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Time and frequency synchronization for CoMP with FBMC

Nicolas Cassiau, Dimitri Kténas and Jean-Baptiste Doré

CEA-Leti, MINATEC Campus, Grenoble, France

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***Abstract*— Filter Bank Multicarrier (FBMC) is a multicarrier modulation technique characterized by a prototype filter that can be optimized either in time or in frequency, depending on the target application and/or on the dispersion of the channel. FBMC is often referred as a non-orthogonal modulation technique as symbols overlap in the time domain and as adjacent subcarriers overlap in the frequency domain. This structural absence of orthogonality at the transmitter implies no need for Cyclic Prefix (CP), increasing the bandwidth efficiency of FBMC systems. In cellular systems, Multi-User (MU) downlink cooperation between cells (CoMP) enables to provide a good quality of service to cell edge users while maintaining a high spectral efficiency in the system. For MU-CoMP, time and frequency synchronization between cooperating Base Stations (BSs) and the User Equipment (UE) is a crucial mandatory step. In this study, where synchronization for CoMP with FBMC is considered, it is demonstrated that the UE can robustly estimate very large delays and arbitrary high carrier frequency offsets. The correction of the delays requires only few bits of feedback and the correction of the frequency offset can be entirely realized at the UE side. These results demonstrate that FBMC is a suitable technique for DL CoMP cellular systems.**

Keywords: FBMC, COMP, Time synchronization, Frequency synchronization

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I. INTRODUCTION

FBMC is a multicarrier modulation technique that was first investigated in the 1960s, prior to OFDM. Today FBMC, whose computational complexity is higher than OFDM, is experiencing renewed interest [1]-[2], due to a significant increase in the processing capacity of electronic equipment. The prototype filter of FBMC can be designed with great flexibility to match time or frequency dispersive channels or to cope with standards constraints. Subcarriers filters can be for example designed with arbitrarily low secondary lobes [1], making FBMC a good candidate for multiple access communications or opportunistic spectrum access (TV White Space applications). FBMC frames are in general longer than OFDM frames, due to the length of the prototype filter; this efficiency loss is however compensated by the lack of Cyclic Prefix in FBMC symbols and by the very good frequency localization of FBMC carriers that allow for smaller frequency guard band than with OFDM.

There has been significant interest towards Multi-User cooperative MultiPoint (MU CoMP) techniques for fourth generation cellular systems and beyond where UEs on the edge of cells are served by several BSs (see [3] and references therein). In such cooperating schemes, a crucial issue is the synchronization between the UE and the serving BSs.

In this paper, time and frequency synchronizations are studied for a DownLink CoMP cellular scheme incorporating FBMC at the physical layer. The model used for cooperation between two cells and the FBMC modulation are first described. Section 3 details the algorithm for the estimation and the correction of the CFO (Carrier Frequency Offset) and section 4 shows how the delay between the signals of the two cells is estimated and compensated. Performance of the whole synchronization process are assessed in section 5.

II. SYSTEM MODEL

A. *Cooperation*

The system model is described by figure 1. The UE, at a distance d of BS1, is equipped with two receive antennas. The whole system forms a virtual 2x2 MIMO transceiver. The two BSs transmit in the same time and frequency Resources Blocks (RBs). The signal from BS2 is received at the UE with a delay τ compared to the signal from BS1, reflecting the over-the-air propagation delay. The UE is assumed perfectly synchronized in time with BS1. BS1 and BS2 are assumed synchronized in frequency but not in phase. The Carrier Frequency Offset $\delta_{\Delta f}$ between the BSs and the UE is normalized to the carrier spacing Δ_f.

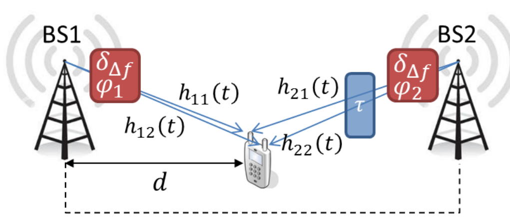

Figure 1 Cooperation between two BSs

B. *FBMC modulation*

A FBMC system with N_c carriers and an overlapping factor K is considered: FBMC symbols are then KN_c samples long (which is the length of the prototype filter in time domain) and each symbol overlap in the time domain with $2(2K-1)$ other symbols. At the receiver side a KN_c point FFT (perfectly aligned with BS1) is realized to recover transmit symbols. In the frequency domain, the m^{th} received FBMC symbol, $m \geq 0$, $\mathbf{R}_m^i \in \mathbb{C}^{1\times KN_c}$, for the link BS i - UE with no delay is given by:

$$\mathbf{R}_m^i = \left(\mathbf{d}_m\widehat{\mathbf{G}}\right)\mathbf{H}_m^i + \mathbf{Z}_m = \mathbf{X}_m\mathbf{H}_m^i + \mathbf{Z}_m \tag{1}$$

$\mathbf{Z}_m \in \mathbb{C}^{1\times KN_c}$ is the white noise (AWGN). $\mathbf{H}_m^i \in \mathbb{C}^{KN_c\times KN_c}$ is the matrix of the channel coefficients. As the channel impulse response is assumed to be short compared to the length of the FFT, $\mathbf{H}_m^i$ can be considered as diagonal. $\mathbf{X}_m = \mathbf{d}_m\widehat{\mathbf{G}}, \in \mathbb{C}^{1\times KN_c}$ are the data after spreading in the frequency domain, $\widehat{\mathbf{G}} \in \mathbb{R}^{N_c\times KN_c}$ is the matrix of filtering vectors given by (2), with $\mathbf{G} \in \mathbb{R}^{1\times 2K-1}$, the filtering vector, i.e. the frequency response of the filter and $\mathbf{d}_m \in \mathbb{C}^{1\times N_c}$ is the vector of transmitted data.

$$\widehat{\mathbf{G}} = \begin{bmatrix} \boxed{G} & & 0 \\ & \boxed{G} & \\ 0 & & \ddots \quad \boxed{G} \end{bmatrix} \tag{2}$$

Note that with FBMC modulation, adjacent carriers in $\mathbf{d}_m$ overlap after frequency spreading, due to the shape of $\widehat{\mathbf{G}}$. This Inter Carrier Interference (ICI) is almost nulled by transmitting alternatively real and pure imaginary numbers on $\mathbf{d}_m$. The data rate loss due to this so-called OQAM modulation is compensated by increasing the symbol rate. Furthermore, FBMC symbols overlap in the time domain, which for sake of clarity is not shown in the signal model. This omission has no impact on the study, once again thanks to the use of OQAM that keeps Inter Symbol Interference (ISI) very low [1].

III. FREQUENCY SYNCHRONIZATION

The CFO between BS i and the UE on received time sample n ($n = 0,1,\dots,KN_C - 1$) of FBMC symbol m ($m \geq 0$) is a phase rotation $e^{j\pi\delta_{\Delta f}m}e^{j\varphi_i}c[n]$ with:

$$c[n] = exp\left(j2\pi\delta_{\Delta f}n/N_c\right) \tag{3}$$

In the frequency domain, the received signal and the CFO convolute. The received symbol $\mathbf{R}_m$ from both BSs affected by the CFO is then (omitting the noise) given by (4).

$$\begin{aligned} \mathbf{R}_m &= \mathbf{R}_m^1\left(e^{j\pi\delta_{\Delta f}m}\mathbf{C}e^{j\varphi_1}\right) + \mathbf{R}_m^2\left(e^{j\pi\delta_{\Delta f}m}\mathbf{C}e^{j\varphi_2}\right) \\ &= e^{j\pi\delta_{\Delta f}m}\mathbf{X}_m\left(\mathbf{H}_m^1 e^{j\varphi_1} + \mathbf{H}_m^2\mathbf{D}_\tau e^{j\varphi_2}\right)\mathbf{C} + \mathbf{Z}_m^{'} \\ &= e^{j\pi\delta_{\Delta f}m}\mathbf{X}_m\mathbf{H}_m\mathbf{C} + \mathbf{Z}_m^{'} \end{aligned} \tag{4}$$

With $\mathbf{D}_\tau \in \mathbb{C}^{KN_C \times KN_C}$ a diagonal matrix such that $D_\tau(i,i) = e^{-2j\pi\tau i}$. $\mathbf{H}_m (= \mathbf{H}_m^1 e^{j\varphi_1} + \mathbf{H}_m^2\mathbf{D}_\tau e^{j\varphi_2}) \in \mathbb{C}^{KN_C \times KN_C}$ is a diagonal matrix and $\mathbf{C} \in \mathbb{C}^{KN_C \times KN_C}$ is the CFO matrix, given by (5). $N = KN_c$.

