Suitable combination of channel coding and space–time schemes for moderate-to-high spectral efficiency MIMO systems

Mohammad-Ali Khalighi\textsuperscript{a,}*\textsuperscript{,}, Jean-François Hélard\textsuperscript{b}, Seyed Mohammad-Sajad Sadough\textsuperscript{c}, Salah Bourennane\textsuperscript{a}

\textsuperscript{a}École Centrale Marseille, Institut Fresnel, UMR CNRS 6133, Marseille, France
\textsuperscript{b}Institut d’Électronique et de Télécommunications de Rennes, Rennes, France
\textsuperscript{c}Faculty of Electrical and Computer Engineering, Shahid Beheshti University, Tehran, Iran

Received 23 September 2008; accepted 30 March 2009

Abstract

In order to propose practical solutions for future communication systems employing a combination of multiple-input multiple-output (MIMO) structures and orthogonal frequency division multiplexing (OFDM), we analyse and compare the performances obtained by orthogonal and non-orthogonal space–time coding schemes. A simple iterative detector based on soft parallel interference cancellation (Soft-PIC) is used in the receiver in order to cope with the interference terms when using non-orthogonal schemes. First, we propose a modification of the log-likelihood ratio (LLR) calculation part in the Soft-PIC detector that provides an improvement of the receiver performance compared to previous implementations of this detector. Then, the performances of different space–time schemes proposed for two, three or four antennas at the transmitter and two antennas at the receiver are compared. In the case of perfect channel estimation, a substantial gain is obtained by using the appropriate non-orthogonal schemes for moderate-to-high spectral efficiency MIMO systems, which justifies the increased complexity of the receiver. This conclusion is conditioned to the presence of sufficient frequency diversity that is the case in a practical system. We also study the case of imperfect channel estimation at the receiver and discuss the choice of the appropriate space–time scheme. The gain offered by the non-orthogonal schemes is confirmed in that case, specifically for high spectral efficiency systems.

© 2009 Elsevier GmbH. All rights reserved.

Keywords: MIMO systems; Space–time coding; OSTBC; LD codes; Iterative detection

1. Introduction

The potential of multiple-input multiple-output (MIMO) systems in providing high spectral efficiencies [1] has made them a promising solution for the ever growing need to high-speed, spectrally efficient, and reliable data transmission. In particular, MIMO structures are suitable candidates for the future cellular mobile radio systems that should provide high throughput and high quality of service. For instance, the application of MIMO systems in the fourth generation of cellular networks has been considered in the European IST-4MORE project [2].

Concerning the practical implementation of MIMO systems, an important aspect is the space–time (ST) coding at the transmitter that aims to exploit effectively the spatial diversity inherent in a multiple-antenna communication system. The choice of a suitable ST scheme consists in distributing appropriately some redundancy in time and in space, that is, among transmit antennas, in order to improve the system
performance. To this date, there has been considerable work on this subject and a variety of ST schemes has been proposed in the literature for MIMO systems (see [3] and the references therein). The two key criteria in the design of ST codes are the multiplexing gain and the diversity gain [4]. Multiplexing gain concerns the increase obtained in the data rate, whereas the diversity gain is related to the mitigation of channel fading. More precisely, designs privileging the diversity gain aim at increasing the reliability of data transmission at the expense of a loss in channel capacity, whereas the ST designs privileging the multiplexing gain aim at increasing the data rate, hence benefiting less from the diversity advantage.

In this paper, we focus on three families of ST schemes: the orthogonal space–time block codes (OSTBC) [5], the pure spatial multiplexing scheme (which is the transmission scheme used in the V-BLAST architecture [6]), and the linear dispersion (LD) codes [7] that maximize the mutual information between the transmitter and the receiver. OSTBCs offer full diversity, spatial multiplexing maximizes the data rate, and LD codes allow flexible rate-diversity trade-off.

Besides code construction, an important practical aspect is the ST decoding at the receiver and its corresponding complexity. When we consider the achievable data rate at the same time, the choice of an ST scheme becomes delicate. Orthogonal designs can be decoded using an optimal decoder with linear complexity. However, they suffer from low rate, especially for a large number of transmit antennas. Non-orthogonal schemes, like spatial multiplexing, on the other hand, offer higher coding rates but their optimal decoder becomes prohibitively complex for a large number of transmit antennas and for large signal constellation sizes. In addition, some of these non-orthogonal schemes cannot be used when there are fewer antennas at transmitter than at the receiver.

In order to compare the performances of different ST schemes, we fix the spectral efficiency, and consider the power efficiency as the figure of merit. In other words, we compare the required signal-to-noise ratio (SNR) for different ST schemes to attain a desired average bit-error-rate (BER) for a given spectral efficiency $\eta$. We also take into account channel coding that is almost always used in communication systems to mitigate the noise, interference, or channel fading. For a given ST scheme, we set accordingly the signal constellation and the channel coding rate so as to achieve a desired $\eta$. In other words, our degrees of freedom are the signal constellation, the channel coding rate, and the choice of the ST scheme. The answer to the question “what is the most suitable combination” is not obvious for moderate-to-high spectral efficiencies. It has been shown in [8] that, for low spectral efficiency applications, an OSTBC together with a powerful turbo-code would be a suitable solution. To attain high spectral efficiencies with OSTBC schemes, however, we have to use large signal constellations and to increase the channel coding rate. Specially, the use of larger signal constellations complicates the tasks of synchronization and detection at the receiver and also results in a higher required signal-to-noise ratio to ensure a desired bit-error-rate.

Higher ST coding rates are offered by non-orthogonal schemes, hence, relaxing the conditions on signal constellation and channel coding. The disadvantage, as we pointed out above, is that the optimal decoding is much more computationally complex than for the orthogonal schemes. One good solution can be to use a simple suboptimal iterative detector for this purpose. Here, we consider MIMO signal detection based on soft parallel interference cancellation (Soft-PIC) [9,10]. This way, we may approach the performance of the optimal detector after few iterations [11]. Nevertheless, the detector remains more complex, as compared to OSTBC case. We should hence investigate if this increased receiver complexity is justified. In other words, we want to see whether or not by using such a detector, we gain in performance with respect to the OSTBC choice, and if this gain is considerable enough to convince us to privilege a non-orthogonal solution. Notice that we are not going to present an exhaustive comparison of all already-proposed non-orthogonal schemes, as this is not the aim of this paper. The question is, in fact, to choose between the suitable orthogonal and the suitable non-orthogonal schemes.

