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Efficient Index Modulation-Based MIMO OFDM Data Transmission and Detection for V2V Highly Dispersive Channels

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Abstract: Vehicle-to-vehicle (V2V) communication networks are based on vehicles that wirelessly exchange data, traffic congestion, and safety warnings between them. The design of new V2V systems requires increasingly energetically and spectrally efficient systems. Conventional multiple-input-multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems have been used successfully for the last decade. However, MIMO-OFDM systems need to be improved to face future communication networks in high-mobility environments. This article proposes an efficient index modulation (IM)-based MIMO-OFDM system for V2V channels. The proposed transmission system is evaluated in high Doppler-spread channels. The results demonstrate that the proposed scheme reduces the required computational complexity in data detection and exhibits gains of up to 3 dB in bit error rate (BER) performance when compared to the conventional MIMO-OFDM system under the same conditions and parameters, along with achieving superior spectral efficiency. The results show the viability of implementing the proposed system in practical applications for high-transmission-rate V2V channels.

Keywords: V2V channels; MIMO-OFDM; efficient index modulation

MSC: 68M10; 68M12; 90B18; 94A05



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1. Introduction

Vehicle-to-vehicle (V2V) communications comprise a wireless network wherein automobiles send each other information about their actions. These data include speed, location, direction of travel, braking, and loss of stability. Traffic accidents worldwide cause around 1.2 million fatal accidents annually. According to the World Health Organization (WHO), this corresponds to a quarter of deaths caused by injuries [1]. New technologies in wireless communications can provide mobile users with the information to anticipate possible accidents. In this way, it is possible to reduce the number of deaths that occur each year due to this cause [2]. In this sense, it is important to understand the characteristics of the communication channel. With this, adequate system design can be obtained. In particular, the characteristics of the V2V channel make proper system design challenging. The V2V channel has apparent differences compared with the traditional cellular communication channel, such as lower antenna height, faster moving speed of transmitters (TXs) and receivers (RXs), different frequency bands, etc. Therefore, it is important to conduct targeted research on the V2V channel [3–7]. In V2V scenarios with high mobility and multipath propagation, two significant factors come into play: the Doppler effect caused by high mobility and the delay spread caused by multipath propagation. When both phenomena impact the received signal, it leads to the so-called dual selective channel (DSC) [8,9].

In multi-carrier systems operating in such a scenario, two types of interference become inevitable: inter-symbol interference (ISI) and inter-carrier interference (ICI). These interferences severely impact the system performance, leading to a degradation in overall system performance. Multiple-input-multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems have been widely recognized to satisfy the increasing demand for high-data-rate wireless communications systems. OFDM is one of the most popular multi-carrier technologies and is part of many standards such as long-term evolution (LTE) and IEEE 802.11x wireless local area network (LAN). MIMO is an antenna technology for wireless communications that use multiple TX antennas and multiple RX antennas, which can be used for transferring more data at the same time (multiplexing) or to create multiple versions of the same signal, increasing the total energy at the receiver (diversity) [2,10–13]. Although systems based on MIMO-OFDM have been used successfully in communication systems such as those mentioned above. It is not known what the performance of these systems could be in more complex scenarios such as high-mobility V2V networks or in systems such as massive MIMO. Future communication systems must be improved by fast, efficient transmission technologies through a high mobility wireless communication medium for [14]. Communication systems of the sixth generation (6G) will be enhanced by the combination of new technologies, such as intelligent reflective surfaces (IRS) in massive MIMO systems, artificial intelligence, tera-hertz (THz) communications, and new cell-free architectures [14,15].

Recently, index modulation (IM) has emerged as one of the promising MIMO solutions for spectral- and energy-efficient next-generation communications systems [16]. The most widely studied IM technique is spatial modulation (SM). In spatial modulation (SM), each Tx and/or Rx antenna can be used as a point in the spatial constellation [17,18]. In this way, in SM, the complete array of Tx antennas is used to convey extra information in addition to transmission of the conventional quadrature amplitude modulation (QAM) symbol transmission. However, in addition to the spatial dimension, other features of a conventional communication system can be used to modulate. Existing IM schemes can be classified into various signal domain IM schemes, which include spatial-domain IM (SD-IM), channel-domain IM (CD-IM), frequency-domain IM (FD-IM), and time-domain IM (TD-IM) transmission schemes [19]. In [20], a simple OFDM-IM is implemented fusing radar and wireless communication systems. In [21], the basic OFDM-IM system uses modulation based on null subcarriers rather than active subcarriers to transmit the information. This system achieves better bit error rate (BER) performance and improved spectral efficiency (SE) as compared to the conventional OFDM-IM system. The SE of the conventional OFDM-IM scheme can be further improved by dividing the total number of subcarriers into subgroups [22]. In [23], an OFDM-IM system is used for orthogonal multiple access (NOMA) operating in dual mode (DM). The DM-OFDM-IM-NOMA increases the SE using indices selection bits and data transmission through all the OFDM subcarriers employing different constellation sets. With the aim of improving the SE of the system and inspired by the quadrature spatial modulation (QSM) technique, in this paper, we use the sub-carriers of the OFDM system to independently transmit the real and imaginary parts of one QAM symbol over the in-phase and quadrature dimensions, respectively. In particular, QSM is an SM-based technique that has the advantage of doubling the number of bits that can be transmitted in the spatial constellation, which results in improved SE compared to the basic SM scheme [24]. In addition to carrying information bits through conventional amplitude/phase modulation (APM)-based transmission, the proposed scheme allows for the harvesting of additional index bits without consuming extra energy [13]. This makes our proposal capable of employing only part of the resources to achieve the same throughput as its conventional counterpart. Results show that since this technique is based on inserting zeros in the transmission, the systems have a reduced detection complexity while consuming substantially less energy. Mainly, our simulations show that the proposed technique is adequate for high-mobility V2V channels. The contributions of this paper can be summarised as follows:

- (i) A novel IM-MIMO-OFDM system based on QSM is proposed for handling doubly selective MIMO-V2V channels.
- (ii) The MIMO receiver developed here maintains a structure and computational complexity lower than the conventional MIMO-OFDM receiver.
- (iii) The proposed system demonstrates high robustness against uncertainties in channel statistics, achieving a low BER even at high levels of Doppler frequency dispersion.

The remainder of this paper is organised as follows. Section 2 presents the system model of the proposed MIMO-OFDM-IM system. Section 3 describes the receiver, channel stimulation, and low-complexity detection for OFDM-IM signals. Section 4 analyses the computational complexity of the proposed system. Section 5 presents the performance analysis of the utilized algorithm and results. Finally, in Section 6, we conclude the work.

Notation

Lowercase (uppercase) bold letters are used for vectors (matrices); $(\cdot)^T$, $(\cdot)^H$, and $\lceil \cdot \rceil$ denote the transpose, Hermitian, and rounding up operators, respectively; $\langle \cdot \rangle_N$ denotes circular shift of modulus N ; $E\{\cdot\}$ is the expected value operator. The operator $(\cdot)^k$ refers to the k -th OFDM symbol being considered.

2. System Model MIMO-OFDM-IM

This article considers a MIMO-OFDM-IM system equipped with N_t transmitting antennas, N_r receiving antennas, and a transmission of N samples during a block period of T_s . The block diagram of the MIMO-OFDM-IM transmitter is depicted in Figure 1. Each system frame is composed of mN_t data bits. These bits are divided into N_t groups for each of the transmitting antennas, each of which contains m bits that will be modulated by the OFDM-IM technique for transmission by the N_t transmitting antennas. The m bits are further divided into N_B groups, each with b bits; hence, $m = N_B b$. The OFDM-IM symbol consists of $N = N_d + N_p + N_g$ subcarriers, where N_d , N_p , N_g are the carriers assigned to data, pilots, and guards, respectively. IM is performed by dividing the N_d subcarriers into $N_B = N_d / N_C$ sub-blocks, where N_C is the number of subcarriers per sub-block.

