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Complexity and Performance Comparison of Filter Bank Multicarrier and OFDM in Uplink of Multicarrier Multiple Access Networks

Hamid Saeedi-Sourck, Yan Wu, Jan W. M. Bergmans, Saeed Sadri, and Behrouz Farhang-Boroujeny

Abstract—We compare filter bank multicarrier (FBMC) and orthogonal frequency-division multiplexing (OFDM) in the uplink of a multiple access network. Our study reveals that the high sensitivity of OFDM to carrier frequency offset (CFO) among different users and the need for interference cancellation methods to reduce this sensitivity leads to very complex and yet not very high performance systems. In FBMC-based networks, on the other hand, near-perfect performance is achieved without any need for interference cancellation, thanks to the excellent frequency localized filters used in the realization of FBMC systems.

Index Terms—Carrier frequency offset (CFO), computational complexity, filter bank multicarrier (FBMC), offset quadrature amplitude modulation (OQAM), orthogonal frequency division multiple access (OFDMA).

I. INTRODUCTION

Multicarrier modulation has recently been recognized as an efficient technique for realization of broadband communication systems. Moreover, adoption of multicarrier technologies for multiuser communications has recently been seriously considered by the industry, as exemplified by activities in the third-generation partnership project (3GPP) long-term evolution (LTE) radio standard working group, where orthogonal frequency-division multiple access (OFDMA) has been suggested and studied extensively [1]. In the uplink of a multiuser network, OFDMA faces a major problem: *high sensitivity to carrier frequency offsets (CFO) among different users* [2]. This may lead to significant *multiple-access interference (MAI)* among users which should be taken care of by using computationally expensive signal processing compensation techniques [3].

The choice of a CFO compensation method in a base station (BS) is closely related to the adopted subcarrier allocation scheme [3]. Block and block-interleaved allocation schemes have been suggested [1], [4], [5]. However, the current trend in the industry, as reflected in the 3GPP LTE documents, is more towards block allocation, where a block of contiguous subcarriers are allocated to each user.

There has been extensive research in CFO compensation methods. The single-user detection (SUD) analyzed in [6] was the first carrier compensation technique that compensates for each user's CFO independently prior to applying a discrete Fourier transform (DFT) demodulator. Since in the SUD, each user requires a distinct DFT demodulator, the BS receiver complexity increases linearly with the number

of users. Choi *et al.* proposed a CFO compensation method after DFT, noting that multiplication of two vectors before DFT is equivalent to the circular convolution of the associated vectors after DFT [7]. This method is usually referred to as CLJL, which stands for the initials of the authors—Choi–Lee–Jung–Lee.

Both SUD and CLJL techniques compensate for each user's CFO without considering any MAI cancellation [3]. Some parallel interference cancellation (PIC) techniques have been proposed to reduce MAI after SUD and CLJL processing, e.g., [8]–[11]. Huang and Letaief [8] proposed the addition of a PIC to the CLJL method. Similarly, a PIC may be adopted to reduce MAI after DFT block in SUD [9]. Successive interference cancellation (SIC) and linear multiuser detectors have also been suggested in [10] and [11], respectively. These result in some performance gain, however, at a cost of significant increase in complexity [3], [10]. In the rest of this paper, we use the names *pre-DFT* to refer to the SUD method of [6] and *post-DFT* to refer to the CLJL method of [7]. Also, the suffix PIC is added to these when a parallel interference cancellation is added.

None of the above techniques can correct the CFO for the uplink of an OFDMA perfectly. The interference cancellation (IC) techniques mitigate MAI to an extent, but not perfectly [3]. They also add significant computational complexity to the BS receiver. This paper studies an alternative method that resolves the problem of MAI in a more natural way. We propose using filter bank multicarrier (FBMC) modulation as an alternative technique to OFDM. The advantages of FBMC over OFDM in cognitive radios have recently been discussed in [12]. In this paper, we emphasize on the advantages of FBMC over OFDM in the uplink of a multiuser network. We note that the low side-lobe filters used in FBMC in a natural way leads to an effective MAI cancellation at virtually no additional cost—no IC is needed. At the same time, we note that although in a point-to-point (i.e., single-user) communication system FBMC is more complex than OFDM, in a multiuser system, the added complexity arising from IC blocks in an OFDMA system makes it significantly more complex than its FBMC counterpart.

