Abstract—In this paper, space–time bit–interleaved coded OFDM (orthogonal frequency division multiplexing) modulated systems are studied. Using multiple antennas at transmitter and receiver side, spatial diversity can be exploited with ordinary channel codes followed by an interleaver. Moreover, the combined use with OFDM modulation enable effective frequency selectivity exploitation in multipath channels. High spectral efficiency can be achieved using high modulation orders and increasing the number of transmitting antennas.

Asymptotical performances are analyzed using pairwise error probability evaluation. It is shown as transmission parameters and radio channel characteristics impact on achievable diversity order and coding gain.

Different channel codes are taken into account. Convolutional codes with several constraint lengths are simulated and results are compared with those of turbo coded solution.

The iterative receiver complexity grows at the increase of modulation order, subcarriers and number of antennas. Simplified detection criterions can be used for signals separation at receiver side. Turbo coded system allows for iterative approaches with reduced complexity when some turbo decoding iterations are performed between successive detection steps.

Index Terms—Space–frequency coding, OFDM, MIMO, coding gain, turbo detection, turbo decoding.

I. INTRODUCTION

The use of multiple antennas is considered as a promising way to increase modern wireless communication systems performances. Multiple–input multiple–output (MIMO) radio interfaces have been carefully studied over the last years as the works in [1] and [2] most of all. MIMO architectures combined with channel coding allow reliable transmissions at high data rate even in fading channels.

A practical design of telecommunication systems have to consider also the complexity required for the implementations. Although, a single carrier system could exploit channel time and space diversities [3], [4], a convenient technique for wideband transmission is OFDM modulation. OFDM allows very simple and effective equalization methods at the cost of partial spectral efficiency loss due to cyclic prefix insertion. Channel frequency selectivity can be well exploited using effective coding techniques working over OFDM subcarriers. Moreover, if channel state information were available at the transmitter (not in this paper) bit and power loading algorithms could be used for an optimal resource allocation over each subcarrier.

The joint use of MIMO and OFDM techniques is straightforward. In this paper we consider bit–interleaved coded modulation extended to MIMO systems with OFDM. We refer to this technique as: space–frequency bit–interleaved coded modulation (SF–BICM). The basic concepts of such scheme have been addressed in [5]. In that paper, an analytical treatment has pointed out as diversity gain improves system performance with respect to the single antenna case. It has been shown that such system is able to exploit spatial and temporal diversity offered by a frequency selective time correlated channel.

In this paper we extend the work in [5] assuming spatial correlation too. Using pairwise error probability (PEP) evaluation in the way used in [6] we show as system parameters (channel code type, interleaver pattern, modulation order and number of OFDM subcarriers) impact on diversity gain. Afterwards, we focus our analysis on the coding gain. For a given transmitter configuration we show that coding gain depends on the channel power delay profile and on the spatial/temporal correlation between channels taps. We relate these factors with the eigenvalue spread of the MIMO channel.

The selection of the channel code is an important aspect to determine overall performances. We here consider convolutional codes with different constraint lengths. Turbo codes [7] with short block length are also included in our analysis.

The transmission system here adopted is very flexible to data rate changes. Considering a fixed bandwidth and number of subcarriers, different spectral efficiencies can be obtained changing transmitter parameters as: coding rate, modulation order and number of transmitting antennas. This distinctive feature makes the system a good candidate for improved radio interfaces. Simulations reported in this paper refer to wireless local area network (WLAN) scenario. This is of topical interest because a high throughput radio interface for new generation WLAN is under investigation within IEEE (working group IEEE 802.11n) in these days.

In addition to performance requirements it is important to try to limit global system complexity. On the other hand, the receiver can get very complex when a lot of antennas and high modulation orders are used. An iterative strategy can be adopted to separate symbols detection to sequence decoding. Detection based on maximum a posteriori (MAP) criterion is a good reference but it is still too much complex [5]. In [8] minimum mean square error (MMSE) criterion has been proposed instead of MAP one. However, detection operations have to be carried out for each iteration when convolutional codes are used. Turbo codes allow instead for a different iterative approach. In fact, within iterative approach, detection can be performed after some turbo decoding iterations when soft information improve their reliability. This could reduce the receiver complexity without heavy performance loss.