$$\mathbf{C} = \begin{bmatrix} C[0] & C[-1] & \cdots & C[-(N-1)] \\ C[1] & C[0] & & C[-(N-1)+1] \\ \vdots & \vdots & \ddots & \vdots \\ C[N-1] & C[N-2] & \cdots & C[0] \end{bmatrix} \tag{5}$$

Coefficients $C[r]$, (6), can be computed thanks to the FFT of the coefficients $c[n]$, for $-(N-1) \leq r \leq N-1$.

$$C[r] = e^{j\pi\left(\frac{N-1}{N}\right)\left(K\delta_{\Delta f} - r\right)} \frac{\operatorname{sinc}\left(K\delta_{\Delta f} - r\right)}{\operatorname{sinc}\left(\left(K\delta_{\Delta f} - r\right)/N\right)} \tag{6}$$

$\mathbf{C}$ can be written $\mathbf{C} = \mathbf{FcF}^H$ with $\mathbf{F}$ the Fourier matrix $F[k,l] = exp(-j2\pi kl/N)$, and $\mathbf{c}$ the diagonal matrix composed of the N coefficients $c[n]$ (3). $\mathbf{C}$ is then a Toeplitz matrix that can be easily computed. It is furthermore worth noticing that $\mathbf{C}^H = \mathbf{C}^{-1}$; indeed $\mathbf{C}^{-1} = (\mathbf{FcF}^H)^{-1} = \mathbf{F}\boldsymbol{c}^{-1}\mathbf{F}^H$ and because $1/c[n] = c^H[n]$, $\boldsymbol{c}^{-1} = \boldsymbol{c}^H$ and $\mathbf{C}^{-1} = \mathbf{C}^H$.

Based on the observation of (6), for $\delta_{\Delta f} > 1/(2K)$, $C[0]$ is no longer the dominant term in the CFO matrix, most of the energy is concentrated on the carriers next to the carriers of interest. In the following the notation of (7) will then be used, with $q \in \mathbb{N}$ the 'integer part' of the CFO and $y \in \mathbb{R}$, $|y| < 1/(2K)$ the 'fractional part' of the CFO.

$$\delta_{\Delta f} = q/K + y \tag{7}$$

The scheme proposed for the estimation of the CFO relies on a specifically designed preamble transmitted by the cooperating BSs. It has the advantage to allow accurate estimation of very high CFOs, entirely in the frequency domain. The preamble is composed of M FBMC symbols in which BS1 and BS2 transmit pilots on the same carriers. Based on (4), with the omission of the noise, and with $\mathbf{R}_m$ and $\mathbf{R}_{m+p}$, $p + m < M$, two received symbols of the preamble:

$$\begin{aligned} \mathbf{R}_m\mathbf{R}_{m+p}^H &= \left(e^{j\pi\delta_{\Delta f}m}\mathbf{X}_m\mathbf{H}_m\mathbf{C}\right)\left(e^{j\pi\delta_{\Delta f}(m+p)}\mathbf{X}_{m+p}\mathbf{H}_{m+p}\mathbf{C}\right)^H \\ &= e^{-jp\pi\delta_{\Delta f}}\mathbf{X}_m\mathbf{H}_m\mathbf{C}\mathbf{C}^H\mathbf{H}_{m+p}^H\mathbf{X}_{m+p}^H \end{aligned} \tag{8}$$

Then, remembering that $\mathbf{C}^H = \mathbf{C}^{-1}$, that $\mathbf{H}_m$ is a diagonal matrix and with the assumptions that the channel is constant during the reception of the preamble - $H_m[f] = H_{m+p}[f]$ - and that carrier frequencies of FBMC symbols m and $m+p$ of the preamble carry the same information - $X_m[f] = X_{m+p}[f]$ - an estimation of the CFO is given by the phase of the product $\mathbf{R}_m \mathbf{R}_{m+p}^H$ divided by $-p\pi$. Nevertheless this product requires lots of operations. To decrease the computational complexity the product could be done on any subset of pilot carriers, consecutive or not, of the N carriers of symbols $\mathbf{R}_m$ and $\mathbf{R}_{m+p}$.

