Preamble-based Channel Estimation in Asynchronous FBMC-OQAM Distributed MIMO Systems

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***Abstract*—This paper investigates downlink channel estimation in a distributed MIMO context employing Filter Bank-based Multi-Carrier Offset QAM (FBMC-OQAM) modulation. Training preambles are constructed based on two different subcarriers assignment schemes (SAS), with the aim of freeing the estimation procedure from the effects of the multi-stream interference (MSI). The Linear Minimum Mean Squared Error (LMMSE) estimator is considered for computing estimates of the channel frequency responses over the entire frequency band and shown to be robust to ill-conditioning associated with the SAS. The application of the different SAS to FBMC-OQAM is evaluated via simulations in an unsynchronized scenario. The performances are compared with the corresponding fully synchronized CP-OFDM scenario. The results demonstrate the robustness of FBMC-OQAM to asynchronism and reveal the advantages of each SAS in various situations.**

Keywords*—*FBMC, preamble, distributed MIMO, channel estimation.

I. Introduction

Cyclic Prefix-based Orthogonal Frequency Division multiplexing (CP-OFDM) is the most popular multicarrier modulation scheme in today's standards, in view of its many advantages, including implementation simplicity, easy handling of the channel frequency selectivity and rather straightforward combination with multiple-input multiple-output (MIMO) technology [1]. However, CP-OFDM is known to suffer from high spectral leakage, which significantly decreases the system flexibility. For instance, it complicates synchronization of the different users of a cell to combat interference. In this and several other respects, Offset-QAM based Filter Bank Multi-Carrier (FBMC-OQAM) has been shown to be an interesting alternative [1]. At the cost of increased complexity and reconstruction delay, FBMC-OQAM allows the use of pulse shaping that enjoys good localization in both time and frequency. This in turn permits the co-existence of asynchronous users at a bandwidth cost (i.e., guard bands) much lower than in a multiuser OFDM context [2].

This paper considers a distributed MIMO downlink system employing FBMC-OQAM. To ensure a Multi-Stream Interference (MSI)-free transmission, a precoding technique can be applied, as shown in [3], [4]. Such techniques require the Channel State Information (CSI) for each Base Station (BS)-user pair to be available. The problem of acquiring CSI in such a context, with the aid of training preambles, is addressed in this paper. The challenges faced in such a multi-user environment are studied, with special emphasis given to the role of the MSI in the pilot subcarriers selection.

To avoid MSI, the subcarriers are distributed among different BS's during the training phase following two different Subcarriers Assignment Schemes (SAS), namely, blocks of pilots or sparsely distributed pilots. This type of preamble training is commonly known as *frequency-division multiplexing* (FDM) and may result in preambles of short duration (not proportional to the number of antennas) [5]. Once CSI has been estimated, each BS is assumed to use the entire frequency range to transmit its useful data signal.

The present scenario has been only partially covered by the literature so far. The authors in [6] were also interested in preamble-based channel estimation for asynchronous access FBMC systems. However, their study is conducted for the single user case only while the present work considers the case of multiple users. Moreover, they do not investigate the potential gains from allocating blocks of subcarriers for training instead of using isolated pilot tones.

Adopting an FDM preamble implies that a receiver can sense only a part of the channel frequency response (CFR) for each transmitter. Depending on the SAS and the channel length, the channel sensing matrix can get ill-conditioned due to, e.g., a lack of pilots or high correlation between neighboring subcarriers as shown in [7]. A classical Least Squares (LS) estimator requiring the inversion of such an ill-conditioned matrix would amplify the noise and is therefore not an appropriate solution. In order to reconstruct the entire CFR while taking care of the possibly ill-conditioned channel sensing matrix, a Linear Minimum Mean Squared Error (LMMSE) estimator is considered in this work.

