IMPROVED INTERFERENCE APPROXIMATION METHOD FOR PREAMBLE-BASED CHANNEL ESTIMATION IN FBMC/OQAM

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ABSTRACT

Filter bank-based multicarrier modulation (FBMC) using offset quadrature amplitude modulation (OQAM), known as FBMC/OQAM (or OFDM/OQAM), provides an attractive alternative to the conventional cyclic prefix-based orthogonal frequency division multiplexing (CP-OFDM), especially in terms of increased robustness to frequency offset and Doppler spread, and high bandwidth efficiency. It suffers, however, from an inherent (intrinsic) imaginary inter-carrier/inter-symbol interference that complicates signal processing tasks such as channel estimation (CE). Recently, the so-called interference approximation method (IAM) was proposed for preamble-based CE. It relies on the knowledge of the pilot's neighborhood to approximate this interference and constructively exploit it in simplifying CE and improving its performance. The IAM preamble with nulls at the neighboring time instants that results in optimum CE performance was recently reported. This paper investigates an extended version of it, which can provide significant improvement through an appropriate exploitation of the interfering symbols from neighboring time instants as well. Simulation results are presented that compare the methods, with each other and with CP-OFDM, in both mildly and highly frequency-selective channels.

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) has become quite popular in both wired and wireless communications [6, 2], mainly because of its immunity to multipath fading, which allows a significant increase in the transmission rate [19]. Using the cyclic prefix (CP) as a guard interval, OFDM manages to turn a frequency selective channel into a set of parallel flat channels with independent noises. This greatly simplifies both the estimation of the channel and the recovery of the transmitted data at the receiver. However, these advantages come at the cost of an increased sensitivity to frequency offset and Doppler spread. This is due to the fact that, although the subcarrier functions are perfectly localized in time, they suffer from spectral leakage in the frequency domain and hence inter-carrier interference (ICI) results. Moreover, the inclusion of the CP entails a waste in transmitted power as well as in spectral efficiency, which, in practical systems, can go up to 25% [2].

Filter bank-based multicarrier modulation (FBMC) using offset quadrature amplitude modulation (OQAM), known as FBMC/OQAM or OFDM/OQAM [13], provides an alternative to CP-OFDM, that can mitigate these drawbacks. FBMC/OQAM employs pulse shaping via an IFFT/FFT-based efficient filter bank, and staggered OQAM symbols, i.e., real symbols at twice the symbol rate of OFDM/QAM, are loaded on the subcarriers [18]. This allows for the pulses to be well localized in *both* the time and the frequency domains while still keeping maximum spectral efficiency. As a consequence, the system's robustness to frequency offsets and Doppler effects is increased and at the same time an enhanced spectral containment, for bandwidth sensitive applications (e.g., cognitive radio [1]), is offered. Moreover, FBMC/OQAM does not require the inclusion of a CP, which may lead to even higher transmission rates [18].

However, the subcarrier functions are now only orthogonal in the *real* field, which means that there is always an *intrinsic* imaginary interference among (neighboring) subcarriers and symbols [11]. As a consequence, the simplicity of the channel estimation (CE) in CP-OFDM is lost in FBMC/OQAM systems, as the real-valued pilots are also "polluted" by this imaginary-valued contribution of their neighbors. A number of training schemes and associated CE methods that take this into account have recently appeared in the literature, including both preamble-based (e.g., [15, 17]) and scattered pilots-based (e.g., [11, 16, 3]) ones. These can be roughly characterized as aiming at cancelling the undesired interference or constructively exploiting it to improve the estimation performance. The latter approach relies on the fact that the interference to a pilot symbol is mainly contributed to by its first-order neighbors, and hence it can be approximated if these neighbors also carry known symbols, a not uncommon situation within a preamble. This has been known as the Interference Approximation Method (IAM). The idea is to combine the real-valued pilot with the interference approximation into a complex-valued *pseudo-pilot* and performing CE as in CP-OFDM, under the assumption of a locally constant channel frequency response. As the pseudo-pilot divides the noise component, its magnitude should be as large as possible. This way, the interference turns into a benefit for the CE task and hence the neighboring pilots should be so chosen as to maximize its effect.

