Performance Evaluation of Turbo Decoding in DFTS-OFDM Systems over V2V Channel

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***Abstract*—Vehicle-to-vehicle (V2V) doubly-dispersive channels cause loss of orthogonality in multi-carrier systems, resulting in inter-carrier interference (ICI). In order to suppress ICI in orthogonal frequency division multiplexing (OFDM) systems, the techniques of forward error correction (FEC) and nonlinear detection of data are usually employed, requiring low coding rates and high decoding computational complexity. In this paper, different configurations for reception of OFDM signals with frequency dispersion, discrete Fourier transform-spread-OFDM (DFTS-OFDM) are explored including FEC techniques and linear/nonlinear detection. The results show that the use of turbo encoding and non-linear detection of ordered successive interference cancellation (OSIC) in the DFTS-OFDM system offers significant gains in terms of performance when compared to conventional OFDM receivers, obtaining a performance in terms of bit error rate (BER) of 10^{-4} with a signal-to-noise ratio (SNR) close to 7 dB over doubly selective channel with 1 kHz Doppler spread. The results found can be applied in the design of communication systems with high mobility.**

***Index Terms*—DFTS-OFDM, ICI mitigation, Double selective channels, Time-varying channels, V2V, Turbo code, OSIC.**

I. Introduction

It is widely thought that vehicle-to-vehicle (V2V) communications will facilitate many future automotive applications related to road safety, diagnostics, and autonomous driving [1]. The mobile environment as well as the high carrier frequency of 5.9 GHz cause the V2V links to experiment doubly selective channels (DSC) [2], [3]. Under these scenarios, variations high Doppler spread alters the orthogonality among carriers, which causes inter-carrier interference (ICI). The ICI impacts the performance of receiver, mainly in the stages of channel estimation and data detection, resulting in higher decoding error rates [4].

In order to mitigate the ICI some detection techniques, e.g. minimum mean squared error (MMSE) [5], [6], maximum likelihood (ML) [7], decision-feedback filter (DFE) [8], and ordered successive interference cancellation (OSIC) [9], have been proposed. The analysis in [5] and [6] is limited to linear detectors which are not designed for ICI mitigation. Approach [7] obtains optimal performance in terms of bit error rate (BER), but it requires a high decoding computational complexity. The approaches [8], [9] achieve a better performance compared to linear detection. However, the approaches in [7]–[9] require higher computational complexity than $\mathcal{O}(N^3)$, where N denotes the OFDM symbol length. Furthermore, these approaches have only been evaluated in conventional OFDM systems without exploiting frequency diversity at signal level in data.

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The forward error correction (FEC) adds redundancy to the transmitted bits in order to compensate bit errors caused by either noise, fading or ICI, decreasing the overall spectral efficiency. There are several works dedicated to analyze the performance of turbo decoder on channels with interference, experimenting for different parameters such as block size, code rate, number of iterations, etc [4], [10], [11].

The combined performance of FEC and nonlinear detection stages on DSC has only been explored in conventional OFDM systems without exploiting the possible inclusion of frequency diversity in the data [10], [11]. In this work, the performance of multi-carrier receivers with different configurations is analyzed, in particular an OFDM system with data diversity (DFTS-OFDM) capable to incorporate the nonlinear detection OSIC and turbo decoding simultaneously is exposed, as a proposal to combat the ICI in scenarios with high mobility. It will be demonstrated that the performance of turbo decoder increases when operating in DFTS-OFDM system with nonlinear detection OSIC, in comparison with a conventional OFDM system with linear detection.

The paper is organized as follow. Section II describes the signal model with frequency diversity in data. In Section III, we expose non-linear detector OSIC for DFTS-OFDM system, used to combine the ICI of the V2V channel. Section IV describes the computational complexity of the common FECs and detectors found in literature. In Section V, we present the simulations. Finally, Section V provides the final conclusions.

Notation: Lower- (upper-)case letters refer to vectors (matrices); $[\cdot]^T$, $(\cdot)^H$, $(\cdot)_N$ and $[\cdot]_T$ are the transpose, Hermitian, circular module slipping N and band truncation operators, respectively; $(\cdot)^k$ refers to the k-th OFDM symbol being considered. The subscripts $(\cdot)_p$ and $(\cdot)_d$ are the sampled versions of the vectors in the positions of the pilots and data: and in the case of matrices, the versions sampled in rows and columns in the positions of pilots and data.

