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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

1. 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 afavorable 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 appendingof 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 proposedin 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 theAlamouti 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.

2. OFDM AND ALAMOUTI

The OFDMTransmultiplexer:The basebandequivalent of a continuous-timemulticarrier OFDMsignal can be expressed as follows [7]:

$$s(t) = \sum_{m=0}^{M-1} \sum_{n \in Z} a_{m,n} g(t - n\tau_0) e^{j2\pi m F_0 t} \nu_{m,n} \quad (1)$$

where$g_{m,n}(t) = g(t - n\tau_0) e^{j2\pi m F_0 t} \nu_{m,n}$with **Z** the set of integers, M = 2N an even umber ofsubcarriers, $F_0 = 1/T_0 = 1/2\tau_0$ the subcarrier spacing, gthe prototype function assumed here to be a real-valued andeven function of time, and $\nu_{m,n}$an additional phase termsuchthat $\nu_{m,n} = j^{m+n} e^{j\varphi 0}$, where φ0 can be chosen arbitrarily.The transmitted data symbols $a_{m,n}$are real-valued. Theyare obtained from a 22K-QAM constellation, taking the realand imaginary parts of these complex-valued symbols ofduration $T_0 = 2\tau_0$, where τ_0 denotes the time offset betweenthe two parts [2, 6, 7, 9].

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

$$R\{\langle g_{m,n}, g_{p,q}\rangle\} = R\{\int g_{m,n}(t) g^*_{p,q}(t) dt\} = \delta_{m,p}\delta_{n,q} \quad (2)$$

Where∗denotes conjugation, <·, ·>denotes the innerproduct, and $\delta_{m,p} = 1$ if m = p and $\delta_{m,p} = 0$ if$m \neq p$. Otherwise said, for $(m,n) \neq (p, q)$, $\langle g_{m,n}, g_{p,q}\rangle$is apure imaginary number. For the sake of brevity, we set$\langle g\rangle^{p,q}_{m,n} = -j\langle g_{m,n}, g_{p,q}\rangle$. The orthogonality condition for theprototype filter can also be conveniently expressed using itsambiguity function

$$A_g(n,m) = \int_{-\infty}^{\infty} g(u - n\tau_0)\, g(u) e^{2j\pi m F_0 u}\, du. \quad (3)$$

It is well-known [7] that to satisfy the orthogonalitycondition (2), the prototype filter should be chosen such that $A_g(2n, 2m) = 0$ if $(n,m) \neq (0, 0)$ and $A_g(0,0) = 1$.

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

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

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

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 prototypefilter length, denoted L_g, is such that $L_g = bM = 2bN$, with bbeing a positive integer. With the discrete time formulation,the real orthogonality condition can also be expressed as:

$$R\langle g_{m,n}, g_{p,q}\rangle\} = R\{\sum g_{m,n}[k]) g^*_{p,q}[k] dt\} = \delta_{m,p}\delta_{n,q} k \in Z \quad (5)$$

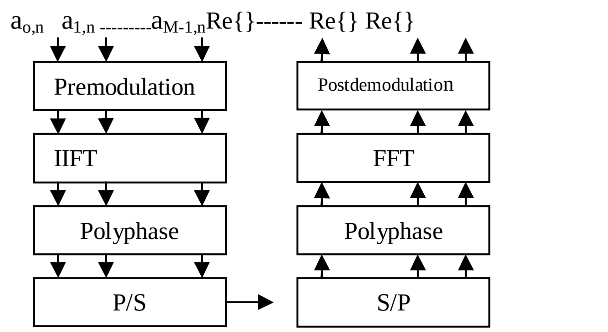


Figure.1: Transmultiplexer scheme for the OFDM modulation.

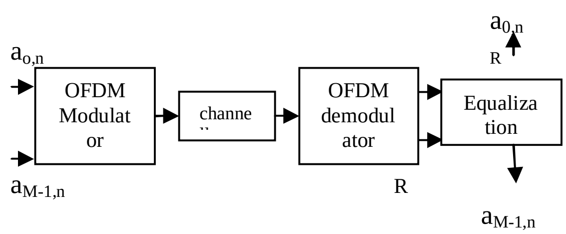


Figure.2: The transmission scheme based on OFDM.

