COMPARISON OF CDMA-OFDM AND CP-OFDM USING ALAMOUTI SCHEME IN MIMO SYSTEMS

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ABSTRACT

This paper deals with a combination of OFDM and the Alamouti scheme for transmission of digital information. First the OFDM modulation scheme is presented. The Alamouti decoding scheme cannot be directly applied to OFDM modulation. The Alamouti coding scheme requires a complex orthogonality property; whereas OFDM only provides real orthogonality. However, under special conditions, a transmission scheme combining CDMA and OFDM can satisfy the complex orthogonality condition. Adding a CDMA component can thus be seen as a solution to apply the Alamouti scheme in combination with OFDM. However, a detailed analysis shows that the CDMA-OFDM combination has to be built taking into account particular features of the transmission channel. The simulation results presented in this paper illustrates the 2×1 Alamouti coding scheme for which CDMA-OFDM and CP-OFDM are compared in two different scenarios: (i) CDMA performed in the frequency domain (ii) CDMA performed in time domain.

Keywords - MIMO, Alamouti, OFDM, CDMA

I. INTRODUCTION

Increasing the transmission rate and/or providing robustness to channel conditions are nowadays two of the main research topics for wireless communications. Indeed, much effort is done in the area of MIMO, where Space Time Codes (STCs) enable to exploit the spatial diversity when using several antennas either at the transmitting side or at the receiving side. One of the most known and used STC technique is Alamouti code [1]. Alamouti code has the property to be simple to implement while providing the maximum channel diversity. On the other hand, multicarrier modulation (MCM) is becoming, popular Orthogonal Frequency Division Multiplexing (OFDM) scheme, and an appropriate modulation for transmission over frequency selective channels. Furthermore, when appending the OFDM symbols with a Cyclic Prefix (CP) longer than the maximum delay spread of the channel to preserve the orthogonality, CP-OFDM has the capacity to transform a frequency selective channel into a bunch of flat fading channels which leads to various efficient combinations of the STC and CP-OFDM schemes. However, the insertion of the CP yields spectral efficiency loss. In addition, the conventional OFDM modulation is based on a rectangular windowing in the time domain which leads to a poor (sinc(x)) behavior in the frequency domain. Thus CP-OFDM gives rise to two drawbacks: loss of spectral efficiency and sensitivity to frequency dispersion. These limitations may be overcome by some other OFDM variants that also use the exponential base of functions. It can be deduced from the Balian-Low theorem, [2], it is not possible to get at the same time (i) Complex orthogonality; (ii) Maximum spectral efficiency; (iii) A well-localized pulse shape in time and frequency. With CP-OFDM conditions (ii) and (iii) are not satisfied, while there are two main alternatives that satisfy two of these three requirements and can be implemented as filter bank-based multicarrier (FBMC) modulations. Relaxing condition (ii) a modulation scheme named Filtered MultiTone (FMT) [3], also named as oversampled OFDM in [4], where the baseband implementation scheme can be seen as the dual of an oversampled filter bank. But if one really wants to avoid the two drawbacks of CP-OFDM the only solution is to relax the complex orthogonality constraint.

In a publication [11], under certain conditions, a combination of Coded Division Multiple Access (CDMA) with OFDM could be used to provide the complex orthogonal property. On the other hand, it has also been shown in [12] that spatial multiplexing MIMO could be directly applied to OFDM. However, in the MIMO case there is still a problem which has not yet found a favorable issue: It concerns the combined use of the popular STBC Alamouti code together with OFDM. Basically the problem is related to the fact that OFDM by construction produces an imaginary interference term.

Unfortunately, the processing that can be used in the SISO case, for cancelling it at the transmitter side (TX) [11] or estimating it at the receiver side (RX) [7], cannot be successfully extended to the Alamouti coding/decoding scheme. Indeed, the solutions proposed so far are not fully satisfactory. The Alamouti-like scheme for OFDM proposed in [8] complicates the RX and introduces a processing delay. The pseudo-Alamouti scheme recently introduced in [12] is less complex but requires the appending of a CP to the OFDM signal which means that condition (ii) is no longer satisfied.

The aim of this paper is to take advantage of the orthogonality property resulting from the CDMA-OFDM combination introduced in [11] to get a new MIMO Alamouti scheme with OFDM. The paper includes some general descriptions of the OFDM modulation and the MIMO Alamouti scheme. Both techniques are combined. However the MIMO decoding process is very difficult because of the orthogonality mismatch between Alamouti and OFDM. A cobmbination of CDMA-OFDM is proposed in order to solve the problem. It is shown that the combination of CDMA and OFDM (CDMA-OFDM) can provide the complex orthogonality property; The, two different approaches with Alamouti coding are proposed, by considering either a spreading in the frequency or in the time domain.3 When spreading in time is considered, strategies of implementing the Alamouti coding is proposed. The simulation results show that, using particular channel assumptions, the Alamouti CDMA-OFDM technique achieves similar performance to the Alamouti CP-OFDM system.

II. OFDM AND ALAMOUTI

The OFDM Transmultiplexer: The baseband equivalent of a continuous-time multicarrier OFDM signal can be expressed as follows [7]:

$$s(t) = \sum_{m-0}^{M-1} \sum_{n \epsilon Z} a_{m,n}\, g(t - n\tau_0) e^{j2\pi m F0t}\, \nu_{m,n} \qquad (1)$$

where $g_{m,n}(t) = g(t - n\tau_0)e^{j2\pi m F0t}\,\nu_{m,n}$ with **Z** the set of integers, M = 2N an even umber of subcarriers, $F_0 = 1/T_0 = 1/2\tau_0$ the subcarrier spacing, g the prototype function assumed here to be a real-valued and even function of time, and $\nu_{m,n}$ an additional phase term such that $\nu_{m,n}$= jm+nejφ0, where φ0 can be chosen arbitrarily. The transmitted data symbols $a_{m,n}$ are real-valued. They are obtained from a 22K-QAM constellation, taking the real and imaginary parts of these complex-valued symbols of duration $T_0 = 2\tau_0$, where τ_0 denotes the time offset between the two parts [2, 6, 7, 9].

Assuming a distortion-free channel, the Perfect Reconstruction (PR) of the real data symbols is obtained owing to the following real orthogonality condition:

$$R\{<g_{m,n},g_{p,q}>\}=R\{\int g_{m,n}(t)g^*_{p,q}(t)dt\}=\delta_{m,p}\delta_{n,q} \qquad (2)$$

Where $*$ denotes conjugation, $<\cdot, \cdot>$ denotes the inner product, and $\delta_{m,p} = 1$ if m = p and $\delta_{m,p} = 0$ if m≠p. Otherwise said, for (m,n) ≠ (p, q), $<g_{m,n},g_{p,q}>$ is a pure imaginary number. For the sake of brevity, we set $<g>^{p,q}_{m,n} = -j<g_{m,n}, g_{p,q}>$. The orthogonality condition for the prototype filter can also be conveniently expressed using its ambiguity function

$$A_g(n,m)= \int_{-\infty}^{\infty} g(u - n\tau_0)\, g(u)e^{2j\pi mF_0 u}\, du. \qquad (3)$$

It is well-known [7] that to satisfy the orthogonality condition (2), the prototype filter should be chosen such that

$A_g(2n, 2m) = 0$ if (n,m) ≠ (0, 0) and $A_g(0,0) = 1$.