A number of variants of IAM have been proposed, corre-

THIS IS PART OF THE WORK DONE BY THE AUTHORS WITHIN THE FP7 PROJECT PHYDYAS (HTTP://WWW.ICT-PHYDYAS.ORG) AS MEMBERS OF THE PARTNER RA-CTI, GREECE, AND IN COLLABORATION WITH THE DEPT. OF INFORMATICS AND TELECOMMUNICATIONS OF THE UNIVERSITY OF ATHENS, GREECE.

sponding to different choices for the preamble structure [14]. Those that have received most of the attention comprise three FBMC/OQAM symbols (i.e., 1.5 complex OFDM/QAM symbols), with the side ones being all zeros. This is to protect the middle vector of pilots from being interfered by the unknown data parts of the previous and current frames. The middle pilot symbols are then chosen so as to maximize the interference contributions from neighboring subcarriers. Recently, such a preamble structure with *imaginary* pilots was proposed in [8] and indepedently in [10] and shown to be the optimum choice within this kind of preambles. It was called IAM-C to signify the presence of complex (imaginary) pilots, in analogy with IAM-R [15], the optimum real-valued preamble. IAM-C improves upon IAM-I [17], a previous IAM variant also employing imaginary pilots, by maximizing the magnitude of *all* pseudo-pilots instead of only one third of them as in IAM-I.

This paper investigates the possible gains from a preamble that has nonzero pilots at its side FBMC/OQAM symbols as well. A preamble srtucture of this kind is proposed that exploits the symmetry of the interference weights and their relative values to obtain pseudo-pilots of a magnitude greater than in IAM-C. As the new method extends IAM-C to incorporate side symbols as well, it will be called Extended IAM-C (E-IAM-C). The superiority of E-IAM-C over both IAM-R and IAM-C is verified through simulations in both mildly and highly frequency selective channels.

The rest of the paper is organized as follows: In Section 2, the discrete-time baseband equivalent model for the FBMC/OQAM system is described. Section 3 reviews the two main preamble structures that include side zero guard symbols, IAM-R and IAM-C. The proposed preamble is developed and discussed in Section 4. Simulation results are reported in Section 5. Section 6 concludes the paper.

2. SYSTEM MODEL

The (QAM or OQAM based) OFDM modulator output is transmitted through a channel of length L_h, which is assumed to be invariant in the duration of the preamble. At the receiver front-end, noise is added, which is assumed Gaussian with zero mean and variance σ^2. For CP-OFDM, the classical configuration is assumed [19]. The discrete-time signal at the output of an FBMC/OQAM synthesis filter bank (SFB) is given by [18]

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

where $d_{m,n}$ are *real* OQAM symbols, and

$$g_{m,n}(l) = g\left(l - n\frac{M}{2}\right) e^{j\frac{2\pi}{M}m\left(l-\frac{L_g-1}{2}\right)} e^{j\varphi_{m,n}},$$

with g being the *real symmetric* prototype filter impulse response (assumed here of unit energy) of length L_g, M being the *even* number of subcarriers, and $\varphi_{m,n} = \varphi_0 + \frac{\pi}{2}(m+n) \bmod \pi$, where φ_0 can be arbitrarily chosen [18]. In this paper, and without loss of generality, $\varphi_{m,n}$ is defined as $(m+n)\frac{\pi}{2} - mn\pi$ as in [18]. The filter g is usually designed to have length $L_g = KM$, with K being the overlapping factor. The double subscript $(\cdot)_{m,n}$ denotes the (m,n)-th frequency-time (FT) point. Thus, m is the subcarrier index and n the OQAM symbol time index.