In summary, this paper investigates the design of preambles for channel estimation in a distributed MIMO downlink system. The performance of two SAS is evaluated via simulations in an unsynchronized scenario using FBMC-OQAM modulation. The resulting mean squared error (MSE) is analyzed and

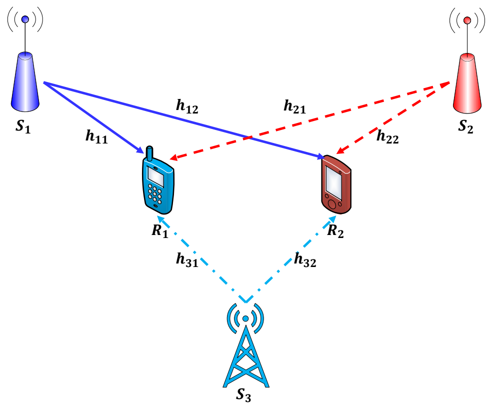

Fig. 1. FBMC-based distributed MIMO system: K_t BS's (transmitters) and Q users (receivers)

compared with that of a corresponding fully synchronized CP-OFDM distributed system.

II. System Model

We consider the downlink of a distributed MIMO system having K_t BS's (transmitters) and Q users (receivers), as depicted in Fig. 1.[1] Each terminal and BS is assumed to be equipped with a single transmit/receive antenna. Both the BS's and the user nodes employ FBMC-OQAM for transmission.

To avoid MSI when estimating CSI, specific SAS ensure that each subcarrier is assigned to at most one BS during the training phase. With M denoting the total number of subcarriers, each user obtains at best $L_p = \lfloor \frac{M}{K_t} \rfloor$ measurements of each BS-user CFR. From the L_p measurements, each user reconstructs the CFR at each of the M frequency points. Each user then sends this CSI back allowing the BS's to apply precoding techniques using the entire band [3], [4]. Two specific SAS are investigated, as depicted in Fig. 2, namely, the *block SAS* and the *equispaced SAS*:

- Equispaced SAS: the subcarriers assigned to the BS's are allocated uniformly over the entire frequency range. To avoid MSI, δ free subcarriers serve as guard bands between neighboring pilot subcarriers. In this scheme, the pilot subcarriers are well spread over the whole band inducing very little (if any) correlation with each other. However, this results in a large number of subcarriers used as guard bands: maximum $L_p = \lfloor \frac{M}{(\delta+1)K_t} \rfloor$ subcarrier pilots can be assigned to each BS.
- Block SAS: several blocks of consecutive subcarriers in the preamble are allocated to each BS. δ free subcarriers adjacent to each block serve as guard bands between the different blocks. The ratio of pilot subcarriers, $L_p = \lfloor \frac{M}{(\delta+L_{\text{Block}})K_t} \rfloor L_{\text{Block}}$, per BS to guard band subcarriers is much more favorable in this case compared to the previous one. However, this scheme may lead to performance degradation given that the subcarriers are much more correlated increasing the ill-conditioning of the channel sensing matrix.

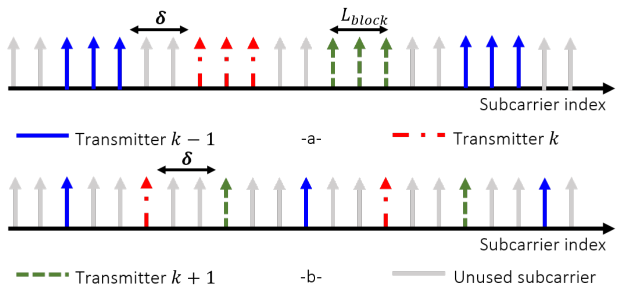

Fig. 2. Subcarriers assignment schemes: -a- block (block size $L_{\text{Block}} = 3$ subcarriers), -b- equispaced, (guard band size $\delta = 2$ subcarriers)

In [8], the authors show that for FBMC-OQAM systems, all equipowered equispaced preambles with a number of pilots L_p equal to or larger than the channel length L_h minimize the MSE of the CFR estimates, subject to a constraint on the total transmitted energy. However, it should be noted that [8] is only concerned with the single-user case and does not address the issues involved in the present distributed MIMO scenario. Furthermore, it assumes there are sufficiently many pilots to be able to recover the entire CFR, *i.e.*, $L_p = \lfloor \frac{M}{(\delta+1)K_t} \rfloor \geq L_h$, which is not always true in our MIMO case, especially for a large number of BS's.