In a first step, we assume perfect channel knowledge at the receiver and present a comparative performance study of different ST schemes. In practice, however, perfect channel knowledge conditions could never be met and we should make sure whether or not the preference of an ST scheme over another is preserved when taking into account channel estimation errors. So, in a second step, we will extend our study to the case of imperfect channel knowledge. As a side contribution of this work, we also propose a modification of the log-likelihood ratio (LLR) calculation part in the Soft-PIC detector that results in a performance improvement, compared to the implementations considered in [12–14].

The organization of the paper is as follows. Section 2 presents our system model and general assumptions. In Section 3 we present different ST schemes that we consider in this work. Details on Soft-PIC data detection are provided in Section 4, where our modification of the Soft-PIC detector is also presented. Simulation results are presented in Section 5 comparing the performance obtained by different ST schemes for several spectral efficiencies. Two cases of perfect channel knowledge and pilot-only-based channel estimation are considered. Finally, Section 6 concludes the paper.

2. System model and general assumptions

Let us consider the downlink transmission in a mobile communication system as the framework of our study. We consider two, three, or four antennas at the transmitter (base station) and two antennas at the receiver (mobile terminal). Notice that the more critical case regarding the
computational complexity is in the downlink, where we should take into consideration serious constraints on the handset power consumption and cost. We denote by $M_T$ and $M_R$ the number of antennas at the transmitter and at the receiver, respectively.

2.1. Transmitter

At the transmitter, channel coding and modulation are performed based on bit-interleaved coded modulation (BICM). As it is the case in most practical implementations of MIMO systems, multi-carrier modulation based on OFDM (orthogonal frequency division multiplexing) is employed in order to simplify the channel equalization task at the receiver. The block diagram of the transmitter is shown in Fig. 1. Binary data $b$ are encoded by the non-recursive non-systematic convolutional (NRNSC) channel code $C$, before being randomly interleaved (the block $II$). Output bits $c$ are then mapped to symbols according to Gray-coded M-QAM modulation with $B$ bits per symbol, $M = 2^B$. Power normalized symbols $s$ are then combined according to a given ST scheme. Next, the resulting symbols $x_t$, corresponding to each antenna $t$, are passed through the OFDM modulator of $N_c$ subcarriers before transmission.

2.2. ST coding

Let $S$ of dimension $(Q \times 1)$ be the vector of data symbols prior to ST coding, $S = [s_1, s_2, \ldots, s_Q]^t$, where $^t$ stands for vector or matrix transpose. By ST coding, $S$ is mapped into a matrix $X$ of dimension $(M_T \times T)$, $T$ being the number of channel uses. We call $X$ the generator matrix of the ST scheme. The ST coding rate is defined as $R_{STC} = Q / T$. A frame of $N$ encoded bits, $N$ being the interleaver size, corresponds to $NT / BQ$ symbols after ST coding. We denote by $N_F$ the number of OFDM symbols per frame, with $N_F = NT / (BQN_c)$.

2.3. Communication channel

The absence of inter-symbol interference and inter-carrier interference is guaranteed by using a guard interval longer than the delay spread of the impulse response of the channel. MIMO channel fading coefficients corresponding to each subcarrier are assumed to be independent and Rayleigh distributed. We assume that the channel coefficients are constant during the $N_F$ OFDM symbols, and change to new independent values from one frame to next.

2.4. Signal detection at the receiver

The block diagram of the receiver is shown in Fig. 2. After performing OFDM demodulation and parallel to serial data conversion on each receive antenna, we perform soft ST decoding, or in other words, MIMO signal detection, followed by soft channel decoding. For the case of non-orthogonal ST schemes, we use a relatively simple iterative detector based on soft parallel interference cancellation [9], shown in
3. Considered space–time schemes

Here, we present the ST schemes that we consider in our study, depending on the number of transmit antennas.

3.1. \( M_T = 2 \)

As the OSTBC scheme, we consider the famous Alamouti code [16] with \( Q = M_T = T = 2 \), described by the following generator matrix. Note that in addition to its simplicity, it has \( R_{STC} = 1 \) and exploits full diversity.

\[
X = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix}.
\]  

(1)

As the non-orthogonal scheme, we firstly consider the simple spatial multiplexing scheme (denoted here by MUX) for which, \( Q = 2, T = 1 \) and \( R_{STC} = 2 \):

\[
X = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix}.
\]  

(2)

It consists, in fact, in spatially multiplexing the data stream on the two transmit antennas.

We also consider an optimized scheme, called the golden code (denoted here by GLD), that is presented in [17]. For this code, described below, we have \( Q = 4, T = 2, \) and \( R_{STC} = 2 \).

\[
X = \frac{1}{\sqrt{5}} \begin{bmatrix} x(s_1 + 3s_2) & \gamma x_3 + \bar{x}_4 \\
\bar{x}(s_3 + 3s_4) & x(s_1 + 3s_2) \end{bmatrix},
\]  

(3)

where \( \theta = (1 + \sqrt{5})/2, x = 1 + j(1 - \theta), \gamma = 1 + j(1 - \bar{\theta}), \gamma = j, j = \sqrt{-1} \). The factor \( 1/\sqrt{5} \) in (3) ensures normalized transmitted power per channel use. This scheme offers full rate diversity.

3.2. \( M_T = 3 \)

For complex signal constellations, there does not exist a full diversity OSTBC scheme with \( R_{STC} = 1 \) for \( M_T > 2 \) [5]. Here, as the orthogonal choice, we consider the following code with \( Q = M_T = 3, T = 4, \) which has \( R_{STC} = \frac{3}{2} \) [18]:

\[
X = \begin{bmatrix} s_1 & -s_2^* & -s_3^* \\ s_2 & s_1^* & 0 \\ s_3 & 0 & s_1^* \\ s_4 & s_2^* & s_3^* \end{bmatrix}.
\]  

(4)

We will denote this code by OR3. As the non-orthogonal scheme, we again consider the simple spatial multiplexing scheme for which \( Q = 3, T = 1, \) and \( R_{STC} = 3 \):

\[
X = \begin{bmatrix} s_1 \\ s_2 \\ s_3 \end{bmatrix}.
\]  

(5)

Notice that, as reported in [11], although \( M_R < M_T, \) the MUX scheme can still be used with iterative Soft-PIC detection for \( M_R > M_T/2 \).