In practical implementations, the baseband signal is directly generated in discrete time, using the continuous time signal samples at the critical frequency, that is, with $F_e = MF_0 = 2NF_0$. Then, based on [9], the discrete-time baseband signal taking the causality constraint into account, is expressed as

$$s(t)= \sum_{m-0}^{M-1} \sum_{n\epsilon Z} a_{m,n}\, g(k - nN)e^{j2\pi m(k-(Lg-1)/2)}\, \nu_{m,n} \qquad (4)$$

where, $gm,n[k] = g(k - nN)e^{j2\pi m(k-(Lg-1)/2)}\, \nu_{m,n}$

The parallel between (1) and (4) shows that the overlapping of duration τ_0 corresponds to N discrete-time samples. For the sake of simplicity, we will assume that the prototype filter length, denoted L_g, is such that $L_g = bM = 2bN$, with b being a positive integer. With the discrete time formulation, the real orthogonality condition can also be expressed as:

$$R<g_{m,n},g_{p,q}>\}=R\{\sum g_{m,n}[k])g^*_{p,q}[k]dt\}=\delta_{m,p}\delta_{n,q}\ k\epsilon Z \qquad (5)$$

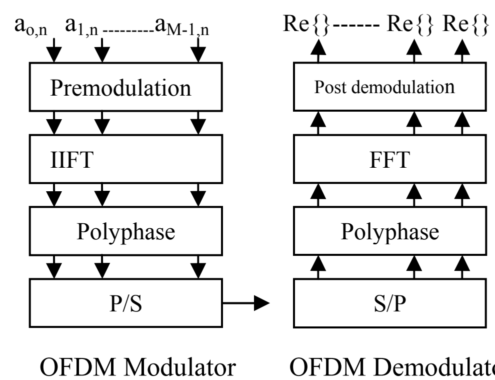

Fig. 1: Transmultiplexer scheme for the OFDM modulation.

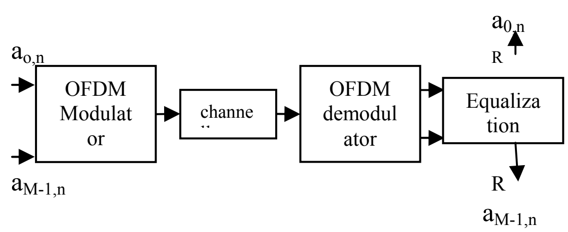

Fig. 2: The transmission scheme based on OFDM.

As shown in [9], a simplified description is provided in Figure 1, where it has to be noted that the pre-modulation corresponds to a single multiplication by an exponential whose argument depends on the phase term $\nu_{m,n}$ and on the prototype length. Note also that in this scheme, to transmit OFDM symbols of a given duration, denoted T_0, the IFFT block has to be run twice faster than for CP-OFDM. The polyphase block contains the polyphase components of the prototype filter g. At the RX side, the dual operations are carried out.

The prototype filter has to be PR, or nearly PR. In this paper, we use a nearly PR prototype filter, with length Lg = 4M, resulting from the discretization of the continuous time function named Isotropic Orthogonal Transform Algorithm (IOTA).

Before being transmitted through a channel the baseband signal is converted to continuous-time. An OFDM modulator delivers a signal denoted s(t), modulator corresponds to an FBMC modulator as shown in Fig 1.

In Fig. 2, OFDM transmission scheme, it is compared to a channel that breaks the real orthogonality condition thus equalization must be performed at the receiver side to restore this orthogonality.

Let us consider a time-varying channel, with maximum delay spread equal to Δ. We denote it by h (t, τ) in time, and it can also be represented by a complex-valued number $H^{(c)}_{m,n}$ for subcarrier m at symbol time n. At the receiver side, the received signal is the summation of the s (t) signal convolved with the channel impulse response and a noise component η(t). For a locally invariant channel, we can define neighborhood, denoted $\Omega_{\Delta m,\Delta n}$, around (m_0,n_0) position, with

$$\Omega_{\Delta m,\Delta n}=\{(p,q)|p|\le\Delta m,|q|\le\Delta n|H^{(c)}_{m0+p,n0+q}\approx H^{(c)}_{m0,n0} \qquad (6)$$

It is defined that $\Omega^*_{\Delta m,\Delta n}=\Omega_{\Delta m,\Delta n}-\{(0,0)\}$.

Also Δn and Δm are chosen according to the time and bandwidth coherence of the channel, respectively. Then, assuming $g(t-\tau-n\tau_0)\approx g(t-n\tau_0)$, for all $\tau\epsilon[0,\Delta]$, the demodulated signal can be expressed as [13, 14, 17]

$$y(c)_{m0,n0}=H(c)_{m0,n0}(a_{m0,n0}+ja^{(i)}_{m0,n0})+J_{m0,n0}+\eta_{m0,n0} \qquad (7)$$

with $\eta_{m0,n0}=\langle\eta,g_{m0,n0}\rangle$ the noise component, $a^{(i)}_{m0,n0}$, the interference created by the neighbor symbols, given by

$$a^{(i)}_{m0,n0}=\sum_{(p,q)\,\epsilon\,\Omega^*_{\Delta m,\Delta n}} a_{m0+p,n0+q}\,\langle g\rangle^{m0,n0}_{m0+p,n0+q}, \qquad (8)$$

and $J_{m0,n0}$ the interference created by the data symbols outside $\Omega_{\Delta m,\Delta n}$.

It can be shown that, even for small size neighborhoods, if the prototype function g is well localized in time and frequency, $J_{m0,n0}$ becomes negligible when compared to the noise term $\eta_{m0,n0}$. Indeed a good time-frequency localization [7] means that the ambiguity function of g, which is directly related to the $\langle g\rangle^{m_0,n_0}_{m0+p,n0+q}$ terms, is concentrated around its origin in the time-frequency plane, that is, only takes small values outside the $\Omega_{\Delta m,\Delta n}$ region. Thus, the received signal can be approximated by $y^{(c)}_{m0,n0}\approx H(c)_{m0,n0}(a_{m0,n0}+ja^{(i)}_{m0,n0})+\eta_{m0,n0}$ (9)

For the rest of our study, we consider (9) as the expression of the signal at the output of the OFDM demodulator.

III. ALAMOUTI SCHEME

General Case. In order to describe the Alamouti scheme [1], let us consider the one-tap channel model described as

$$y_k = h_{k,u}s_{k,u} + n_k, \qquad (10)$$

where, at time instant k, $h_{k,u}$ is the channel gain between the transmit antenna u and the receive antenna and n_k is an additive noise. We assume that $h_{k,u}$ is a complex valued Gaussian random process with unitary variance. In SISO model we consider coherent detection in which the receiver has a perfect knowledge of $h_{k,u}$.

The Alamouti scheme is implemented with 2X1 antennas. Let us consider s_{2k} and s_{2k+1} to be the two symbols to transmit at time (time and frequency axis can be permuted in multicarrier modulation.) instants 2k and 2k + 1, respectively. At time instant 2k, the antenna 0 transmits s2k/√2 whereas the antenna 1 transmits s2k+1/√2. At time instant 2k + 1, the antenna 0 transmits $-(s_{2k+1})^*/\sqrt{2}$ whereas the antenna 1 transmits s*2k/√2. The 1/√2 factor is added to normalize the total transmitted power. The received signal samples at time instants 2k and 2k + 1 are given by

$$y_{2k}=1/\sqrt{2}\,(h_{2k,0}\,s_{2k}+h_{2k,1}\,s_{2k+1})+n_{2k}, \qquad (11)$$
$$y_{2k+1}=1/\sqrt{2}(-h_{2k+1,0}(s_{2k+1})^*+h_{2k+1,1}(s_{2k})^*)+n_{2k+1}.$$

Assuming the channel to be constant between the time instants 2k and 2k + 1, we get

$$\begin{bmatrix} y_{2k} \\ y^*_{2k+1}\end{bmatrix}=1/\sqrt{2}\underbrace{\begin{bmatrix} h_{2k,0} & h_{2k,1} \\ (h_{2k,1})^* & -(h_{2k,0})^*\end{bmatrix}}_{H_{2k}}\begin{bmatrix} s_{2k} \\ s_{2k+1}\end{bmatrix}+\begin{bmatrix} n_{2k} \\ n^*_{2k+1}\end{bmatrix} \qquad (12)$$