As shown in [9],a simplified description is provided in Figure 1, where it hasto be noted that the pre-modulation corresponds to a single multiplicationby an exponential whose argument dependson the phase term $\nu_{m,n}$and on the prototype length. Notealso that in this scheme, to transmit OFDM symbols of a givenduration, denoted T_0, the IFFT block has to be run twicefaster than for CP-OFDM. The polyphase block contains thepolyphase components of the prototype filter g. At the RXside, the dual operations are carried out.

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

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

In Figure. 2, OFDM transmission scheme, it is compared toa channel that breaks the real orthogonalitycondition thus equalization must be performed at thereceiver side to restore this orthogonality.

Let us consider a time-varying channel, with maximumdelay 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 receiverside, the received signal is the summation of the s(t) signalconvolved with the channel impulse response and a noisecomponent η(t). For a locally invariant channel, we can define neighborhood, denoted $\Omega_{\Delta m,\Delta n}$, around (m_0,n_0)position, with

$\Omega_{\Delta m,\Delta n}=\{(p,q)|p|\leq\Delta m,|q|\leq\Delta n|H^{(c)}_{m0+p,n0+q}\approx H^{(c)}_{m0,n0}$ (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 thetime and bandwidth coherence of the channel, respectively.Then, assuming$g(t-\tau-n\tau_0)\approx g(t-n\tau_0)$, for all τϵ[0,Δ],the demodulated signal can be expressed as [13, 14, 17]

$y(c)_{m0,n0}=H(c)_{m0,n0}(a_{m0,n0}+ja^{(i)}_{m0,n0)})+J_{m0,n0}+\eta_{m0,n0}$ (7) with $\eta_{m0,n0}=<\eta,g_{m0,n0}>$the noise component, $a^{(i)}_{m0,n0}$, theinterference created by the neighbor symbols, given by $a^{(i)}_{m0,n0}=\sum a_{m0+p,n0+q}<g>^{m0,n0}$

$(p,q)\ \epsilon\Omega*_{\Delta m,\Delta n m0+p,n0+q}$, (8)and $J_{m0,n0}$ the interference created by the data symbolsoutside $\Omega_{\Delta m,\Delta n}$.

It can be shown that, even for small size neighborhoods,if the prototype function g is well localized in time andfrequency, $J_{m0,n0}$ becomes negligible when compared to thenoise term $\eta_{m0,n0}$. Indeed a good time-frequency localization[7] means that the ambiguity function of g, which is directlyrelated to the $<g>^{m_0,n_0}_{m0+p,n0+q}$ terms, is concentrated around itsorigin in the time-frequency plane, that is, only takes smallvalues outside the $\Omega_{\Delta m,\Delta n}$region. Thus, the received signalcan 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 theexpression of the signal at the output of the OFDMdemodulator.

3. ALAMOUTI SCHEME

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

$y_k = h_{k,u}s_{k,u}+ n_k$, (10)where, at time instant k, $h_{k,u}$is the channel gain betweenthe transmit antenna u and the receive antenna and $n_k$is an additive noise. We assume that $h_{k,u}$is a complexvaluedGaussian 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} tobe the two symbols to transmit at time (time and frequencyaxis can be permuted in multicarrier modulation.) instants2k and 2k + 1, respectively. At time instant 2k, the antenna 0transmits 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 isadded to normalize the total transmitted power. The receivedsignal samples at time instants 2k and 2k + 1 are given by

$y_{2k}=1/\sqrt{2}\,(h_{2k,0}s_{2k}+h_{2k,1}s_{2k+1})+n_{2k}$, (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 timeinstants 2k and 2k + 1, we get

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

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 matrixof size (2, 2) and H stands for the transpose conjugateoperation. Thus, using the Maximum Ratio Combining(MRC) equalization, the estimatess^1_{2k}ands^1_{2k+1} are obtainedas

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

$$= \begin{bmatrix} s_{2k} \\ s_{2k+1} \end{bmatrix}+\begin{bmatrix} \mu_{2k} \\ \mu_{2k+1} \end{bmatrix} \quad (13)$$

where,

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

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 aQPSK modulation, based on [18], the bit error probability,denoted p_b, is given by

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

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

4. OFDM WITH ALAMOUTI SCHEME.

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

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

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

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

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

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

$a_{m,2n+3,1} = I\{d^{(c)}\}_{m,2n}{}^*\} = -I\{d^{(c)}\}_{m,2n} = -a_{m,2n+1,0}.$ (16)

We also assume that in OFDM the channel gainis a constant between the time instants 2n and 2n + 3. Letus denote the channel gain between the transmit antenna i and the receive