II. System model

The parts that conform the transmitter are shown in Figure 1. Consider a DFTS-OFDM transmission on a doubly selective V2V channel, the sequence of bits of information $a_j \in \{0,1\}$ for $j = 0,1,...,J-1$ from top layers are coded using a FEC of R_c code rate. The FEC generates the $b_i \in \{0,1\}$ codified bits for $i = 0,1,...,I-1$. The codified bits are interleaved in blocks of m elements, for the purpose of minimizing burst error in the decoder. The c_m bits are grouped in blocks of M elements, each block is mapped to the k-th symbol belonging to a set of finite alphabet Ω with cardinality equal to $|\Omega| = 2^M$. The data symbols are grouped in blocks of length N_d, forming the data vector $\mathbf{s}_d$. Through the discrete Fourier transform (DFT) the vector of data $\mathbf{s}_d$ is linearly precoded, with the aim of exploiting the frequency diversity. χ^k denotes the k-th precoded data symbol composed of the N_d samples in the frequency domain (FD). The pilot and guard symbols are assigned to the subcarrier index indicated by the standard 8702.11p [12], forming the vector $\mathbf{s}$ composed of the symbols $s[m]$ for $m = 0,1,...,N-1$. For the OFDM transmission, the k-th OFDM symbol in the time domain (TD) is obtained using the inverse discrete Fourier transform (IDFT) of size N, finally a cyclic prefix (CP) of length $\nu = (\frac{1}{16})N$ is inserted.

Let $x^k[n]$ be the n-th time samples of k-th symbol OFDM in time domain (TD) without including the CP, can be expressed by:

$$x^k[n] = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} s^k[m] e^{j2\pi nm/N}, \quad n = 0,...,N-1. \quad (1)$$

Once the CP exclusion is done, the received signal for the k-th symbol in the receiver in its complex baseband representation can be described with the following discrete circular convolution:

$$y^k[n] = \sum_{l=0}^{L-1} h^k[n,l] x^k[\langle n-l \rangle_N] + w^k[n], \quad (2)$$

where, $n = \{0,1,...,N-1\}$, $l = \{0,1,...,L-1\}$, L denotes the baseband channel impulse response (CIR) length, $h^k[n,l]$ is the CIR of the k-th block in the n instant for an impulse function as input in the l previous samples, and $w^k[n]$ is the complex additive white Gaussian noise (AWGN), with zero mean and variance $\sigma_w^2 = N_0/2$. The circular convolution (2) between the CIR and $x^k[n]$ can be rewritten in matrix form as:

$$\mathbf{y}^k = \mathbf{H}^k \mathbf{x}^k + \mathbf{w}^k, \quad (3)$$

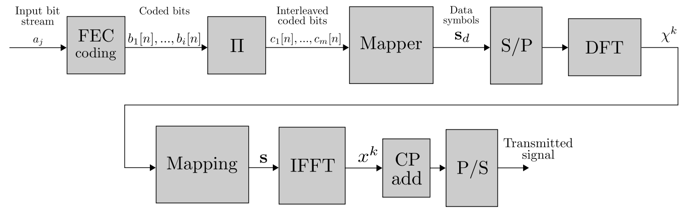

Fig. 1: DFTS-OFDM transmitter.

where:

$$\begin{aligned}
\mathbf{y}^k &= \begin{bmatrix} y^k[0], & y^k[1], & \cdots, & y^k[N-1] \end{bmatrix}^T, \\
\mathbf{x}^k &= \begin{bmatrix} x^k[0], & x^k[1], & \cdots, & x^k[N-1] \end{bmatrix}^T, \\
\mathbf{w}^k &= \begin{bmatrix} w^k[0], & w^k[1], & \cdots, & w^k[N-1] \end{bmatrix}^T,
\end{aligned}$$