FBMC-OQAM transmission

The discrete-time signal transmitted at a given FBMC-OQAM BS i is

$$s^i[l] = \sum_{n=-\infty}^{+\infty} \sum_{k \in \mathcal{F}_i} d_k^i[n] g_{k,n}[l], \tag{1}$$

where $d_k^i[n]$ are the real-valued PAM symbols transmitted on subcarrier k at FBMC symbol period n. $\mathcal{F}_i$ stands for the subcarriers assigned to the i-th BS for CSI estimation. $g_{k,n}[l]$ is a time/frequency-shifted version of the prototype filter impulse response $g[l]$,

$$g_{k,n}[l] = g[l - nM/2] e^{j\frac{2\pi}{M}k\left(l - \frac{L_g - 1}{2}\right)} e^{j\varphi_{k,n}}, \tag{2}$$

where $\varphi_{k,n} = \frac{\pi}{2}(k+n) - \pi k n$ is introduced to ensure a phase shift of $\frac{\pi}{2}$ between the adjacent transmitted PAM symbols along time and frequency [1]. $L_g = KM$ is the length of g, where K is the overlapping factor.

All signals propagate through multipath channels. The Channel Impulse Response (CIR) between the i-th BS and the q-th user is given by $h^{q,i}[l]$, $l = 0, 1, \ldots, L_h - 1$. The composite signal received at a given user q can be written as

$$y^q[l] = \sum_{i=1}^{K_t} s^i[l] \star h^{q,i}[l] + \eta^q[l], \tag{3}$$

where $\star$ stands for convolution and $\eta^q[l]$ is the additive white Gaussian noise (AWGN) at the q-th user with zero mean and variance $\sigma^2 = N_0$. Assuming a relatively (to the filter bank size) low selective channel, the CFR can be viewed (as it is common) to be constant over the neighborhood Ω_{k_0,n_0} of

[1]For the sake of clarity, only two users are shown.

(k_0, n_0) in which the intrinsic interference is non-negligible [5]. Then, the demodulated signal of the q_0-th user at the k_0-th subcarrier and n_0-th signaling period becomes

$$y_{k_0}^{q_0}[n_0] = \sum_{i=1}^{K_t} H_{k_0}^{q_0,i} \sum_{(k,n)\in\Omega_{k_0,n_0}} d_k^i[n] t_{k-k_0,n-n_0} + \eta_{k_0}^{q_0}[n_0]$$

where $t_{k-k_0,n-n_0} = \sum_l g_{k,n}[l] g_{k_0,n_0}^*[l]$ and $H_{k_0}^{q_0,i}$ is the k_0-th component of the CFR between the i_0-th BS and the q_0-th user. Moreover, the demodulated signal is assumed free from MSI thanks to the guard-bands:

$$y_{k_0}^{q_0}[n_0] = H_{k_0}^{q_0,i_0} \underbrace{\sum_{(k,n)\in\Omega_{k_0,n_0}} d_k^{i_0}[n] t_{k-k_0,n-n_0}}_{c_{k_0}^{i_0}[n_0]} + \eta_{k_0}^{q_0}[n_0]. \tag{4}$$

$c_{k_0}^{i_0}[n_0]$ is referred to as the *pseudo-pilot* and can be approximated based on the knowledge of the symbols in the Ω_{k_0,n_0} neighborhood [5]. A part of each BS-user CFR can then be estimated at pilot subcarriers as

$$\hat{H}_k^{q_0,i} = \frac{y_k^{q_0}[n_0]}{c_k^i[n_0]} = H_k^{q_0,i} + \frac{\eta_{k_0}^{q_0}[n_0]}{c_k^i[n_0]} \tag{5}$$

where $k \in \mathcal{F}_i$ and $i = 1, 2, ..., K_t$.

