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A combining design of precoder and equalizer based on shared redundancy to improve performance of ISI MIMO systems

Bui Quoc Doanh[1] · Do Thanh Quan[1] · Pham Thanh Hiep[1] · Ta Chi Hieu[1]

Abstract

There have been a number of researches on block transmission systems via MIMO channels due to really high data transmission rate. However, because of the existence of inter-symbol interference (ISI) in the systems, guard intervals are added to eliminate the ISI, leading to a reduction in channel energy and bandwidth efficiency. In order to optimize these systems, there are many solutions in which a combining design of precoder and equalizer appears as a potential candidate. In this paper, a jointly optimal design for precoder and equalizer for ISI multiple input multiple output (MIMO) channels based on sharing redundancy is proposed. Theory analysis and simulation results demonstrate that the proposed design produces a significant improvement in the system performance such as a reduction in bit error rate, a decrease in channel energy loss and an increase in system throughput.

Keywords Block transmission · Precoder · Equalizer · ISI MIMO systems · BER · Leading zero · Trailing zero · Sharing redundancy

1 Introduction

There have been burning issues on the 5G mobile communication system owing to really high data rate, absolutely low latency and high quality of service (QoS) [1–3]. Moreover, the 5G system also can become a fundamental communication infrastructure part of or cooperate efficiently with advanced electronic systems based on the Internet of things (IoT) as in [4, 5]. One of the main technical characteristics of the 5G system and many the state of the art communication systems is block transmission via MIMO channel. As a result, there have been a number of initial researches on block transmission through MIMO channel and a study on this idea is a continuous work in order to meet requirements of the future communication systems [6–10]. However, the block transmission via single input single output (SISO) or MIMO channel has to face with the ISI which may cause a reduction in bandwidth efficiency because of adding guard intervals in the form of zero padding (ZP) or cyclic prefix (CP) intervals. In order to deal with this matter and improve system performance, there have been many solutions that may be clarified into three groups such as basically focusing on the precoder, mainly concentrating on the equalizer and jointly designing for both precoder and equalizer.

Firstly, there have been a number of solutions basically related to the precoder as in papers [11–23] which also can be clarified into three subgroups as following reviews. In the first subgroup [11–15], the authors proposed many direct approaches to improve the precoder. For example, in [11], the authors proposed a method to improve the MIMO precoder based on the minimum euclidean distance. In [12], the authors applied the singular value decomposition algorithm for single carrier MIMO transmission. In [13], the authors improved zero forcing (ZF) and minimum mean square error (MMSE) precoders based on modified cost functions. In [14], a novel method to precoding design was proposed based on the joint precoding of legacy and

✉ Bui Quoc Doanh
buiquocdoanh@tcu.edu.vn

Do Thanh Quan
dtquan82@gmail.com

Pham Thanh Hiep
phamthanhhiep@gmail.com

Ta Chi Hieu
hieunda@mta.edu.vn

[1] Le Quy Don Technical University, Hanoi, Vietnam

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new users, minimizing the normalization power while not making perturbations. In [15], the linear precoder design was proposed to maximize the average mutual information with that the transmitter has statistical channel state information. In the second subgroup, the authors proposed methods to optimize the system based on the view of work in a pair of precoder and decoder as in [16, 17]. For instance, a design concentrated on minimization of the symbol mean square error (MSE) and maximization of the minimum distance between symbol hypotheses was proposed in [16]. In [17], the authors optimized the MSE under a low complexity of implementation. In the papers [18–20], basing on a condition that each pair of transmit and receive users has information about other pairs, the interference alignment (IA) precoding technique is utilized to optimize system parameters, such as the transmit power, sum rate, or computational complexity. In the third subgroup, some analyses of the system performance were explored in [21, 22], and effective MU-MIMO precoding techniques for WiMAX were proposed in [23].