The rest of the paper is organized as follow. In Section II we review the system model. Transmitter, channel model and receiver are described. In Section III we analyze system performance using pairwise error probability. Particular emphasis is laid on diversity order and coding gain computation for frequency selective channels. Section IV contains simulation results useful for analysis validation. System behavior is put on for different channel codes and propagation conditions. Finally, Section V concludes the paper.
II. COMMUNICATION SYSTEM DESCRIPTION

A. Transmitter Model

The block diagram of the transmission chain is depicted in Fig. 1(a). A binary information sequence \( b = \{b_0, b_1, \ldots, b_{N-1}\} \) is passed through a channel code with coding rate \( R_c \). The resulting binary codewords \( c = \{c_0, c_1, \ldots, c_{N_{c}-1}\} \), \( N = N_c R_c \), is then interleaved before multiplexing over \( N_t \) branches. Each branch sequence is likewise mapped using canonical constellation types, e.g. M-QAM, with zero-mean and average symbol energy \( E_s \). The symbols vector \( \mathbf{A}^t = \{A_1^t, A_2^t, \ldots, A_{P}^t\}, t = 1, 2, \ldots, N_t \), is transmitted over the \( t \)-th antenna after OFDM modulation and digital-to-analog conversion. For simplified equalization purpose the OFDM symbol, with \( F \) subcarriers, is properly extended using a cyclic prefix whose length is fixed to \( C \) samples. The transmission filter \( g(t) \) is a square root raised cosine pulse with roll-off factor \( \rho \). The frame size is given by \( FN_t N_c \log_2(M) = N \) bits, where \( N_s \) is a whole number of consecutive OFDM symbols.

B. Channel Model

We make use of frequency-selective channel models for the radio links description between the \( t \)-th transmitting antenna and \( r \)-th receiving one, \( r = 1, 2, \ldots, N_r \). A widespread discrete model assumes wide-sense-stationary channel impulse response where a tapped delay line describes the multipath components. Therefore the baseband channel impulse response can be written as [3]

\[
h_{t,r}^*(t) = \sum_{p=1}^{P} h_{t,r}^{*p} \delta(t - \tau_p),
\]

where \( P \) is the number of separably paths and \( \delta(t) \) is the Kronecker delta operator. The attenuation of the \( p \)-th path delayed by \( \tau_p \), \( h_{t,r}^{*p} \), is assumed to be Rayleigh distributed. For a simplified notation we assume equals power delay profiles for each multipath link:

\[
\sigma_p^2 = E[|h_{t,r}^{*p}|^2], \quad t = 1, \ldots, N_t, \quad r = 1, \ldots, N_r,
\]

under power constraint \( \sum_{p=1}^{P} \sigma_p^2 = 1 \). We don’t exclude temporal (\( p \) direction) and spatial (\( t \) and \( r \) directions) correlation among channel impulse responses. It is deducible from (2) that we have assumed a static channel over \( N_s \) OFDM symbols. This hypothesis is usually referred to block fading assumption. In this case the channel coherence time is greater than \( N_s \) times OFDM symbol duration. We could choose \( N \) in order to comply with that.

C. Receiver Model

The signal on the \( r \)-th receiving antenna is analog-to-digital converted, Fig. 1(b). We perform optimal synchronization with respect to oscillators frequency offsets and propagation time delays. Then, the cyclic prefix is optimally removed from the match filter output. Perfect channel state information is also assumed. For a given subcarrier \( k \), \( k = 1, 2, \ldots, F \), the OFDM demodulated signal reads:

\[
y_k = \sum_{t=1}^{N_t} H_k^{t,r} A_k^t + n_k^r,
\]

where the additive noise term \( n_k^r \) is a sample of a filtered stationary complex zero mean white gaussian process independent of \( k \) and \( r \) with power spectral density \( N_0/2 \) per dimension. The complex tap \( H_k^{t,r} \) is the channel frequency response of the multipath radio channel \( h_{t,r}^*(t) \) corresponding to the \( k \)-th subcarrier:

\[
H_k^{t,r} = \sum_{p=1}^{P} h_{t,r}^{*p} e^{-j2\pi k \tau_p/r} = h_{t,r} w_k^r,
\]

where we have defined

\[
w_k = [e^{-j2\pi k \tau_1/r}, e^{-j2\pi k \tau_2/r}, \ldots, e^{-j2\pi k \tau_P/r}]^T, \quad \text{and} \quad h_{t,r} = [h_{t,r}^{1}, h_{t,r}^{t,r}, \ldots, h_{t,r}^{N_r}]^T.
\]