There are a few choices for FBMC. The emphasis of this paper is on a type of FBMC that has been widely studied in the literature and often referred to as OFDM/OQAM where OQAM stands for offset quadrature amplitude modulation [13]. The word offset reflects the fact that, in OQAM, the in-phase and quadrature components of each data symbol have a time-offset of half a symbol interval. In [13] it has been noted that the term *staggered QAM* has also been used to refer to OQAM and accordingly suggests the more concise name *staggered modulated multitone (SMT)*. Throughout this paper, we consider SMT for implementation of FBMC systems, and accordingly use the name SMTMA (SMT multiple access) when reference is made to FBMC-based multiple access networks.

Fusco *et al.* [14] have also noted the significance of SMTMA in the uplink of multiuser networks. A sensitivity analysis of SMTMA when different users are subject to different CFOs was presented in [14]. The emphasis of this paper, on the other hand, is on the complexity comparison of SMTMA and OFDMA in the uplink of a network. Here both pre-DFT and post-DFT CFO correction techniques are studied in detail. Our study of post-DFT CFO correction leads to a reduced complexity implementation of the SMTMA system. Simulation results that show the performance comparison between SMTMA and OFDMA, with different CFO compensation techniques, confirm superior performance of the former.

This paper is organized as follows. The system model is presented in Section II. To facilitate the complexity comparison of SMTMA

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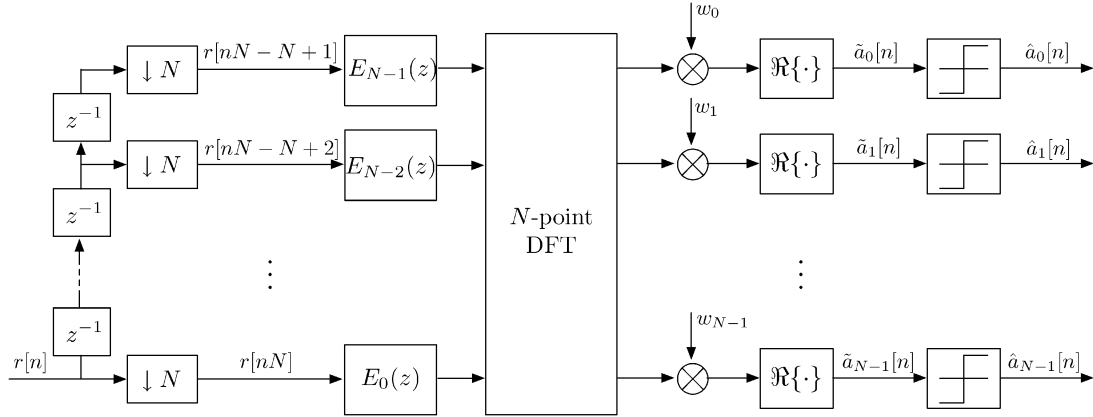


Fig. 1. The in-phase part of an SMT receiver. The quadrature part has a similar structure with the following minor differences: i) The input is backward delayed by one half of a symbol interval ($N/2$ samples), and ii) the $\Re\{\cdot\}$ blocks are replaced by $\Im\{\cdot\}$ blocks. $\Re\{\cdot\}$ denotes the real part of and $\Im\{\cdot\}$ means the imaginary part of.

and OFDMA, a polyphase structure of SMT receiver is reviewed in Section III. In Section IV, the CFO compensation methods that could be applied to SMTMA are developed. The computational complexity comparison of OFDMA and SMTMA methods are presented in Section V. Simulation results are presented in Section VI and the conclusions of the paper are drawn in Section VII.