$y_{m,2n+2}=h_{m,2n,0}(a_{m,2n+2,0}+ja^{(i)}_{m,2n+2,0})+h_{m,2n,1}(a_{m,2n+2,1}+ja^{(i)}_{m,2n+2,1})+ n_{m,2n+2,0},$
$y_{m,2n+3}=h_{m,2n,0}(a_{m,2n+3,0}+ja^{(i)}_{m,2n+3,0})+h_{m,2n,1}(a_{m,2n+3,1})+ ja^{(i)}_{m,2n+3,1})+ n_{m,2n+3,1}.$ (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 issimilar to the one found in (12) for the conventional 2 × 1Alamouti 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 noiseand assuming a distortion-free channel, we cannot achievea good error probability since $K_{2n}x_{2n}$ is an inherent “noiseinterference” component that, differently from the

antenna at subcarrier m and time instant nby $h_{m,n,i}$. Therefore, at the single receive antenna we have

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

oneexpressed in (9), cannot be easily removed. (In a particularcase, where $h_{m,2n,0}= h_{m,2n,1}$, one can nevertheless get rid ofthe interference terms.)

To tackle this drawback some research studies are beingcarried 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 oftheoretical maximum spectral efficiency, that is, does notsatisfy condition (ii). The one we propose hereafter is basedon a combination of CDMA with OFDM and avoidsthese two shortcomings.

5. CDMA-OFDM AND ALAMOUTI

CDMA-OFDM. In this section we summarizethe results obtained, assuming a distortion-free channel, in [19] and [11] for CDMA-OFDM schemes transmittingreal and complex data symbols, respectively. Then, weshow how this latter scheme can be used for transmissionover a realistic channel model in conjunction with Alamouticoding.

Transmission of Real Data Symbols. We denote byNc the length of the CDMA code used and assume thatN_S = M/Nc is an integer number. Let us denote by cu =$[c_{0,u} \cdot \cdot \cdot c_{Nc-1,u}]^T$, where $(\cdot)T$ stands for the transposeoperation, the code used by the uth user. When applyingspreading in the frequency domain such as in pure MCCDMA(Multi-Carrier-CDMA) [20], for a user u0 at agiven time n0, NS different data are transmitted denotedby: $d_{u0,n0,0}$, $d_{u0,n0,1}$, . . . , $d_{u0,n0,NS-1}$. Then by spreading withfrequency of real data the $c_u$codes, we get the real symbol $a_{m0,n0}$ transmitted atfrequency 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, thereconstruction of $d_{u,n0,p}$(for p €[

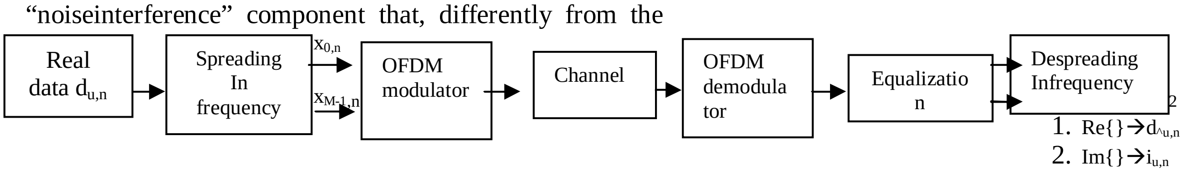
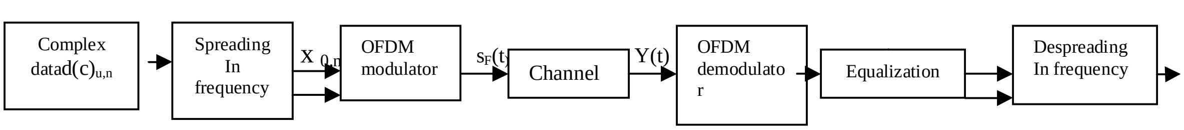


Figure. 3: Transmission scheme for the CDMA-OFDM system with spreading in frequency of real data.