Secondly, there have been also some papers proposed methods concentrating on equalizer as in [24–28]. In the [24], the finite impulse response (FIR) MMSE equalizer was optimized without requiring a guard to be larger than or equal to the channel order. In the [25], a low complexity equalizer was proposed by using a combination of space-frequency MMSE filter and a pre-whitened maximum likelihood detector (MLD) while a block interactive decision feedback equalizer with noise prediction was proposed in [26]. Moreover, some analyses of diversity and BER performance of system using equalizer were illustrated in [27, 28].

Finally, in order to eliminate the ISI in the block transmission system, guard intervals in form of the ZP or CP intervals are added. However, this also makes a part of channel energy to be lost during the cancellation of guard intervals, therefore, the spectrum efficiency is reduced, especially for the channels with long impulse response. As a result, there were some papers that propose combining designs of precoder and equalizer in order to optimize the system [29–38]. In the paper [29], for the ISI SISO channels, a jointly optimal precoder and equalizer designs were proposed under conditions of maximizing output signal-to-noise ratio (SNR), the MMSE and ZF criteria, and fixed transmitted power while an optimization of information rate over dispersive channels is derived in [30]. Using the weighted MMSE criterion, proposed designs were developed based on the maximum information rate and QoS while the unweighted MMSE and equal-error design was produced for fixed rate systems [31]. In the paper [32], under a condition that unique optical power constraints, the authors proposed a method to choose proper precoder and equalizer to minimize the MSE for multi-user multi-cell MIMO visible light communication (VLC) systems. When the signal is non-negativity and total average power constraints, two precoders based on the geometric mean decomposition and uniform channel decomposition for the MIMO VLC systems are analyzed and compared [33]. Moreover, some other designs using oversampled filter banks (OSFBs), a Wiener filter banks, the signal-to-leakage (SLR), Tomlinson–Harashima precoding and frequency domain equalization and sharing redundancy were proposed as in [34–38], respectively.

In this paper, by combining the ideas reported in [29, 31, 38], an efficient design of a jointly optimal precoder and equalizer for frequency selective MIMO channels is developed based on sharing redundancy during the transmission. The detail ideas and contributions of this paper will be explained as followings. The redundancy is generally understood as the length of the guard intervals as in [39, 40] and defined as the difference between the input symbol blocks and the transmitted or received symbol blocks. In addition, in order to optimize the linear precoder and equalizer of the SISO channel, the ISI needs to be eliminated basically by trailing zero (TZ) and leading zero (LZ) as in [29, 30]. While the last L rows of the precoder matrix are set to zero in the TZ method, the first L columns of the equalizer matrix are set to zero in the LZ method in which L is the order of FIR. In fact, the LZ and TZ show the same performance and are used separately. Moreover, the key idea of sharing redundancy in paper [38] for the SISO channel is that instead of setting the last L rows of the precoder matrix to zero, the authors only set the last $K = \lfloor \frac{L}{2} \rfloor$ rows to zero at the transmitter and the first $(L-K)$ columns of equalizer matrix are also set to zero at the receiver. As a result, this helps to reduce the loss in channel energy, consequently, and achieve better BER performance than that in previous scheme. Basing on above analysis, in this paper, we developed and extended the LZ method as in [29], and proposed sharing redundancy method as in [38] for the ISI MIMO channel in which the *water-pouring* algorithm as in [31] is used for signal processing in the proposed system.

The rest of the paper is organized as follows: Sect. 2 will illustrate the system model, Sect. 3 will demonstrate the developed LZ method and proposed sharing redundancy method for the ISI MIMO channels. Section 4 will give the simulation results and the conclusion will be drawn in Sect. 5. In this paper, notations are used as following: boldface font is used for vector and matrix; The set of complex numbers is denoted by symbol $\mathbb{C}$; $(\cdot)^H$ is the Hermitian transpose operation and $(\cdot)^T$ is transpose operation.

2 System model

In this paper, the authors investigate system model of block transmission based on linear precoding and equalization for ISI MIMO channels illustrated as in Fig. 1a. The ISI MIMO channel with T transmit and R receive antennas is assumed to be stationary, frequency selective fading and has FIR with order of L. The channel impulse response (CIR) is given by matrices $\mathbf{H}[0], \mathbf{H}[1], \ldots, \mathbf{H}[L]$, where $\mathbf{H}[l] \in \mathbb{C}^{T \times R}, (l = 0, \ldots, L)$.