II. SYSTEM SETUP

We consider the uplink of a multicarrier multiuser network where P active users are communicating with a BS. This network may be based on OFDMA or SMTMA. We assume that there are $N = P \cdot Q$ subcarriers, where Q is the number of subcarriers allocated to each user, including both active and null subcarriers. The set of Q subcarriers assigned to the p th user is denoted by S_p , and we assume $\bigcup_{p=0}^{P-1} S_p = \{0, 1, \dots, N-1\}$ and $S_p \cap S_q = \emptyset$, for $\forall p \neq q$. The transmitted signal from the p th user is represented by $x_p[n]$, and the channel response between the p th user and the BS is denoted by the sequence $c_p[n]$. It is assumed that $c_p[n]$ is nonzero only for $n = 0, 1, \dots, D-1$, where D is the maximum channel delay spread. Accordingly, the received signal can be written as

$$r[n] = \sum_{p=0}^{P-1} (x_p[n] \star c_p[n]) e^{j2\pi\varepsilon_p n/N} + \nu[n] \quad (1)$$

where \star denotes linear convolution, ε_p , $p = 0, \dots, P-1$ is the normalized CFO (with respect to the carrier spacing) corresponding to the p th user, and $\nu[n]$ is an additive white Gaussian noise.

III. POLYPHASE STRUCTURE OF SMT RECEIVER

Several polyphase structures have been proposed for the implementations of SMT systems, e.g., [15]–[17]. The structures proposed in [15] and [16] have the same computational complexity. On the other hand, [17] proposes a structure with half computational complexity of those in [15] and [16]. However, unfortunately, careful examination of the structure in [17] reveals that certain symmetry properties used to simplify the implementation do not hold in the presence a generic channel. This renders the proposed structure unusable in practical implementation of SMT receivers. For the purpose of our study, in this paper we have chosen the polyphase structure of [15].

Fig. 1 presents the in-phase part of the polyphase structure proposed in [15]. This structure assumes that there is no CFO and thus it con-

tains no CFO compensation block. $E_0(z)$ through $E_{N-1}(z)$ are the polyphase components of a prototype filter $H(z)$ based on which the SMT is implemented. We assume that $H(z)$ has a length of $L \cdot N$, thus, each polyphase component has a length of L . The DFT block takes care of the demodulation part of the system. Although in SMT, in general, the equalizer at each subcarrier may be a multi-tap transversal filter [18], in this work single-tap equalizers (the multipliers w_0 through w_{N-1}) are used. This is a reasonable assumption, as in practice one may always increase N so that each subcarrier channel can be well approximated by a flat gain [19].

It is apparent that Fig. 1 is different from a conventional analysis filter bank as appears in the literature where the polyphase components are often followed by an IDFT, [15], [20]. This difference arises, simply, because here we have chosen to feed the input signal from the bottom of a tapped delay line, i.e., opposite to the common practice where the tapped delay line is fed from the top. This puts the input samples in a reversed order and accordingly the IDFT has to be replaced by DFT. This arrangement is chosen, here, because it matches the common practice in presentation of OFDM receiver.

The $N \times 1$ input vector to the DFT block at the time instant n may be written as

$$\mathbf{y}[n] = \sum_{i=1}^L \mathbf{y}^{(i)}[n] \quad (2)$$

where

$$\mathbf{y}^{(i)}[n] = \mathbf{r}[n+i] \odot \mathbf{h}[L-i],$$

$$\mathbf{h}[l] = \begin{bmatrix} h[lN+N-1] \\ h[lN+N-2] \\ \vdots \\ h[lN] \end{bmatrix}, \quad \mathbf{r}[n] = \begin{bmatrix} r[nN-N+1] \\ r[nN-N+2] \\ \vdots \\ r[nN] \end{bmatrix},$$

and \odot denotes point-wise multiplication of vectors.

Using (2) and following the block diagram of Fig. 1, one finds that the input vector to the slicers is given by

$$\tilde{\mathbf{a}}[n] = \Re \{ \mathbf{w} \odot \mathcal{F}(\mathbf{y}[n]) \} = \Re \left\{ \mathbf{w} \odot \mathcal{F} \left(\sum_{i=1}^L \mathbf{y}^{(i)}[n] \right) \right\} \quad (3)$$

where $\mathbf{w} = [w_0 \ w_1 \ \cdots \ w_{N-1}]^T$ is the column vector of the equalizer coefficients, \mathcal{F} denotes the DFT operation, and $\hat{\mathbf{a}}[n]$ is the vector of the recovered data symbols $\hat{a}_k[n]$, $k = 0, \dots, N-1$, before being passed to the slicers whose outputs are the final decisions $\hat{a}_k[n]$ of the transmitted data symbols. In a multiple access system, each set of subcarriers belongs to one user, hence, each set of elements of $\hat{\mathbf{a}}[n]$ belongs to a different user.