This method only allows an accurate estimation of CFOs $|\delta_{\Delta f}| < 1/(2K)$. For estimating higher CFOs, the first step in the estimation is then to scan the frequencies next to the pilot frequencies to find the carriers with the highest energy. This operation allows the estimation of the value q. The second step is the estimation of y thanks to (8) on carriers $\tilde{q}$ carriers apart from the pilot carriers (the tilde symbol $\tilde{}$ stands for the estimation).

The correction of the CFO can also be realized in the frequency domain. The most obvious way to correct it is to multiply the corrupted received symbol $\mathbf{R}_m$ (4) by $e^{-j\pi\delta_{\Delta f}m}\tilde{\mathbf{C}}^{-1} = e^{-j\pi\delta_{\Delta f}m}\tilde{\mathbf{C}}^{H}$.

IV. TIME SYNCHRONIZATION

Cooperative MIMO-OFDM deals with Inter Symbol Interference thanks to the use of a CP: when delay difference between the cooperative BSs is shorter than the CP, all ISI that may deteriorate the quality of service at the receiver are suppressed. For cooperative MIMO-OFDM, due to high propagation distances, the CP must be longer than for non-cooperative OFDM. In LTE 10 MHz the short CP is 72 samples long and the long CP has the duration of 256 samples. Contrary to OFDM symbols, FBMC symbols structurally overlap in the time domain at the transmitter, the use of a CP thus becomes useless. Figure 2 shows the influence of the delay on the BER at the UE for the scheme of figure 1, for 16QAM modulation with convolutional code of rate 3/4. The distance d is 250 m. The CFO is null and neither estimated nor corrected. At the UE, on each antenna j the equivalent channel $h_{eq,j}(t) = h_{1j}(t) + h_{2j}(t-\tau)$ is either supposed perfectly known (dotted line curve) or estimated (solid line curve). The estimation is realized thanks to preamble symbols with one carrier over four being a pilot carrier.

It first must be noted that with perfect channel estimation, performance does not decrease even for very high delays. It is then shown that FBMC, with real channel estimation, can cope with delays up to 120 times samples without any correction. For the parameters of LTE with 10 MHz channel bandwidth, this corresponds to a distance of 2340 m.

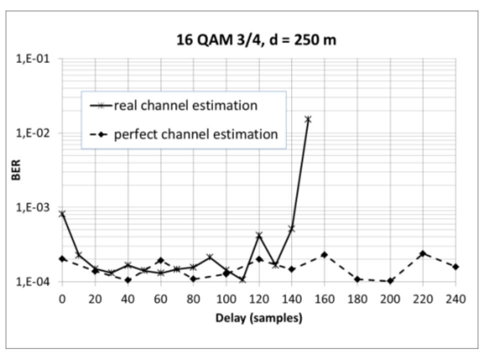

Figure 2 BER of CoMP FBMC with MRC at the receiver for different delays τ between BSs.

The causes of the deterioration of the BER for a real channel estimation with high delays are the phase rotations due to the delay τ that increase the frequency selectivity of the equivalent channel $h_{eq,j}(t)$. Indeed when the coherence bandwidth of the channel is much smaller than the pilot spacing, channel interpolation cannot be reliably computed.

Table 1 shows the 50% coherence bandwidth [4] of the equivalent channel, computed thanks to Monte-Carlo simulations. The channel between a BS i and the UE is modeled by the exponential model, characterized by the maximum excess delay $\gamma_{max,i}$ [5], with random phases for each path.

Table 1 Smallest 50 % coherence bandwidth over 100 draws for each set of parameters. In number of carriers.