0,N_S − 1]) is insuredthanks to the orthogonality of the code, that is, $c^T_{u1cu2}=\delta_{u1,u2}$;



$x_{M-1,n}$ $\hat{a}_{M-1,n}$

$Z^{(c)}_{u,n}$

Figure. 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, thetransmission of these spread real data ($d_{u,n0,p}$) can be insuredat a symbol rate which is more than twice the one usedfor transmitting complex MC-CDMA data. Figure 3 depictsthe realCDMA-OFDM transmission scheme forreal data and a maximum spreading length (limited bythe number of subcarriers), where after the dispreading operation, only the real part of the symbol is kept whereasthe imaginarycomponent 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 theinterference term is proposed in [11] assuming that theCDMA codes are Walsh-Hadamard (W-H) codes of lengthM = 2N = 2^n, with n an integer. The prototype filterbeing of length L_g = bM, its duration is also given by theindicating function $I_{|n-n0|}<2b$, equal to 1 if $|n - n0| < 2b$ and0 elsewhere. Then, the scalar product of the base functionscan be expressed as

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

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 whentransmitting 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 ≤ M/2 spreading codes areproperly selected then the $i_{u,n}$interference is cancelled. TheW-H matrix being of size M = 2N = 2^ncan be divided intotwo subsets of column indices, $S^n_1$and S^n_2, with cardinal equalto M/2 making a partition of all the index set. To guaranteethe absence of interference between users, the constructionrule 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, thetwo 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 identicalsize 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, asfollows:

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

Applying this rule one can check that for n = 5, as anexample, 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 ofU users up to M/2. The complete proof given in [11] takesadvantage of three properties ofW-H codes.

Transmission of Complex Data Symbols. As the imaginarycomponent can be cancelled when transmitting realdata through a distortion-free channel when using CDMA-OFDM, one can imagine extending this scheme tothe transmission of complex data. Indeed, the transmissionsystem being linear, real and imaginary parts will notinterfere if the previous rule is satisfied.

Then, denoting by $d^{(c)}_{n,u}$the complex data to transmit,the OFDMsymbols transmitted at time $n\tau_0$ over thecarrier 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 correspondingcomplex CDMA-OQAM transmission scheme is depicted inFigure 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 termis $\nu_{n,m} = j^{n+m}(-1)^{nm}$, that is, $\varphi 0 = \pi nm$. Then, if the U codesare all in S^n_1, or S^n_2, the interference terms are cancelled and weget

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

This CDMA-OFDMscheme satisfiesa complex orthogonality condition, that is, the back-tobacktransmultiplexer is a PR system for the transmissionof complex data. The maximum number of users is M/2instead of M. In both cases the overall data rate is thereforethe same.In the presence of a channel, equalization must beperformed before the despreading since the signal at theoutput of the equalization block is supposed to be free fromany channel distortion or attenuation. Then, the signal at theequalizer output is somewhat equivalent to the one obtainedwith a distortion-free channel. Then, despreading operationwill recover the complex orthogonality.

Alamouti with CDMA-OFDMwith Spreading inthe FrequencyDomain. In a realistic transmission scheme thechannel is no longer distortion-free. So, we assume now thatwe are in the case of a wireless Down-Link (DL) transmissionand perfectly synchronized.

6. PROBLEM STATEMENT

Before trying to apply Alamoutischeme to CDMA-OFDM, it is notedthat the channel equalization process is replaced by theAlamouti decoding. When adapting Alamouti scheme toCDMA-OFDM, the equalizer component, depictedin Figure 4, must be replaced by the Alamouti decodingprocess and the despreading operation must be carried outjust after the OFDM modulator. Then, contraryto the DL conventional MC-CDMA case, the dispreading operation must be performed before the Alamouti decoding.Indeed, with OFDM, we can only recover a complexorthogonality property at the output of the dispreading block. The point complex orthogonality hold in CDMA-OFDM if weperform despreading operation before equalizationleads to the following problem:let us consider complex quantities t_i, β_i, λ_i. Does it soundpossible 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 ismaterialized by $e_i = t_i/\lambda_i$and the despreading operationby $\sum_{i=0}^{M-1}\beta i(e_i)$. 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 constantchannel over frequencies. Indeed, only in this case theorder of the equalization and despreading operations can beexchanged without impairing the transmission performance.Conversely, applying despreading before equalization shouldhave an impact in terms of performance for a channelbeing nonconstant in frequency. So, let us consider at firsta flat channel. Then the subset of subcarriers where a givenspreading code is applied will be affected by the same channelcoefficient.