We can also construct a block-orthogonal scheme by performing Alamouti coding over each antenna pair alternatively. We hence obtain a simple LD code, given by (6), which is obtained from a symmetrical concatenation of three Alamouti schemes [7]. For this code that we denote by\(^1\) LD3–1, \( Q = T = 6, \) and \( R_{STC} = 1 \).

\[
X = \sqrt{\frac{3}{2}} \begin{bmatrix} s_1 & -s_2^* & 0 & 0 & s_5 & -s_6^* \\ s_2 & s_1^* & s_3 & -s_4^* & 0 & 0 \\ 0 & 0 & s_4 & s_3^* & s_6 & s_5^* \end{bmatrix}.
\]  

(6)

At last, we consider an optimized LD code, proposed in [7], for \( M_T = 3, \) with \( Q = 9, T = 3, \) and \( R_{STC} = 3, \) that we denote by LD3–3. The generator matrix of this code is given by Eq. (31) in [7] that we do not present here due to space limitations. The optimization for this scheme is done in the sense of maximizing the mutual information transfer through the channel.

3.3. \( M_T = 4 \)

Again an OSTBC code with full diversity and \( R_{STC} = 1 \) does not exist as we consider complex signal constellations. Here, as the orthogonal choice, we consider a simple scheme by which we perform Alamouti coding alternatively on one pair of antennas, while turning the other pair off. This scheme is called time-switched Alamouti and denoted here by Sw-Al. For the Sw-switched Alamouti scheme, described by (7), we have \( Q = T = 4 \) and \( R_{STC} = 1 \).

\[
X = \sqrt{2} \begin{bmatrix} s_1 & -s_2^* & 0 & 0 \\ s_2 & s_1^* & 0 & 0 \\ 0 & 0 & s_3 & -s_4^* \\ 0 & 0 & s_4 & s_3^* \end{bmatrix}.
\]  

(7)

The first non-orthogonal scheme that we consider for \( M_T = 4 \) is the simple double-Alamouti code, described by (8), that we denote by D-Al. As it is seen from (8), Alamouti coding is performed simultaneously on each of the two antenna pairs. For the D-Al scheme, we have \( Q = 4, T = 2, \) and \( R_{STC} = 2 \).

\[
X = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \\ s_3 & -s_4^* \\ s_4 & s_3^* \end{bmatrix}.
\]  

(8)

\(^1\) Subscript 3–1 indicates the number of transmit antennas \( M_T = 3 \) followed by the ST coding rate \( R_{STC} = 1 \).
We also consider the LD code proposed in [7], optimized for\( M_T = 4 \) and \( M_R = 2 \) by maximizing the mutual information between the transmitted and received signals. For this code, denoted here by LD_{4,2}, we have \( Q = 12 \), \( T = 6 \) and \( R_{STC} = 2 \). Its generator matrix is not presented here due to space limitations. Note that the simple MUX scheme cannot be used for \( (M_T = 4, M_R = 2) \) because the iterative Soft-PIC detector does not converge for this case.

4. MIMO detection: ST decoding

4.1. Model simplification

With the assumptions made in Section 2, we can ignore the ensemble of the operations of demultiplexing, OFDM modulation and demodulation, and multiplexing. Projecting hence the frequency dimension onto the time dimension, the MIMO OFDM channel can equivalently be considered as a single-carrier block fading channel. This way, we can describe our channel by a matrix \( H \) of dimension \((M_R \times M_T)\), invariant over a block of \( N_F \) channel uses (remember our notations from Section 2). A frame of \( N \) bits corresponds hence to \( N_F \) blocks with independent fades, each block corresponding to a given subcarrier in the original OFDM model. Now, corresponding to an ST encoded \((M_T \times T)\) matrix \( X \), we receive the \((M_R \times T)\) matrix \( Y \):

\[
Y = HX + N ,
\]

where \( N \) represents the receiver noise; its entries are zero-mean complex Gaussian random variables.

4.2. General formulation of LD codes

In order to provide a general formulation for the MIMO detector irrespective of the underlying ST scheme, we adopt the formulation of LD codes, proposed in [7]. This formulation applies also to OSTBC and MUX schemes.

Let \( x_q \) and \( \beta_q \) be the real and imaginary parts of a symbol \( s_q \), i.e., \( s_q = x_q + j\beta_q \). The main parameters of the code are its \((M_T \times T)\) dispersion matrices \( A_q \) and \( B_q \), \( q = 1, \ldots, Q \). The ST generator matrix is then

\[
X = \sum_{q=1}^{Q} (x_q A_q + j\beta_q B_q) ,
\]

\( A_q \) and \( B_q \) are in general complex matrices, but like in [7], we consider them of real entries. We further separate the real and imaginary parts of the entries of \( S \) and \( X \) and stack them row-wise in vectors. Remember from Section 2.2 that \( S \) is the \((Q \times 1)\) vector of data symbols prior to ST encoding. We obtain hence the vectors \( \mathcal{S} \) of dimension \((2Q \times 1)\) and \( \mathcal{X} \) of dimension \((2M_T T \times 1)\):

\[
\mathcal{S} = \{ \Re{s_1}, \ldots, \Re{s_Q}, \Im{s_Q} \}^T ,
\]

\[
\mathcal{X} = \{ \Re{X(1,1)}, \ldots, \Re{X(M_T, T)}, \Im{X(M_T, T)} \}^T ,
\]

where \( \Re{} \) and \( \Im{} \) denote the real and imaginary part operators, respectively. The LD ST encoder can now be considered as a \((2M_T T \times 2Q)\) matrix \( \mathcal{F} \) such that

\[
\mathcal{X} = \mathcal{F} \mathcal{S} ,
\]