The pulse g is designed so that the associated subcarrier functions $g_{m,n}$ are orthogonal in the *real* field [18]. This implies that even in the absence of channel distortion and noise, and with perfect time and frequency synchronization, there will be some intercarrier (and/or intersymbol) interference at the output of the analysis filter bank (AFB), which is purely imaginary (the notation in [15] is adopted):

$$\sum_{l} g_{m,n}(l) g^{\star}_{p,q}(l) = j\langle g\rangle^{p,q}_{m,n}, \tag{2}$$

and known as *intrinsic* interference [11]. Making the common assumption that the channel is (approximately) frequency flat at each subcarrier and constant over the duration of the prototype filter [15], which is true[1] for practical values of L_h and L_g and for well time-localized g's, one can express the AFB output at the pth subcarrier and qth FBMC/OQAM symbol as:

$$y_{p,q} = H_{p,q} d_{p,q} + \underbrace{j \underbrace{\sum_{m=0}^{M-1} \sum_{n}}_{(m,n)\neq(p,q)} H_{m,n} d_{m,n} \langle g\rangle^{p,q}_{m,n}}_{I_{p,q}} + \eta_{p,q} \tag{3}$$

where $H_{p,q}$ is the channel frequency response (CFR) at that FT point, and $I_{p,q}$ and $\eta_{p,q}$ are the associated interference and noise components, respectively. $\eta_{p,q}$ has been shown to be stationary (in fact, also Gaussian with zero mean and variance σ^2) and correlated among adjacent subcarriers (see [7]). This correlation will be henceforth assumed negligible, a valid assumption for well frequency-localized prototype filters.

A common assumption is that, with a well time-frequency localized pulse, contributions to $I_{p,q}$ only come from the first-order neighborhood of (p,q), namely $\Omega_{p,q} = \{(p\pm 1, q\pm 1), (p, q\pm 1), (p\pm 1, q)\}$. If, moreover, the CFR is almost constant over this neighborhood, one can write (3) as

$$y_{p,q} \approx H_{p,q} c_{p,q} + \eta_{p,q} \tag{4}$$

where

$$c_{p,q} = d_{p,q} + j \underbrace{\sum_{(m,n)\in\Omega_{p,q}} d_{m,n} \langle g\rangle^{p,q}_{m,n}}_{u_{p,q}} = d_{p,q} + j u_{p,q} \tag{5}$$

is the *virtual* transmitted symbol at (p,q), with

$$u_{p,q} = \sum_{(m,n)\in\Omega_{p,q}} d_{m,n} \langle g\rangle^{p,q}_{m,n} \tag{6}$$

being the imaginary part of the interference from the neigboring FT points. When known pilots are transmitted at that FT point and its neighborhood $\Omega_{p,q}$, the quantity in (5) can be approximated. This can then serve as a *pseudo pilot* [15] to compute an estimate of the CFR at the corresponding FT point, as, for example,

$$\hat{H}_{p,q} = \frac{y_{p,q}}{c_{p,q}} \approx H_{p,q} + \frac{\eta_{p,q}}{c_{p,q}} \tag{7}$$

[1] This is not an accurate subchannel model in case M, the number of subcarriers, is not large enough with respect to the frequency selectivity of the channel (see Section 5 for such examples). This is also the case in high mobility scenarios, where M should be small, to cope with ICI from frequency dispersion.

This observation underlies the IAM schemes to be desribed in detail in the sequel.

3. IAM PREAMBLES WITH ZERO GUARD SYMBOLS

One can see from (7) that the preamble should be so structured as to result in pseudo-pilots of *maximum magnitude* [15]. Therefore, the training symbols surrounding the pilot $d_{p,q}$ should be such that all terms in (6) have the same sign so that they add together. Moreover, this should happen for *all* frequencies p.