7. IMPLEMENTATION SCHEME

In a SISO configuration, ifwe denote by hn,i the single channel coefficient between thetransmit antenna i and the single receive antenna at timeinstant n, the despreaded signal is given by:

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

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

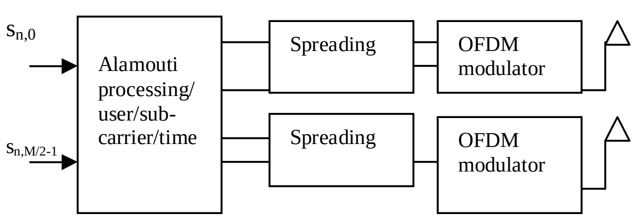


Figure. 5: An Alamouti CDMA-OFDM transmitter.

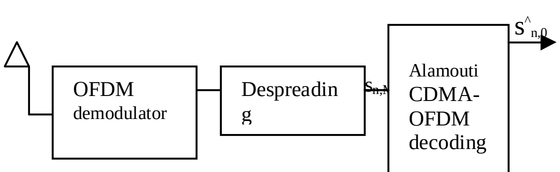


Figure.6: An Alamouti CDMA-OFDM receiver.

with 2 antennas with indexes 0 and 1, respectively, and if weapply Alamouti coding scheme to every user u data, denotingby 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 despreadedsignal for user u is given by

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

Hence,

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

This is the same decoding equation as in the Alamoutischeme presented in Section 2.2. Hence, the decoding could

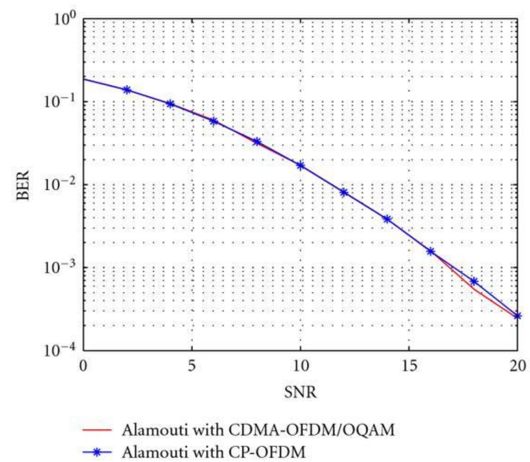


Figure. 7: BER for the complex version of the Alamouti CDMA OFDM with spreading in frequency domain, versus Alamouti CP-OFDM for transmission over a flat fading channel.

be performed in the same way. Figures 5 and 6 present theAlamouti CDMA-OFDM transmitter and receiverrespectively.

8. PERFORMANCE EVALUATION

We compare the proposedAlamouti CDMA-OFDM scheme with the AlamoutiOFDM using the following parameters:

(i) QPSK modulation
(ii) M = 128 subcarriers
(iii) maximum spreading length, implying that the W-Hspreading codes are of length Nc = 128,

(iv)flat fading channel (one single Rayleigh coefficient forall 128 subcarriers);
(v) the IOTA prototype filter with length 512,
(vi) zero forcing one tap equalization for both transmissionschemes,

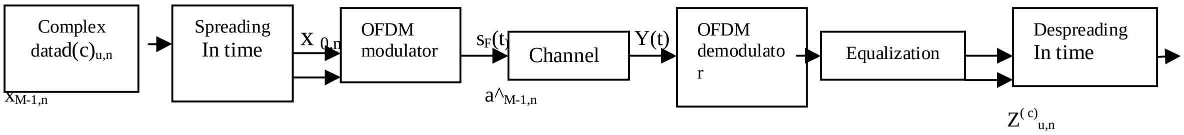


Figure. 8: Transmission scheme for the CDMA-OFDM system with spreading in time of complex data.

(vii) no channel coding.
Figure 7 gives the performance results. As expected, bothsystems perform the same.

Alamouti and CDMA-OFDMwithTime Domain Spreading

In this section, we keep the same assumptions as the onesused for the transmission of complex data with a spreadingin frequency. Firstly, we again suppose that the prototypefunction is a real-valued symmetric function and also thatthe W-H codes are selected using the procedure recalled inSection 3.1.2.