III. Channel Estimation Algorithms

From now on, we will only consider the channel between the i_0-th BS and the q_0-th user and we drop those indexes for the sake of simplicity. Stacking the CFR estimates $\hat{H}_k^{q_0,i_0}$, $k \in \mathcal{F}_{i_0}$ in an $L_p \times 1$ vector $\hat{\boldsymbol{H}}_{L_p}$ and the estimation error samples $\eta_{k_0}^{q_0}[n_0]/c_k^i[n_0]$, $k \in \mathcal{F}_{i_0}$ in an $L_p \times 1$ vector $\boldsymbol{\eta}_{L_p}$, we obtain from (5)

$$\hat{\boldsymbol{H}}_{L_p} = \boldsymbol{F}_{L_p\times L_h}\boldsymbol{h} + \boldsymbol{\eta}_{L_p}, \tag{6}$$

where the $L_h \times 1$ vector $\boldsymbol{h}$ is the true CIR and $\boldsymbol{F}_{L_p\times L_h}$ is the $L_p \times L_h$ sub-matrix of the full $M \times M$ DFT matrix $\boldsymbol{F}$ consisting of its L_h first columns and its rows corresponding to the indexes in $\mathcal{F}_{i_0}$.

A. Linear minimum mean squared error estimator

Depending on the SAS and the channel length, the channel sensing matrix can get ill-conditioned due to e.g. a lack of pilots or high correlation between neighboring pilots as shown in [7] and [9]. Therefore, a linear minimum mean squared error (LMMSE) estimator instead of the LS one, is considered because of its regularization capacity, that is, $\hat{\boldsymbol{h}} = \boldsymbol{G}\hat{\boldsymbol{H}}_{L_p}$, where

$$\boldsymbol{G} = \arg\min \mathbb{E}(\|\boldsymbol{h} - \hat{\boldsymbol{h}}\|^2) \tag{7}$$

which gives (by the orthogonality principle)

$$\boldsymbol{G} = \boldsymbol{C}_h \boldsymbol{F}_{L_p\times L_h}^H \left(\boldsymbol{F}_{L_p\times L_h}\boldsymbol{C}_h\boldsymbol{F}_{L_p\times L_h}^H + \boldsymbol{C}_\eta\right)^{-1} \tag{8}$$

where $\boldsymbol{C}_h = \mathbb{E}(\boldsymbol{h}\boldsymbol{h}^H)$ and $\boldsymbol{C}_\eta = \mathbb{E}(\boldsymbol{\eta}_{L_p}\boldsymbol{\eta}_{L_p}^H)$, with E_{Pilot} denoting the energy of the (pseudo-)pilots, assumed the same for all pilot tones. With non-adjacent pilot tones, $\boldsymbol{C}_\eta$ reduces to $\boldsymbol{C}_\eta = \frac{\sigma^2}{E_{\text{Pilot}}}\boldsymbol{I}_{L_p}$. When pilots are placed at adjacent subcarriers, the estimation error samples become correlated. However, we will neglect this correlation to simplify the expressions.[2] Moreover, the channel taps are assumed uncorrelated making $\boldsymbol{C}_h$ diagonal and with power normalized such that $\text{tr}(\boldsymbol{C}_h) = 1$. As it is common for channels of unknown power delay profile (PDP), a uniform PDP approximation will be assumed here [10], that is, $\boldsymbol{C}_h = \frac{1}{L_h}\boldsymbol{I}_{L_h}$. We then obtain

$$\begin{aligned}\hat{\boldsymbol{h}} &= \boldsymbol{F}_{L_p\times L_h}^H \left(\boldsymbol{F}_{L_p\times L_h}\boldsymbol{F}_{L_p\times L_h}^H + L_h\boldsymbol{C}_\eta\right)^{-1} \hat{\boldsymbol{H}}_{L_p} \\ &= \left(\boldsymbol{F}_{L_p\times L_h}^H\boldsymbol{F}_{L_p\times L_h} + \frac{L_h\sigma^2}{E_{\text{Pilot}}}\boldsymbol{I}_{L_h}\right)^{-1} \boldsymbol{F}_{L_p\times L_h}^H\hat{\boldsymbol{H}}_{L_p}. \end{aligned} \tag{9}$$