To this end, one needs to know the interference weights $\langle g\rangle_{m,n}^{p,q}$ for the neighbors $(m,n) \in \Omega_{p,q}$ of each FT point (p,q) of interest. These can be *a priori* computed based on the prototype filter g employed. Moreover, it can be shown that, for *any* choice of g, these weights follow a specific pattern, which, for the definition of φ_0 adopted here, and for *all* q, can be written as[2]

$$\begin{array}{ccc} (-1)^p\delta & -\beta & (-1)^p\delta \\ -(-1)^p\gamma & d_{p,q} & (-1)^p\gamma \\ (-1)^p\delta & \beta & (-1)^p\delta \end{array} \tag{8}$$

with the horizontal direction corresponding to time and the vertical one to frequency. The above quantities can be shown to be given by

$$\beta = e^{-j\frac{2\pi}{M}\frac{L_g-1}{2}} \sum_{l=0}^{L_g-1} g^2(l) e^{j\frac{2\pi}{M}l} \tag{9}$$

$$\gamma = \sum_{l=\frac{M}{2}}^{L_g-1} g(l) g\left(l-\frac{M}{2}\right) \tag{10}$$

$$\delta = -je^{-j\frac{2\pi}{M}\frac{L_g-1}{2}} \sum_{l=M/2}^{L_g-1} g(l) g\left(l-\frac{M}{2}\right) e^{j\frac{2\pi}{M}l} \tag{11}$$

and are positive and, clearly, smaller than one. Generally, $\beta,\gamma > \delta$. For example, $\gamma = 0.5644$, $\beta = 0.2393$, and $\delta = 0.2058$ for the g's employed in Section 5. This observation, along with the symmetry of the above pattern, will be seen to underly the development of the IAM method proposed in Section 4.

3.1 IAM-R

In order to simplify the task of generating pseudo-pilots of large magnitude, it was suggested in [15, 14] to place nulls at the first and third FBMC/OQAM symbols of the preamble, that is, $d_{p,0} = d_{p,2} = 0$, $p = 0,1,\ldots,M-1$. Then, the imaginary part of the pseudo-pilot at $(p,1)$ will only come from the symbols at the positions $(p\pm 1,1)$. Obviously, these pilots must be OQAM symbols of maximum modulus, d. Moreover, in view of the pattern (8), they should satisfy the condition $d_{p+1,1} = -d_{p-1,1}$ for all p, in order to yield pseudo-pilots of maximum magnitude, namely $|d_{p,1} + j2\beta d_{p+1,1}| = d\sqrt{1+4\beta^2}$. An example for $M = 8$ and OQPSK modulation is shown in Fig. 1(a). This method is called IAM-R to signify the fact that its pilot symbols are real.

[2] This is without loss of generality as analogous patterns result for other definitions of φ_0.

$$\begin{array}{ccc} 0 & 1 & 0 \\ 0 & -1 & 0 \\ 0 & -1 & 0 \\ 0 & 1 & 0 \\ 0 & 1 & 0 \\ 0 & -1 & 0 \\ 0 & -1 & 0 \\ 0 & 1 & 0 \end{array} \qquad \begin{array}{ccc} 0 & 1 & 0 \\ 0 & -j & 0 \\ 0 & -1 & 0 \\ 0 & j & 0 \\ 0 & 1 & 0 \\ 0 & -j & 0 \\ 0 & -1 & 0 \\ 0 & j & 0 \end{array} \qquad \begin{array}{ccc} j & 1 & -j \\ -1 & -j & 1 \\ -j & -1 & j \\ 1 & j & -1 \\ j & 1 & -j \\ -1 & -j & 1 \\ -j & -1 & j \\ 1 & j & -1 \end{array}$$

(a) (b) (c)

Figure 1: Preamble structures for (a) IAM-R , (b) IAM-C, and (c) E-IAM-C methods. $M = 8$. OQPSK modulation is assumed.