The overall received signal associated to \( k \)-th subcarrier can be rearranged in matrix form as follow

\[
y_k = \begin{bmatrix} y_k^1 \\
\vdots \\
\vdots \\
\vdots \\
y_k^{N_r} \end{bmatrix} = \begin{bmatrix} H_k^{1,1} & \cdots & H_k^{1,N_r} \\
\vdots & \ddots & \vdots \\
\vdots & \ddots & \vdots \\
H_k^{N_r,1} & \cdots & H_k^{N_r,N_r} \end{bmatrix} \begin{bmatrix} A_k^1 \\
\vdots \\
\vdots \\
\vdots \\
A_k^{N_r} \end{bmatrix} + \begin{bmatrix} n_k^1 \\
\vdots \\
\vdots \\
\vdots \\
n_k^{N_r} \end{bmatrix}
\]

or more briefly

\[
y_k = H_k A_k + n_k.
\]

Thanks to OFDM modulation proprieties, symbols carried over distinct subcarriers are orthogonal. The receiver has to separate and to equalize symbols transmitted by different antennas and carried on a given OFDM subchannel. Moreover, OFDM modulation allows inter–symbol interference free transmissions. Section IV reports more details about receiver structure.

Let’s analyze some theoretical limits achievable with an optimum receiver.

III. ASYMPTOTICAL PERFORMANCE EVALUATION

An ordinary mathematical tool used for asymptotical performance analysis is the pairwise error probability (PEP) [2], [3], [6]. In the sequel we have a look at PEP expression for our system.

Let’s define the complex symbol distance vector between the transmitted symbol \( A_k \) and its estimation \( \hat{A}_k \), for the \( k \)-th subcarrier, as

\[
e_k = \hat{A}_k - A_k.
\]

The conditional PEP is the probability that the receiver erroneously decides for a vector of symbols \( \hat{A} \) when \( A = [A_1^1, \ldots, A_1^{N_r}, \ldots, A_F^1, \ldots, A_F^{N_r}]^T \) has been transmitted for a given channel realization. It can be rewritten as:

\[
P(\hat{A} \rightarrow A_k | H_k, k = 1, \ldots, F) = Q(\sqrt{D_{A_k} E_r^i / 2 N_0}),
\]

where \( Q(\cdot) \) is the 2-dimensional Frobenius norm. \( E_r^i \) indicates the expectation operation. \( \lambda(G) \), \( \lambda_1(G) \), \( \lambda_{\min}(G) \), and \( \lambda_{\max}(G) \) are respectively the set of eigenvalues, the \( i \)-th eigenvalue, the minimum nonzero eigenvalue, and the maximum eigenvalue of matrix \( G \). \( I_n \) is the identity matrix with dimension \( n \times n \).
where $Q(x)$ is the Gaussian $Q$-function and the distance metric is given by

$$D_{AA} = \sum_{k=1}^{F} \| H_k e_k \|^2. \quad (11)$$

Using Chernoff bound, $Q(x) \leq \exp(-x^2/2)$, the PEP is approximated by

$$P(A \rightarrow \hat{A}|\{H_k\}, k = 1, ..., F) \leq \exp\left(-\frac{E_s}{4N_0} \sum_{k=1}^{F} \| H_k e_k \|^2\right). \quad (12)$$

Let’s focus the attention on the expression of the distance $D_{AA}$. We collect the $r$-th row of $H_k$, whose elements are the channel frequency responses associated to the $k$-th subcarrier for each radio link connected to $r$-th receiving antenna:

$$H'_k = [h_{1,r}, \ldots, h_{N_r,r}] (I_{N_r} \otimes w_k) = h_r (I_{N_r} \otimes w_k). \quad (13)$$

The operator $\otimes$ represents the matrix direct product operator (Kronecker product). The distance metric can be rewritten as

$$\sum_{k=1}^{F} \| H_k e_k \|^2 = \sum_{k=1}^{F} \sum_{r=1}^{N_r} \| H'_k e_k \|^2 = \sum_{k=1}^{F} \sum_{r=1}^{N_r} \| h_r e_k \|^2 = \| \hat{h} E \|^2 \quad (14)$$

where $\hat{h} = [h_1, h_2, \ldots, h_{N_r}]$ contains the channel impulse responses of all multipath links among transmitting and receiving antennas. Let $\hat{e}_k = e_k \otimes w_k$ and $\hat{e} = [\hat{e}_1, \hat{e}_2, \ldots, \hat{e}_P]$, we have denoted in (14) $E = I_{N_r} \otimes \hat{e}$ that is a diagonal block matrix.