CDMA-OFDM with Spreading in the TimeDomain. Let us first consider a CDMA-OFDMsystem, carrying out a spreading in the time domain, thatis, on each subcarrier m the data are spread over the timeduration frame length. Let us consider Lf the length of theframe, that is, the frame is made of M data in the frequencydomain and Lf data in time domain. $N_c$is the length of thespreading code. We assume that $N_S = L_f / N_c$is an integernumber. Let us denote by: $c_u = [c_{0,u} \cdot\cdot\cdot c_{Nc-1,u}]^T$thecode used by the uth user. Then, for a user u0 at a givenfrequency m0, NS different data are transmitted denotedby: $d_{u0,m0,0}$, $d_{u0,m0,1}$, . . . , $d_{u0,m0,NS-1}$. By spreading with the $c_u$codes, we get the real symbol $a_{m0,n0}$ transmitted at frequencym_1 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, thereconstruction of du,m0,p (for p€$[0,N_S - 1]$) is insuredthanks 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 complexdata, denoted $d^{(c)}_{m,u,p}$, using U well chosen W-H codes. Inorder to establish the theoretical features of this complexCDMA-OFDM scheme, we suppose that the transmissionchannel is free of any type of distortion. Also, forthe sake of simplicity, we now assume a maximum spreadinglength (in time domain, $L_f = N_c$). We denote by $d^{(c)}_{m,u}$thecomplex data and by $a^{(c)}_{m,n,u} = c_{n,u}d^{(c)}_{m,u}$the complex symboltransmitted at time $n\tau_0$ over the carrier m and for the codeu. As usual, the length of the W-H codes are supposed to bea power of 2, that is, Lf = 2L = 2q with q an integer.

The block diagram of the transmitter is depicted inFigure 8. For a frame containing 2L OFDM datasymbols, the baseband signal spread in time, can be writtenas

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

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

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 satisfiesthe real orthogonality condition (2) and is real-valued andsymmetric, that is, $g(t) = g(-t)$. To express the complexinner product of the base functions $g_{m,n}$, using a similarprocedure that led to (24), we get

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

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:$y^{(c)m0,n0} = \langle y, g_{m0,n0}\rangle$. (42)

In this configuration, the demodulation operation onlytakes place when the whole frame is received. Then, thedespreading operation gives us the despreaded data for thecode u_0 as

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

Replacing $x_{m,n}$and $\langle g_{m,n}, g_{m0,q}\rangle$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 equalto 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

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

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

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

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 followingrequirements:

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

(ii) Since g is a real-valued function, $A_g(n, 0)$ is realvalued and the ambiguity function of the prototypefunction 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 areindependent of the phase term and thus are satisfied in thecase of the CDMA-OFDM system with spreading intime. It can also be shown that the modification of the phaseterm vm,n leads to the substitutions n → m and p + m →n + n0, in obtaining (48) from (41). Accordingly the secondterm 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 inTime. Now, if we consider the CDMA-OFDM withspreading in time, contrary to the case of a spreading infrequency domain, as long as the channel is constant duringthe spreading time duration, we can perform dispreading before equalization. At the equalizer output we will have acomplex orthogonality. Indeed, considering at first a SISOcase, if we denote by $h_{m,i}$the channel coefficient betweena single transmit antenna i and the receive antenna atsubcarrier 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 transmittedat subcarrier m by antenna i. Thus, we can easily apply theAlamouti decoding scheme knowing the channel is constantfor each antenna at each frequency. Otherwise said, themethod becomes applicable for a frequency selective channel.Actually two strategies can be envisioned.(1) Strategy 1. Alamouti performed over pairs of frequencies. Ifwe 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 datafor 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 despreadedsignal for user u is given by

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

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

That means, when assuming the channel to be flat over twoconsecutive subcarriers, that is, $h_{2m,i} = h_{2m+1,i}$for all i, wehave exactly the same decoding equation as the Alamoutischeme presented in Section 3.2, by permuting the frequencyand time axis. Then, the decoding is performed in the sameway.(2) Strategy 2. Alamouti performed over pairs of spreadingcodes. In this second strategy, we apply the Alamouti schemeon 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 complexdata for user pair (u0, u1). At subcarrier m, antennas 0 and 1transmit

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

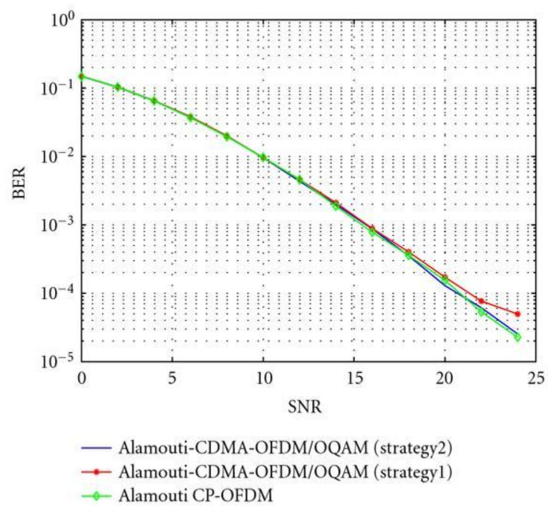


Figure.9: BER for two complex versions of the Alamouti CDMAOFDM/OQAM with spreading in time domain, versus AlamoutiCP-OFDM for transmission over the 4-path frequency selectivechannel.