In addition, $\mathbf{H}^k$ is the channel matrix of size $N \times N$ whose elements are formed by the coefficients of the CIR with the following assignment:

$$\left[\mathbf{H}^k\right]_{n,n'} = h^k\left[n, \langle n-n' \rangle_N\right], \quad (4)$$

where $n, n' = \{0,1,...,N-1\}$ and CIR is assumed to be zero for $\langle n-n' \rangle_N > L-1$. The OFDM symbol received in FD is obtained by multiplying both sides of (3) by the matrix of the normalized discrete Fourier transform (DFT):

$$[\mathbf{F}]_{n,n'} = \frac{1}{\sqrt{N}} e^{(-j2\pi nn'/N)}, \quad (5)$$

which gives the result:

$$\mathbf{u}^k = \mathbf{F}\mathbf{H}^k\mathbf{x}^k + \mathbf{z}^k, \quad (6)$$

where $\mathbf{u}^k$ is the OFDM symbol received in the FD and $\mathbf{z}^k$ is the DFT of the noise vector. Since matrix $\mathbf{F}$ is unitary, the equation (6) can be rewritten in the following way:

$$\begin{aligned}
\mathbf{u}^k &= \mathbf{F}\mathbf{H}^k\mathbf{F}^H\mathbf{F}\mathbf{x}^k + \mathbf{z}^k, \\
&= \mathbf{F}\mathbf{H}^k\mathbf{F}^H\mathbf{s}^k + \mathbf{z}^k, \quad (7) \\
&= \mathbf{G}^k\mathbf{s}^k + \mathbf{z}^k, \quad (8)
\end{aligned}$$

where, $\mathbf{s}^k$ is composed of N_d precoded symbols belonging to vector $\chi^k = \mathbf{F}\mathbf{s}_d^k$, N_p pilot symbols and N_g guard symbols. $\mathbf{G}^k = \mathbf{F}\mathbf{H}^k\mathbf{F}^H$ is the channel frequency matrix (CFM). If the CIR is time varying, then $\mathbf{G}^k$ is a non-diagonal matrix implying ICI. By truncating the model to the transmitted and received vectors in data positions, the following observation model is obtained:

$$\mathbf{u}_d^k = \mathbf{G}_d^k \chi_d^k + \mathbf{z}_d^k. \quad (9)$$

III. Symbol detection

The signal model in (9) can be rewritten to allow its simple fitting in the nonlinear detection algorithm in the following form:

$$\begin{aligned}
\mathbf{F}^H\mathbf{u}_d^k &= \mathbf{F}^H\mathbf{G}_d^k\chi_d^k + \mathbf{F}^H\mathbf{z}_d^k, \\
&= \mathbf{F}^H\mathbf{G}_d^k\mathbf{F}\mathbf{F}^H\chi_d^k + \mathbf{F}^H\mathbf{z}_d^k,
\end{aligned}$$

$$\mathbf{y}_d^k = \mathbf{C}\mathbf{s}_d^k + \mathbf{w}_d^k, \tag{10}$$

where, $\mathbf{C} = \mathbf{F}^H \mathbf{G}_d^k \mathbf{F}$ and the vectors $\mathbf{y}_d^k$, $\mathbf{s}_d^k$ and $\mathbf{w}_d^k$ are the sampled versions of $\mathbf{y}^k$, $\mathbf{s}^k$ and $\mathbf{w}^k$ in data positions.

A. QR decomposition

It has been shown that Zero-Forcing (ZF) and MMSE detection can be represented in terms of the QR decomposition of the channel matrix [9], [13], having $\mathbf{C} = \mathbf{Q}\mathbf{R}$, where, $\mathbf{Q}$ is an orthogonal matrix of unit norm and $\mathbf{R}$ is an upper triangular matrix, both of size $N_d \times N_d$. Multiplying the received signal $\mathbf{y}_d^k$ by $\mathbf{Q}^H$ the following signal model is obtained:

$$\tilde{\mathbf{y}} = \mathbf{R}\mathbf{s}_d^k + \tilde{\mathbf{w}}, \tag{11}$$

where $\tilde{\mathbf{y}} = \mathbf{Q}^H \mathbf{y}_d^k$ and $\tilde{\mathbf{w}} = \mathbf{Q}^H \mathbf{w}_d^k$. Since matrix $\mathbf{Q}$ is unitary, the statistics of the noise vector $\mathbf{w}_d^k$ are not altered. As in [9] an algorithm of Sorted-QR was used for the calculation of the QR decomposition of $\mathbf{C}$, with MMSE criteria. Below is presented a nonlinear detection scheme to the signal model described by the equation (11).