The ISI MIMO channel model with precoder and equalizer is shown as in Fig. 1b. In this work, we will shortly describe this system model as following. Firstly, the input symbol stream $s[n]$ is converted into symbol vectors $s[i]$ with the size of $N \times 1$ by the serial-to-parallel converter. Secondly, the symbol vector $s[i]$ is transferred to the precoder that generates symbol vectors $x[i]$ with the size of $PT \times 1$. Finally, the symbol vectors $x[i]$ are then divided into P vectors with the size of $T \times 1$ and then passed through the ISI MIMO channel.

At the channel output, the shaded serial-to-parallel converter forms symbol vectors $y[i]$ with the size of $PR \times 1$ from P received symbol vectors in which each has the size of $R \times 1$. Due to effect of noise, a received symbol vector $y[i]$ consists of two parts, an information symbol vector $r[i]$ and a noise sample vector $v[i]$. In the receiver, the symbol vectors $y[i]$ are transferred to the equalizer in order to generate symbol vectors $\hat{\mathbf{s}}[i]$ with the size of $N \times 1$. As a result, the output symbol stream $\hat{s}[n]$ is obtained by the parallel-to-serial converter from the symbol vectors $\hat{\mathbf{s}}[i]$.

In the two previous paragraphs, with the input symbol stream $\mathbf{s}[n]$ and the sampled version of received signal $\hat{\mathbf{s}}[n]$, the operation of the system is expressed obviously and logically. Moreover, the terms defined as $\mathbf{s}[i]$, $\mathbf{x}[i]$, $\mathbf{y}[i]$, $\hat{\mathbf{s}}[i]$ and $\mathbf{v}[i]$ can be illustrated mathematically as in following Eqs. (1)–(5), respectively

$$\mathbf{s}[i] = [s[iN], s[iN+1], \ldots, s[iN+N-1]]^T \quad (1)$$

$$\mathbf{x}[i] = [x[iPT], x[iPT+1], \ldots, x[iPT+PT-1]]^T \quad (2)$$

$$\mathbf{y}[i] = [y[iPR], y[iPR+1], \ldots, y[iPR+PR-1]]^T \quad (3)$$

$$\hat{\mathbf{s}}[i] = [\hat{s}[iN], \hat{s}[iN+1], \ldots, \hat{s}[iN+N-1]]^T \quad (4)$$

$$\mathbf{v}[i] = [v[iPR], v[iPR+1], \ldots, v[iPR+PR-1]]^T. \quad (5)$$

3 Performance analysis

3.1 The leading zero method

In this paper, we will firstly develop and extend the LZ method used in the SISO channel for the ISI MIMO channel. In the papers [29, 30], the authors had analyzed the channel model, proofed and built equations in the case of the ISI SISO channels. These equations will be analyzed and extended for the ISI MIMO channels in which the output symbol vectors $\mathbf{s}[\hat{i}]$ can be calculated in the case of $P \geq L$ by following

$$\hat{\mathbf{s}}[i] = \mathbf{G}\mathbf{H}_0\mathbf{F}\mathbf{s}[i] + \mathbf{G}\mathbf{H}_1\mathbf{F}\mathbf{s}[i-1] + \mathbf{G}\mathbf{v}[i] \quad (6)$$

where $\mathbf{F} \in \mathbb{C}^{PT \times N}$ and $\mathbf{G} \in \mathbb{C}^{N \times PR}$ are used to express operation and function of the precoder and the equalizer, respectively. In these works, the authors assumed that $\mathbf{v}[i] \backsim CN(0, 1)$ represents the additive white Gaussian noise (AWGN) sample vector with the size of $PR \times 1$. Moreover, $\mathbf{H}_0$ and $\mathbf{H}_1$ are the lower and upper Toeplitz matrices with the same size of $PR \times PT$ and they can be illustrated by following