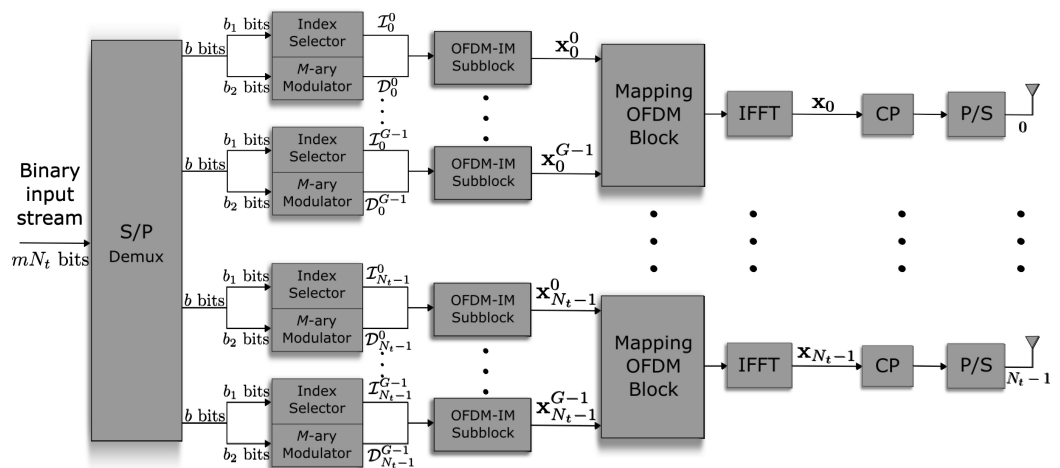


Figure 1. MIMO-OFDM-IM transmitter proposed.

Carrier index modulation (IM) transmits a different symbol inspired by the technique known as QSM and considers the N_C subcarriers within a sub-block as transmitting antennas. This allows for the transmission of $b = 2 \log_2(N_C) + \log_2(M)$ bits for each sub-block of OFDM-IM symbol, where M represents the size of the constellation Ω used with cardinality equal to $|\Omega| = 2^M$. OFDM-IM separates the b bits of each sub-block into two parts for different purposes: the first part with $b_1 = 2 \log_2(N_C)$ bits is used to select K active subcarriers, denoted by the following set:

$$\mathcal{I}_t^g = \{i_t^g(0), \dots, i_t^g(K-1)\}, \quad (1)$$

where $g = \{0, \dots, N_B - 1\}$, $i_t^g(k) \in \{0, \dots, K - 1\}$ for $k = \{0, \dots, N_c - 1\}$, and the active elements of $\mathcal{I}_t^g$ are assigned using a spatial constellation among the N_c subcarriers. The second part with $b_2 = \log_2(M)$ bits is mapped to a symbol $\mathcal{D}_t^g$ using an M -ary modulation. $\{\Re(\mathcal{D}_t^g(n))\}_{n \in \mathcal{I}_t^g}$ and $\{\Im(\mathcal{D}_t^g(n))\}_{n \in \mathcal{I}_t^g}$ are transmitted by the same or different subcarriers indicated by $\mathcal{I}_t^g$, where $\mathcal{D}_t^g$ is extracted from a complex alphabet, and we assume that $E\{|\mathcal{D}_t^g|^2\} = 1$ for signal normalization. Therefore, the OFDM-IM symbol element of the g -th sub-block in the t -th transmitting antenna can be expressed as $\mathbf{s}_t^g = [s_t^g(0) \ s_t^g(1) \ \dots \ s_t^g(N_c - 1)]^T$, where:

$$s_t^g(n) = \begin{cases} \Re(\mathcal{D}_t^g(n)), & n \in \mathcal{I}_t^g \\ \Im(\mathcal{D}_t^g(n)), & n \in \mathcal{I}_t^g \\ 0, & \text{otherwise} \end{cases} \quad (2)$$

Therefore, each sub-block of the OFDM-IM symbol contains one or two active carriers whose positions carry information through the subcarrier indices. Table 1 shows the carrier index assignment for a single sub-block with $N_c = 2$ using 4-QAM constellation. For example, in Table 1, for the input sequence $[0 \ 0 \ 0 \ 0]$, the real and imaginary parts of the symbol ($\mathcal{D} = +1 + j$) are transmitted by the first carrier of the sub-block, while the second carrier remains inactive. A detailed explanation of the QSM scheme can be found in [25].

Table 1. A look-up table example for $N_c = 2, M = 4$, and $b = 4$.

b Input Bits $\mathbf{a} = [a_0 \ a_1 \ a_2 \ a_3]$	$N_c = 2, M = 4$		Sub-Blocks
	Carrier Index	Transmit Symbol	
$[0 \ 0 \ 0 \ 0]$	$\{0, 0\}$	$+1 + j$	$[+1 + j \ 0]^T$
$[0 \ 0 \ 0 \ 1]$	$\{0, 0\}$	$-1 + j$	$[-1 + j \ 0]^T$
$[0 \ 0 \ 1 \ 0]$	$\{0, 0\}$	$+1 - j$	$[+1 - j \ 0]^T$
$[0 \ 0 \ 1 \ 1]$	$\{0, 0\}$	$-1 - j$	$[-1 - j \ 0]^T$
$[0 \ 1 \ 0 \ 0]$	$\{0, 1\}$	$+1 + j$	$[+1 \ +j]^T$
$[0 \ 1 \ 0 \ 1]$	$\{0, 1\}$	$-1 + j$	$[-1 \ +j]^T$
$[0 \ 1 \ 1 \ 0]$	$\{0, 1\}$	$+1 - j$	$[+1 \ -j]^T$
$[0 \ 1 \ 1 \ 1]$	$\{0, 1\}$	$-1 - j$	$[-1 \ -j]^T$
$[1 \ 0 \ 0 \ 0]$	$\{1, 0\}$	$+1 + j$	$[+j \ +1]^T$
$[1 \ 0 \ 0 \ 1]$	$\{1, 0\}$	$-1 + j$	$[+j \ -1]^T$
$[1 \ 0 \ 1 \ 0]$	$\{1, 0\}$	$+1 - j$	$[-j \ +1]^T$
$[1 \ 0 \ 1 \ 1]$	$\{1, 0\}$	$-1 - j$	$[-j \ -1]^T$
$[1 \ 1 \ 0 \ 0]$	$\{1, 1\}$	$+1 + j$	$[0 \ +1 + j]^T$
$[1 \ 1 \ 0 \ 1]$	$\{1, 1\}$	$-1 + j$	$[0 \ -1 + j]^T$
$[1 \ 1 \ 1 \ 0]$	$\{1, 1\}$	$+1 - j$	$[0 \ +1 - j]^T$
$[1 \ 1 \ 1 \ 1]$	$\{1, 1\}$	$-1 - j$	$[0 \ -1 - j]^T$

Each OFDM-IM symbol contains G OFDM-IM sub-blocks in each transmitter branch, denoted by $\mathbf{s}_t$. Each OFDM-IM block is transformed to the time domain using the normalized inverse discrete Fourier transform (IDFT) matrix of size $N \times N$ for transmission.