B. OSIC detection

The combination of the ordered successive interference cancellation (OSIC) with the sorted QR decomposition allows for the implementation of a suboptimal detector. Due to the triangular structure of $\mathbf{R}$, the k-th element of the vector $\tilde{\mathbf{y}}$ can be obtained as follows:

$$\tilde{\mathbf{y}}_k = \mathbf{R}_{kk}\mathbf{s}_j + \sum_{i=k+1}^{N_D} \mathbf{R}_{ki}\mathbf{s}_i + \tilde{\mathbf{w}}_k. \tag{12}$$

The detection of the k-th received symbol is completed sequentially in the order $k = N_D, N_D - 1, \cdots, 1$ using the following expression:

$$\hat{s}_k = \mathcal{Q}\left[\frac{\tilde{\mathbf{y}}_k - \sum_{i=k+1}^{N_D} \mathbf{R}_{ki}\hat{\mathbf{s}}_i}{\mathbf{R}_{kk}}\right], \tag{13}$$

where, $\mathcal{Q}[\cdot]$ is a decision operator that maps its argument to the closest point in the constellation Ω and $\hat{s}_k$ is the k-th estimated symbol. The contribution of noise $\tilde{\mathbf{w}}$ is not considered due to the fact that its variance was not affected by the $\mathbf{Q}$ matrix' orthonormal property.

C. LLR calculation

The detected complex symbols of the data vector $\mathbf{s}$ are converted to Log-Likelihood Ratio (LLR) values, approximated by the equations reported in [14]:

$$D_{I,j} = \begin{cases} s_I[j], & j = 1 \\ -|D_{I,j-1}| + d_{I,j} & j > 1 \end{cases} \tag{14}$$

$$LLR(b_I, j) = \sigma^2 \cdot D_{I,j}, \quad j \geq 1, \tag{15}$$

and

$$D_{Q,j} = \begin{cases} s_Q[j], & j = 1 \\ -|D_{Q,j-1}| + d_{Q,j} & j > 1 \end{cases} \tag{16}$$

$$LLR(b_Q, j) = \sigma^2 \cdot D_{Q,j}, \quad j \geq 1, \tag{17}$$

TABLE I: Computacional complexity for signal decoding.

Decoder method	Complexity
Viterbi algorithm	$\mathcal{O}(K^2 T)$
Log-MAP (turbo decoder)	$\mathcal{O}(T 2^\eta (\eta + 1))$

K is number of combinations of states ($K = 2^r$), r is numeber of appliances, T is number of sampling point and η is number of bits of convolutional code [16].

TABLE II: Computational complexity in terms of complex products per OFDM symbol required for signal detection.

Detection method	Complexity
Full MLD	$\mathcal{O}(\Omega^N N_d)$
Conventional QR-MLD	$\mathcal{O}(\Omega^N N_d)$
LMMSE	$\mathcal{O}(N_d^3)$
OSIC	$\mathcal{O}(N_d^3)$

where the subindexes I and Q represent the phase and quadrature of the detected symbol, σ^2 is the SNR and $-|D_{I,j-1}|$ (or $-|D_{Q,j-1}|$) is the distance between $s(j)$ and the closest partition limit.

IV. Computational Complexity

This section contains a comparison of the most representative decoding and detection algorithms used on doubly selective channels. Table I compares the computational complexity required by decoding algorithms Log-MAP and Viterbi; it should be noted that both algorithms have similar computational complexity. Table II [15] indicates that the nonlinear detector OSIC has the same computational complexity as the conventional LMMSE detector.

V. Simulations and Results

Figure 2 shows the structure of the proposed receiver consisting of nonlinear detection and turbo decoding algorithms. This paper studies the performance of the turbo decoding with soft decision, comparing its performance with linear and nonlinear data detection. In the same way it compares the turbo decoding performance on the DFTS-OFDM and conventional OFDM systems. Furthermore, performance of the turbo decoding against the Viterbi decoding with hard/soft decision is also included. The experiments were completed in an 802.11p simulation environment through the multi-trajectory Reyleigh channel replicating the non-line-of-sight (NLOS) V2V scenario at $v = 100$ km/h, with a delay power profile exponentially decreasing, delay equal to $\tau_{RMS} = 0.4\mu s$ and Doppler spreading frequency of $f_D = 1$ kHz. Table III reports the system's configuration parameters [12]. The channel estimator implements the bidimensional basis expansion model (BEM-2D) reported in [6]. The turbo decoder operates over frames with soft decoded bits, in order to evaluate its fixed-point performance, the soft LLR data were quantized in 5 bits: one bit for the sign, another one for the integer part and three for the fractional part. To maintain the spectral efficiency, the same code ratio of $R_c = 1/3$ was used for the turbo codes and convolutional codes.