Note that H_{2k} is an orthogonal matrix with $H_{2k}H^H_{2k}=(1/2)(|h_{2k,0}|^2+|h_{2k,1}|^2)I_2$, where I_2 is the identity matrix of size (2, 2) and H stands for the transpose conjugate operation. Thus, using the Maximum Ratio Combining (MRC) equalization, the estimates s^1_{2k} and s^1_{2k+1} are obtained as

$$\begin{bmatrix} s^1_{2k} \\ s^1_{2k+1}\end{bmatrix}=\frac{\sqrt{2}}{(|h_{2k,0}|^2+|h_{2k,1}|^2}\begin{bmatrix} h^*_{2k,0} & h_{2k,1} \\ (h_{2k,1})^* & -(h_{2k,0})^*\end{bmatrix}\begin{bmatrix} y_{2k} \\ y_{2k+1}\end{bmatrix}$$
$$=\begin{bmatrix} s_{2k} \\ s_{2k+1}\end{bmatrix}+\begin{bmatrix} \mu_{2k} \\ \mu_{2k+1}\end{bmatrix} \qquad (13)$$

where,

$$\begin{bmatrix} \mu_{2k} \\ \mu_{2k+1}\end{bmatrix}=\frac{\sqrt{2}}{(|h_{2k,0}|^2+|h_{2k,1}|^2}\begin{bmatrix} h^*_{2k,0} & h_{2k,1} \\ (h_{2k,1})^* & -(h_{2k,0})^*\end{bmatrix}\begin{bmatrix} n_{2k} \\ n^*_{2k+1}\end{bmatrix} \qquad (14)$$

Since the noise components n_{2k} and n_{2k+1} are uncorrelated, $E(|\mu_{2k}|^2)=E(|\mu_{2k+1}|^2)=2N_0/(|h_{2k,0}|^2+|h_{2k,1}|^2)$, where N_0 denotes the monolateral noise density. Thus, assuming a QPSK modulation, based on [18], the bit error probability, denoted p_b, is given by

$$p_b = Q\sqrt{(|h_{2k,0}|^2+|h_{2k,1}|^2)/2)\,SNR_t} \qquad (15)$$

where SNR= denotes the Signal-to-Noise Ratio (SNR) at the transmitter side. When the two channel coefficients are uncorrelated, we will have a diversity gain of two [18].

IV. OFDM with Alamouti Scheme.

Equation (9) indicates that we can consider the transmission of OFDM on each subcarrier as a flat fading transmission. Moreover, recalling that in OFDM each complex data symbol, $d^{(c)}_{m,n}$, is divided into two real symbols, $R\{d^{(c)}_{m,n}\}$ and $I\{d^{(c)}_{m,n}\}$, transmitted at successive time instants, transmission of a pair of data symbols, according to Alamouti scheme, is organized as follows:

$a_{m,2n,0} = R\{d^{(c)}_{m,2n}\}$, $a_{m,2n,1} = R\{d^{(c)}_{m,2n+1}\}$,
$a_{m,2n+1,0} = I\{d^{(c)}_{m,2n}\}$, $a_{m,2n+1,1} = I\{d^{(c)}_{m,2n+1}\}$,
$a_{m,2n+2,0} = -R\{d^{(c)}_{m,2n+1}*\} = -R\{d^{(c)}_{m,2n+1}\} = -a_{m,2n,1}$,
$a_{m,2n+2,1} = R\{d^{(c)}_{m,2n}*\} = R\{d^{(c)}\}_{m,2n} = a_{m,2n,0}$,
$a_{m,2n+3,0} = -I\{d^{(c)}_{,2n+1}*\} = I\{d^{(c)}\}_{m,2n+1} = a_{m,2n+1,1}$,
$a_{m,2n+3,1} = I\{d^{(c)}\}_{m,2n}*\} = -I\{d^{(c)}\}_{m,2n} = -a_{m,2n+1,0}$.

(16)

We also assume that in OFDM the channel gain is a constant between the time instants 2n and 2n + 3. Let us denote the channel gain between the transmit antenna i and the receive antenna at subcarrier m and time instant n by $h_{m,n,i}$. Therefore, at the single receive antenna we have

$y_{m,2n} = h_{m,2n,0}(a_{m,2n,0} + ja^{(i)}_{m,2n,0}\} + h_{m,2n,1}(a_{m,2n,1} + ja^{(i)}_{m,2n,1}) + n_{m,2n,0}$,
$y_{m,2n+1} = h_{m,2n,0}(a_{m,2n+1,0} + ja^{(i)}_{m,2n+1,0}) + h_{m,2n,1)}a_{m,2n+1},1 + ja^{(i)}_{m,2n+1,1}) + n_{m,2n+1,1}$,
$y_{m,2n+2} = h_{m,2n,0}(a_{m,2n+2,0} + ja^{(i)}_{m,2n+2,0}) + h_{m,2n,1}(a_{m,2n+2,1} + ja^{(i)}_{m,2n+2,1}) + n_{m,2n+2,0}$,
$y_{m,2n+3} = h_{m,2n,0}(a_{m,2n+3,0} + ja^{(i)}_{m,2n+3,0}) + h_{m,2n,1}(a_{m,2n+3,1}) + ja^{(i)}_{m,2n+3,1}) + n_{m,2n+3,1}$. (17)

Setting

$z_{m,2n} = y_{m,2n} + j\, y_{m,2n+1}$,
$z_{m,2n+1} = y_{m,2n+2} + j\, y_{m,2n+3}$, (18)

and using (16), we obtain

$z_{m,2n} = h_{m,2n,0}d^{(c)}_{m,2n} + h + d^{(c)}_{m,2n+1} + h_{m,2n,0}x_{m,2n,0} + h_{m,2n,1}x_{m,2n,1} + \kappa_{m,2n,0}$,
$z_{m,2n+1} = -h_{m,2n,0}(d^{(c)}_{m,2n+1}*) + h_{m,2n,1}(d^{(c)}_{m,2n}*) - h_{m,2n,0}(x_{m,2n+2,0}*) + h_{m,2n,1}(x_{m,2n+2,1}*) + \kappa_{m,2n+2,0}$,

(19)

where, $x_{m,2n,0} = -a^{(i)}_{m,2n+1,0} + ja^{(i)}_{m,2n,0}$,
$x_{m,2n,1} = -a^{(i)}_{m,2n+1,1} + ja^{(i)}_{m,2n,1}$,
$\kappa_{m,2n,0} = n_{m,2n,0} + jn_{m,2n+1,0}$,
$\kappa_{m,2n,0} = n_{m,2n+2,0} + jnm_{,2n+3,0}$,
$x_{m,2n+2,0} = a^{(i)}_{m,2n+3,0} + ja^{(i)}_{m,2n+2,0}$,
$x_{m,2n+2,1} = -a^{(i)}_{m,2n+3,1} - ja^{(i)}_{m,2n+2,1}$. (20)

This result in

$$\begin{bmatrix} z_{m,2n} \\ (z_{m,2n+1}*) \end{bmatrix} = \begin{bmatrix} h_{m,2n,0} & h_{m,2n,1} \\ (h_{m,2n,1}*) & -(h_{m,2n,0}*) \end{bmatrix} \begin{bmatrix} d^{(c)}_{m,2n} \\ d^{(c)}_{m,2n+1} \end{bmatrix} + \begin{bmatrix} h_{m,2n,0} & h_{m,2n,1} & 0 & 0 \\ 0 & 0 & (h_{m,2n,1}*) & -(h_{m,2n,0}*) \end{bmatrix} \begin{bmatrix} x_{m,2n,0} \\ x_{m,2n,1} \\ x_{m,2n+2,0} \\ x_{m,2n+2,1} \end{bmatrix} + \begin{bmatrix} \kappa_{m,2n+1} \\ \kappa_{m,2n+1} \end{bmatrix} \quad (21)$$

We note that Q2n is an orthogonal matrix which is similar to the one found in (12) for the conventional 2 × 1 Alamouti scheme. However, the $K_{2n}x_{2n}$ term appears which an interference term is due to the fact that OFDM has only a real orthogonality. Therefore, even without noise and assuming a distortion-free channel, we cannot achieve a good error probability since $K_{2n}x_{2n}$ is an inherent "noise interference" component that, differently from the one expressed in (9), cannot be easily removed. (In a particular case, where $h_{m,2n,0} = h_{m,2n,1}$, one can nevertheless get rid of the interference terms.)