The overall channel autocorrelation matrix $R = E[\hat{h}^H \hat{h}]$ takes account of channels power delay profiles and spatial/temporal correlations. $R$ is a positive definite matrix and allows for Cholesky factorization: $R = F F^H$, where $F$ is a lower triangular matrix. For simplicity we have assumed $R$ full-rank. When it isn’t, we could extend our results using eigenvalue decomposition.

Using (14) in (12) and averaging the conditioned PEP over the complex gaussian vector $\hat{h}$ we obtain

$$P(A \rightarrow \hat{A}) = \frac{1}{\det(I_{PN_r} + \gamma F E E^H F^H)}, \quad (15)$$

where $\gamma = E_s/4N_0$ and $EE^H = (I_{N_r} \otimes \hat{e} \hat{e}^H)$. The identity matrix $I_{PN_r}$ is hermitian definite positive and $EE^H F^H$ is hermitian as well. Then, for high signal–to–noise ratios the bounded PEP can be expressed as

$$P(A \rightarrow \hat{A}) \leq \gamma^{-qN_r} \left( \prod_{i=1}^{N_r} \lambda_i \right)^{-1}, \quad (16)$$

where $q = \text{rank}(\hat{e} \hat{e}^H)$ and $\lambda_i$ are the nonzero eigenvalues of $EE^H R$. From (16) we observe that the maximum achievable diversity order for transmission over frequency selective channels with $P$ taps is equal to $qN_r$. The coding gain $\prod_{i=1}^{N_r} \lambda_i$ depends on $E$ and $R$. For instance, in absence of space correlation, $R$ is a diagonal block matrix and we’ll have $p$ eigenvalues with multiplicity $N_r$.

In the sequel we report some considerations on the achievable diversity order and on the coding gain.

**A. On the diversity order**

The rank $q$ can be rewritten as $q = \text{rank}(\hat{e} \hat{e}^H) = \text{rank}(\hat{e})$. Let $L$ the number of nonzero $e_k$, i.e. not entirely zero $e_k$, $k = 1, 2, ..., F$, it is straightforward to verify that the vector space generated by the columns of $\hat{e}$ has dimension $L$. Moreover, the dimension of the kernel generated by the rows of $\hat{e}$ is no greater than $PN_r$. The row rank is maximum when symbols transmitted on different antennas are independent, i.e. perfectly interleaved. We can conclude that $q = \min\{L, PN_r\}$. Based on the value of $q$ two transmission scenarios can be separately studied:

- $L \geq PN_r$. That occurs when the channel time delay spread is small, i.e. there isn’t enough frequency selectivity over the OFDM bandwidth. In this case the diversity order is equal to $PN_r N_r$ only if complex symbols are perfectly interleaved. The channel behavior sets a limit to the code capability. Diversity gain can be increased using more antennas at the transmitter and/or receiver sides.
- $L < PN_r$. This occurs when high channel time delay spreads are measured, that it is typical of WLAN applications. The diversity gain become $L N_r$. $L$ depends on the transmitter design parameters: coding, mapping, number of subcarriers, and interleaver pattern. Diversity gain is increased with the number of receiving antennas $N_r$ but it is independent by the number of transmitting antennas $N_t$. However spectral efficiency increase with $N_t$. The transmitter parameters set a limit to diversity exploitation.

These results could be intuitively understood considering the channel coherence bandwidth $B_{coh}$. In fact for a transmission frequency band $B$ and a discrete channel model with $P$ taps equally spaced by $1/B$, we have $B_{coh} \approx B/P$ [3].

We emphasize that for high signal–to–noise ratios the effect of the channel power delay profiles disappear. The only thing that matters is the number of discrete taps in the channel impulse response.

These results are in agreement with [5] where only time correlation has been considered. Furthermore, we remark that in our analysis the diversity level depends on the (minimum) complex symbol distance and not on the product distance. This last result is in agreement with what obtained for single antenna, single carrier, bit-interleaved coded modulation, e.g. [3], [9].

**B. On the Coding Gain**

The autocorrelation matrix $R$ takes part on the eigenvalues calculation in (16). The power delay profile shape impacts on the coding gain as we are going to explain in this subsection.