$$\mathbf{x}_t = \mathbf{F}^H \mathbf{s}_t, \quad (3)$$

where $\mathbf{F}$ is the matrix of the normalized discrete Fourier transform (DFT):

$$[\mathbf{F}]_{n,n'} = \frac{1}{\sqrt{N}} e^{(-j2\pi nn'/N)}, \quad (4)$$

The complex baseband model of the received signal at the r -th receiving antenna at time instant n can be expressed as follows:

$$y_r[n] = \sum_{t=0}^{N_t-1} \sum_{l=0}^{L-1} h[r, t; n, l] x_t[\langle n-l \rangle_N] + w_r[n], \quad (5)$$

where $n = \{0, 1, \dots, N-1\}$, $r = \{0, 1, \dots, N_r-1\}$, $x_t[n]$ represents the signal obtained from the t -th transmitting antenna, L denotes the number of channel taps, and $w_r[n]$ represents the circular complex delta correlation function of the additive white Gaussian noise on the r -th receiver antenna. We can express Equation (5) in a matrix–vector form as shown below:

$$\mathbf{y}_r = \sum_{t=0}^{N_t-1} \mathbf{H}(t, r) \mathbf{x}_t + \mathbf{w}_r \quad (6)$$

where

$$\begin{aligned} \mathbf{y}_r &= [y_r[0] \quad y_r[1] \quad \cdots \quad y_r[N-1]]^T, \\ \mathbf{x}_t &= [x_t[0] \quad x_t[1] \quad \cdots \quad x_t[N-1]]^T, \\ \mathbf{w}_r &= [w_r[0] \quad w_r[1] \quad \cdots \quad w_r[N-1]]^T, \end{aligned}$$

Additionally, $\mathbf{H}(t, r)$ is the MIMO channel matrix that contains the channel impulse response (CIR) from the t -th transmitting antenna to the r -th receiving antenna, the elements of which are formed by the coefficients of the CIR as follows:

$$[\mathbf{H}(t, r)]_{n, n'} = h[t, r; n, \langle n-n' \rangle_N], \quad (7)$$

Assuming that $n, n' = \{0, 1, \dots, N-1\}$, and the channel impulse response (CIR) is assumed to be zero for $\langle n-n' \rangle_N > L-1$, the received OFDM symbol in the frequency domain (FD) can be obtained by multiplying both sides of Equation (6) by the matrix $\mathbf{F}$. This operation yields the following result:

$$\mathbf{u}_r = \mathbf{F} \sum_{t=0}^{N_t-1} \mathbf{H}(t, r) \mathbf{x}_t + \mathbf{z}_r, \quad (8)$$

where $\mathbf{u}_r$ is the DFT of the $\mathbf{y}_r$ and $\mathbf{z}_r$ is the DFT of the noise sequence. Since matrix $\mathbf{F}$ is unitary, (8) can be expressed as:

$$\mathbf{u}_r = \sum_{t=0}^{N_t-1} \mathbf{G}(t, r) \mathbf{s}_t + \mathbf{z}_r, \quad (9)$$

In the described system, $\mathbf{s}_t = \mathbf{F} \mathbf{x}_t$ represents the OFDM symbol transmitted from the t -th transmitter antenna in the frequency domain. The matrix $\mathbf{G} = \mathbf{F} \mathbf{H} \mathbf{F}^H$ is the channel frequency matrix (CFM). When there is insignificant Doppler propagation, $\mathbf{G}$ becomes a diagonal matrix, resulting in an inter-carrier interference (ICI)-free system. However, in V2V (vehicle-to-vehicle) environments, where both the transmitter and the receiver experience high mobility, the Doppler dispersion becomes significant. As a result, the matrix $\mathbf{G}$ contains energy in components outside the diagonal, leading to the presence of ICI. Figure 1 illustrates the block diagram of the proposed MIMO-OFDM-IM (multiple-input–multiple-output orthogonal frequency division multiplexing with index modulation) transmitter. This design integrates index modulation (IM) into a conventional MIMO-OFDM system.

3. Receiver

3.1. Channel Estimation

To address the challenges posed by the reduced number of pilots in a single OFDM-IM symbol and the presence of inter-carrier interference (ICI), the task of estimating the time-varying channel impulse response (CIR) is made more complex. In order to overcome these difficulties, the observation model of the channel estimator is extended to incorporate a

sliding window that encompasses adjacent OFDM-IM symbols. This extension is expressed as follows:

$$\begin{bmatrix} \mathbf{u}_r^{k-1} \\ \mathbf{u}_r^k \\ \mathbf{u}_r^{k+1} \end{bmatrix} = \begin{bmatrix} \mathbf{G}(t, r)^{k-1} \mathbf{s}_t^{k-1} \\ \mathbf{G}(t, r)^k \mathbf{s}_t^k \\ \mathbf{G}(t, r)^{k+1} \mathbf{s}_t^{k+1} \end{bmatrix} + \begin{bmatrix} \mathbf{z}_r^{k-1} \\ \mathbf{z}_r^k \\ \mathbf{z}_r^{k+1} \end{bmatrix}, \quad (10)$$

the superscript associated with each variable indicates its position relative to the current k symbol. To perform channel estimation using this observation model, an extension of the algorithm proposed in [26,27] is utilized. The basis expansion model (BEM) is employed to obtain a concise representation of the CIR for each transmitter–receiver coupling within the three consecutive OFDM-IM symbols in the interval. This extension can be expressed as follows:

$$h[r, t; n, l] = \sum_{r=0}^{M_\tau-1} \sum_{q=0}^{M_D-1} \rho_{q,r} \phi_q^I[n] \phi_r^{II}[l] + \varepsilon[n, l]. \quad (11)$$

In the given expression, $\rho_{q,r}$ represents the coefficients, $m = \{0, \dots, 3N_b - 1\}$, M_τ and M_D denote the number of functions used to expand the delay-time domain and the time domain, respectively, and $\{\phi_q^I[n], \forall q \in [0, M_D - 1]\}$ and $\{\phi_r^{II}[l], \forall r \in [0, M_\tau - 1]\}$ are the functions that expand the time domain and the delay-time domain, respectively. In V2V scenarios, where there is a wide range of statistical variations in Doppler and delay dispersion, discrete prolate spheroidal sequences (DPSS) are used as base functions. DPSS functions optimally concentrate energy within a finite time and bandwidth window. The modelling error in this representation within subspaces is concentrated in the term $\varepsilon[n, l]$.

To determine the number of functions required in each of the CIR domains, the approach proposed in [26,27] is used.

$$M_\tau = \lceil F_s \tau_{max} \rceil + 1, \quad (12)$$

$$M_D = \lceil 2f_D 3N_b / F_s \rceil + 1. \quad (13)$$

In the provided equation, $\lceil \cdot \rceil$ represents the ceiling rounding operator, F_s corresponds to the system's bandwidth, τ_{max} denotes the maximum time delay spread, and f_D represents the maximum frequency Doppler spread.