To tackle this drawback some research studies are being carried out. However, as mentioned in the introduction, the first one [15] significantly increases the RX complexity, while the second one [16] fails to reach the objective of theoretical maximum spectral efficiency, that is, does not satisfy condition (ii). The one we propose hereafter is based on a combination of CDMA with OFDM and avoids these two shortcomings.

V. CDMA-OFDM AND ALAMOUTI

CDMA-OFDM. In this section we summarize the results obtained, assuming a distortion-free channel, in [19] and [11] for CDMA-OFDM schemes transmitting real and complex data symbols, respectively. Then, we show how this latter scheme can be used for transmission over a realistic channel model in conjunction with Alamouti coding.

Transmission of Real Data Symbols. We denote by Nc the length of the CDMA code used and assume that N_S = M/Nc is an integer number. Let us denote by cu = $[c_{0,u} \cdots c_{Nc-1,u}]^T$, where (·)T stands for the transpose operation, the code used by the uth user. When applying spreading in the frequency domain such as in pure MCCDMA (Multi-Carrier-CDMA) [20], for a user u0 at a given time n0, NS different data are transmitted denoted by: $d_{u0,n0,0}$, $d_{u0,n0,1}$, . . . , $d_{u0,n0,NS-1}$. Then by spreading with frequency of real data the c_u codes, we get the real symbol $a_{m0,n0}$ transmitted at frequency m_0 and time n_0 by

$$a_{m0,n0} = \sum_{u=0}^{U-1} c_{m0/Nc\,,u} d_{u,n0,[m_0/N_c]}, \quad (22)$$

where U is the number of users, / the modulo operator, and [ ] the floor operator. From the $a_{m0,n0}$ term, the reconstruction of $d_{u,n0,p}$ (for p € [

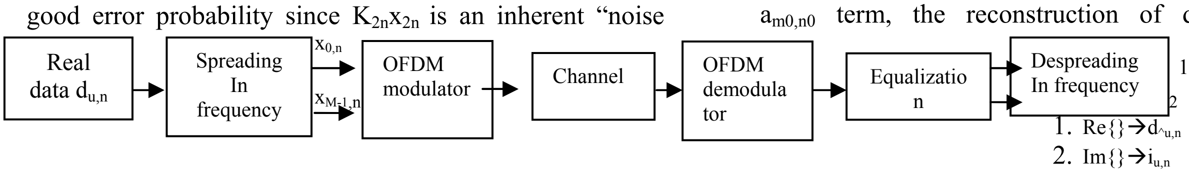

Fig. 3: Transmission scheme for the CDMA-OFDM system with spreading in frequency of real data.

$0, N_S - 1]$) is insured thanks to the orthogonality of the code, that is, $c^T_{u1}c_{u2} = \delta_{u1,u2}$;

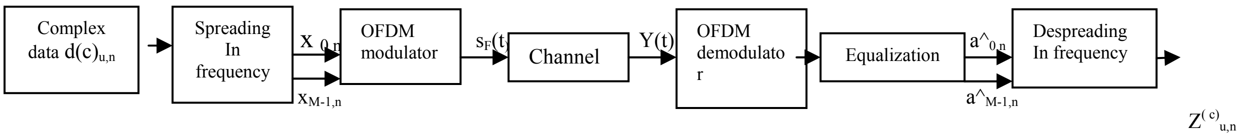

Fig. 4: Transmission scheme for the CDMA-OFDM system with spreading in frequency of complex data.

Therefore, noise taken apart, the de-spreading operator leads to

$$\hat{d}_{u,n0,p} = \sum^{Nc-1} c_{m,u} a_p N_{C+m,n0}\ . \quad (23)$$

In [19], it is shown that, since no CP is inserted, the transmission of these spread real data ($d_{u,n0,p}$) can be insured at a symbol rate which is more than twice the one used for transmitting complex MC-CDMA data. Figure 3 depicts the real CDMA-OFDM transmission scheme for real data and a maximum spreading length (limited by the number of subcarriers), where after the dispreading operation, only the real part of the symbol is kept whereas the imaginary component iu,n is not detected. This scheme satisfies a real orthogonality condition and can work for a number of users up to M.

Interference Cancellation. A closer examination of the interference term is proposed in [11] assuming that the CDMA codes are Walsh-Hadamard (W-H) codes of length $M = 2N = 2^n$, with n an integer. The prototype filter being of length $L_g = bM$, its duration is also given by the indicating function $I_{|n-n0|}<2b$, equal to 1 if $|n - n0| < 2b$ and 0 elsewhere. Then, the scalar product of the base functions can be expressed as

$$\langle g_{m,n}, g_{p,n0}\rangle = \delta_{m-p,n-n0} + j\gamma^{(p,n0)}_{m,n}\, I_{|n-n0|<2b}, \quad (24)$$

where $\gamma^{(p,n0)}_{m,n}$ is given by

$$\gamma^{(p,n0)}_{m,n} = I\{(-1)^{m(n+n0)}\, j^{m+n-p-n0} A_g(n-n0, m-p)\}. \quad (25)$$

For a maximum spreading length, that is, $M = 2N = N_c$, based on [11, Equation (18)], the interference term when transmitting real data can be expressed as

$$i_{u,n} = \sum_{u=0}^{U-1} \sum_{n=-2b+1,\, n\neq 0}^{2b-1} d_{n+n0,u}\Big(\sum_{p=0}^{2N-1}\sum_{m=0}^{2N-1} c_{p,u0} c_{m,u} \gamma^{(p,n0)}_{m,n+n0}\Big) \quad (26)$$

It is shown in [11] that if $U \le M/2$ spreading codes are properly selected then the $i_{u,n}$ interference is cancelled. The W-H matrix being of size $M = 2N = 2^n$ can be divided into two subsets of column indices, S^n_1 and S^n_2, with cardinal equal to M/2 making a partition of all the index set. To guarantee the absence of interference between users, the construction rule for theses two subsets is as follows.

For $n_0 = 1$, each subset is initialized by setting: $S^1_1 = \{0\}$ and $S^1_2 = \{1\}$.

Let us now assume that, for a given integer $n = n_0$, the two subsets contain the following list of indices:

$$S^{n0}_1 = \{i_{1,1}, i_{1,2}, i_{1,3}, \ldots, i_{1,2n0-1}\}$$
$$S^{n0}_2 = \{i_{2,1}, i_{2,2}, i_{2,3}, \ldots, i_{2,2n0-1}\} \quad (27)$$

These subsets are used to build two new subsets of identical size such that

$$S^{n0}_1 = \{i_{2,1} + 2^{n0}, i_{2,2} + 2^{n0}, i_{2,3} + 2^{n0}, \ldots, i_{2,2n0-1} + 2^{n0}\},$$
$$S^{n0}_2 = \{i_{1,1} + 2^{n0}, i_{1,2} + 2^{n0}, i_{1,3} + 2^{n0}, \ldots, i_{1,2n0-1} + 2^{n0}\} \quad (28)$$

Then, we get the subsets of higher size, $n = n_0 + 1$, as follows:

$$S^{n0+1}_1 = S^{n0}_1 \cup S^{n0}_1\ ,\ S^{n0+1}_2 = S^{n0}_2 \cup S^{n0}_2\ . \quad (29)$$

Applying this rule one can check that for n = 5, as an example, we get

$$S^5_1 = \{1,4,6,7,10,11,13,16,18,19,21,24,25,28,30,31\},$$
$$S^5_2 = \{2,3,5,8,9,12,14,15,17,20,22,23,26,27,29, 32\}. \quad (30)$$

Hence, for a given user and at a given time, we get $\hat{d}_{u,n} = d_{u,n}$ and $i_{u,n} = 0$ and these equalities hold for a number of U users up to M/2. The complete proof given in [11] takes advantage of three properties of W-H codes.