IV. CFO COMPENSATION METHODS FOR SMTMA

We assume that the receiver (the BS) has perfect knowledge of CFOs of all users. As in SMT each subcarrier may receive MAI only from the nearby bands, thanks to the very low side lobes of its prototype filter, no IC is required. MAI is avoided, simply, by adding a few null subcarriers at the edges of each set of subcarriers that have been allocated to each user.

A. Pre-DFT Compensation

This is equivalent to the SUD method in OFDMA. The CFO of each user is compensated before applying the analysis filter bank/the demodulator. For the p th user, the CFO compensated signal is formed as

$$y_p[n] = r[n]e^{-j2\pi\hat{\varepsilon}_p n/N} \quad (4)$$

where $\hat{\varepsilon}_p$ is an estimate of the CFO of the p th user. The CFO compensated signal $y_p[n]$ is then analyzed to extract the data transmitted from the p th user. In other words, detection of the data from each user requires a separate polyphase filter bank. Implementation of this polyphase structure is mathematically summarized as

$$\hat{\mathbf{a}}_p[n] = \Re \left\{ e^{-j2\pi\hat{\varepsilon}_p n} \mathbf{w} \odot \mathcal{F} \left(\sum_{i=1}^L e^{-j2\pi\hat{\varepsilon}_p i} \mathbf{c}(\hat{\varepsilon}_p) \odot \mathbf{y}^{(i)}[n] \right) \right\} \quad (5)$$

where

$$\mathbf{c}(\hat{\varepsilon}_p) = \begin{bmatrix} e^{j2\pi\hat{\varepsilon}_p(N-1)/N} \\ e^{j2\pi\hat{\varepsilon}_p(N-2)/N} \\ \vdots \\ 1 \end{bmatrix}, \quad (6)$$

and the subscript p has been added to $\hat{\mathbf{a}}[n]$ to indicate that it contains the recovered data of the p th user. Clearly, one is only interested in the elements of $\hat{\mathbf{a}}_p[n]$ that correspond to the subcarrier outputs associated with the present user.

B. Post-DFT Compensation

In the pre-DFT CFO compensation, detection of the data from each user requires a separate analysis filter bank. This increases the computational complexity as the number of users increases. To resolve this problem, as proposed in [7], for OFDMA, one may compensate CFO after DFT. We follow the same philosophy here and develop a post-DFT CFO compensation method for SMTMA.

Straightforward manipulations of (5) lead to

$$\hat{\mathbf{a}}_p[n] = \Re \left\{ e^{-j2\pi\hat{\varepsilon}_p n} \mathbf{w} \odot \left[\left(\sum_{i=1}^L e^{-j2\pi\hat{\varepsilon}_p i} \mathcal{F}(\mathbf{y}^{(i)}[n]) \right) \odot \mathcal{F}(\mathbf{c}(\hat{\varepsilon}_p)) \right] \right\} \quad (7)$$

where \odot denotes circular convolution. This result shows how one may apply CFO compensation for each user, after converting each column vector $\mathbf{y}^{(i)}[n]$ to the frequency domain. One may notice that while (5) involves one DFT operation per user, (7) has to perform L DFT operations that can be shared among all users. However, we need to remove the phase component $2\pi\hat{\varepsilon}_p i$ for each user separately. This adds some complexity to the receiver which increases with the number of users. Additional complexity is required to obtain $\mathcal{F}(\mathbf{c}(\hat{\varepsilon}_p))$. To resolve this problem, in the following, we develop an alternative structure.

We refer to (5) and note that the vector

$$\mathbf{z}[n] = \sum_{i=1}^L e^{-j2\pi\hat{\varepsilon}_p i} \mathbf{c}(\hat{\varepsilon}_p) \odot \mathbf{y}^{(i)}[n] \quad (8)$$