γmax,1	γmax,2	τ (samples)													
(µs)	(µs)	5	25	30	35	40	65	75	85	95	105	115	125	135	185
3	3	>32	13	11	10	9	6	5	4	4	4	3	3	3	2
3	0,2	>32	16	13	11	10	6	5	5	4	4	4	3	3	2
1,6	3	>32	13	11	10	8	6	5	4	4	4	3	3	3	2
1,6	0,2	>32	15	13	11	9	6	5	5	4	4	4	3	3	2
0,2	3	>32	12	10	9	8	5	5	4	4	4	3	3	3	2
0,2	0,2	>32	14	12	10	9	6	5	5	4	4	3	3	3	2

Table 1 first illustrates that with the pilot pattern considered in this study (one carrier over four), the frequency interpolation used to estimate channel cannot recover the channel when the delay is higher than around 115 samples (this result is coherent with figure 2). It is also shown that the channel can be considered flat over a RB (12 consecutive carriers) if the delay τ seen at the UE between the signals from the two BSs is smaller than 30 samples.

In this aim, an algorithm is proposed to estimate and compensate the delay at the UE. The estimation of the channel in the time domain is first computed at the UE (thanks to the IFFT of the estimate in the frequency domain). The UE then seeks for a peak in the channel response. The returned value corresponds to the arrival time of the signal from BS2. **Figure 3** shows an example for two values of the delay. Note that the range of the estimation is limited by the aliasing of the IFFT; i.e. the peeks at $\tau = 0$ and $\tau = 256$ both correspond to the arrival time of signal from BS1.

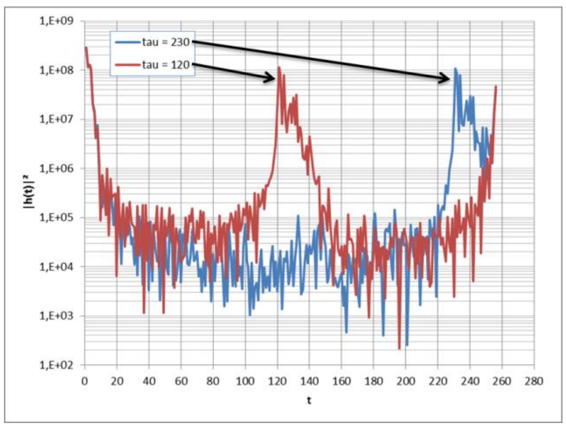

Figure 3 Time estimation of the equivalent channel. d=250 m, example.

Based on table 2, the goal of the delay detection algorithm is to return a value $\tilde{\tau}$ such that $0 \leq \tau - \tilde{\tau} \leq 30$, with τ the actual delay. BS2 is then asked to rotate each carrier k of the transmitted data with a factor $exp(2j\pi k\tilde{\tau}/N)$, with N the number of carriers of the symbol. The virtual delay seen by the UE is then $\hat{\tau} = \tau - \tilde{\tau}$. The set of values taken by $\tilde{\tau}$ must be as small as possible in order to limit the feedback from UE to BS.

The algorithm schematically consists in filtering $|h(n)|^2$ presented on figure 3 and finding the index of the maximal value. Based on this index the algorithm outputs a value $\tilde{\tau}$ chosen in the set $\{0, 20n + 5\}, n = 1,2, \dots, 11$. This set is composed of 12 values, it then requires 4 bits of feedback, plus the identifier of the BS that must rotate the carriers.

V. PERFORMANCE EVALUATION

Figure 4 shows the chronology of the operations to be realized for the synchronization of the BSs and the UE.

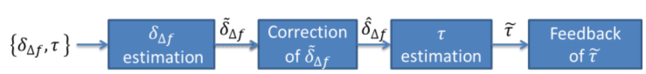

Figure 4 Chronology of the synchronization process.

A. Estimation of the delay with high CFO

It is first established that the CFO must be estimated and compensated before the estimation of the delay. The algorithm presented in section 4 is used for the estimation of the delay. Based on table 1, remembering that the channel is required to be flat on 12 consecutive carriers:

The probability of non detection of a delay (p_{nd}) is defined as the probability that $\hat{\tau} = \tau - \tilde{\tau} \geq 30$.

The probability of false alarm of a delay (p_{fa}) is defined as the probability that $\hat{\tau} = \tau - \tilde{\tau} < 0$.

Figures 5 left and right respectively illustrate p_{nd} and p_{fa} for different values of the delay and of the CFO. The CFO is not corrected. It is demonstrated that the delay cannot be estimated accurately when the CFO is high, which is likely to happen at the beginning of the synchronization process.