and, for instance, \( A_{q}^{\Re}(m, t) \) denotes the \((m, t)\)th entry of \( A_q^{\Re} \). Receiving the matrix \( Y \), corresponding to a transmitted matrix \( X \), we construct the vector \( \mathcal{Y} \) from \( Y \) as we did to obtain \( \mathcal{X} \). We can write \( \mathcal{Y} = \mathcal{H} \mathcal{X} + \mathcal{N} \), where \( \mathcal{N} \) is the vector of real AWGN of zero mean and variance \( \sigma_n^2 \), and the matrix \( \mathcal{H} \) of dimension \((2M_R T \times 2M_T T)\) is constructed from the \( \Re{} \) and \( \Im{} \) parts of the entries \( H_{ij} \) of the initial matrix \( H \). It is composed of segments \( \mathcal{H}_{ij}, i = 1, \ldots, M_R, j = 1, \ldots, M_T \), each one of dimension \((2T \times 2T)\):

\[
\mathcal{H}_{ij} = \begin{bmatrix}
H_{ij} & 0 & \cdots & 0 \\
0 & H_{ij} & \cdots & 0 \\
\vdots & \ddots & \ddots & \vdots \\
0 & \cdots & 0 & H_{ij}
\end{bmatrix} ,
\]

where \( H_{ij} \) is obtained from each entry \( H_{ij} \) of the initial matrix \( H \):

\[
H_{ij} = \begin{bmatrix}
\Re{H_{ij}} & -\Im{H_{ij}} \\
\Im{H_{ij}} & \Re{H_{ij}}
\end{bmatrix} .
\]

Now, we can consider an equivalent channel matrix \( \mathcal{H} \mathcal{eq} \) of dimension \((2M_R T \times 2Q)\) such that

\[
\mathcal{Y} = \mathcal{H} \mathcal{eq} \mathcal{S} + \mathcal{N} = \mathcal{H} \mathcal{eq} \mathcal{F} + \mathcal{N} .
\]

4.3. MIMO soft detection

The detection problem is to find the transmitted data vector \( \mathcal{S} \), given the vector \( \mathcal{Y} \). For this purpose, we perform iterative Soft-PIC detection, as shown in Fig. 2, that we describe in the following in more detail.
4.3.1. PIC detector
Let us denote by \( \gamma_p \) the \( p \)th entry of \( \mathcal{S} \), \( p = 1, \ldots, 2Q \). At the first iteration, where we have no information on the transmitted symbols, to obtain the estimate of \( \gamma_p \), \( \hat{\gamma}_q \), we apply the MMSE (minimum mean-square error) filter \( w_p \) to \( \mathcal{S} \) \cite{5}:

\[
\hat{\gamma}_p = w_p^T \mathcal{S},
\]

where

\[
w_p^T = h_p^T (\mathcal{H}_p \mathcal{H}_p^T + \sigma_n^2 I)^{-1}.
\]

Here, \( h_p \) of dimension \( (2M_RT \times 1) \) is the \( p \)th column of \( \mathcal{H}_p \). From the second iteration, we calculate soft estimates of the transmit symbols \( \hat{\mathcal{S}} \) using the SISO decoder outputs and use them in the PIC detector to perform interference cancellation followed by simplified MMSE detection:

\[
\begin{align*}
\text{Interference cancellation:} & \quad \hat{\mathcal{S}}_p = \mathcal{S} - \mathcal{H}_p \hat{\mathcal{S}}_p, \\
\text{MMSE filtering:} & \quad \hat{\gamma}_p = \frac{w_p^T \hat{\mathcal{S}}_p}{h_p^T h_p + \sigma_n^2},
\end{align*}
\]

where \( \hat{\mathcal{S}}_p \) of dimension \( ((2Q - 1) \times 1) \) is \( \hat{\mathcal{S}} \) with its \( p \)th entry removed, and \( \mathcal{H}_p \) of dimension \( (2M_RT \times (2Q - 1)) \) is \( \mathcal{H}_p \) with its \( p \)th column removed. In fact, (21) is a suboptimal solution to the detection problem, and has a considerable reduced computational complexity, compared to the exact solution, proposed in \cite{9,19}. We have verified by simulations that, when used in an iterative scheme, this suboptimal detector mostly converges to the optimal solution at high enough SNR and when there is enough diversity available, e.g., when the number of receive antennas is large enough (see Section 5.3).

For the case of orthogonal ST schemes, we perform maximum likelihood (ML) detection once (without iteration).

4.3.2. Conversion to LLR

As we consider QAM modulation with \( B \) (assumed to be an even number) bits per symbol, we attribute \( m = B/2 \) bits to real and imaginary parts of each symbol. Let, for instance, the bit \( c_i \) correspond to \( \gamma_p \). Let also \( a_{1,j} \) and \( a_{0,j} \) denote the real (or imaginary) part of the signal constellation points, corresponding to \( c_i = 1 \) and \( 0 \), respectively. Remember that we consider power-normalized signal constellation. The LLR corresponding to \( c_i \) is calculated as follows:

\[
\begin{align*}
\text{LLR}_i & = \log \frac{\sum_{j=1}^{2m-1} \exp \left( -\frac{1}{2\sigma_p^2} (\hat{\gamma}_p - \sigma_p a_{1,j})^2 + L_j,1 \right)}{\sum_{j=1}^{2m-1} \exp \left( -\frac{1}{2\sigma_p^2} (\hat{\gamma}_p - \sigma_p a_{0,j})^2 + L_j,0 \right)} \\
i & = 1, \ldots, m,
\end{align*}
\]

where \( \exp(L_j,1) \) (resp. \( \exp(L_j,0) \)) denotes the \textit{a priori} probability of the transmission of \( a_{1,j} \) (resp. \( a_{0,j} \)), calculated for a given \( j \) from the \( (m - 1) \) \textit{a posteriori} LLRs at the SISO decoder output at the previous iteration, excluding the corresponding LLR on the bit \( c_i \) itself. Obviously, at the first iteration, these LLRs are equal to zero. Also, \( \sigma_p = w_p^T h_p \) is the weight given to \( \gamma_p \) after PIC detection, and \( \sigma_p^2 \) is the variance of noise plus residual interference (RI)\(^2\) corresponding to \( \hat{\gamma}_p \). In fact, since the detection is performed on blocks of \( Q \) complex symbols, or in other words on blocks of \( 2Q \) real symbols in our model, the RI comes in fact from \((2Q - 1)\) other real symbols in the corresponding block. In (22), we have in fact assumed Gaussian RI and independent symbols. The calculation of \( \sigma_p^2 \) is left to Section 4.3.4.