can be obtained by defining the L times taller vectors

$$\mathbf{c}'(\hat{\varepsilon}_p) = \begin{bmatrix} e^{-j2\pi\hat{\varepsilon}_p} \mathbf{c}(\hat{\varepsilon}_p) \\ e^{-j4\pi\hat{\varepsilon}_p} \mathbf{c}(\hat{\varepsilon}_p) \\ \vdots \\ e^{-j2\pi\hat{\varepsilon}_p L} \mathbf{c}(\hat{\varepsilon}_p) \end{bmatrix} = \begin{bmatrix} e^{-j2\pi\hat{\varepsilon}_p/N} \\ e^{-j4\pi\hat{\varepsilon}_p/N} \\ \vdots \\ e^{-j2\pi\hat{\varepsilon}_p L/N} \end{bmatrix},$$

$$\mathbf{y}'[n] = \begin{bmatrix} \mathbf{y}^{(1)}[n] \\ \mathbf{y}^{(2)}[n] \\ \vdots \\ \mathbf{y}^{(L)}[n] \end{bmatrix},$$

and $\mathbf{z}'[n] = \mathbf{c}'(\hat{\varepsilon}_p) \odot \mathbf{y}'[n]$, and then adding the successive length N partitions of $\mathbf{z}'[n]$. This last step may be viewed as an aliasing in the time domain. Its equivalent in the frequency domain is to take the L -fold decimated samples of the DFT of $\mathbf{z}'[n]$.

From the above result, we conclude that (5) can be rearranged as

$$\hat{\mathbf{a}}_p[n] = \Re \left\{ e^{-j2\pi\hat{\varepsilon}_p n} \mathbf{w} \odot [\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p) \odot \mathbf{y}'[n])]_{\downarrow L} \right\} \quad (9)$$

where the subscript $\downarrow L$ means L -fold decimation. Noting that

$$\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p) \odot \mathbf{y}'[n]) = \mathcal{F}(\mathbf{y}'[n]) \otimes \mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p)),$$

(9) can be written as

$$\hat{\mathbf{a}}_p[n] = \Re \left\{ e^{-j2\pi\hat{\varepsilon}_p n} \mathbf{w} \odot [\mathcal{F}(\mathbf{y}'[n]) \otimes \mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))]_{\downarrow L} \right\}. \quad (10)$$

V. COMPUTATIONAL COMPLEXITY

Table I summarizes the computational complexity of OFDMA, as presented in [7] and [9], and also that of SMTMA for different CFO compensation techniques. It is assumed that there are P users and N/P subcarriers are assigned to each user from a total of N . All operations involve complex numbers and the complexity expressions refer to the number of complex multiplications (CMs) for each case. Also, to give an idea of what these expressions lead to in practical systems, the respective computational complexities are calculated and listed in Table II, for two typical cases of $N = 256$ and 2048 , following the IEEE 802.16e specifications [21]. As seen, while OFDMA without PIC

TABLE I

COMPUTATIONAL COMPLEXITY OF SMTMA AND OFDMA WITH THE CFO COMPENSATION TECHNIQUES. N : NUMBER OF SUBCARRIERS, P : NUMBER OF USERS, m : STAGE INDEX/NUMBER OF CANCELLATION ITERATIONS, NL : PROTOTYPE FILTER LENGTH, M : NUMBER OF NON ZERO ELEMENTS IN $\mathcal{C}'(\hat{\varepsilon}_p)$

Compensator	Computational complexity
pre-DFT OFDMA, [7]	$\frac{NP}{2} \log_2 N - \left\{ \frac{NP}{2} \log_2 P - \frac{3}{2} N(P-1) \right\}$
post-DFT OFDMA, [7]	$\frac{N}{2} \log_2 N + \frac{N^2}{P}$
pre-DFT OFDMA with PIC, [9]	$\frac{NP}{2} \log_2 N - \left\{ \frac{NP}{2} \log_2 P - \frac{3}{2} N(P-1) \right\} + (m-1) \left(N^2 - \frac{N^2}{P} \right)$
post-DFT OFDMA with PIC, [9]	$\frac{N}{2} \log_2 N + \frac{N^2}{P} + (m-1) \left(N^2 + \frac{N^2}{P} \right)$
pre-DFT SMTMA	$NLP + NP \log_2 N - (NP \log_2 P - 2N(P-1))$
post-DFT SMTMA	$NL + NL \log_2 NL + 2NM$