3.2 IAM-C

Nevertheless, it can be verified from (5) that one can do better in maximizing the magnitude of the pseudo-pilots if *imaginary* pilot symbols $jd_{p,1}$ of maximum modulus d are also allowed in the preamble. The reason is that the corresponding pseudo-pilot would then be imaginary and hence of larger magnitude than in IAM-R, namely $|d_{p,1} + u_{p,1}|$, provided that the signs of the pilot symbol and its neighbors are properly chosen so that $d_{p,1}u_{p,1} > 0$. Such an IAM scheme was first proposed in [17], and was called IAM-I to signify the presence of imaginary pilots. The middle preamble vector consists of triplets, with each of them following the above principle, and the pilots in each triplet are otherwise selected in random and independently of the other triplets. Hence, imaginary pseudo-pilots result in only one third of the subcarriers, whereas the rest of them deliver complex pseudo-pilots of smaller magnitude, $d|(1+\beta)+j\beta|$.

With a slight but important modification, one can improve upon this and obtain pseudo-pilots that are either purely real or imaginary at *all* the subcarriers. The idea is to simply set the middle FBMC/OQAM symbol equal to that in IAM-R but with the pilots at the odd subcarriers multiplied by j, as shown in [8] and independently in [10]. *All* of the resulting pseudo-pilots are then of maximum magnitude, $d|1+2\beta|$. Again, for the $M = 8$ OQPSK example, this will result in the preamble shown in Fig. 1(b). This scheme was called IAM-C. Note that this preamble is, strictly speaking, not OQAM. Nevertheless, it can still be fed to the synthesis filter bank and perfectly reconstructed at the analysis bank.

4. IAM-C EXTENDED: E-IAM-C

It was shown in [10] that the pseudo-pilots generated by IAM-C are of the maximum possible magnitude. However, this optimality is restricted to the 3-symbol preambles with nulls at their sides.[3] The property mentioned at the end of Section 2 suggests that an alternative preamble structure which employs the side symbols as well could yield pseudo-pilots that are stronger than those of IAM-C. It turns out that such a preamble can indeed be formed in an analogous way as previously, by employing the left- and right-hand sides of the neighborhood (8) as well.[4] Thus, at an odd-indexed subcarrier p, with the pilot $\pm dj$ in the middle, one should place

[3] In fact, one can readily see that this IAM-C preamble follows the general structure of an optimal preamble of this kind, recently derived in [12].

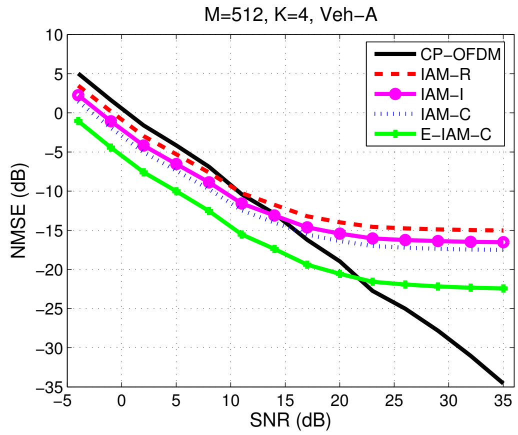
[4] This idea seems to have first been investigated in [9, Section 5.4], as pointed out to the authors by one of the reviewers. It must be noted however, that, for a number of pulse types (including those considered in this paper, where $\gamma > \beta > \delta$ and $\delta < 0.5$), E-IAM-C results in pseudo-pilots of a larger magnitude than in the preambles developed in [9].

$\mp d$ at the right hand side, and its negative, $\pm d$ at the left hand side. These will then contribute $\pm d\gamma j$ each to the interference. In an analogous way, at an even-indexed subcarrier p, the middle pilot, say $\pm d$, will get a total interference of $\pm d\gamma$ from its $(p,0)$ and $(p,2)$ neighbors, if these are chosen as $\pm dj$ and $\mp dj$, respectively. It can then be verified that, because of (8), the interference components from the corner neighbors cancel each other, and hence have no contribution to the pseudo-pilot. However, this loss is not significant since δ is significantly smaller than γ. The resulting pseudo-pilots are again either purely real or imaginary, with magnitude $d|1+2(\beta+\gamma)|$, which is clearly larger than that achieved by IAM-C. For the example filters employed here, the corresponding magnitudes are $2.7274d$ and $1.5986d$. An example of the preamble configuration for $M=8$ and OQPSK modulation is given in Fig. 1(c). One can see that the left hand column is built by repeating each quadruple of the middle pilots in reverse order, while the right hand column is simply the negative of the left hand one. Note that since all three FBMC symbols of this preamble are nonzero, it will require more power to be transmitted than the previous ones. This fact is *fairly* taken into account in the simulation experiments.