We use the approximation \( \log(e^{x_1} + e^{x_2}) \approx \max(x_1, x_2) \) in the calculation of LLR, in (22). This simplifies considerably the LLR calculation for the cases of 16-QAM and 64-QAM modulations. In fact, that is why we use the Max-Log-MAP SISO channel decoder instead of Log-MAP decoder. The performance loss by using this approximation is practically negligible.

4.3.3. Soft estimation of transmit symbols

Each entry \( \hat{\gamma}_p \) of the vector \( \hat{\mathcal{S}} \) is obtained by taking a summation over all possible values of the real (or imaginary) part of the signal constellation, multiplied by the corresponding probability calculated using the \textit{a posteriori} LLRs at the SISO decoder output. Remember that \( \gamma_p \) corresponds to the real (imaginary) part of a symbol in \( S \). Using our notations of the previous subsection, we have

\[
\hat{\gamma}_p = E[\gamma_p] = \sum_{j=1}^{2B/2} a_j P[\gamma_p = a_j],
\]

where \( E[\cdot] \) and \( P[\cdot] \) denote the expected value and probability, respectively, and \( a_j \) is the \( j \)th possible real (imaginary) part of the signal constellation points. Let us denote by \( c_i^j \), \( i = 1, \ldots, B/2 \) the bits corresponding to \( a_j \). The probability \( P[\gamma_p = a_j] \) is calculated using the probabilities \( P[c_i^j] \) at the decoder output:

\[
P[\gamma_p = a_j] = K^{B/2} \prod_{i=1}^{B/2} P[c_i^j],
\]

where \( K \) is a normalization factor. The probabilities \( P[c_i^j] \) can simply be obtained by the conversion of the corresponding LLRs that we denote by \( \text{LLR}_i^j \):

\[
P[c_i^j] = \frac{\exp(\text{LLR}_i^j)}{1 + \exp(\text{LLR}_i^j)}.
\]

Notice that, although from a theoretical point of view, we would have to use extrinsic LLRs for this purpose, in practice, use of a posteriori LLRs results in a better and faster convergence of the iterative detector \cite{19}.

\(^2\) Note that this is only the case for non-orthogonal ST schemes. For orthogonal schemes we have no RI after signal detection.
4.3.4. Calculation of the variance of noise plus RI

As we explained in Section 4.3.2, to convert a detected symbol at the output of the PIC detector, \( r_p \), to LLRs on its constituting bits, we need the variance of noise plus RI, \( \sigma_p^2 \) (see (22)). The parameter \( \sigma_p^2 \) has an important impact on the detector performance.

At the first iteration, this variance can easily be calculated analytically. As we perform no interference canceling, here, the RI comes directly from the other \((2Q - 1)\) symbols in \( \mathcal{Q} \). So, \( \sigma_p^2 \) is calculated using the following expression (more details are provided in the Appendix):

\[
\sigma_p^2 = w_p^i \left[ \frac{1}{2} H_{eq} H_{eq}^t + \sigma_n^2 I \right] w_p - \frac{1}{2} (w_p^i H_p)^2.
\]  

(25)

If we denote by \( H_p \) the matrix \( H_{eq} \) with its \( p \)th column removed, we can simplify (25) as follows:

\[
\sigma_p^2 = \frac{1}{2} w_p^i H_p H_p^t w_p + \sigma_n^2 w_p^i w_p.
\]  

(26)

To calculate \( \sigma_p^2 \) from the second iteration, it has been proposed to use the following expression \([10,20]:\)

\[
\sigma_p^2 = \frac{1}{2} \beta_p^2 (1 - \beta_p),
\]  

(27)

where \( \beta_p = w_p^i H_p \). This approach assumes Gaussian distributed RI-plus-noise at the output of the MMSE detector. Otherwise, as a simple solution, we may neglect the RI and to use the approximation of \( \sigma_p^2 \approx \sigma_n^2 \), as proposed in \([12,13] \).

These solutions give satisfying results for relatively simple ST schemes and for the case of QPSK modulation. However, for more complex ST schemes where the RI is more considerable due to larger \( Q \), as well as for larger constellation sets, these approaches are not suitable and can result in a considerable performance degradation. We propose here a new method for calculating \( \sigma_p^2 \). Using a similar approach as in \([20] \), it can be shown that, from the second iteration, \( \sigma_p^2 \) is given by (see the Appendix)

\[
\sigma_p^2 = \sigma_n^2 w_p^i w_p + w_p^i H_p \hat{\Lambda}_p H_p^t w_p,
\]  

(28)

where \( \hat{\Lambda}_p \) is a \((2Q - 1) \times (2Q - 1)\) diagonal matrix whose diagonal entries are soft-estimates of the symbol variances calculated using the SISO decoder soft outputs, in the same way as \( \mathcal{Q} \).

5. Simulation results

After explaining the iterative detector, we can now present the comparative study of the different ST schemes presented in Section 3. For a given number of transmit antennas \( M_T \) and different ST schemes, we set the signal constellation and the channel coding rate \( R_c \), so as to obtain the same spectral efficiency \( \eta \) (in bps/Hz). This ensures a fair comparison of the different schemes considered in this paper. Then, the performance comparison is made in terms of the bit-error-rate as a function of SNR. The SNR is considered in the form of \( E_b/N_0 \) and includes the receiver array gain, \( M_R \). \( E_b \) is the average received energy per information bit and \( N_0 \) is the noise power spectral density. Also, the 64-state NRNSC channel code (133, 171)\text{q} is considered where without puncturing, \( R_c = \frac{1}{2} \).

We set the number of OFDM subcarriers \( N_c \) to 32 and assume independent fades on different subcarriers. Hence, we have a frequency diversity order of 32. Also, \( N_F \) is set to 24. Different ST schemes for \( M_T = 2, 3, 4 \) are resumed in Table 1 To perform a fair comparison between different MIMO structures, we have set the spectral efficiency proportional to \( M_T \), and have chosen accordingly the (smallest possible) signal constellations and the channel coding rates. We have also specified in Table 1 the interleaver length \( N \). The interleaver is of pseudo-random type.