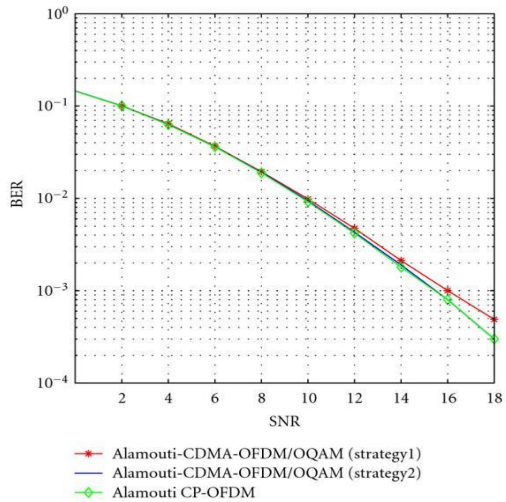


Figure. 10: BER for two complex versions of the Alamouti CDMAOFDM with spreading in time domain, versus AlamoutiCP-OFDM for transmission over the 7-path frequency selective channel.

Then, we do not need to consider the channel constantover two consecutive subcarriers. We have exactly the samedecoding equation as the Alamouti scheme presented inSection 3.2. Hence, the decoding is performed in the sameway.

We have tested two different channels considering eachtime 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 thedelay spread profiles corresponding to a 4-path and to a 7-path channel. The 4-path channel is characterized by thefollowing 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 theframe duration (32 complex OQAM symbols),
(vi) number of CDMA W-H codes equals to 16 incomplex OFDM, with symbol duration τ0and this corresponds to 32 codes in OFDM, withsymbol duration 2τ0, leading to the same spectralefficiency
(vii) zero forcing, one tap equalization,
(viii) no channel coding.

In Figures 9 and 10, the BER results of the AlamoutiCDMA-OFDM technique for the two proposedstrategies are presented.

The two strategies perform the same until a BER of10^{-3} or 10^{-2} for the 4 and 7-path channel, respectively.

For lower BER the strategy 2 performs better than thestrategy 1. This could be explained by the fact that strategy1 makes the approximation that the channel is constantover two consecutive subcarriers. This approximation leadsto a degradation of the performance whereas the strategy2 does not consider this approximation. If we compare theperformance of Alamouti CDMA-OFDM strategy2 with the Alamouti CP-OFDM, we see that both systemperform approximately the same. It is worth mentioningthat however the corresponding throughput is higher for theOFDM solutions (no CP). Indeed, it is increasedby approximately 4 and 7% for the 4 and 7-path channels,respectively.

9. CONCLUSION

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

If the CDMA spreading iscarried out in the frequency domain, the Alamouti decodingscheme can only be applied if the channel is assumedto be flat. For a frequency selectivechannel, the CDMA spreading component has to beapplied in the time domain.

For the Alamouti scheme withtime spreading CDMA-OFDM, two strategies are suggested for implementing the MIMO space-time codingscheme. Strategy 1 implements the Alamouti over pairs ofadjacent frequency domain samples whereas the strategy 2processes the Alamouti coding scheme over pairs ofspreading codes from two successive time instants. Strategy2 appears to be more appropriate since it requires lessrestrictive assumptions on the channel variations acrossthe frequencies.

The performance comparisonswith Alamouti CP-OFDM show that under the channel hypothesis, the combination of Alamoutiwith complex CDMA-OFDM is possible withoutincreasing the complexity of the Alamouti decoding process. In the case of a frequency selective channel, OFDM keeps its intrinsic advantage with a SNRgain in direct relation with the CP length.

To find asimpler Alamouti scheme, that is, without adding a CDMAcomponent, remains an open problem.Naturally, some otheralternative transmit diversity schemes for OFDM, asfor instance cyclic delay diversity deserve furtherinvestigations.

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