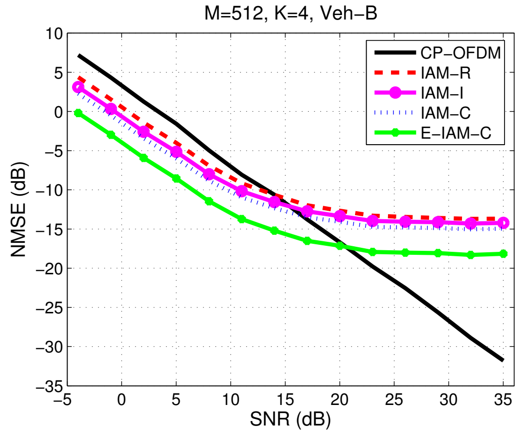
5. SIMULATION RESULTS

In this section, simulation results are reported that demonstrate the estimation performances of the IAM schemes discussed above. For the sake of completeness, CP-OFDM is also included, with the minimum possible CP length (=channel order). A realistic scenario where the preambles are followed by pseudo-random data was considered. This is of importance in the results to be presented, since the assumption used above that the middle preamble FBMC symbol receives negligible interference from the following data is not exact, even with the well localized prototype filters employed in the experiments. Thus, the late part of the preamble signal contains a portion of the front tail of the data burst as well, and *this was taken into account when estimating the transmit power required.* Note that this problem is not present in CP-OFDM and is due to the long tails of the prototype filter impulse response. It should also be pointed out here that the SFB output signal due to the 3-symbol FBMC/OQAM preamble is of length $(K+1)M$, which is about $K+1$ times as long as that of CP-OFDM. One could consider shortening the preamble burst by eliminating its tails as it was recently described in [5]. As for the time invariance assumption for the channel and its validity throughout the preamble signal, one can see that, for scenarios similar to that of WiMAX for example [2], and for reasonable mobile speeds, the channel coherence interval spans many more than $K+1$ M-blocks.

The experiments utilized prototype filters designed as in [4]. QPSK modulation was adopted. Training in CP-OFDM was based on a single-symbol complex QPSK pseudo-random preamble. The normalized MSE, $\frac{\|\mathbf{H}-\hat{\mathbf{H}}\|^2}{\|\mathbf{H}\|^2}$, where $\mathbf{H}$ is the CFR vector and $\hat{\mathbf{H}}$ its estimate, is plotted with respect to the signal to noise ratio (SNR). The latter is defined as the ratio of the channel input (i.e., SFB output) power to the power of the noise. Note that in order to be fair with respect to the power transmission requirements, all the preambles are appropriately scaled so as to result in the *same* power *at the SFB output*. It must be emphasized that these powers turn out to be much more different than they are before the SFB. (In fact, earlier comparison studies seem not to have taken this into account and only equalized powers at the SFB input.) As it was already pointed out in [14], this kind of IAM preambles result in high PAPR signals at the SFB output due to their deterministic/periodic nature. In fact, this effect is more evident in IAM-C than in IAM-R,[5] while E-IAM-C exhibits even higher peaks than these two. These peaks were included in the average power estimates of the corresponding channel inputs.



(a)



(b)

Figure 2: MSE performance for an FBMC/OQAM system with $M=512$, $K=4$, and (a) Veh-A and (b) Veh-B channel models.