The information obtained from the BEM of the channel in the frequency and frequency Doppler domains is condensed in a doubly-indexed matrix, which can be expressed as follows:

$$\Phi_{q,r}^k = \mathbf{F} \mathbf{B}_{q,r}^k \mathbf{F}^H, \quad (14)$$

with

$$[\mathbf{B}_{q,r}^k]_{n,n'} = \phi_q^I[n + (k+1)(N + N_g)] \phi_r^{II}[(n - n')_N] \quad (15)$$

and the BEM coefficients of each Tx-Rx coupling given by:

$$\rho_{q,r} = \sum_{m=0}^{3N-1} \sum_{l=0}^{L-1} h[r, t; m, l] \bar{\phi}_q^I[m] \bar{\phi}_r^{II}[l]. \quad (16)$$

Substituting the channel for its BEM in (9) and considering only Tx-Rx coupling yields the expression:

$$\mathbf{u}^k = \sum_{q=0}^{M_D-1} \sum_{r=0}^{M_\tau-1} \rho_{q,r} \Phi_{q,r}^k \mathbf{s}^k + \mathbf{z}^k. \quad (17)$$

By incorporating this representation of the channel into Equation (10) and considering only the positions where the transmitted and received pilots are present, we obtain the following expression:

$$\mathbf{u}_P = \mathbf{A} \mathbf{p} + \mathbf{e}_P, \quad (18)$$

where:

$$\mathbf{\Lambda}^T = \begin{bmatrix} \mathbf{\Lambda}^{k-1T} & \mathbf{\Lambda}^{kT} & \mathbf{\Lambda}^{k+1T} \end{bmatrix}^T, \quad (19)$$

$$\mathbf{\Lambda}^k = \begin{bmatrix} \mathbf{\Phi}_{1,\mathcal{P}}^k \mathbf{s}_{\mathcal{P}}^k & \mathbf{\Phi}_{2,\mathcal{P}}^k \mathbf{s}_{\mathcal{P}}^k & \mathbf{\Phi}_{I,\mathcal{P}}^k \mathbf{s}_{\mathcal{P}}^k \end{bmatrix}, \quad (20)$$

$$\mathbf{u}_{\mathcal{P}} = \begin{bmatrix} \mathbf{u}_{\mathcal{P}}^{k-1T} & \mathbf{u}_{\mathcal{P}}^{kT} & \mathbf{u}_{\mathcal{P}}^{k+1T} \end{bmatrix}^T, \quad (21)$$

$$\boldsymbol{\rho} = [\rho_0 \quad \rho_1 \quad \cdots \quad \rho_i \quad \cdots \quad \rho_I]^T. \quad (22)$$

In the given context, the subscript $\mathcal{P}$ pertains to the subset of vectors and matrices that correspond to the positions of the pilots. To simplify the notation, we introduce the indexed variable $i = q + M_D(p - 1)$, where $0 \leq p \leq M_\tau$ and $0 \leq q \leq M_D - 1$. This indexing allows us to represent the pilots' positions more conveniently. Furthermore, the vector $\epsilon_{\mathcal{P}}$ is introduced to encompass the contributions from noise, modelling errors, and intersymbol interference, with the purpose of simplifying the expressions.

The assumption is made that the receiver possesses the matrix $\mathbf{\Lambda}$ and the received vector $\mathbf{u}_{\mathcal{P}}$. With these available, the estimation of the channel's coefficient vector for each transmitter–receiver coupling can be obtained using the least squares (LS) algorithm.

$$\hat{\boldsymbol{\rho}} = (\mathbf{\Lambda}^H \mathbf{\Lambda})^{-1} \mathbf{\Lambda} \mathbf{u}_{\mathcal{P}}. \quad (23)$$

Once these coefficients are obtained, various representations of the channel, such as the time-varying impulse response or the frequency response, can be calculated directly by computing the weighted sum of the base functions. This allows for the determination of the frequency Doppler and frequency response matrices for each transmitter–receiver coupling. The calculation can be performed using the following expression:

$$\hat{\mathbf{G}}^k(t, r) = \sum_{i=0}^{I-1} \hat{\rho}_i \mathbf{\Phi}_i^k. \quad (24)$$

3.2. Low-Complexity Detection for OFDM-IM Signals

Due to resource limitations, the high complexity required by optimal maximum likelihood (ML) detectors makes them impractical for real-world implementations in vehicular systems. This section proposes a low-complexity detection algorithm suitable for the proposed MIMO-OFDM-IM system to overcome this restriction. To simplify the notation during detection, we rewrite the signal model described in Equation (8) as follows:

$$\mathcal{U} = \mathcal{G}\mathcal{S} + \mathcal{Z} \quad (25)$$

where:

$$\begin{aligned} \mathcal{U} &= \begin{bmatrix} \mathbf{u}_0^T & \mathbf{u}_1^T & \cdots & \mathbf{u}_{N_r-1}^T \end{bmatrix}^T, \\ \mathbf{u}_r &= \begin{bmatrix} u[0] & u[1] & \cdots & u[N_d - 1] \end{bmatrix}^T, \\ \mathcal{S} &= \begin{bmatrix} \mathbf{s}_0^T & \mathbf{s}_1^T & \cdots & \mathbf{s}_{N_t-1}^T \end{bmatrix}^T, \\ \mathbf{s}_t &= \begin{bmatrix} s[0] & s[1] & \cdots & s[N_d - 1] \end{bmatrix}^T, \\ \mathcal{Z} &= \begin{bmatrix} \mathbf{z}_0^T & \mathbf{z}_1^T & \cdots & \mathbf{z}_{N_r-1}^T \end{bmatrix}^T, \\ \mathbf{z}_r &= \begin{bmatrix} z[0] & z[1] & \cdots & z[N_d - 1] \end{bmatrix}^T. \end{aligned}$$

The block matrix $\mathcal{G}$ is composed of channel matrices $\mathbf{G}(t, r)$ according to the following assignment:

$$\mathcal{G} = \begin{bmatrix} G_{(0,0)} & G_{(0,1)} & \cdots & G_{(0,N_t-1)} \\ G_{(1,0)} & \ddots & & G_{(1,N_t-1)} \\ \vdots & & \ddots & \vdots \\ G_{(N_r-1,0)} & G_{(N_r-1,1)} & \cdots & G_{(N_r-1,N_t-1)} \end{bmatrix}, \quad (26)$$

The ordered successive interference cancellation (OSIC) detection technique used in V-BLAST systems emphasizes the importance of the data detection order, as an optimal order reduces the risk of error propagation during symbol estimation. An extension of the modified Gram–Schmidt (MGS) algorithm [25] is employed in [28] to obtain an ordered QR decomposition of the channel matrix $\mathcal{G}$. Rearranging the columns of the $\mathcal{G}$ matrix before each orthogonalization results in an $\mathbf{R}$ matrix with rows in decreasing order with respect to the SNR. Complexity reduction in the QR decomposition process is achieved by precomputing the norm of each column only once and updating it according to the MGS algorithm described in [25]. The complete algorithm used for MMSE-SQR-IM decomposition is shown in Algorithm 1. Note that the computational overhead due to ordering is insignificant.

Algorithm 1: Sorted QR decomposition.

Input: The channel matrix $\mathcal{G}$;

Output: The unity matrix $\mathbf{Q}_1$, the upper triangular matrix $\mathbf{R}$, and the detection order $\mathbf{p}$;

Initialization::

$\mathbf{Q} = \mathcal{G}$, $\mathbf{R} = \mathbf{0}$, $\mathbf{p} = [0, 1, \dots, N_r N_d - 1]^T$, $l = 0$, and $m = 0$;

for $i = 0 : 1 : N_r N_d - 1$ **do**

 Compute the norm of the n -th column vector in $\mathbf{Q}$ as $\mathbf{Norm}_i = \|\mathbf{q}_i\|^2$;

end

Givens Rotations Procedure::

for $i = 0 : 1 : N_r N_d - 1$ **do**

if $\text{mod}(i, 2) \neq 0$ **then**

 Find $[l, \text{block}] = (\mathbf{Norm}_i)$;

 Exchange the i -th and l -th columns in $\mathbf{R}$, $\mathbf{p}$ and $\mathbf{Norm}$ with the first $N_d + i$ rows in $\mathbf{Q}$, and exchange the elements $\mathbf{Q}_{N_B+i,i}$ and $\mathbf{Q}_{N_B+l,l}$;

 Compute $\mathbf{R}_{i,i} = \sqrt{\mathbf{Norm}_i}$ and update the i -th column in $\mathbf{Q}$ as

$\mathbf{Q}_{:,i} = \mathbf{Q}_{:,i} / \mathbf{R}_{i,i}$;

for $k = 0 : 1 : N_r N_d - 1$ **do**

 Compute $\mathbf{R}_{i,k} = \|\mathbf{q}_i\|^2$ and update the k -th column vector in $\mathbf{Q}$ as

$\mathbf{Q}_{:,k} = \mathbf{Q}_{:,k} - \mathbf{R}_{i,k} \mathbf{Q}_{:,i}$ and $\mathbf{Norm}_k = \mathbf{Norm}_k - \|\mathbf{R}_{i,k}\|^2$;

end

end

end

OSIC-MMSE Sorted Detection

The detection process consists of two stages. Firstly, the channel matrix $\mathcal{G}$ undergoes a sorted QR decomposition. Then, utilizing this decomposition, the proposed OSIC-MMSE detector with low computational complexity is executed.