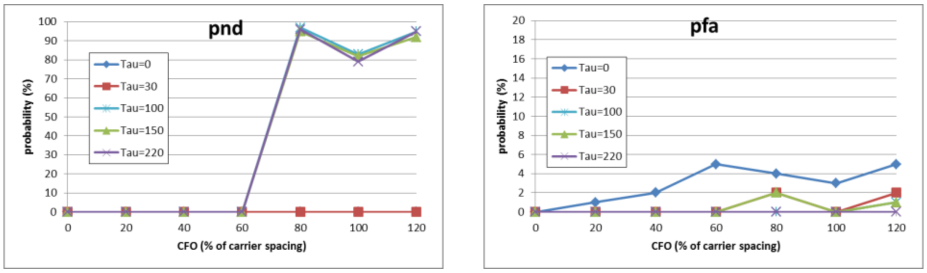

Figure 5 Estimation of the delay before CFO correction: probabilities of non detection and false alarm. *d*=250 m.

B. *Estimation of the CFO with high delay*

The first step of the frequency synchronization consists in finding the integer part q of the CFO by observing a received preamble symbol $\mathbf{R} \in \mathbb{C}^{1 \times KN_c}$ in the frequency domain. In the following the positions (frequency number) of the 150 pilots in the symbol are $I = \{177: 4K: 2577\}$ with $K = 4$. Let $I_p, p = 0,1, \ldots 4K - 1$, be the set of values $I_p = \{177 + p: 4K: 2577\}$ and $\langle I_p \rangle$ the number of values in I_p (I_0 is then the set of pilot positions). The estimated integer part $\tilde{q}$ of the CFO is given by (9). This method allows for estimation of CFOs up to $100(4K-1)/K$ % of the carrier spacing.

$$\tilde{q} = \underset{p=0,1,\ldots,4K-1}{\operatorname{argmax}} \left(\frac{1}{\langle I_p \rangle} \sum_{f \in I_p} |R[f]|^2 \right) \tag{9}$$

Extensive simulations were run with delays up to 220 samples, with randomly drawn phases φ_1 and φ_2 (figure 1), in a frequency selective channel. They have shown that the method in (9) always give an accurate value of $\tilde{q}$.

Once the integer part of the CFO have been estimated, it is necessary to estimate the fractional part y. Figures 6 and 7 show the residual CFO after estimation $\hat{\delta}_{\Delta f} = |\tilde{\delta}_{\Delta f} - \delta_{\Delta f}|$ on 200 draws, with $\tilde{\delta}_{\Delta f}$ computed with the algorithm proposed in section 3. The actual CFO $\delta_{\Delta f}$ is equal to 12.5 %, which is the highest possible value of the fractional part of the CFO y (here the overlapping factor K is 4) and $d = 250$ m and 160 m. The estimation $\tilde{\delta}_{\Delta f}$ is done on $N_0 \times (2K - 1)$ pilot carriers, averaged on the two receive antennas.

First it can be noted that the performance of the algorithm is nearly the same for BSs perfectly synchronized in time ($\tau = 0$) at the receiver or with a high time desynchronization between the BSs at the receiver ($\tau = 220$). When the SNR at the UE is small (d=250 m), a quite high number of carriers are necessary to compute a good estimation of the CFO: with $N_0 = 40$, the residual CFO is most of the time lower than 0.13 % of the carrier spacing, which is a reasonable value. In an easier situation, d=160 m, few carriers are enough to get a good estimation: with $N_0 = 20$, the residual CFO is always lower than 0.15 % of the carrier spacing.

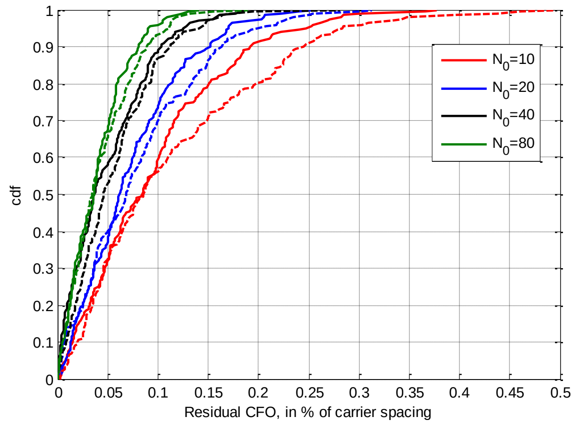

Figure 6 Residual CFO, in % of carrier spacing. $\delta_{\Delta f} = 12.5$ %, d=250 m. Solid lines: τ=220 samples, dashed lines: τ=0 sample.