TABLE II

COMPUTATIONAL COMPLEXITY OF SMTMA AND OFDMA WITH THE CFO COMPENSATION TECHNIQUES. $m = 3$, $L = 3$, $M = 20$

Type	$N = 256$			$N = 2048$		
	$P = 2$	$P = 4$	$P = 16$	$P = 2$	$P = 4$	$P = 16$
pre-DFT OFDMA	2,176	4,224	13,956	23,552	46,080	160,768
post-DFT OFDMA	33,792	17,408	5,120	2,108,416	1,059,840	273,408
pre-DFT OFDMA with PIC	67,712	102,528	136,832	4,217,856	6,337,536	8,025,088
post-DFT OFDMA with PIC	230,400	181,248	144,384	14,691,328	11,545,600	9,186,304
pre-DFT SMTMA	5,632	10,752	36,352	57,344	110,592	389,120
post-DFT SMTMA	18,369	18,369	18,369	165,390	165,390	165,390

offers a lower complexity than SMTMA, the addition of PIC to the former leads to a significantly higher complexity.

The computational complexity expressions for SMTMA are obtained as follows. In the case of pre-DFT compensation, P DFTs, each of size N , have to be performed, for each of the phase and quadrature parts of the receiver. Assuming that N is a power of 2 and FFT technique is used, this leads to $2 \times P \times (N/2) \log_2 N = NP \log_2 N$ CMs. NP CMs are needed for the CFO compensation. Other terms are minor adjustment to the number of operations whose details are left to the interested readers.

In the case of post-DFT compensation, the major operations are two DFTs, each of size NL , which has to be performed to recover a block of N QAM data symbols. This requires $NL \log_2 NL$ CMs. The implementation of each circular convolution will lead to the complexity number $N^2 L^2$. This should be repeated $2P$ times (for phase and quadrature parts of each of P users) and leads to a total of $2PN^2 L^2$ CMs. This surpasses the complexity of all the other systems listed in Table I. However, a closer look at the actual output samples which are necessary for each user reveal that this reduces to $2PNL \times (N/P) = 2N^2 L$ CMs. Further computational saving is made by noting that for typical values of NL (in the order of 1000 or larger), a large percentage of the elements of $\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))$ are very close to zero, and thus can be substituted by zero, [7], [8]. This special structure of $\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))$ comes from the fact that $\mathbf{c}'(\hat{\varepsilon}_p)$ is a complex sine-wave with the frequency $\hat{\varepsilon}_p$, thus, the non-zero samples of its Fourier transform are concentrated around zero. Numerical evaluations reveal that, for the cases of interest (such as those in Table II), the number of significant elements of $\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))$, that we denote by M , is approximately independent of N . For instance, the choice of $M = 20$, for both cases of $N = 256$ and 2048, assures that for any value of $\hat{\varepsilon}_p$ in the range of -0.5 to $+0.5$, 98% of the energy of samples of $\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))$ remain within its first $M/2$ and last $M/2$ samples. This choice of M in Table II has been on this basis and the fact that this value of M

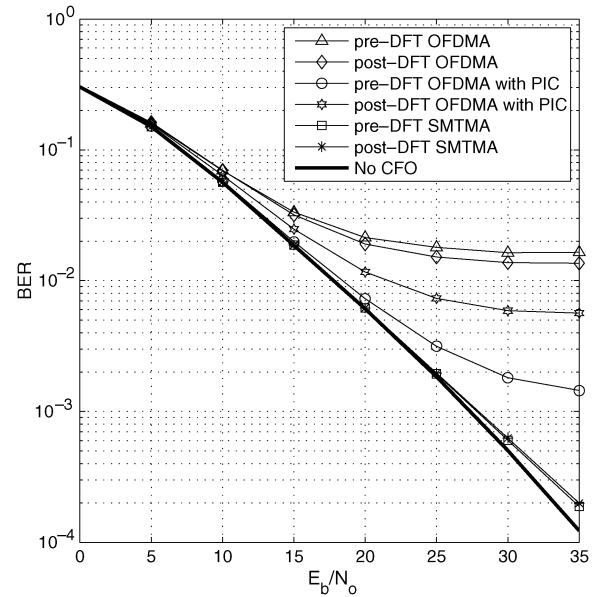


Fig. 2. Coded BER performance for SMTMA and OFDMA when all users reach the BS with the same power.

incurs no noticeable loss in the BER curves of post-DFT SMTMA; see Fig. 2, below. Excluding the near zero elements of $\mathcal{F}(\mathbf{c}'(\hat{\varepsilon}_p))$ in performing the circular convolutions, further reduces the number of CMs from $2N^2 L$ to $2MN$.