To enhance the detection process based on the minimum mean square error (MMSE) criterion, QR decomposition is performed on the extended channel matrix $\underline{\mathcal{G}}$. This extended channel matrix is defined as follows:

$$\underline{\mathcal{G}} = \begin{bmatrix} \mathbf{P}\mathcal{G} \\ \sigma_w \mathbf{I} \end{bmatrix} = \underline{\mathbf{Q}}\underline{\mathbf{R}} = \begin{bmatrix} \mathbf{Q}_1 \\ \mathbf{Q}_2 \end{bmatrix} \underline{\mathbf{R}} = \begin{bmatrix} \mathbf{Q}_1 \underline{\mathbf{R}} \\ \mathbf{Q}_2 \underline{\mathbf{R}} \end{bmatrix}. \quad (27)$$

In the context of this article, we consider a matrix $\mathbf{Q}$ of dimensions $(N_t + N_r)N_d \times N_tN_d$ consisting of orthonormal columns. This matrix is partitioned into two submatrices: $\mathbf{Q}_1$, with dimensions $N_rN_d \times N_tN_d$, and $\mathbf{Q}_2$, with dimensions $N_rN_d \times N_tN_d$. Additionally, we have a superior triangular matrix denoted as $\mathbf{R}$ and a permutation matrix $\mathbf{P}$ that reorders the columns based on the signal-to-interference ratio.

To achieve this decomposition, we employ the method based on Givens unit rotations, as described in [25]. During the orthonormalization process for determining the $\mathbf{Q}$ matrix, column permutations are performed to ensure that the resulting rows in the $\mathbf{R}$ matrix are ordered according to their signal-to-interference ratios. Utilizing unit rotations in the QR decomposition preserves the original energy of all elements in the initial $\mathcal{G}$ matrix, allowing for a dynamic range of variables used in the process. This characteristic facilitates the implementation of this method in real-time devices using fixed-point arithmetic.

In this article, we modify the algorithm proposed in [28] (see Algorithm 2) to incorporate the ordered successive interference cancellation (OSIC) process for detecting the IM scheme. Once the preprocessing of the received signal and the QR decomposition of the estimated channel matrix are completed, the subsequent step involves data detection.

Algorithm 2: Proposed OSIC algorithm.

Input: $\tilde{\mathbf{U}}, \mathbf{R}, N_B, N_C, \Omega_{IM}, \mathbf{P}$

Output: $\hat{\mathbf{s}}$

for $k = N_B - 1 : -1 : 0$ **do**

$ind = N_p N_B$;

Select the k -th submatrix $\mathbf{R}_B$ from $\mathbf{R}$;

Select the k -th subblock $\tilde{\mathbf{U}}_B$ from $\tilde{\mathbf{U}}$;

Calculate the metric $d_B^2 = \|\tilde{\mathbf{U}}_B - \mathbf{R}_B \tilde{\mathbf{S}}_B\|^2$, for all possible values of $\mathbf{S} \in \Omega_{IM}$;

Find the minimum value d_B^2 with its corresponding position in matrix Ω_{IM} ,
and store the values in d_B and $\mathbf{index}(k)$;

$\hat{\mathbf{s}}_B = \Omega_{IM}(\mathbf{index}(k))$;

$\tilde{\mathbf{U}} = \tilde{\mathbf{U}} - \mathbf{R}(:, ind - N_C + 1 : ind) \hat{\mathbf{s}}_B$;

end

Permute $\hat{\mathbf{s}}$ according $\mathbf{index}$;

return $\hat{\mathbf{s}}$

The OSIC algorithm, known for its effectiveness in cancelling inter-carrier interference (ICI), is suitable for our proposed model described in Equation (25). When combined with the aforementioned QR decomposition, it allows for suboptimal data detection with significantly reduced computational complexity. By substituting this decomposition into Equation (25), we obtain the following expression:

$$\mathbf{U} = \mathbf{Q}_1 \mathbf{R} \mathbf{P}^T \mathbf{S} + \mathbf{Z}. \quad (28)$$

Premultiplying both sides of this equation by $\mathbf{Q}_1^H$ elicits the following system of equations:

$$\tilde{\mathbf{U}} = \mathbf{R} \tilde{\mathbf{S}} + \mathbf{Q}_1^H \mathbf{Z}, \quad (29)$$

where $\tilde{\mathbf{U}} = \mathbf{Q}_1^H \mathbf{U}$, and $\tilde{\mathbf{S}} = \mathbf{P}^T \mathbf{S}$ is the data vector ordered in decreasing order with respect to its contained energy. The $\mathbf{Z}$ noise vector maintains its statistics due to the fact that $\mathbf{Q}_1$ is unitary. The OSIC detector was proposed to operate at the sub-block symbol level considering the look-up table with N_C subcarrier and constellation size M . The look-up table has 2^b rows (see Table 1). Then, Equation (29) is rewritten as a block system:

$$\tilde{\mathbf{U}}_B = \mathbf{R}_B \tilde{\mathbf{S}}_B + \tilde{\mathbf{Z}}_B, \quad (30)$$

where the subscript $\mathcal{B} = \{0, 1, \dots, N_t N_B - 1\}$ refers to the subsample of the vectors and matrices in the rows and columns corresponding to the sub-block symbol position. The block matrix $\mathbf{R}_{\mathcal{B}} \in \mathbb{C}^{N_t \times N_r}$ is called a block matrix IM. The $\mathcal{B}$ -th sub-block of $\tilde{\mathcal{U}}_{\mathcal{B}}$, $\tilde{\mathcal{S}}_{\mathcal{B}}$ and $\tilde{\mathcal{Z}}_{\mathcal{B}}$ for $\mathcal{B} = \{0, 1, \dots, N_t N_B - 1\}$ is equal to:

$$\begin{aligned}\tilde{\mathcal{U}}_{\mathcal{B}} &= [\tilde{u}[\mathcal{B}N_C] \quad \tilde{u}[1 + \mathcal{B}N_C] \quad \cdots \quad \tilde{u}[(N_C - 1) + \mathcal{B}N_C]]^T, \\ \tilde{\mathcal{S}}_{\mathcal{B}} &= [\tilde{s}[\mathcal{B}N_C] \quad \tilde{s}[1 + \mathcal{B}N_C] \quad \cdots \quad \tilde{s}[(N_C - 1) + \mathcal{B}N_C]]^T, \\ \tilde{\mathcal{Z}}_{\mathcal{B}} &= [\tilde{z}[\mathcal{B}N_C] \quad \tilde{z}[1 + \mathcal{B}N_C] \quad \cdots \quad \tilde{z}[(N_C - 1) + \mathcal{B}N_C]]^T.\end{aligned}$$