Transmission of Complex Data Symbols. As the imaginary component can be cancelled when transmitting real data through a distortion-free channel when using CDMA-OFDM, one can imagine extending this scheme to the transmission of complex data. Indeed, the transmission system being linear, real and imaginary parts will not interfere if the previous rule is satisfied.

Then, denoting by $d^{(c)}_{n,u}$ the complex data to transmit, the OFDM symbols transmitted at time $n\tau_0$ over the carrier m and for the code u are complex numbers, that is, $a^{(c)}_{m,n,u} = c_{m,u} d^{(c)}_{n,u}$ are complex symbols. The corresponding complex CDMA-OQAM transmission scheme is depicted in Figure 4. The baseband equivalent of the transmitted signal, with a spreading in frequency, can be written as

$$s_F(t) = \sum_{n\in Z}\sum_{m=0}^{2N-1} x_{m,n} g_{m,n}(t) \text{ with } x_{m,n} = \sum_{u=0}^{U-1} a^{(c)}_{m,n,u} \quad (31)$$

In this expression, as in [11], we assume that the phase term is $\nu_{n,m} = j^{n+m}(-1)^{nm}$, that is, $\varphi 0 = \pi nm$. Then, if the U codes are all in S^n_1, or S^n_2, the interference terms are cancelled and we get

For all n, u, $z^{(c)}_{n,u} = d^{(c)}_{n,u}$. (32)

This CDMA-OFDM scheme satisfies a complex orthogonality condition, that is, the back-to back transmultiplexer is a PR system for the transmission of complex data. The maximum number of users is M/2 instead of M. In both cases the overall data rate is therefore the same. In the presence of a channel, equalization must be performed before the despreading since the signal at the output of the equalization block is supposed to be free from any channel distortion or attenuation. Then, the signal at the equalizer output is somewhat equivalent to the one obtained with a distortion-free channel. Then, despreading operation will recover the complex orthogonality.

Alamouti with CDMA-OFDM with Spreading in the Frequency Domain. In a realistic transmission scheme the channel is no longer distortion-free. So, we assume now that we are in the case of a wireless Down-Link (DL) transmission and perfectly synchronized.

VI. PROBLEM STATEMENT

Before trying to apply Alamouti scheme to CDMA-OFDM, it is noted that the channel equalization process is replaced by the Alamouti decoding. When adapting Alamouti scheme to CDMA-OFDM, the equalizer component, depicted in Figure 4, must be replaced by the Alamouti decoding process and the despreading operation must be carried out just after the OFDM modulator. Then, contrary to the DL conventional MC-CDMA case, the dispreading operation must be performed before the Alamouti decoding. Indeed, with OFDM, we can only recover a complex orthogonality property at the output of the dispreading block. The point complex orthogonality hold in CDMA-OFDM if we perform despreading operation before equalization leads to the following problem: let us consider complex quantities t_i, β_i, λ_i. Does it sound possible to obtain $\sum_{i=0}^{M-1} \beta i(ti/\lambda i)$ (equalization + despreading) $\sum_{i=0}^{M-1} \beta i(ti)$ (despreading). Here, equalization is materialized by $e_i = t_i/\lambda_i$ and the despreading operation by $\sum_{i=0}^{M-1} \beta i(e_i)$. If all the λi are the same, that is, $\lambda i = \lambda j = \lambda$. This is the case if we are in the presence of a constant channel over frequencies. Indeed, only in this case the order of the equalization and despreading operations can be exchanged without impairing the transmission performance. Conversely, applying despreading before equalization should have an impact in terms of performance for a channel being nonconstant in frequency. So, let us consider at first a flat channel. Then the subset of subcarriers where a given spreading code is applied will be affected by the same channel coefficient.

VII. IMPLEMENTATION SCHEME

In a SISO configuration, if we denote by hn,i the single channel coefficient between the transmit antenna i and the single receive antenna at time instant n, the despreaded signal is given by:

$$z^{(c)}_{n0,u0} = h_{n0,i}\, d^{(c)}_{n0,u0,i}, \quad (33)$$

where $d^{(c)}_{n0,u0,i}$ is the complex data of user u_0 being transmitted at time instant n0 by antenna i. Now, if we consider a system

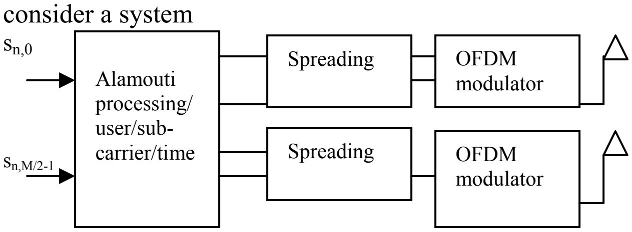

Fig. 5: An Alamouti CDMA-OFDM transmitter.

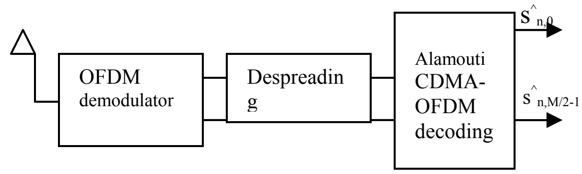

Fig. 6: An Alamouti CDMA-OFDM receiver.

with 2 antennas with indexes 0 and 1, respectively, and if we apply Alamouti coding scheme to every user u data, denoting by sk,uthe main stream of complex data for user u, we have

at time 2k, $d^{(c)}_{2k,u,0} = s_{2k,u}/\sqrt{2}$, $d^{(c)}_{2k,u,1} = s_{2k+1,u}/\sqrt{2}$;
at time 2k + 1, $d^{(c)}_{2k+1,u,0} = -(s_{2k+1,u})^*/\sqrt{2}$,
$d^{(c)}_{2k+1,u,1} = s_{2k}^*/\sqrt{2}$ (34)

For a flat fading channel, ignoring noise, the despreaded signal for user u is given by

$$z^{(c)}_{n,u} = h_{n,0} d^{(c)}_{n,u,0} + h_{n,1} d^{(c)}_{n,u,1}. \quad (35)$$

Hence,

$$\begin{bmatrix} z^{(c)}_{2k,u} \\ (z^{(c)}_{2k+1,u,})^* \end{bmatrix} = 1/\sqrt{2} \begin{bmatrix} h_{2k,0} & h_{2k,1} \\ (h_{2k+1,1})^* & -(h_{2k+1,0})^* \end{bmatrix} \begin{bmatrix} s_{2k,u} \\ s_{2k+1,u} \end{bmatrix} \quad (36)$$

This is the same decoding equation as in the Alamouti scheme presented in Section 2.2. Hence, the decoding could be performed in the same way. Figures 5 and 6 present the Alamouti CDMA-OFDM transmitter and receiver respectively.