Finally, the CFR at all frequencies is found from the estimated CIR as $\hat{\boldsymbol{H}} = \boldsymbol{F}_{M\times L_h}\hat{\boldsymbol{h}}$.

1) Asymptotic behavior at low SNR: At very low SNR, $\hat{\boldsymbol{h}}$ becomes

$$\hat{\boldsymbol{h}}^{\text{low}} = \frac{E_{\text{Pilot}}}{L_h\sigma^2}\boldsymbol{F}_{L_p\times L_h}^H\hat{\boldsymbol{H}}_{L_p}. \tag{10}$$

The LMMSE estimator is then seen to effectively amount to a scaled IDFT of the CFR measurements. It is as if an IDFT is performed with the assumption that CFR is zero at the non-pilot positions. Notably, this avoids the inversion of a probably ill-conditioned matrix which would lead to noise amplification. Indeed, for large pilot block sizes L_{Block}, the rows of the matrix $\boldsymbol{F}_{L_p\times L_h}^H\boldsymbol{F}_{L_p\times L_h}$ become more correlated and its determinant approaches zero.

2) Asymptotic behavior at high SNR: At high SNR and when $L_p \geq L_h$ (no lack of pilots), the estimator reduces to the pseudo inverse of the DFT matrix $\boldsymbol{F}_{L_p\times L_h}$, yielding

$$\hat{\boldsymbol{h}}^{\text{high}} = (\boldsymbol{F}_{L_p\times L_h}^H\boldsymbol{F}_{L_p\times L_h})^{-1}\boldsymbol{F}_{L_p\times L_h}^H\hat{\boldsymbol{H}}_{L_p}. \tag{11}$$

This estimate will approach the true CIR at sufficiently high SNR values.

In summary, the performance of the LMMSE estimator will directly depend on the SNR level and the conditioning of the matrix $\boldsymbol{F}_{L_p\times L_h}^H\boldsymbol{F}_{L_p\times L_h}$. Regarding its complexity, the LMMSE estimator requires a matrix inversion. However, this can always be performed off-line and/or using a low-rank (SVD-based) implementation [10].

B. Analytical comparison of equispaced and block SAS

The following analysis is conducted in the high SNR regime and making the assumption that there is no lack of pilots ($L_p \geq L_h$), giving rise to the estimator in (11). The frequency domain MSE can then be calculated as

$$\begin{aligned}\text{MSE} &= \mathbb{E}(\|\boldsymbol{H} - \hat{\boldsymbol{H}}\|^2) \\ &= \frac{\sigma^2 L_p}{E_T}\text{tr}\left(\boldsymbol{F}_{M\times L_h}(\boldsymbol{F}_{L_p\times L_h}^H\boldsymbol{F}_{L_p\times L_h})^{-1}\boldsymbol{F}_{M\times L_h}^H\right)\end{aligned} \tag{12}$$

where we assumed a fixed total preamble energy E_T leading to $\boldsymbol{C}_\eta = \frac{\sigma^2}{E_{\text{Pilot}}}\boldsymbol{I}_{L_p} = \frac{\sigma^2 L_p}{E_T}\boldsymbol{I}_{L_p}$.[3] This expression can be

[2]Simulations showed that neglecting these correlation terms induces a negligible performance degradation.

[3]This expression is accurate for equispaced SAS. It holds approximately with a block SAS.