Because of the triangular structure of the $\mathbf{R}_{\mathcal{B}}$ matrix, the j -th element of the $\tilde{\mathcal{U}}_{\mathcal{B}}$ vector can be expressed individually as:

$$\tilde{u}_j = r_{jj}\tilde{s}_j + \sum_{i=j+1}^{N_C} r_{ji}\tilde{s}_i + \tilde{z}_j. \quad (31)$$

where $j = \{0, 1, \dots, N_C - 1\}$ and r_{ab} denotes the element of matrix $\mathbf{R}_{\mathcal{B}}$ in the a -th row and in the b -th column. The sub-blocks are estimated in regressive sequence using OS detection. This way, the detection of each of the data points can be obtained iteratively using the following expression:

$$\hat{s}_j = \mathcal{Q} \left\{ \frac{\tilde{u}_j - \sum_{i=j+1}^{N_C} r_{ji}\tilde{u}_i}{r_{jj}} \right\}, \quad (32)$$

where the operator $\mathcal{Q}\{\cdot\}$ is a decision operator that maps its arguments to the closest point in the constellation Ω_{IM} used by the transmitter. A full description of the OSIC detector is shown in Algorithm 2. Assuming that in each iteration the previous decisions are correct, the interference of the previously detected symbols can be subtracted before the current symbol detection is executed.

4. Computational Complexity

Figure 2 illustrates the block diagram with the proposed structure of the MIMO-IM-OFDM receiver. In comparison to a conventional MIMO-OFDM receiver, the proposed receiver simplifies the implementation of the channel estimator by utilizing matrix-vector multiplication. Additionally, it offers lower-complexity nonlinear detection and mitigates ICI through the use of IM and OSIC detection techniques. The channel estimation required for data detection is only performed on carrier positions using the two-dimensional Slepian BEM estimator [reference] at the OFDM block level. The proposed OSIC-IM detection stage has a complexity of $\mathcal{O}((B\mathcal{V})^3)$, where $\mathcal{V} = N_d/N_c$ represents the ratio of non-zero carriers per sub-block. As a result, the computational complexity of the proposed technique is lower than that of a conventional MIMO-OFDM system with linear or nonlinear detection. Table 2 provides a comparison of receiver complexities.

Table 2. Computational complexity in terms of complex products.

	OFDM-MMSE	OFDM-ML	OFDM-OSIC	Proposed
Demodulation	$\mathcal{O}(N \log_2 N)$	N/A	$\mathcal{O}(N_d \log_2 N_d)$	$\mathcal{O}(B \log_2 B)$
CE	$\mathcal{O}(N^2)$	$\mathcal{O}(N^2)$	$\mathcal{O}(N_d^2)$	$\mathcal{O}((B\mathcal{V})^2)$
Detection	$\mathcal{O}(N^3)$	$\mathcal{O}(\Omega^N N)$	$\mathcal{O}(N_d^3)$	$\mathcal{O}((B\mathcal{V})^3)$

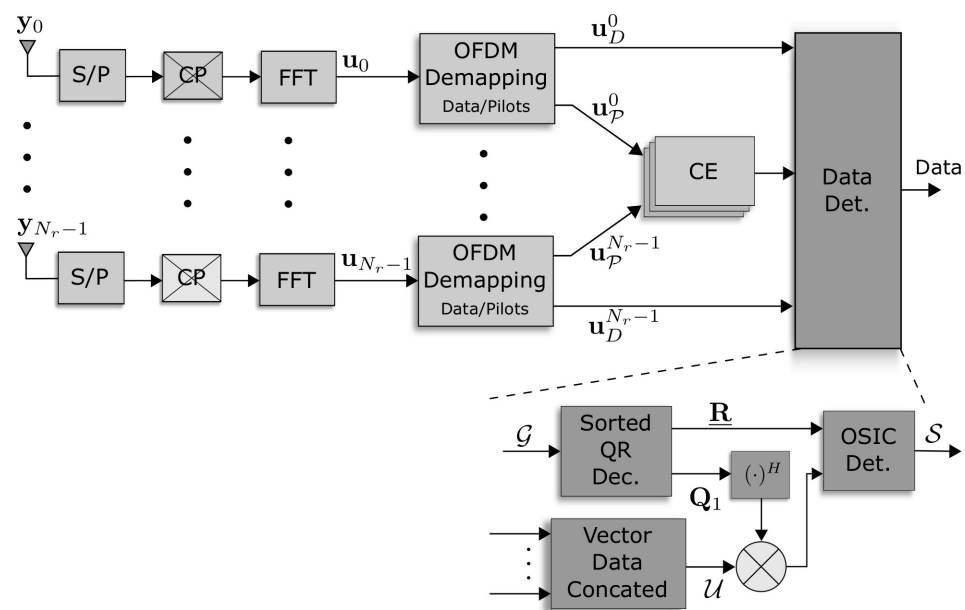


Figure 2. Block diagram of proposed MIMO-OFDM-IM receiver.

5. Performance Analysis

The performance of the proposed IM-MIMO-OFDM system was evaluated through Monte Carlo simulations to obtain achievable BER metrics and compare them with MIMO-OFDM systems. The simulations considered a 2×2 MIMO system (Experiments 1, 2, and 3) and a 4×4 MIMO system (Experiments 4 and 5). A MIMO-V2V channel with selective Rayleigh fading in frequency with a maximum delay spread (τ_{rms}) of $0.4 \mu s$ and Doppler frequency (f_D) of 1 KHz was considered. Each OFDM-IM symbol block consists of $N = 64$ subcarriers with the addition of a cyclic prefix of size $CP = 16$. For proper comparison, channel state information was estimated using a two-dimensional orthogonal basis expansion model (2D-BEM) for all evaluated receivers. The remaining system and channel parameters are shown in Table 3.

Table 3. System and channel parameters.

Parameter (Units)	Experiments 1, 2, and 3	Experiments 3 and 4
$\{N, N_d, CP, N_p\}$ (Samples)	$\{64, 48, 16, 8\}$	$\{64, 48, 16, 8\}$
Bandwidth (MHz)	10	10
Tx antennas N_t	2	4
Rx antennas N_r	2	4
Constellation size M	$\{4, 4, 16\}$	$\{4, 64\}$
N_C (Samples)	$\{2, 2, 4\}$	$\{2, 2\}$
N_B (Samples)	$\{24, 24, 12\}$	$\{24, 24\}$
PDP $E\{h^2[n, l]\}$	$\lambda e^{-0.4l}$	$\lambda e^{-0.4l}$
Number of multipaths	6	8
f_D (kHz)	1	1.2
τ_{rms} (μs)	0.4	0.4
Channel estimation	2D-BEM	-
$\{M_D, M_\tau\}$	$\{2, 5\}$	-
Spectral efficiency (bit/s/Hz)	1.2	2.4

5.1. Experiment 1

The proposed channel estimator's performance was evaluated in the first experiment. For each OFDM symbol, there are $N_p = 8$ pilots available for channel estimation. The estimation performance was assessed in both the time and frequency domains using the proposed extended 2D-BEM model. The results shown in Figure 3 demonstrate that in vehicular scenarios with a high Doppler dispersion frequency ($f_D = 1$ kHz), the conventional

OFDM-MIMO system fails to estimate the channel parameters accurately. The mean square error (MSE) exhibits a floor error at low SNR levels. However, in the case of the proposed system, specifically for $N_t = 2, N_r = 2$, the MSE error behaves uniformly without any floor error in the reported SNR region. Additionally, for $N_t = 4, N_r = 4$, the MSE error exhibits a floor error at high SNR levels. This can be explained by the fact that $N_p = 8$ remains constant, and a larger number of parameters need to be estimated with the same amount of information.