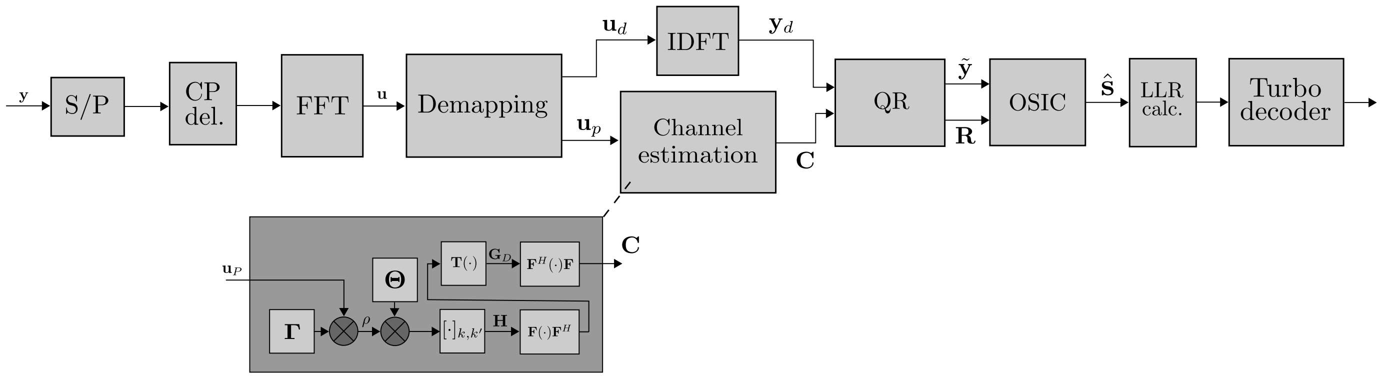

Fig. 2: Proposed receiver configuration with turbo decoder and OSIC detector.

TABLE III: Simulation parameters.

Parameter	Value
Modulation	OFDM, DFTS-OFDM
Number of OFDM subcarriers	$N = 64$
Number of data subcarriers	$N_d = 48$
Number of pilot subcarriers	$N_p = 4$
Bandwidth	$BW = 10$ MHz
Samples in the cyclic prefix	$CP = 16$
Sampling time	$T_s = 100ns$
Symbol period	$T_{OFDM} = 8\mu s$
Data modulation	QPSK
Type decoding	Turbo decoder, Viterbi
Coding rate	$R_c = 1/3$
Frame length	37 OFDM symbols
Channel model	Exponentially decaying , Rayleigh fading (rms delay spread = 0.4 μs, $f_D = 1$ kHz)
Channel estimation	BEM-2D [6]

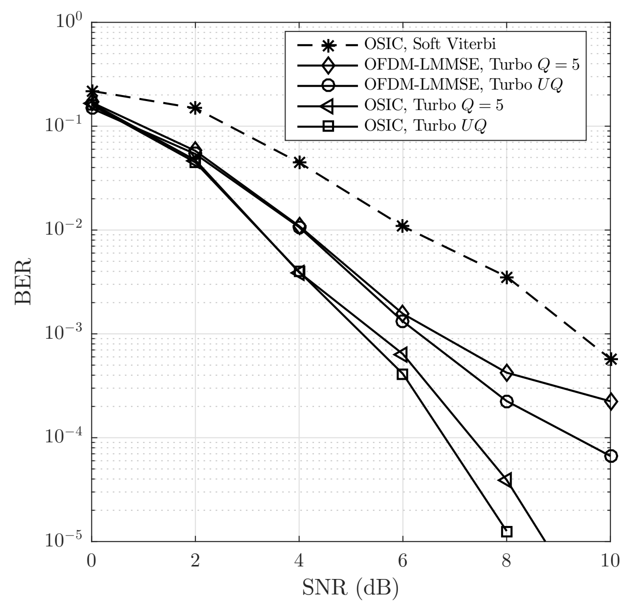

Fig. 3: BER vs SNR performance test of the turbo decoding of data detected by OSIC. The channel model uses a Jake's Doppler profile for each channel tap. The power delay profile is exponentially decaying with RMS delay spread of $0.4\mu s$ and frequency Doppler spread of 1 kHz modeling a Rayleigh fading NLOS scenario.