Fig. 2 shows the MSE performance of the methods under study for $M=512$ subcarriers with an overlapping factor of $K=4$. Channels of the Veh-A and Veh-B types are considered in Figs. 2(a) and (b), respectively. Observe that in both cases E-IAM-C outperforms the other schemes in the whole SNR range considered. Moreover, all the IAM methods perform better than CP-OFDM for low to moderate SNR values. At higher SNRs, CP-OFDM takes over while the MSE curves of the IAM schemes exhibit an error floor. This is a well known phenomenon in FBMC/OQAM and is due to the fact that the approximation (4) is not exact for channels of significant time dispersion. Hence there is residual intrinsic interference, which is hidden by the noise at low SNRs and shows up in the weak noise regime. It should also be noticed that IAM methods are less well performing with Veh-B channels and the crossing point with the CP-OFDM curve

[5] It is much less severe in IAM-I due to its pseudorandom symbols.

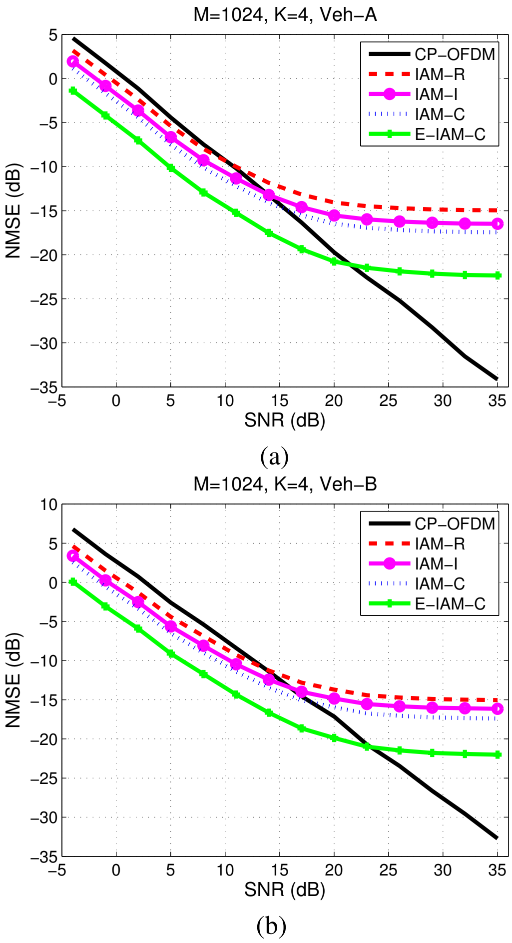


Figure 3: As in Fig. 2, with $M = 1024$.

appears earlier in that case (see Fig. 2(b)). This behavior is again explained by the fact that, in the highly frequency selective Veh-B case, the assumption of a locally flat CFR is much less valid. Increasing the number of subcarriers however, subchannels become closer to being frequency flat and the crossing point with CP-OFDM moves to the right as one can see in Fig. 3, where prototype filters of the same type but with $M = 1024$ were employed.

6. CONCLUSIONS

The problem of preamble-based CE in FBMC/OQAM systems was revisited in this paper, with a focus on the interference approximation idea. The two most well known and efficient IAM variants, IAM-R and IAM-C, were reviewed along with their associated preamble structures. Emphasis was put on IAM-C and its optimality within the class of 3-symbol preambles with all zeros side FBMC symbols. The possible gains from relaxing the latter constraint of all zeros at the sides were investigated, on the basis of fully exploiting the symmetries in the interference weights of the first-order FT neighbors. A novel preamble structure of this type, E-IAM-C, was developed as an extension of the IAM-C idea and shown to result in pseudo-pilots of significantly larger magnitude. The superiority of E-IAM-C was demonstrated via results of simulation in a realistic (including interference from data) and fair scenario and for both mildly and highly frequency-selective channels. Further investigation is required to assess the loss in performance from tail truncation and mitigate the high PAPR problem in these preambles.

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