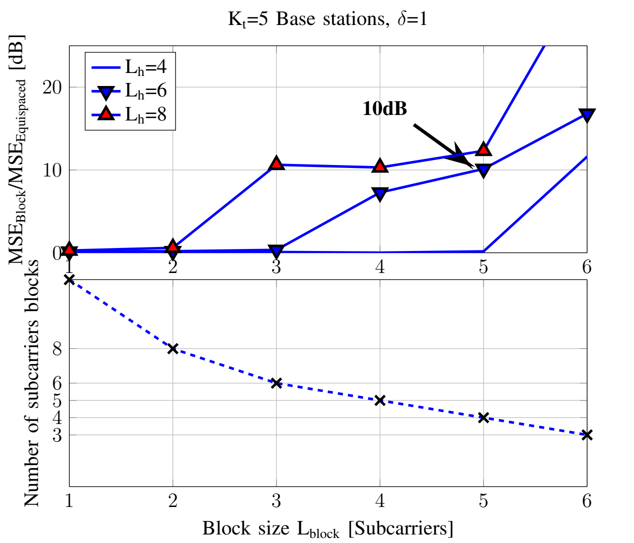

Fig. 3. Analytical comparison of block and equispaced SAS ($L_p \geq L_h$). Equispaced SAS is optimal. However block SAS is also optimal up to the point where a lack of pilot blocks occurs: $\lfloor \frac{M}{(\delta+L_{\text{Block}})K_t} \rfloor < L_h$.

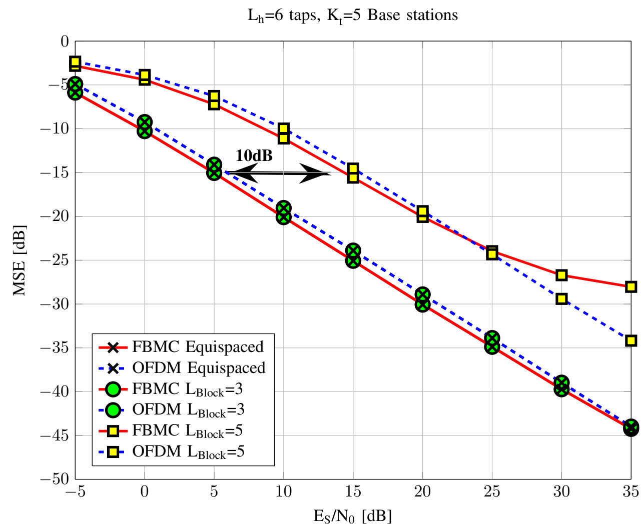

Fig. 4. The equispaced SAS and the $L_{\text{Block}} = 3$ SAS are optimal. However, the $L_{\text{Block}} = 5$ SAS encounters a lack of pilot blocks $\lfloor \frac{M}{(\delta+L_{\text{Block}})K_t} \rfloor \geq L_h$, which degrades the performance as anticipated in Fig. 3.

further simplified for the equispaced SAS by noting that[4] $\boldsymbol{F}^H_{L_p \times L_h} \boldsymbol{F}_{L_p \times L_h} \approx L_p \boldsymbol{I}_{L_h}$, which results in

$$\text{MSE}_{\text{Equispaced}} = \frac{\sigma^2 M L_h}{E_T}. \tag{13}$$

Fig. 3 shows the ratio $\frac{\text{MSE}_{\text{Block}}}{\text{MSE}_{\text{Equispaced}}}$ for an increasing block size L_{Block} with one subcarrier guard band ($\delta = 1$) and for different channel lengths L_h in a scenario with 5 BS's. As suggested in [8], the equispaced SAS is always optimal meaning that this ratio is always greater than one. However, the block SAS in Fig. 3 performs as well as the equispaced SAS up to a certain point depending on L_h. The bottom subfigure demonstrates that this degradation actually occurs when the number of pilot blocks is smaller than the channel length: $\lfloor \frac{M}{(\delta+L_{\text{Block}})K_t} \rfloor < L_h$. In other words, the lack of pilot blocks worsens the conditioning of the channel sensing matrix $\boldsymbol{F}^H_{L_p \times L_h} \boldsymbol{F}_{L_p \times L_h}$.