Figure 3 shows the BER-vs-SNR achieving DFTS-OFDM system with OSIC nonlinear detection. For the specific case in SNR = 5 dB, the OSIC detection with turbo decoding surpassed these nonlinear detection with Soft-Viterbi decoding by approximately 5 dB. For the level of SNR = 6 dB a superior performance of 4 dB was obtained with respect to the use of the turbo-decoder in a conventional OFDM system with LMMSE detection. It is observed that the turbo decoding in a conventional OFDM system and LMMSE detection has a noticeable error floor beyond the 10dB SNR level, this because the LMMSE detector is not suitable to mitigate ICI, causing the error of bits in the turbo decoding increase. The difference in performance of the turbo decoder with quantized (Q) data and unquantized (UQ) data is approximately 0.5 dB starting from SNR = 7 dB, with a BER inferior to 10^{-4} in both cases.

Figure 4 shows the BER vs SNR achieved by the turbo decoding with soft decision and OSIC detection, in comparison with the Viterbi decoding with hard/soft decision and linear/nonlinear detection, both in a system DFTS-OFDM . It can be observed that the detection proposed OSIC with soft decision Viterbi decoding surpass by 2 dB the performance achieved by the same detection with hard decision Viterbi. For the specific case of SNR = 13 dB, the OSIC detection with soft decision Viterbi surpass by 4 dB to the LMMSE detection with hard decision Viterbi, in addition to not presenting the floor error shown in the LMMSE detection in high SNR. The performance of the turbo decoder is superior to the Viterbi decoder, which justifies the extra computational cost in the receiver. The proposed OSIC detection in conjunction with the turbo decoding achieves a 6 dB reduction of SNR with respect to the OSIC detection with Viterbi decoding and an 8 dB reduction with LMMSE detection with Viterbi decoding to achieve a rate of BER = 10^{-4}.

VI. Conclusion

This paper compares distinct OFDM receivers for V2V channels. It was found that the DFTS-OFDM configuration with turbo decoding and OSIC detection achieves the best performance in BER at low levels of SNR in addition to a computational complexity which is close to that of conven-

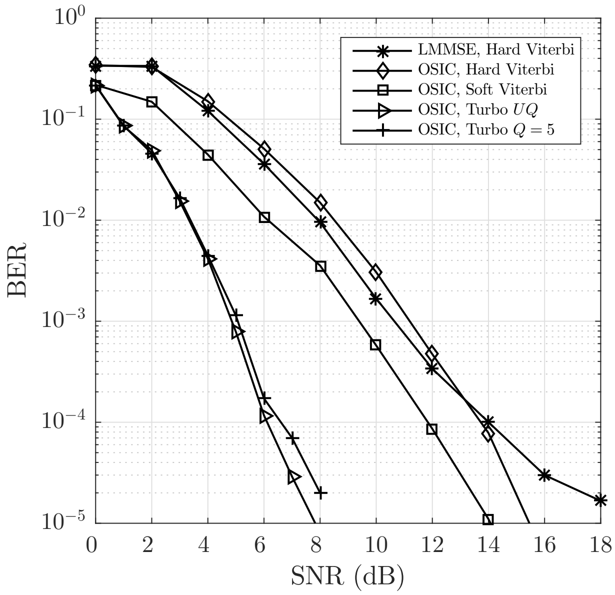

Fig. 4: BER vs SNR comparison of Viterbi decoding performance with detection: LMMSE and OSIC. The channel model uses a Jake's Doppler profile for each channel tap. The power delay profile is exponentially decaying with RMS delay spread of $0.4\mu s$ and frequency Doppler spread of 1 kHz modeling a Rayleigh fading NLOS scenario.

tional OFDM receivers. The results show that turbo decoding outperforms the Viterbi decoding by at least 6 dB when using hard/soft decision. Furthermore, the combination of turbo coding and nonlinear detection significantly outperforms turbo coding with linear detection, so that low complexity nonlinear detector become relevant. These results can be used to design multicarrier communication systems for efficient ICI mitigation in highly dispersive V2V channels.

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