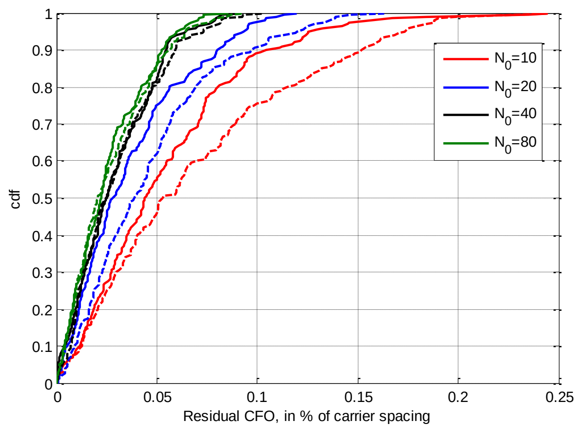

Figure 7 Residual CFO, in % of carrier spacing. $\delta_{\Delta f} = 12.5$ %, d=160 m. Solid lines: τ=220 samples, dashed lines: τ=0 sample.

C. Estimation of the delay with CFO corrected

In this section the performance of the estimation of the delay after the compensation of the CFO is assessed (block 'τ estimation' on figure 4). The CFO is estimated and corrected thanks to the algorithm presented in section 5.2.

Similarly to section 5.1, p_{nd} and p_{fa} were computed for different values of τ (0 to 220 samples) and of $\delta_{\Delta f}$ (0 to 120% of carrier spacing). Those probabilities were shown to be null for all tested values. High delays can then be accurately estimated once the CFO has been estimated and corrected.

VI. CONCLUSION

In this paper time and frequency synchronizations between a UE and two cooperative BSs using FBMC modulation were studied. FBMC modulation was demonstrated to be very resistant to time propagation differences between signals from the two BSs, due to its overlapping structure. Delays up to 120 samples, i.e. 7.8 µs (2340 meters) can be tolerated at the UE without any estimation nor correction, with the proposed pilot scheme. To the contrary of OFDM, FBMC does not need any Cyclic Prefix to cope with time propagation differences; FBMC spectral efficiency is then preserved without impact on time synchronization. Delays higher than 7.8 µs are unlikely to occur in most of cooperative systems. Nevertheless, in order to decrease the effects of phase rotations of the channel due to high delays, an algorithm was proposed to estimate the delay. The algorithm was demonstrated to be very robust to non detections and false alarms with very limited feedback information.

Frequency synchronization (estimation and compensation of the CFO) with FBMC can be entirely realized at the UE in the frequency domain. Thanks to the very good frequency localization of FBMC carriers, the most part of the CFO can be easily and accurately estimated thanks to a simple energy detection algorithm on the preamble carriers. The residual part of the CFO after this estimation is lower than 100/(2K)% of carrier spacing. The same method applied to OFDM would lead to a much higher residual CFO of 50 % at maximum. The algorithm proposed to estimate the residual part was shown to have good performance with reduced complexity. The correction of the CFO requires no feedback from the UE to the BSs.

Future work will assess the performance of FBMC in a Multi-User MIMO CoMP transmission system with precoding at the BSs based on the Channel State Information provided by the UEs.

REFERENCES

[1] M.G. Bellanger et al, “FBMC physical layer : a primer”, 06/2010

[2] Y. Medjahdi, M. Terré, D. Le Ruyet, D. Roviras, A. Dziri, “Performance Analysis in the Downlink of Asynchronous OFDM/FBMC Base Multi-Cellular Networks”, IEEE Transactions on Wireless Communications, vol. 10, N°8, August 2011

[3] WINNER+ project, “D1.4: Initial Report on Advanced Multiple Antenna Systems”, available at http:// projects.celtic- initiative.org /winner+/ deliverables_winnerplus.html

[4] T.S. Rappaport, Wireless Communications, Prentice Hall, 2002.

[5] Jean Paul Linnartz, “Wireless Communication, Jean Paul Linnartz' Reference Website”, http://www. wirelesscommunication.nl/reference/about.htm