<table>
<thead>
<tr>
<th>( \eta ) (bps/Hz)</th>
<th>( M_T \times M_R ) schemes with spectral efficiency</th>
<th>STC</th>
<th>( R_{STC} )</th>
<th>Mod.</th>
<th>( R_c )</th>
<th>( N )</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>(2 \times 2) Alamouti 1 16-QAM 1/2 3072</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(3 \times 2) OR3 3/4 64-QAM 2/3 3456</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(4 \times 2) Sw-Al 1 64-QAM 2/3 4608</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>(2 \times 2) MUX 2 QPSK 1/2 3072</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(3 \times 2) LD\text{q}_3 1 16-QAM 3/4 3400</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(4 \times 2) D-Al 2 16-QAM 1/2 4608</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>(2 \times 2) GLD 2 QPSK 1/2 3072</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(3 \times 2) LD\text{q}_4 3 QPSK 1/2 4608</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>(4 \times 2) LD\text{q}_4 2 16-QAM 1/2 4608</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

5.1. Perfect channel knowledge at the receiver

At first, we assume that perfect channel knowledge is available at the receiver. We consider different cases of \( M_T \) separately in the following.

5.1.1. \( M_T = 2 \)

Performance curves are shown in Fig. 3. Only one iteration is processed for the case of orthogonal codes. For MUX and GLD schemes we have shown BER curves after two iterations, as well as after the almost full convergence of the detector, i.e., after four iterations. We see that a substantial gain in SNR is obtained by using non-orthogonal schemes. For instance, compared to Alamouti coding, at BER = 10\text{−}4 we gain about 2 and 2.65 dB in SNR by using the MUX and GLD schemes, respectively, after four iterations of the receiver. We note also that even when for the reasons of complexity and/or latency reduction, only two receiver iterations are processed, the gain in SNR compared to the Alamouti scheme remains significant: it is about 1.8 and 2.25 dB at BER = 10\text{−}4 for MUX and GLD, respectively. The interesting point is that, because MUX cannot exploit the full
diversity, the corresponding performance after four iterations is almost equivalent to that of the full diversity GLD scheme after two receiver iterations.

Comparing these results with those presented in [12,13], we notice that the estimation of the variance $\sigma_p^2$ of noise plus RI, which is used in (22), has an important impact on the overall performance of the receiver. Actually, we notice a considerable performance improvement as compared to the approximative approaches proposed in [12,13] (the reader is referred to the corresponding references).

5.1.2. $M_T = 3$

Fig. 4 compares BER performances for different ST schemes in this case. There is a negligible performance improvement after the first iteration for the code LD$_3$. For MUX and LD$_3$ schemes, four iterations are required to achieve the almost full convergence of the receiver. We have, however, shown the BER curves after two receiver iterations too. We notice that, although MUX and LD$_3$ should normally be used for $M_R \geq M_T$, thanks to iterative detection, they provide substantial performance improvement, compared to OR and LD$_3$ schemes for ($M_T = 3$, $M_R = 2$). The best performance is, however, obtained for MUX; although LD$_3$ seems to catch it up at very low BER. Note that, in addition, the detector is more computationally complex at the first iteration for LD$_3$ ($Q = 9$), because we have to calculate a $(2Q \times 2Q)$ matrix inversion in (20). The better convergence of MUX compared to LD$_3$ can also be explained by the larger $Q$ for the latter, which results in a more important RI. The gain in SNR of MUX compared to OR$_3$ at BER $= 10^{-4}$ is about 4.2 and 6 dB after two and four iterations, respectively.

We do not reproduce those results here for the sake of conciseness.

5.1.3. $M_T = 4$

In this case, Fig. 5 compares BER curves for different ST schemes. Concerning D-Al and LD$_4$ schemes, again about four iterations are sufficient to attain the almost full detector convergence. We have also shown the performance curves after two iterations. We notice that D-Al has a better performance, compared to LD$_4$, but at large enough SNR (corresponding to BER $< 10^{-5}$) their performance become almost equivalent. However, as for LD$_4$ we have $Q = 12$, the choice of the D-Al scheme appears to be a good compromise between receiver complexity and performance. From Fig. 5, the gain achieved by D-Al compared to Sw-Al scheme is about 4.1 and 4.85 dB, after two and four iterations, respectively.
5.2. Performance comparison with imperfect channel knowledge at the receiver

Up to now, we have assumed perfect channel knowledge at the receiver. In practice, however, the channel has to be estimated at the receiver before coherent signal detection. It is important, hence, to consider the presence of channel estimation errors, which affect the performance of Soft-PIC detector [21]. Actually, lower-rate orthogonal schemes could be more sensitive to channel estimation errors as, in general, they have to use a larger signal constellation to attain a desired spectral efficiency. Non-orthogonal schemes could in turn be more sensitive to these errors, as they affect the iterative detector convergence, and the sensitivity to the estimation errors is not the same for different ST schemes.

To study the effect of channel estimation errors, we consider pilot-only-based channel estimation. Using \( N_p \) pilot bits, we devote \( N_p/(BM_T) \) channel uses to the transmission of mutually orthogonal QPSK pilot sequences from \( M_T \) transmit antennas [22]. Like data symbols, pilot symbols are normalized in power. We assume that \( N_0 \) is known at the receiver and pilots are used only for the estimation of \( H \). We have shown in Figs. 6 and 7 the BER after four iterations versus \( N_p \), for the two cases of \( M_T = 2 \) and 4, respectively. Note that the minimum value for \( N_p \) satisfying the channel identifiability depends on \( M_T \) (see [22] for details). The \( E_b/N_0 \) for each ST scheme is set to what results in BER \( \approx 10^{-4} \) in the case of perfect channel knowledge.

From Fig. 6 and considering MUX and GLD schemes, we notice an almost equivalent sensitivity to the channel estimation errors. This comparison makes sense as the SNRs for these schemes are close to each other. So, we can confirm the preference of using the GLD scheme over MUX. On the other hand, we see that the Alamouti scheme has the lowest sensitivity. This is due to the orthogonal structure of the code, and the fact that the SNR is higher, compared to those for MUX and GLD schemes, and as a result, the quality of channel estimate is much better.

Similarly, from Fig. 7 we notice that the D-Al scheme has a considerably lower sensitivity to the estimation errors than LD\(_{4-2}\). This confirms the preference of using D-Al rather than LD\(_{4-2}\). Again, the orthogonal Sw-Al scheme has the lowest sensitivity to estimation errors, which is due to the higher SNR and its orthogonal structure.