In [10] has been shown as coding gain decreases more and more with temporal and spatial channel correlation. Under the hypothesis of full diversity exploitation, the authors have analyzed the determinant of $R$ for different transmission conditions.

In our system when $L \geq PN_r$, $R$ is full–rank, and $\gamma$ is high, the bounded PEP (15) can be rewritten as

$$P(A \rightarrow \hat{A}) \leq \frac{\gamma^{-qN_r}}{\det(EE^H)^{-1}}; \quad (17)$$

as in [10] the coding gain will be increased if $\det(R)$ is increased as well. When $L < PN_r$, $EE^H$ is not still full-rank and the previous bound fails. However, for any $L$, $P$, $N_r$, and $N_t$, the matrix $F$ is square and nonsingular so we can exploit the Ostrowski’s theorem [11] and write:

$$P(A \rightarrow \hat{A}) \leq \gamma^{-qN_r} \prod_{\lambda_i(EE^H) \neq 0} \vartheta_i \lambda_i(EE^H) \lambda_i(EE^H)^{-1}, \quad (18)$$

where the coefficients $\vartheta_i$ owing to Ostrowski’s factorization belong to: $\min_i \{\lambda_i(R)\} \leq \vartheta_i \leq \max_i \{\lambda_i(R)\}$. Using the following
λ1 \left(EE^H\right) \geq \left(\lambda_{\min}(R)\right)^{\eta N_c} \prod_{\lambda_i \left(EE^H\right) \neq 0} \lambda_i \left(EE^H\right), \tag{19}

we can conclude that the PEP bound is minimized when the minimum eigenvalue of $R$, $\lambda_{\min}(R)$, is maximized. For instance, the minimum PEP value is achieved when $\lambda_{\min}(R) = \lambda_{\max}(R)$. This is the case of full temporal and spatial uncorrelation with a uniform channel power delay profile.

We observe that the maximum and minimum eigenvalues of the correlation matrix are closely link up to the eigenvalue spread of $R$: $\chi(R) = \lambda_{\max}(R) / \lambda_{\min}(R)$, and to the power spectral density of the overall channel impulse response $h$ as well [12]. For $\chi(R) = 1$ there is no mutual correlation among channels taps and the channel power delay profiles are uniform. $\chi(R)$ increases both with a no uniform power delay profile and with correlation. If temporal correlation (Doppler effect) or/and spatial correlation (small distance between antennas) is increased, coding gain will decrease. The eigenvalue spread has been used in [13] for the analysis of the MIMO system capacity.

From the point of view of space-frequency coding, the product distance $EE^H$ determines the coding gain.

IV. ANALYSIS VALIDATION THROUGH SIMULATION

We have shown in Section II that a per-tone demapper and equalizer can be used for symbol detection. This block has to extract a soft reliability value for each coded bit. An iterative process can be established (see Fig. 1(b)) when soft values are exchanged between detector and decoder. Detector can be implemented using either MAP (maximum a posteriori) criterion or MMSE (minimum mean square error). The second one allows for poorer performance but also for a lower complexity implementation with respect to MAP solution. We refer to [5] and [8] for a more detailed explanation of these criterions applied to SF–BICM.

A. System parameters

In our simulations we look at WLAN applications. System modulation parameters are chosen in agreement with IEEE 802.11a standard. Briefly, the transmission bandwidth is $B = 20$ MHz, $F_o = 52$ subcarriers are used for data transmission while $F_v = 12$ are virtual subcarriers, the prefix cyclic length is equals to $C = 16$. The transmission filter is a square root raised cosine pulse with roll-off factor $\rho = F_c / (F_a + F_v)$. The channel models have an exponential decaying power delay profile with an average root-mean square $t_{rms}$ [3]. Two antennas are used at the transmitter and receiver, $N_t = N_r = 2$. We define the average signal-to-noise (SNR) at the receiver side as $SNR = E_s N_t N_r / N_0$.

The channel code can be a convolutional code or a turbo code. The coding rate is selected, $R_c = 1/2$, so the system spectral efficiency roughly reads: $R_t = R_c N_t \log_2(M)$, i.e. $R_t = 2$ bit/s/Hz for QPSK and $R_t = 4$ bit/s/Hz for 16-QAM.

MAP criterion [14] is used for decoding. Log–likelihood ratios (LLRs) are extracted for both information and parity bits.