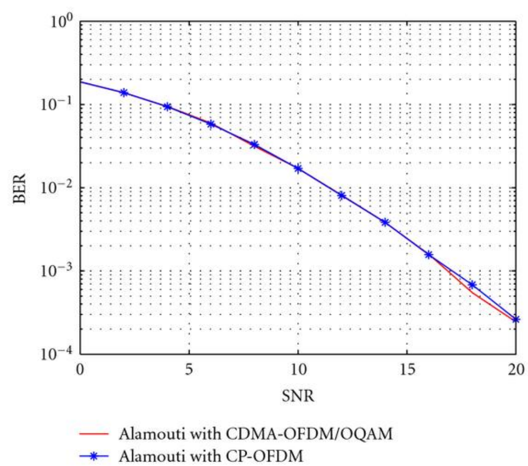

Fig. 7: BER for the complex version of the Alamouti CDMA OFDM with spreading in frequency domain, versus Alamouti CP-OFDM for transmission over a flat fading channel.

VIII. PERFORMANCE EVALUATION

We compare the proposed Alamouti CDMA-OFDM scheme with the Alamouti OFDM using the following parameters:

(i) QPSK modulation

(ii) M = 128 subcarriers

(iii) maximum spreading length, implying that the W-H spreading codes are of length Nc = 128,

(iv)flat fading channel (one single Rayleigh coefficient for all 128 subcarriers);

(v) the IOTA prototype filter with length 512,

(vi) zero forcing one tap equalization for both transmission schemes,

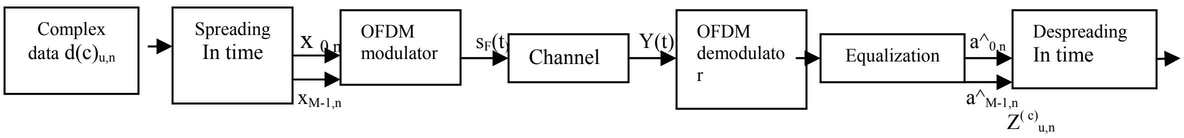

Fig. 8: Transmission scheme for the CDMA-OFDM system with spreading in time of complex data.

(vii) no channel coding.

Figure 7 gives the performance results. As expected, both systems perform the same.

Alamouti and CDMA-OFDM with Time Domain Spreading

In this section, we keep the same assumptions as the ones used for the transmission of complex data with a spreading in frequency. Firstly, we again suppose that the prototype function is a real-valued symmetric function and also that the W-H codes are selected using the procedure recalled in Section 3.1.2.

CDMA-OFDM with Spreading in the Time Domain. Let us first consider a CDMA-OFDM system, carrying out a spreading in the time domain, that is, on each subcarrier m the data are spread over the time duration frame length. Let us consider Lf the length of the frame, that is, the frame is made of M data in the frequency domain and Lf data in time domain. N_c is the length of the spreading code. We assume that $N_S = L_f / N_c$ is an integer number. Let us denote by: $c_u = [c_{0,u} \cdots c_{Nc-1,u}]^T$ the code used by the uth user. Then, for a user u0 at a given frequency m0, NS different data are transmitted denoted by: $d_{u0,m0,0}$, $d_{u0,m0,1}$, . . . , $d_{u0,m0,NS-1}$. By spreading with the c_u codes, we get the real symbol $a_{m0,n0}$ transmitted at frequency m_l and time n_0 by

$$a_{m0,n0} = \sum_{u=0}^{U-1} c_{n0/N_c,u} d_{u,m0,[n0/Nc]} \quad (37)$$

where U is the number of users. From the $a_{m0,n0}$ term, the reconstruction of du,m0,p (for p€ $[0,N_S - 1]$) is insured thanks to the orthogonality of the code, that is, $c^T_{u1}c_{u2} = \delta_{u1,u2}$, see [21] for more details. Therefore, the dispreading operator leads to

$$d^\wedge_{u,m0,p} = \sum_{n=0}^{Nc-1} c_{n,u} a_{m0,pNc+n}. \quad (38)$$

We now propose to consider the transmission of complex data, denoted $d^{(c)}_{m,u,p}$, using U well chosen W-H codes. In order to establish the theoretical features of this complex CDMA-OFDM scheme, we suppose that the transmission channel is free of any type of distortion. Also, for the sake of simplicity, we now assume a maximum spreading length (in time domain, $L_f = N_c$). We denote by $d^{(c)}_{m,u}$ the complex data and by $a^{(c)}_{m,n,u} = c_{n,u} d^{(c)}_{m,u}$ the complex symbol transmitted at time $n\tau_0$ over the carrier m and for the code u. As usual, the length of the W-H codes are supposed to be a power of 2, that is, Lf = 2L = 2q with q an integer.

The block diagram of the transmitter is depicted in Figure 8. For a frame containing 2L OFDM data symbols, the baseband signal spread in time, can be written as

$$s^T(t) = \sum_{n=0}^{2L-1} \sum_{m=0}^{2N-1} x_{m,n} g_{m,n}(t)$$

$$\text{with } x_{m,n} = \sum_{u=0}^{U-1} a^{(c)}_{m,n,u} = \sum_{u=0}^{U-1} c_{n,u} d^{(c)}_{m,u}. \quad (39)$$

In (39), we assume that the phase term is $\nu_{m,n} = j^{m+n}$ as in [7]. Let us also recall that the prototype function g satisfies the real orthogonality condition (2) and is real-valued and symmetric, that is, g(t) = g(−t). To express the complex inner product of the base functions $g_{m,n}$, using a similar procedure that led to (24), we get

$$<g_{m,n}, g_{p,n0}> = \delta_{m-p,n-n0} + j\lambda^{(p,n0)}_{m,n} I_{|n-n0|<2b}, \quad (40)$$

where $\lambda^{(p,n0)}_{m,n}$ is given by

$$\lambda^{(p,n0)}_{m,n} = I\{(-1)^{n(p+m)} j^{m+n-p-n0} Ag(n - n0, m - p)\} \quad (41)$$

As the channel is distortion-free, the received signal is y(t) = s(t) and the demodulated symbols are obtained as follows: $y^{(c)m0,n0} = <y, g_{m0,n0}>$. (42)

In this configuration, the demodulation operation only takes place when the whole frame is received. Then, the despreading operation gives us the despreaded data for the code u_0 as

$$z^{(c)}_{m0,u0} = \sum_{q=0}^{2L-1} c_{q,u_0} y^{(c)}_{m0,q} = \sum_{q=0}^{2L-1} c_{q,u} \sum_{n=0}^{2L-1} \sum_{m=0}^{2N-1} x_{m,n} <g_{m,n}, g_{m0,q}> \quad (43)$$

Replacing $x_{m,n}$ and $<g_{m,n}, g_{m0,q}>$ by their expression given in (39) and (40), respectively, we get:

$$z^{(c)}_{m0,u0} = \sum_{q=0}^{2L-1} c_{q,u_0}$$

$$= \sum_{n=0}^{2L-1} \sum_{m=0}^{2L-1} \sum_{u=0}^{2N-1} c_{n,u} d^{(c)}_{m,u} (\delta_{m-m0,n-q} + j\lambda^{(m0,q)}_{m,n}) \quad (44)$$

Splitting the summation over m in two parts, with m equal to m_0 or not to m_0, (44) can be rewritten as:

$$z^{(c)}_{m0,u0} = \sum_{u=0}^{U-1} d^{(c)}_{m0,u} \sum_{q=0}^{2L-1} c_{p,u0} c_{p,u} +$$

$$J \left(\sum_{u=0}^{U-1} \sum_{m=0,m\neq m0}^{2N-1} d^{(c)}_{m,u} \left(\sum_{q=0}^{2L-1} \sum_{n=0}^{2L-1} c_{q,u0} c_{n,u} \lambda^{(m0,q)}_{m,n} \right. \right. \quad (45)$$

Considering theW-H codes, we obtain

$$z^{(c)}_{m0,u0} = d^{(c)}_{m0,u} + J \left(\sum_{u=0}^{U-1} \sum_{m=0,m\neq m0}^{2N-1} d^{(c)}_{m,u} \left(\sum_{q=0}^{2L-1} \sum_{n=\square 0}^{2L-1} c_{q,u0}\, c_{n,u} \lambda^{(m0,q)}_{m,n} \right.\right. \quad (46)$$

In [11], for W-H codes of length 2L, we have shown that for n=/ n0,

$$\sum_{p=0}^{2L-1} \sum_{m=0,m\neq m0}^{2L-1} c_{p,u0}\, c_{m,u} \gamma^{(p,n0)}_{m,n+n0} = 0, \quad q=0 \;\; n=\square 0 \quad (47)$$

where $\gamma^{(p,n0)}_{m,n}$ is given by

$$\gamma^{(p,n0)}_{m,n} = I\{(-1)^{m(n+n0)} j^{m+n-p-n0} Ag(n-n_0, m-p. \quad (48)$$

To prove the result given in (47), we had the following requirements:

(i) W-H codes satisfy the set of mathematical properties that are proved in [11].