5.3. Impact of frequency diversity order

In the results presented so far, we have assumed that we have independent fading across the \( N_c \) OFDM subcarriers. In practice, however, this assumption of perfect interleaving and independent fading is not realistic. It is important hence to investigate the impact of correlated fading, or in other words, the frequency diversity order, on the conclusions we made previously. Instead of considering the case of correlated fading across subcarriers, we consider here different numbers of subcarriers \( N_c \) while keeping the assumption of independent fading across them. Indeed, a smaller \( N_c \) can equivalently be regarded as corresponding to more fading correlation across subcarriers. We have compared in Fig. 8 the \( E_b/N_0 \) to attain a BER of \( 10^{-4} \) for (2 \times 2) and (4 \times 2) systems for different \( N_c \). To perform a fair comparison in terms of BER, the frame length (interleaver size) is kept the same for any \( N_c \). For the case of (2 \times 2) system, we have considered the Alamouti and GLD schemes, and for the case of (4 \times 2) system, the Sw-Al and D-Al schemes. The preference of non-orthogonal over orthogonal schemes is obvious for relatively large \( N_c \), notably, for \( N_c \geq 4 \). This is not the case, however, for small \( N_c \), and particularly for \( N_c = 1 \).

To understand the results of Fig. 8, one should take into account the frequency diversity factor. Consider the (2 \times
2) structure, for instance. With \( N_c = 1 \), the Alamouti and GLD schemes benefit from an overall diversity of \( M_T M_R = 4 \), thanks to the codes’ structures. For the GLD scheme, however, the suboptimal PIC detection appears to be highly penalizing, and the iterative processing does not permit to take advantage of this code. The obtained performance is close to that of the Alamouti scheme, as seen from Fig. 8. By increased \( N_c \), each of the ST schemes benefits from an additional frequency diversity factor. Particularly, for \( N_c \geq 4 \), the GLD scheme outperforms clearly the Alamouti scheme. It reveals that the turbo-PIC receiver becomes efficient when enough diversity is available.

Note that we use the BICM scheme at the transmitter. When a large frequency diversity order is available, the diversity gain equals the Hamming distance \( d_{\text{free}} \) of the channel code [23]. However, for a fading channel with finite frequency selectivity, the diversity of BICM is limited by the degree of frequency selectivity, rather than \( d_{\text{free}} \) of the binary code [24]. In our case study, we considered NRNSC code of \( d_{\text{free}} = 10 \) [25]. So, by increased \( N_c \), the effective diversity order approaches \( d_{\text{free}} \). This explains the saturation effect in Fig. 8 by increasing \( N_c \). In other words, we may call \( N_c \) the “potential” diversity order that we can exploit more by choosing a channel code of larger Hamming distance.

6. Conclusions

For MIMO systems working at moderate-to-high spectral efficiencies, we considered the choice of appropriate ST schemes in combination with channel coding. We showed that a substantial gain can be obtained by using appropriate non-orthogonal schemes and a simple iterative detector at the receiver, compared to orthogonal ST coding. The price paid is the increased receiver complexity. We proposed to use an iterative detection based on Soft-PIC which has a much lower complexity than the optimal detector and provides satisfying results when enough diversity is available, either in the form of time, frequency or space (i.e., larger \( M_R \)). Meanwhile, we proposed a modification of the LLR calculation part in the iterative Soft-PIC detector, that results in a considerable performance improvement (for non-orthogonal schemes), especially for large number of antennas and large signal constellation sizes. When the space diversity is limited (here, \( M_R = 2 \)) and we cannot exploit any other source of diversity, the use of iterative Soft-PIC becomes too penalizing and a simple OSTBC becomes the appropriate solution. In practice, however, by employing OFDM, we have enough frequency diversity available, even with some fading correlation across subcarriers. The following conclusions assume the availability of enough diversity in the system.

As we considered downlink transmission, we may process few iterations in order to keep the latency and the complexity of the mobile terminal reasonable. We noticed that even when only two iterations are performed at the receiver, non-orthogonal schemes outperform orthogonal ones. We also discussed the case of imperfect channel estimation and contrasted the performances of different ST schemes. The presented results confirmed the preference of the non-orthogonal schemes, especially for high spectral efficiency systems. Among the ST schemes considered in this paper, the GLD, MUX (spatial multiplexing), and D-Al schemes appear to be appropriate choices for \( M_T = 2, 3 \), and 4, respectively. For \( M_T = 3 \) and 4, the preference of the MUX and D-Al schemes to LD3-3 and LD4-2, respectively, can be explained by the fact that, conversely to the hypothesis considered in their design [7], here we have \( M_R < M_T \) and that we use a suboptimal detector at the receiver.

Acknowledgement

The authors would like to thank the anonymous reviewers for their comments and suggestions that resulted in a significant improvement of the paper.

Appendix A. Details on the derivation of the variance of noise + RI

Here we provide details on the derivation of formulas (20) and (21).

- First iteration: We have \( \gamma = \mathcal{H}_p \mathcal{I} + N \), and for the first iteration, \( \gamma_p = w_p \gamma \), where \( w_p = \mathcal{H}_p (\mathcal{H}_p^t \mathcal{H}_p + \sigma_p^2 I)^{-1} \). Let us decompose \( \mathcal{I} \) into \( \gamma_p \) (the symbol to be detected) and \( \mathcal{I}_p \) which is the vector \( \mathcal{I} \) with its \( p \)th entry removed.
We have
\[
\hat{y}_p = w_p^{t} \mathcal{Y} = w_p^{t}(h_p \hat{\gamma}_p + h_p \mathcal{F} p) + w_p^{t} \mathcal{N} \\
= w_p^{t} h_p \hat{\gamma}_p + (w_p^{t} h_p \mathcal{F} p + w_p^{t} \mathcal{N}),
\]
(29)
where \(Z_p\) denotes noise plus RI intervening in the detection of \(\hat{\gamma}_p\). The variance of \(Z_p\), conditioned to \(\gamma_p\) and \(\mathcal{H}\) eq is then
\[
\sigma_p^2 = E[Z_p^{t} Z_p] \\
= \sigma_n^2 w_p^{t} w_p + w_p^{t} h_p E[\mathcal{F}_p \mathcal{F}_p^{t}] h_p^{t} w_p.
\]
(30)