The numerical values presented in Table II show that pre- and post-DFT OFDMA systems without PIC have the lowest complexity. However, as demonstrated in the next section, these methods perform very poorly. Addition of PIC to these systems increase the complexity by an order of magnitude, when $N = 256$, and two orders of

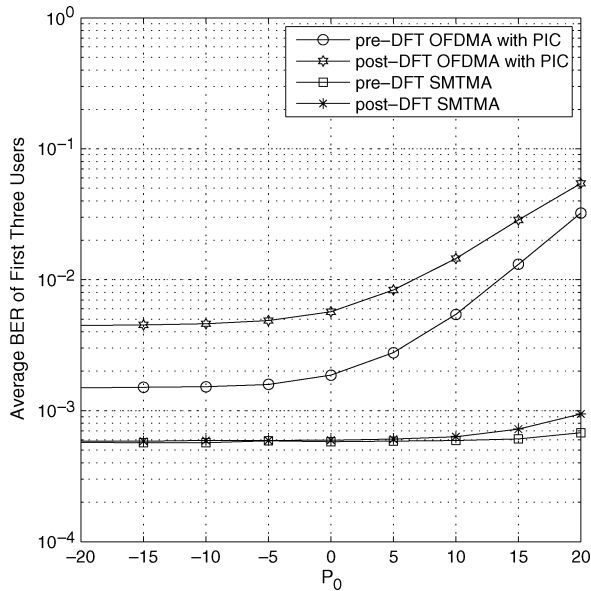


Fig. 3. Coded BER performance for SMTMA and OFDMA when signals from three of four users reach the BS with the same power and the power of the fourth user deviates by P_0 dB.

magnitude, when $N = 2048$. The complexity of pre- and post-DFT SMTMA systems lies between those of OFDMA with and without PIC, and usually closer to the latter. Pre-DFT SMTMA has a lower complexity than its post-DFT counterpart, for smaller numbers of users, P . However, the situation reverses when $P > 6$.

VI. SIMULATION RESULTS

Computer simulations are performed to compare the performance of SMTMA and OFDMA. We let $N = 256$ and assume that there are $P = 4$ users in the network, hence, $Q = N/P = 64$ subcarriers per user. The multipath channel SUI-2 proposed by the IEEE802.16 broadband wireless access working group [22] is considered. The CFO values are chosen randomly and independently for each user from a uniform distribution in the interval $-0.5 < \varepsilon < 0.5$. In all cases, one guard subcarrier is inserted between each pair of adjacent users's bands. Perfect power control is assumed. SMTMA systems use a SR-Nyquist prototype filter with a length of $3N$, designed following [23]. The channel code is a rate 1/2 convolutional code with a constraint length of 5 and data symbols are from a 16-QAM constellation. The bit error rate (BER) results are presented in Fig. 2, where E_b/N_0 denotes bit power over noise power. As seen, OFDMA without PIC performs poorly. An error floor of around 10^{-2} is observed. PIC helps to reduce the error floor, but the performance still remains relatively poor. SMTMA systems, on the other hand, remain solid and follow the reference curve, obtained when all users are perfectly carrier synchronized.

To demonstrate the effect of a possible imperfect power control situation, we run the following experiment. We assume that from the four users, three have perfect power control, thus their respective signals reach the BS with the same power. However, the fourth user's power differs from the others by P_0 dB. Fig. 3 presents the average BERs of the first three users for the case where for them $E_b/N_0 = 30$ dB, and P_0 varies from -20 to $+20$ dB. As seen, in this case, while SMTMA performance remains almost unaffected by variation of P_0 , OFDMA degradation is significant.

VII. CONCLUSION

A study of FBMC for implementation of the uplink of a multicarrier multiuser network was presented. We noted that although OFDMA has been proposed as a candidate in the majority of present standards, an FBMC-based implementation leads to a much superior performance and a lower computational complexity. This is a consequence of the fact that FBMC uses near perfect filters to separate subcarriers/users, thus, avoids the need for any interference cancellation.

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