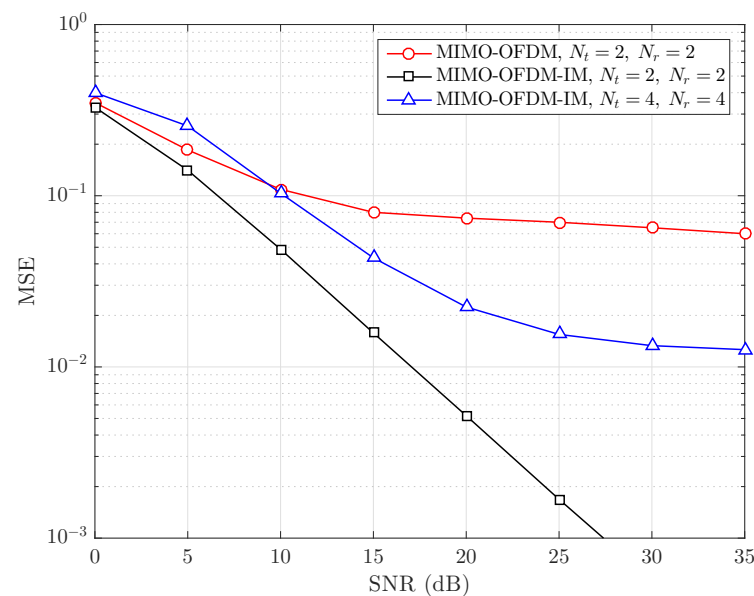


Figure 3. Comparison of the MSEs for the channel estimator in the proposed MIMO-OFDM-IM system versus a conventional MIMO-OFDM.

5.2. Experiment 2

In this case study, the BER performance of the IM-MIMO-OFDM system is compared against the conventional MIMO-OFDM system using MMSE detectors. The proposed system's performance using an OSIC detector is also compared. The tests were conducted using $N_t = 2, N_r = 2$ for all systems, a sub-block size of $N_C = 2$ for the MIMO-OFDM-IM system, and 4-QAM modulation for data in the conventional MIMO-OFDM system, while maintaining a spectral efficiency of 1.2 for both systems. The simulation parameters are shown in Table 3.

The results in Figure 4 show that for the given channel, the conventional MIMO-OFDM system has inferior performance compared to the MIMO-OFDM-IM system. The BER achieved at low E_b/N_0 is similar for the MIMO-OFDM-IM receivers with MMSE detection and the proposed OSIC detection, which is significant considering that the OSIC detector has the advantage of significantly reduced computational complexity. The conventional MIMO-OFDM system experiences a loss of approximately 3 dB for the medium-to-high E_b/N_0 range. The main reason for this is that for the frequency-selective channel, the IM technique is more suitable for mitigating high Doppler dispersion, primarily due to the utilization of null subcarriers in transmission. Finally, it can be observed that the OSIC nonlinear detector does not exhibit an error floor at high E_b/N_0 levels.

5.3. Experiment 3

The following experiment was designed to assess the trade-off between computational complexity and performance in terms of bit error rate (BER) in the proposed system as the sub-block size of OFDM-IM increases. The proposed system is compared against a MIMO-OFDM-IM system with MMSE detection and a conventional MIMO-OFDM system with 4-QAM data modulation. To maintain the same spectral efficiency, the OFDM-IM

modulation was configured with $N_C = 4$ and $M = 16$. The simulation parameters are described in Table 3.

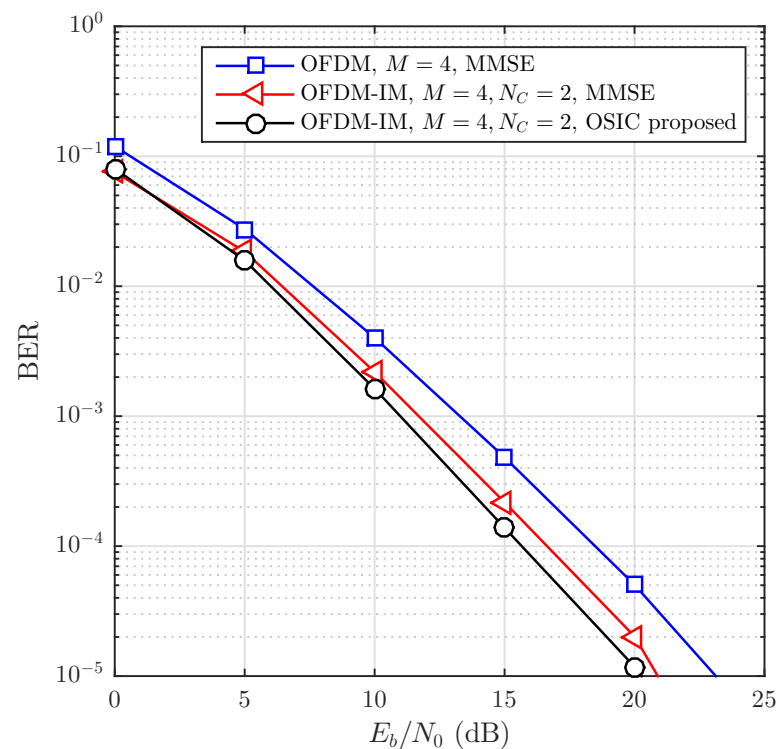


Figure 4. BER-vs- E_b/N_0 comparison of different detection algorithms for MIMO-OFDM-IM and MIMO-OFDM with $N_t = 2$, $N_r = 2$, and 4-QAM data modulation.

From the results in Figure 5, the following observations can be made: (a) Close performance is maintained at low E_b/N_0 values between the MIMO-OFDM-IM systems with OSIC and MMSE detection. However, above 20 dB, OSIC detection exhibits a consistently superior trend of approximately 2 dB compared to MMSE detection. (b) Increasing the constellation size degrades the performance of the proposed system, resulting in an approximately 2 dB performance difference compared to the conventional OFDM system above $E_b/N_0 = 15$ dB. However, an increase in the number of subcarriers per sub-block results in a significant reduction in the required computational complexity. (c) The use of IM technique in combination with low-complexity OSIC detection is an attractive solution to address the issues of frequency selectivity and high mobility present in vehicular links.

5.4. Experiment 4

Next, we increased the spatial diversity of the data by using four antennas for both transmission (N_t) and reception (N_r). We compared the BER performance of the proposed MIMO-OFDM-IM system with OSIC detection against the MIMO-OFDM-IM system with MMSE detection and the conventional MIMO-OFDM system. The simulation parameters used for the experiment are listed in Table 3.

The results from Figure 6 in low E_b/N_0 regions demonstrate close performance between the proposed MIMO-OFDM-IM system with OSIC detection and the MIMO-OFDM-IM system with MMSE detection. However, the proposed detector requires lower computational complexity. Compared to the conventional MIMO-OFDM system, the proposed system is not affected by burst errors caused by ICI, resulting in significant BER improvement of approximately 2.5 dB above 10 dB of E_b/N_0 . This result not only highlights a significant trade-off between computational complexity and achieved BER by increasing the number of antennas in the system but also demonstrates the feasibility of implementing the proposed system.

