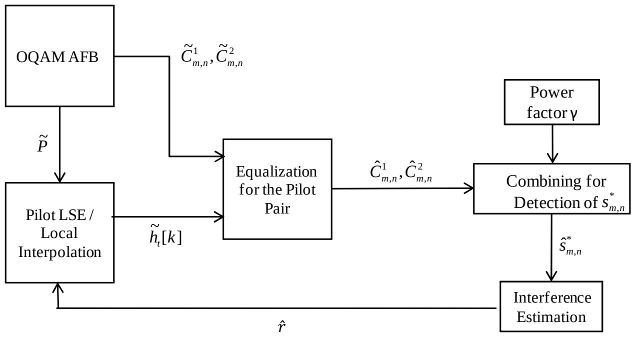

Fig. 3. Proposed receiving procedure

TABLE II. POWER OVERHEAD FOR AUP AND CPP

Algorithm	Compared to normal PAM payload
AUP	3.3 dB
CPP	-2.5 dB

4) Update the intrinsic interference by

$$\tilde{P}^{(i+1)}_{m,n} = \tilde{P}^{(i)}_{m,n} - \frac{s^{*(i)}_{m,n}}{2}[\tilde{A}^{(i)}_{m,n,m,n+1} + \tilde{A}^{(i)}_{m,n,m,n-1}] \quad (17)$$

5) Repeat 1) until convergence or the maximum iteration number is reached.

B. Power Overhead

All the methods (including AUP [10], [12] and the proposed CPP) that are pre-canceling the imaginary interference will introduce a power overhead. The level of power overhead is dependent on the position of inserting the pre-canceling code and how scrambling is done. As the proposed CPP scrambles between the most influential neighboring symbols in orthogonal manner, the power overhead compared to AUP algorithm is substantially lower. The comparison is given in Table II.

To make a fair comparison, we define the excessive power boosting level as the power of the scattered pilots to the power of complex symbol payload. For FBMC, the power of scattered pilots are counted as the sum of the power of main pilot and the power overhead. For the case of same power boosting level 5dB, due to the substantially lower of power overhead for the proposed algorithm, ca. 1.4 dB more stronger is the main pilot power than in the AUP case.

Another major concern for the conventional AUP algorithm is that, in some positions, overloaded AUP power will decrease the peak-to-average-power-ratio (PAPR) performance, we also made PAPR performance evaluation for the proposed CPP algorithm. Since the excessive power of CPP is much lower than AUP, the PAPR is proved to be not an issue. A detailed illustration is given in Figure 5.

C. Convergence and Complexity

The proposed CPP method and the corresponding receiving algorithm base the interference cancellation on the substraction of the influence from the two neighboring symbols. This yields clearly a sub-optimal channel estimation. However, the convergence is very fast, and typically two loops can already lead to a very good performance. This is due to

- the accurate channel estimation model based on (5),
- the combining detection for symbol $s^*_{m,n}$.

Compared to the reference iterative algorithm [13], the complexity of proposed receiving algorithm for each loop is much lower since only symbol $s^*_{m,n}$ is hard detected and its interference is estimated. The major complexity is thus coming from the convolution with the ambiguity function and downsampling, which can be reduced by the pruned FFT.

TABLE III. 3GPP ETU CHANNEL PARAMETER [14]

Delays	[0, 0.05, 0.12, 0.2, 0.23, 0.5, 1.6, 2.3, 5] μ s
Gains	[-1, -1, -1, 0, 0, 0, -3, -5, -7] dB

TABLE IV. SIR OF MAIN PILOT SYMBOL

SIR of $P_{m,n}$	Javaudin et al's AUP	Proposed CPP
AWGN	24.2dB	25.5dB
ETU	19.2dB	23.7dB

IV. PERFORMANCE EVALUATION

A. Simulation Setup

To evaluate the performance, a frequency division duplex (FDD) LTE-like FBMC downlink simulation platform is established, with 5 MHz bandwidth, FFT size $M = 512$, and the subcarrier spacing of 15KHz. The position of pilots are analogous to CRS definition in LTE resource grid (with ca. 10 percentage overhead) and we assume a constant 5 dB (excess) power boosting for the pilot symbols [14]. At the receiver, 1-tap zero-forcing (ZF) equalizer and detector for symbol is used, as described in (12). Linear interpolation is used for the channel interpolation in both frequency and time domain. For the CPP algorithm, the maximum number of iterations for the intrinsic interference update is limited to two.