Given that we have assumed power-normalized (complex) symbols, the variance of \(\mathcal{R}\) and \(\mathcal{I}\) parts of the symbols equal \(\frac{1}{2}\). So, \(E[\mathcal{F}_p \mathcal{F}_p^{t}] = \frac{1}{2} I_{(2Q-1)}\), where \(I_{(2Q-1)}\) denotes the \((2Q-1) \times (2Q-1)\) identity matrix. As a result,
\[
\sigma_p^2 = \sigma_n^2 w_p^{t} w_p + \frac{1}{2} w_p^{t} h_p h_p + \sigma_n^2 w_p^{t} w_p,
\]
(31)

- **Next iterations**: For the next iterations, we perform interference cancellation \(\hat{\mathcal{Y}}_p = \mathcal{Y} - h_p \hat{\mathcal{F}} \), followed by simplified MMSE detection \(\hat{\gamma}_p = w_p^{t} \hat{\mathcal{Y}}_p\), where \(w_p = h_p / (h_p h_p + \sigma_n^2)\). We have
\[
\hat{\gamma}_p = w_p^{t} \hat{\mathcal{Y}}_p = w_p^{t}(\mathcal{Y} - h_p \hat{\mathcal{F}}) \\
= w_p^{t} h_p \hat{\gamma}_p + h_p (\hat{\mathcal{F}} - \hat{\mathcal{F}}) + \mathcal{N} \\
= w_p^{t} h_p \hat{\gamma}_p + (w_p^{t} h_p \mathcal{F}_p + w_p^{t} \mathcal{N}),
\]
(32)
where again \(Z_p\) corresponds to noise and RI. The variance \(\sigma_p^2\) of \(Z_p\), conditioned to \(\gamma_p\) and \(\mathcal{H}\) eq is
\[
\sigma_p^2 = w_p^{t} h_p \mathcal{F}_p E[(\mathcal{F}_p - \hat{\mathcal{F}}_p)(\mathcal{F}_p - \hat{\mathcal{F}}_p)^{t}] h_p^{t} w_p \\
+ \sigma_n^2 w_p^{t} w_p.
\]
(33)
The expectation term can be simplified as follows:
\[
E[(\mathcal{F}_p - \hat{\mathcal{F}}_p)(\mathcal{F}_p - \hat{\mathcal{F}}_p)^{t}] \\
= E[\mathcal{F}_p \mathcal{F}_p^t] - E[\mathcal{F}_p] E[\mathcal{F}_p^t] + \mathcal{F}_p \mathcal{F}_p^t.
\]
(34)

We have \(E[\mathcal{F}_p] = \mathcal{F}_p\). We furthermore define the \((2Q-1) \times (2Q-1)\) matrix \(\mathcal{A}_p \triangleq E[\mathcal{F}_p \mathcal{F}_p^t] - \mathcal{F}_p \mathcal{F}_p^t\). We hence obtain
\[
\sigma_p^2 = w_p^{t} h_p \mathcal{A}_p h_p^{t} w_p + \sigma_n^2 w_p^{t} w_p.
\]
(35)
The off-diagonal terms of \(\mathcal{A}_p\) are null and its diagonal entries are the soft-estimates of the square of the symbols in \(\mathcal{F}_p\), calculated using the output LLRs of the SISO decoder, in the same way that is done for calculating \(\hat{\mathcal{F}}\).

## References


Mohammad-Ali Khalighi received his Ph.D. in electrical engineering from Institut National Polytechnique de Grenoble (INPG), France, in 2002. From 2002 to 2005, he has been with GIPSA-lab, Télécom Paris-Tech (ENST), and IETR-lab, as a post-doctoral research fellow. He joined Institut Fresnel and Ecole Centrale Marseille, in 2005 as an assistant professor. His main research interests include coding, signal detection, and channel estimation for high data rate communication systems.

Jean-François Hélard received his Dipl.-Ing. and his Ph.D. in electronics and signal processing from the National Institute of Applied Sciences (INSA) in Rennes, France, in 1981 and 1992, respectively. From 1982 to 1997, he was research engineer and then head of channel coding for the digital broadcasting research group at CCETT (France Telecom Research Center) in Rennes. In 1997, he joined INSA, where he is currently professor and deputy director of the Rennes Institute for Electronics and Telecommunications (IETR), created in 2002 in association with the CNRS. His research interests lie in signal processing techniques for digital communications, such as space-time and channel coding, multi-carrier modulation, as well as spread-spectrum and multi-user communications. He is involved in several European and national research projects in the fields of digital video terrestrial broadcasting, mobile radio communications and cellular networks, powerline and ultra-wide-band communications. Prof. J.-F. Hélard is a senior member of IEEE, author and co-author of more than 115 technical papers in international scientific journals and conferences, and holds 13 European patents.

Seyed Mohammad-Sajad Sadough was born in Paris, France, in 1979. He received the B.Sc. degree in electrical engineering (electronic) from Shahid Beheshti University, Tehran, Iran, in 2002 and the M.Sc. and Ph.D. degrees in electrical engineering (telecommunications) from Paris Sud 11 University, Orsay, France, in 2004 and 2008, respectively. From 2004 to 2007, he was with the National Engineering School in Advanced Techniques (ENSTA), Paris, and the Laboratory of Signals and Systems (LSS), Supelec CNRS, Gif-sur-Yvette, France. He was a lecturer with the Department of Electronics and Computer Engineering, ENSTA, where his research activities were focused on improved reception schemes for ultrawideband communication systems. From December 2007 to September 2008, he was a Postdoctoral Researcher with the LSS, Supelec-CNRS, where he was involved in research projects with Alcatel-Lucent on satellite mobile communication systems. Since October 2008, he has been with the Faculty of Electrical and Computer Engineering, Shahid Beheshti University, where he is currently an assistant professor with the Department of Telecommunication. His current research interests include signal processing for wireless communications, with particular emphasis on multi-carrier and MIMO systems, joint channel estimation and decoding, iterative reception schemes, and interference cancellation under partial channel state information.

Salah Bourennane received his Ph.D. degree from Institut National Polytechnique de Grenoble, France, in 1990 in signal processing. Currently, he is a full professor at Ecole Centrale Marseille, and the head of Multidimensional Signal Processing team at Institut Fresnel, Marseille, France. His research interests include statistical signal processing, array processing, image processing, multidimensional signal processing, telecommunications, and performances analysis.