One should note that the comparison in Fig. 3 relies on the strong assumption that there is no lack of pilots ($L_p \geq L_h$). Moreover, it does not include low SNR situations where the LMMSE estimator is not equal to the pseudo inverse. However, it gives a good intuition of the SAS behaviors and facilitates the interpretation of the simulation results in the next section.

IV. Simulation Results

A uniform PDP is assumed for the channel and is assumed (as it is common) to be sufficiently slowly varying so as to be constant over the preamble duration. The preamble is assumed protected from the data by a sufficient guard time [5]. The subcarrier spacing is set to $\Delta f = 15$kHz with $M = 128$ subcarriers. The prototype filter used is the PHYDYAS filter [11] with roll-off factor $\rho = 1$ and an overlapping factor of $K = 4$. The preambles for each scheme are scaled so as to ensure equal transmit powers. In the simulations, a CP length of $M/4$ is considered. The preambles are constructed with random ± 1 for the equispaced SAS and using the MSE-optimal IAM-C preamble [5] for the block SAS.

Each BS transmits its signal with a specific delay Δ_i which is assumed to be an integer multiple of the sampling period T_s, uniformly distributed over the interval $[0, M]$ and known by the users. Each user employs a different receiver per BS and each receiver has a perfect frame synchronization with their corresponding BS.[5]

The orthogonality in FBMC-OQAM is only restricted to the real domain. Therefore, guard bands are needed between different blocks or subcarriers assigned to different BS's to ensure orthogonality. Since the prototype filter has a roll-off factor of $\rho = 1$, FBMC-OQAM will not suffer from MSI thanks to the well localized prototype filter and the guard band size $\delta = 1$.

For the sake of comparison, the performance of a corresponding fully synchronized CP-OFDM system is also shown. One should note however that establishing this synchronization would require a large overhead which is not needed in FBMC-OQAM. If the BS's transmission are synchronized, there is no need for CP-OFDM to insert guard bands between the different blocks or subcarriers because the orthogonality is ensured in the complex domain. However, for the sake of comparison, the same SAS (and the same guard band sizes) are assumed for CP-OFDM and FBMC-OQAM in the simulations.

Fig. 4 validates the analytical results shown in Fig. 3. The equispaced and block SAS are both optimal up to the point where a lack of pilots or pilot blocks occurs. The 10 dB gap of the $L_{\text{Block}} = 5$ SAS against equispaced SAS is in accordance with the results of Fig. 3, which demonstrates the accuracy of the analytical study.

[4]The strict equality actually only holds if $(\delta+1)K_t$ divides M. If it does not, it will result in a non-perfectly diagonal matrix.

[5]The way receivers are implemented to realize this assumption is not of interest here and may vary. One could use multiple AFB's per user or a single AFB and process the signals coming from the BS's one at a time.