The comparison with the related work is done in a fair way, using the same frame structure (bitrate, pilot neighborhood, etc.). The parameters of the considered highly frequency selective channel with zero Doppler frequency are given in Table III.

B. Simulation Performance

From Table IV, we see that in AWGN case, the SIR of proposed CPP algorithm is 1.4 dB better than the conventional AUP due to the power overhead reduction from CPP. In frequency selective ETU channel, the SIR gain of CPP is amplified to over 4 dB due to the interference cancellation. Similar performance can be seen in Fig. 4, where channel estimation MSE is substantially lower in high SNR regime.

PAPR performance in Fig. 5 has shown that the proposed CPP will not lead to higher PAPR, since the power overhead is lower than AUP. Meanwhile, the uncoded bit-error performance in Fig. 6 has shown that the proposed CPP is very close to the perfect channel estimation performance.

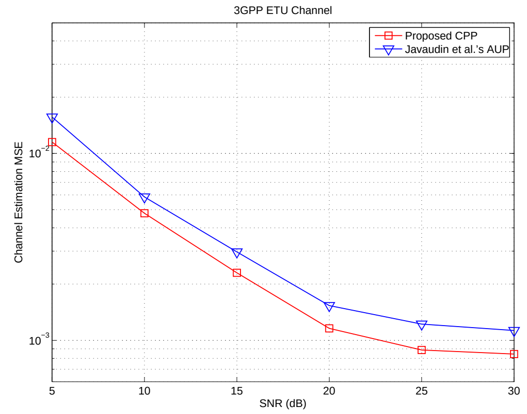

Fig. 4. Channel Estimation MSE performance

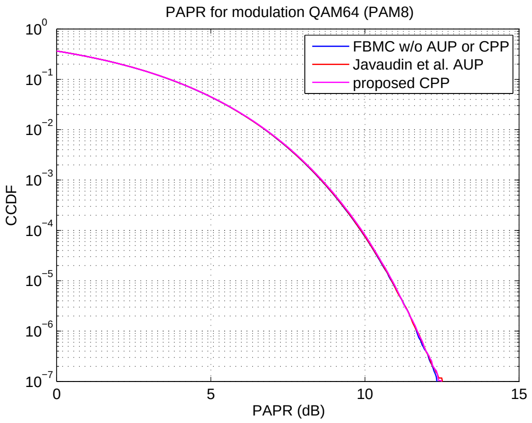

Fig. 5. PAPR performance for transmitted signal

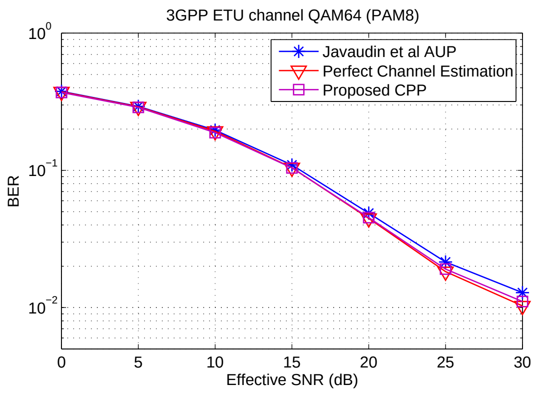

Fig. 6. Uncoded BER for ETU channel

V. Conclusion

The paper proposes a new CPP pilot design and channel estimation algorithm for FBMC/OQAM. The CPP exploits the localization property of the prototype filter and avoids the inefficient iterative estimation due to an inaccurate channel model. The performance results show that the proposed CPP method outperforms the conventional AUP method in highly frequency selective channels and approaches the perfect channel information performance. Moreover, the proposed algorithm has low complexity and does not involve extensive estimation with the local symbols, which makes it directly applicable to mobile systems like LTE and compatible to other scattered pilots in the system.

Future work will be to apply the proposed CPP method to facilitate the channel estimation for FBMC-MIMO transmission [9]. We will also consider even more selective channels such as the SFN scenario channels [8], where the interference due to channel distortion is coming not only from the neighboring symbols, but also from the neighboring subcarriers.

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