The interleaver selected is pseudo-random with length $N = 1248$ for all configurations. We remark that interleaver has the double duty of achieving diversity and coding advantage as discussed in Section III and furthermore it guarantee independence of errors between detector and decoder.

B. On the effect of the channel

In Fig. 2(a) the frame error rate (FER) is shown for different transmission environments. The convolutional code (5, 7), constraint length $K = 3$, is used. Spatial correlation arises when the distance between transmitting or receiving antennas is not large enough. For the simulation of this phenomena (no time correlation is introduced) we use the model proposed in [15], where the autocorrelation matrix $R$ has been splitted between transmitter and receiver in the form: $R = R_{Tx} \otimes R_{Rx}$. We have:

$R_{Tx} = \begin{bmatrix} 1 & -0.7275 + j0.2990 \\ -0.7275 + j0.2990 & 1 \end{bmatrix}$

$R_{Rx} = \begin{bmatrix} 1 & 0.1728 + j0.3288 \\ 0.1728 + j0.3288 & 1 \end{bmatrix}$

In Fig. 2(a) it is evident the diversity gain obtained with respect to the flat fading case ($P = 1$) when the channel becomes frequency
selective. Moreover for $\tau_{\text{fs}} = 30\text{ns}$ and $\tau_{\text{ms}} = 100\text{ns}$ the system falls in the case $L < PN$ and the diversity gain doesn’t change. The channel eigenvalue spread is greater for $\tau_{\text{ms}} = 30\text{ns}$ than for $\tau_{\text{fs}} = 100\text{ns}$, and as predicted in Section III coding gain increases. Spatial correlation doesn’t impact on diversity order but only on coding gain.

C. On the channel code

The choice of the channel code, interleaver and modulation format impact on the matrix $E$ hence on the $L$ value. In Fig. 2(b) simulations results are reported for convolutional codes with different constraint lengths and a turbo code with two equal constituent codes with $K = 4$. For a fixed interleaver and modulation order, we observe as coding gain increases with $K$. Best performance are obtained with the turbo codes though the turbo code interleaver is rather small, i.e 618 bits, and parity bits are alternatively punctured.

The use of turbo codes allows different iterative strategies. In Fig.2(b) the notation $\lceil n+m \rceil$ used in the legend means that for each turbo detection iteration, with a maximum of $n$ $m$ turbo decoding iteration are performed, Fig. 1(b). For instance, in Fig.2(b) turbo code with MAP detection $\lceil 1+6 \rceil$ performs as convolutional code $\lceil 133, 171 \rceil$ with the strategy $\lceil 6+1 \rceil$.

D. Simplified iterative approaches using turbo codes

The iterative decoding for turbo codes enables sub-optimal simplified detection approaches. A turbo detection iteration can be performed after some turbo decoding iterations, Fig. 1(b). In our analysis we have to take account that per-tone demapping and equalization heavily impacts on system complexity. Using MAP criterion, the complexity increase exponentially with the number of antennas and modulation order, and linearly with the number of subcarriers. It becomes soon prohibitive from implementation point of view. Instead of MAP we can use the simpler MMSE criterion [8]. In Fig. 3 some simulation results are reported with different choice of $\lceil n+m \rceil$ for QPSK and 16–QAM modulations. Increasing the modulation order the performance loss between $\lceil 6 + 1 \rceil$ and $\lceil 1 + 6 \rceil$ increases. The strategy $\lceil 2 + 3 \rceil$ is halfway between the other two. One drawback of turbo codes with respect to convolutional ones is given by their higher sensitivity to channel correlation and on the potential additive latency due to the internal interleaver.

Recently (and after the acceptance of this paper), in [16] this strategy has been proposed for MIMO applications with a large number of antennas and it has been called "double-iterative decoding".

V. Conclusion

In this paper, space–time bit–interleaved coded modulation with OFDM have been discussed. In such a scheme, transmission parameters can be easily changed in order to obtain flexible data rates with different levels of reliability. It has been shown as, for a given channel code with prefect interleaving, the achievable diversity order and coding gain depend on the channel impulse response length and on the power delay profile shape. The analysis reported in this paper connects system performance with transmitter design and radio channel characteristics.

Receiver complexity is a great drawback. To overcome this, suboptimal MMSE detection can be performed instead of the MAP one. Turbo codes, with respect to convolutional codes, allow for different mixed iterative approaches when some turbo decoding iteration are performed before a new detection step.

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