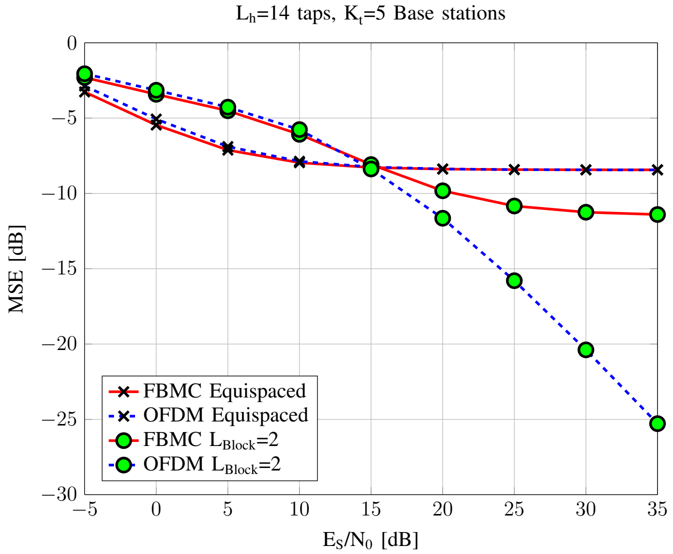

Fig. 5. The equispaced SAS lacks pilots ($\lfloor\frac{M}{(\delta+1)K_t}\rfloor < L_h$) due to the $\delta = 1$ subcarrier guard band. Then, the block SAS at high SNR performs better because its total number of pilots is larger than the channel length, $L_p = \lfloor\frac{M}{(\delta+L_{\text{Block}})K_t}\rfloor L_{\text{Block}} > L_h$.

Unsynchronized FBMC-OQAM performs slightly better than fully synchronized CP-OFDM because it does not waste energy transmitting the CP. As discussed before, FBMC-OQAM does not suffer from MSI thanks to the well localized prototype filter and the guard band size $\delta = 1$. At high SNR however, an error floor appears in FBMC-OQAM due to the inaccuracy of the simplifying assumption that underlies (4) [5]. One can further check that the FBMC-OQAM error floor occurs at lower SNR for $L_{\text{Block}} = 5$ suggesting that not only the noise but also the inaccuracy (residual interference) of the system model (4) is amplified by the ill-conditioning of $\boldsymbol{F}^H_{L_p\times L_h}\boldsymbol{F}_{L_p\times L_h}$.

In Fig. 5, longer channels are considered so that the equispaced SAS is short of pilots due to the many guard bands $\delta = 1$ ($\lfloor\frac{M}{(\delta+1)K_t}\rfloor < L_h$). There are not sufficiently many pilots to estimate the entire CFR and hence a performance floor occurs at high SNR. The result is that the block SAS performs better after a certain SNR value. Indeed, even if it lacks pilot blocks $\lfloor\frac{M}{(\delta+L_{\text{Block}})K_t}\rfloor < L_h$, it has still a total number of pilot tones sufficiently large ($\lfloor\frac{M}{(\delta+L_{\text{Block}})K_t}\rfloor L_{\text{Block}} > L_h$) to be able to compute the entire CFR while the equispaced SAS can only compute a part of it. In other words, in the equispaced SAS case, the matrix $\boldsymbol{F}^H_{L_p\times L_h}\boldsymbol{F}_{L_p\times L_h}$ is non invertible since $L_p < L_h$. The LMMSE estimator will not converge towards the pseudo inverse at high SNR. In the block SAS case, since $L_p \geq L_h$, the matrix $\boldsymbol{F}^H_{L_p\times L_h}\boldsymbol{F}_{L_p\times L_h}$ is invertible even if its determinant is low and the LMMSE estimator will converge towards the pseudo inverse at high SNR as derived in (11).

The error floor appearing at quite low SNR in FBMC-OQAM for the $L_{\text{Block}} = 2$ SAS not present in CP-OFDM comes from the fact that the ill-conditioning amplifies the effect of the inaccuracy of the system model (4) as previously observed in Fig. 4.

V. Conclusion

This paper investigated channel estimation in FBMC-OQAM based distributed MIMO systems. Two different pilot SAS were considered and a LMMSE method was used (because of its regularization capacity) to estimate all BS-user CFR's from their pilot-based measurements. It was shown that block SAS can perform as well as equispaced SAS if there is no lack of pilot blocks. When a lack of pilots occurs for the equispaced SAS, the block SAS performs better at high SNR because it wastes less subcarriers as guard bands. The well-known robustness of FBMC to asynchronism was confirmed. There was no performance degradation due to MSI and, in low SNR situations, FBMC-OQAM outperformed CP-OFDM.

In this paper, the possibly sparse CIR structure was not exploited and the commonly made simplifying assumption of a relatively low channel frequency selectivity was adopted. Coping with strong frequency selectivity and exploiting CIR sparsity to save pilots (or pilot blocks) are possible subjects of future research.

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