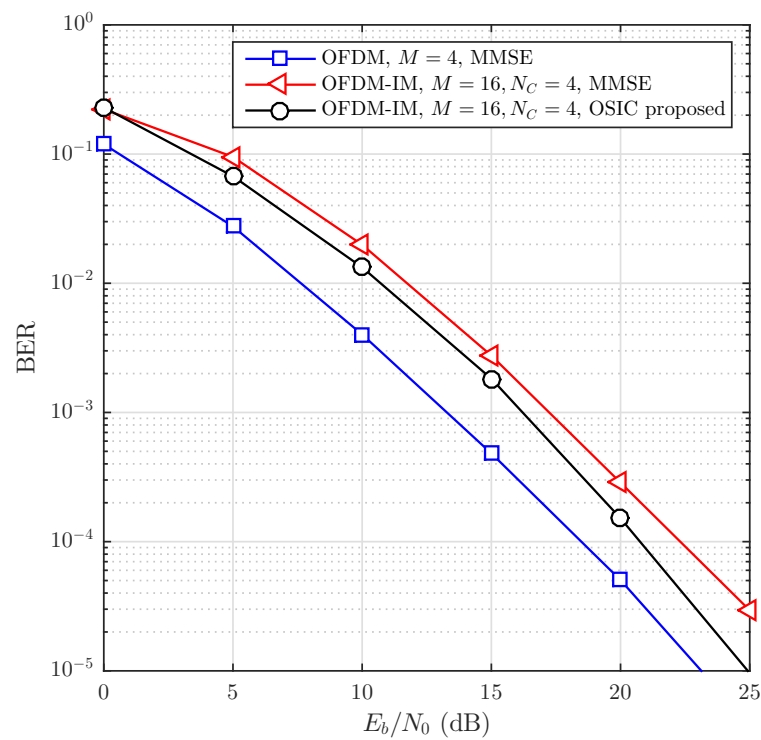


Figure 5. BER-vs- E_b/N_0 comparison of different detection algorithms for MIMO-OFDM-IM and MIMO-OFDM with $N_t = 2$, $N_r = 2$.

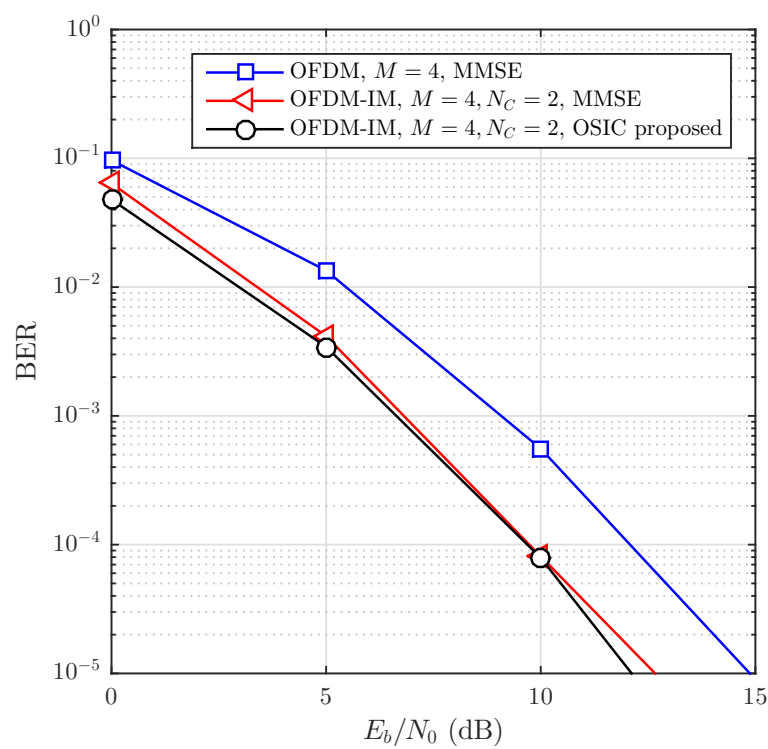


Figure 6. BER-vs- E_b/N_0 comparison of different detection algorithms for MIMO-OFDM-IM and MIMO-OFDM with $N_t = 4$, $N_r = 4$.

5.5. Experiment 5

The following experiment was designed to assess the performance of the proposed system under conditions of high transmission rates and vehicular channel selectivity. Table 3 shows the system configuration parameters. The proposed system is compared against a conventional MIMO-OFDM system with two different constellation sizes: $M = 16$ and $M = 64$.

The results presented in Figure 7 illustrate the main advantages and disadvantages of the proposed system. The primary advantage is the comparable performance between the proposed system with $M = 64$ and $N_c = 2$ and the OFDM-IM system with MMSE detection. However, it should be noted that the proposed system requires lower computational complexity. Compared to the conventional OFDM system with $M = 64$, the proposed system with $N_c = 2$ achieves a gain of approximately 2 dB with reduced complexity. However, achieving this performance requires a sacrifice of 1.2 in spectral efficiency. Additionally, for the case of $N_c = 8$, the proposed system achieves a gain of approximately 2.5 dB, indicating that an increase in the symbol block N_c helps mitigate ICI in the detection process.

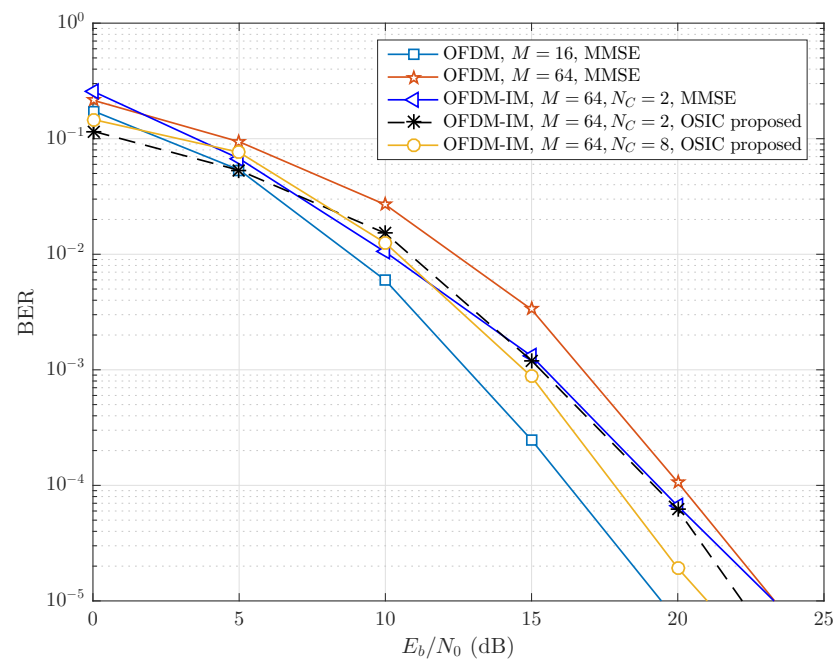


Figure 7. BER-vs- E_b/N_0 comparison of different detection algorithms for MIMO-OFDM-IM and MIMO-OFDM with $N_t = 4$, $N_r = 4$ under conditions of high transmission rates.

6. Conclusions

The structure of a MIMO-OFDM system for V2V channels was proposed as a result of the generalization of the IM developed in this study. Its performance in terms of BER is superior to that of more-advanced approaches requiring complex products $O(N^2)$ thanks to the incorporation of IM. The integration achieved in the modulation, channel estimation, and data detection stages of the proposed system has not been reported elsewhere in the literature, establishing this approach as a suitable solution that achieves the mitigation of ICI and the exploitation of spatial diversity for MIMO-OFDM in V2V channels with reduced computational complexity and improved BER performance compared to a conventional MIMO-OFDM system. Possible areas for future work include studying these techniques in massive and non-orthogonal MIMO systems.

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