(ii) Since g is a real-valued function, A_g (n, 0) is real valued and the ambiguity function of the prototype function g also satisfies the identities $A_g(-n,m)=(-1)^{nm}A_g(n,m)$ and $A_g(n,m)=A^*g\ (n,-m)$.

Using these results, (47) can be proved straight forwardly. It is worth mentioning that the above requirements are independent of the phase term and thus are satisfied in the case of the CDMA-OFDM system with spreading in time. It can also be shown that the modification of the phase term νm,n leads to the substitutions n → m and p + m →n + n0, in obtaining (48) from (41). Accordingly the second term on the right hand side of (46) vanishes and we obtain for all

$$m_0, u_0, z^{(c)}_{m0,u0} = d^{(c)}_{m0,u0} . \quad (49)$$

Alamouti with CDMA-OFDM with Spreading in Time. Now, if we consider the CDMA-OFDM with spreading in time, contrary to the case of a spreading in frequency domain, as long as the channel is constant during the spreading time duration, we can perform dispreading before equalization. At the equalizer output we will have a complex orthogonality. Indeed, considering at first a SISO case, if we denote by $h_{m,i}$ the channel coefficient between a single transmit antenna i and the receive antenna at subcarrier m, the despreaded signal is given by

$$z_{(c)m,u} = h_{m,i} d^{(c)}_{m,u,i,} \quad (50)$$

where $d^{(c)}_{m,u,i,}$is the complex data of user u being transmitted at subcarrier m by antenna i. Thus, we can easily apply the Alamouti decoding scheme knowing the channel is constant for each antenna at each frequency. Otherwise said, the method becomes applicable for a frequency selective channel. Actually two strategies can be envisioned. (1) Strategy 1. Alamouti performed over pairs of frequencies. If we consider a system with 2 transmit antennas, 0 and 1, and

if we apply the Alamouti coding scheme to every user u data, that is, if we denote by sm,u the main stream of complex data for user u, then we have the following at subcarrier 2m:

$$d^{(c)}_{2m,u,0} = s_{2m,u} / \sqrt{2}\ ;\ d^{(c)}_{2m,u,1} = s_{2m+1,u} / \sqrt{2} \quad (51)$$

and at subcarrier 2m + 1, $d^{(c)}_{2m+1,u,0} = -(s_{2m+1,u})^*/\sqrt{2}$; $d^{(c)}_{2m+1,u,} = s^*_2/\sqrt{2}$(52)

Then, considering a flat fading channel, the despreaded signal for user u is given by

$$z^{(c)}_{m,u} = h_{m,0} d^{(c)}_{m,u,0} + h_{m,1} d^{(c)}_{m,u,1} \quad (52)$$

$z^{(c)}_{2m,u}$ $h_{2m,0}$ $h_{2m,1}$ $s_{2m,u}$ $= 1/\sqrt{2}$

That means, when assuming the channel to be flat over two consecutive subcarriers, that is, $h_{2m,i} = h_{2m+1,i}$ for all i, we have exactly the same decoding equation as the Alamouti scheme presented in Section 3.2, by permuting the frequency and time axis. Then, the decoding is performed in the same way. (2) Strategy 2. Alamouti performed over pairs of spreading codes. In this second strategy, we apply the Alamouti scheme on pairs of codes, that is, we divide the U codes in two groups (assuming U to be even). That is, we process the codes by pair (u0, u1). We denote by $s_{m,u0,u1}$ the main stream of complex data for user pair (u0, u1). At subcarrier m, antennas 0 and 1 transmit

$$d^{(c)}_{m,u0,0} = s_{m,u0,u1}/\sqrt{2}\ ;\ d^{(c)}_{m,u0,1} = s_{m+1,u0,u1}/\sqrt{2};$$
$$d_{(c)m,u1,0} = -(s_{m+1,u0,u1})^* / \sqrt{2};$$
$$d^{(c)}_{m,u1,1} = s^*_{m,u0,u1}/\sqrt{2} \quad (5)$$

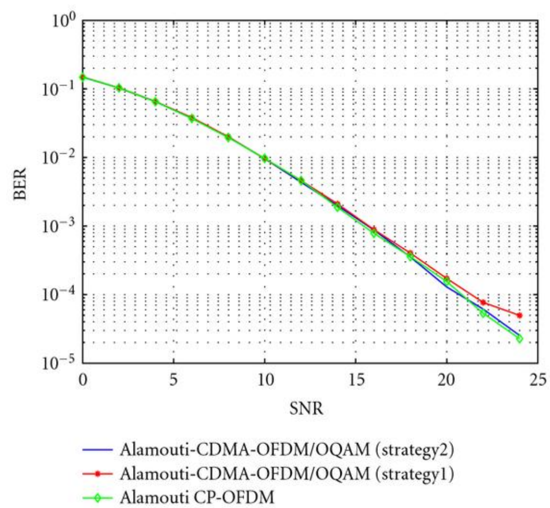

Fig. 9: BER for two complex versions of the Alamouti CDMAOFDM/ OQAM with spreading in time domain, versus Alamouti CP-OFDM for transmission over the 4-path frequency selective channel.

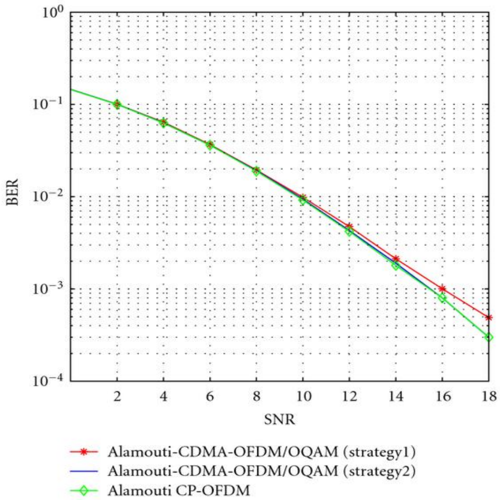

Fig. 10: BER for two complex versions of the Alamouti CDMAOFDM with spreading in time domain, versus Alamouti CP-OFDM for transmission over the 7-path frequency selective channel.

Then, we do not need to consider the channel constant over two consecutive subcarriers. We have exactly the same decoding equation as the Alamouti scheme presented in Section 3.2. Hence, the decoding is performed in the same way.

We have tested two different channels considering each time the same channel profile, but with different realizations, between the 2 transmit antennas and one receive antenna.

The Guard Interval (GI) is adjusted to take into account the delay spread profiles corresponding to a 4-path and to a 7-path channel. The 4-path channel is characterized by the following parameters:

(i) power profile (in dB): 0, −6, −9, −12,
(ii) delay profile (in samples): 0, 1, 2, 3,
(iii) GI for CP-OFDM: 5 samples,

and the 7-path by

(i) power profile (in dB): 0, −6, −9, −12, −16, −20, −22,
(ii) delay profile (in samples): 0, 1, 2, 3, 5, 7, 8,
(iii) GI for CP-OFDM: 9 samples;

We also consider the following system parameters:

(i) QPSK modulation,
(ii) M = 128 subcarriers,
(iii) time invariant channel (no Doppler),
(iv) the IOTA prototype filter of length 512,
(v) spreading codes of length 32, corresponding to the frame duration (32 complex OQAM symbols),
(vi) number of CDMA W-H codes equals to 16 in complex OFDM, with symbol duration $\tau 0$ and this corresponds to 32 codes in OFDM, with symbol duration $2\tau 0$, leading to the same spectral efficiency
(vii) zero forcing, one tap equalization,
(viii) no channel coding.

In Figures 9 and 10, the BER results of the Alamouti CDMA-OFDM technique for the two proposed strategies are presented.

The two strategies perform the same until a BER of 10^{-3} or 10^{-2} for the 4 and 7-path channel, respectively.

For lower BER the strategy 2 performs better than the strategy 1. This could be explained by the fact that strategy 1 makes the approximation that the channel is constant over two consecutive subcarriers. This approximation leads to a degradation of the performance whereas the strategy 2 does not consider this approximation. If we compare the performance of Alamouti CDMA-OFDM strategy 2 with the Alamouti CP-OFDM, we see that both system perform approximately the same. It is worth mentioning that however the corresponding throughput is higher for the OFDM solutions (no CP). Indeed, it is increased by approximately 4 and 7% for the 4 and 7-path channels, respectively.

IX. CONCLUSION

From the analysis carried out in this paper the following can be concluded that the well-known Alamouti decoding scheme cannot be directly applied to the OFDM modulation. A combination of the MIMO Alamouti coding scheme with CDMA-OFDM is more suitable.

If the CDMA spreading is carried out in the frequency domain, the Alamouti decoding scheme can only be applied if the channel is assumed to be flat. For a frequency selective channel, the CDMA spreading component has to be applied in the time domain.

For the Alamouti scheme with time spreading CDMA-OFDM, two strategies are suggested for implementing the MIMO space-time coding scheme. Strategy 1 implements the Alamouti over pairs of adjacent frequency domain samples whereas the strategy 2processes the Alamouti coding scheme over pairs of spreading codes from two successive time instants. Strategy 2 appears to be more appropriate since it requires less restrictive assumptions on the channel variations across the frequencies.

The performance comparisons with Alamouti CP-OFDM show that under the channel hypothesis, the combination of Alamouti with complex CDMA-OFDM is possible without increasing the complexity of the Alamouti decoding process. In the case of a frequency selective channel, OFDM keeps its intrinsic advantage with a SNR gain in direct relation with the CP length.

To find a simpler Alamouti scheme, that is, without adding a CDMA component, remains an open problem. Naturally, some other alternative transmit diversity schemes for OFDM, as for instance cyclic delay diversity deserve further investigations.

X. REFERENCES

[1] S. M. Alamouti, “A simple transmit diversity technique for wireless communications,” IEEE Journal on Selected Areas in Communications, vol. 16, no. 8, pp. 1451–1458, 1998.

[2] H. B¨olcskei, “Orthogonal frequency division multiplexing based on offset QAM,” in Advances in Gabor Analysis, Birkh¨auser, Boston, Mass, USA, 2003.

[3] G. Cherubini, E. Eleftheriou, and S. O¨ lc¸er, “Filtered multitone modulation for very high-speed digital subscriber lines,” IEEE Journal on Selected Areas in Communications, vol. 20, no. 5, pp. 1016–1028, 2002.

[4] R. Hleiss, P. Duhamel, and M. Charbit, Oversampled OFDM systems,” in Proceedings of the International Conference on Digital Signal Processing (DSP ’97), vol. 1, pp. 329–331, Santorini, Greece, July 1997.

[5] B. R. Saltzberg, “Performance of an efficient parallel data transmission system,” IEEE Transactions on Communication Technology, vol. 15, no. 6, pp. 805–811, 1967.

[6] B. Hirosaki, “An orthogonally multiplexed QAM system using the discrete Fourier transform,” IEEE Transactions on Communications Systems, vol. 29, no.7, pp. 982–989, 1981.

[7] B. Le Floch, M. Alard, and C. Berrou, “Coded orthogonal frequency division multiplex,” Proceedings of the IEEE, vol. 83, pp. 982–996, 1995.

[8] T. Karp and N. J. Fliege, “MDFT filter banks with perfect reconstruction,” in Proceedings of IEEE International Symposium on Circuits and Systems (ISCAS ’95), vol. 1, pp. 744–747, Seattle,Wash, USA, May 1995.

[9] P. Siohan, C. Siclet, and N. Lacaille, “Analysis and design of OFDM systems based on filterbank theory,” IEEE Transactions on Signal Processing, vol. 50, no. 5, pp. 1170–1183, 2002.

[10] B. Farhang-Boroujeny and R. Kempter, “Multicarrier communication techniques for spectrum sensing and communication in cognitive radios,” IEEE Communications Magazine, vol. 46, no. 4, pp. 80–85, 2008.

[11] C. L´el´e, P. Siohan, R. Legouable, and M. Bellanger, “CDMA transmission with complex OFDM,” EURASIP Journal on Wireless Communications and Networking, vol. 2008, Article ID 748063, 12 pages, 2008.

[12] M. El Tabach, J.-P. Javaudin, and M. H´elard, “Spatial data multiplexing over OFDM modulations,” in Proceedings of the IEEE International Conference on Communications (ICC ’07), pp. 4201–4206, Glasgow, Scotland, June 2007.

[13] J.-P. Javaudin, D. Lacroix, and A. Rouxel, “Pilot-aided channel estimation for OFDM,” in Proceedings of the IEEE Vehicular Technology Conference (VTC ’03), vol. 3, pp. 1581– 1585, Jeju, South Korea, April 2003.

[14] C. L´el´e, J.-P. Javaudin, R. Legouable, A. Skrzypczak, and P Siohan, “Channel estimation methods for preamble-based OFDM modulations,” in Proceedings of the 13th European Wireless Conference (EW ’07), Paris, France, April 2007.

[15] M. Bellanger, “Transmit diversity in multicarrier transmission using OQAM modulation,” in Proceedings of the 3rd International Symposium on Wireless Pervasive Computing (ISWPC ’08), pp. 727–730, Santorini, Greece, May 2008.

[16] H. Lin, C. L´el´e, and P. Siohan, “A pseudo alamouti transceiver design for OFDM modulation with cyclic prefix,” in Proceedings of the IEEE Workshop on Signal Processing Advances in Wireless Communications (SPAWC ’09), pp. 300– 304, Perugia, Italy, June 2009.

[17] D. Lacroix-Penther and J.-P. Javaudin, “A new channel estimation method for OFDM,” in Proceedings of the 7th International OFDM Workshop (InOWo ’02), Hamburg, Germany, September 2002.

[18] D. Tse and P. Viswanath, Fundamentals of Wireless Communication, Cambridge University Press, Cambridge, Mass, USA, 2004.

[19] C. L´el´e, P. Siohan, R. Legouable, and M. Bellanger, “OFDM for spread spectrum transmission,” in Proceedings of the International Workshop on Multi-Carrier Spread-Spectrum (MCSS ’07), Herrsching, Germany, May 2007.

[20] K. Fazel and L. Papke, “On the performance of convolutionnally- coded CDMA/OFDM for mobile communication EURASIP Journal on Advances in Signal Processing 13 system,” in Proceedings of the Symposium on Personnal, Indoor and Mobile Radio Communications (PIMRC ’93), pp. 468–472, Yokohama, Japan, September 1993.

[21] E. H. Dinan and B. Jabbari, “Spreading codes for direct sequence CDMA and wideband CDMA cellular networks,” IEEE Communications Magazine, vol. 36, no. 9, pp. 48–54