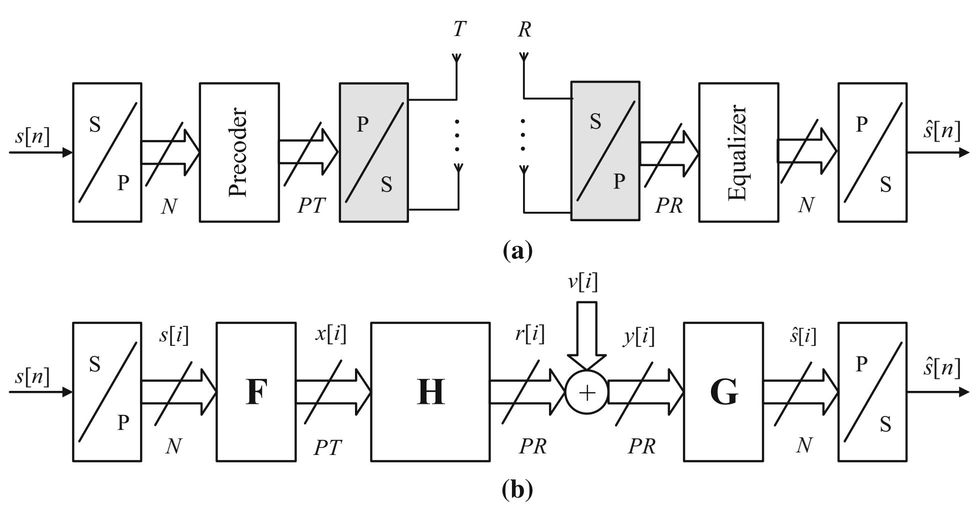

Fig. 1 The ISI MIMO channel model with precoder and equalizer. **a** General system block diagram; **b** equivalent system model with channel matrix

$$\mathbf{H}_0 = \begin{bmatrix} \mathbf{H}[0] & \mathbf{0} & \mathbf{0} & \cdots & \mathbf{0} \\ \vdots & \mathbf{H}[0] & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{H}[L] & \cdots & \ddots & \cdots & \vdots \\ \vdots & \ddots & \cdots & \cdots & \mathbf{0} \\ \mathbf{0} & \cdots & \mathbf{H}[L] & \cdots & \mathbf{H}[0] \end{bmatrix}, \quad (7)$$

$$\mathbf{H}_1 = \begin{bmatrix} \mathbf{0} & \cdots & \mathbf{H}[L] & \cdots & \mathbf{H}[1] \\ \vdots & \ddots & \mathbf{0} & \ddots & \vdots \\ \mathbf{0} & \cdots & \ddots & \cdots & \mathbf{H}[L] \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \mathbf{0} & \cdots & \mathbf{0} & \cdots & \mathbf{0} \end{bmatrix}. \quad (8)$$

In the Eq. (6), $\mathbf{GH}_1\mathbf{Fs}[i-1]$ takes place and characterizes for the ISI in the ISI MIMO channel model. With the assumption of $PT = M + LR$, $(M \geq N)$, in order to optimize the linear precoder and equalizer, the ISI needs to be eliminated. In fact, there are two approaches to cancel the ISI called as trailing zero (TZ) and leading zero (LZ) as in [29]. While the last LT rows of the precoder $\mathbf{F}$ are set to zero in the TZ method, the first LR columns of the equalizer $\mathbf{G}$ are set to zero in the LZ method, as shown in (9) and (10) respectively. In the previous papers, the TZ and LZ methods show the same performance and they are used separately. In this paper, we consider the LZ method only in order to compare the LZ method to the proposed method

$$\mathbf{F} = \begin{bmatrix} \mathbf{F}_0[0] & \mathbf{F}_1[0] & \cdots & \mathbf{F}_{M-1}[0] \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{F}_0[M-1] & \mathbf{F}_1[M-1] & \cdots & \mathbf{F}_{M-1}[M-1] \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{0} \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{0} \end{bmatrix}, \quad (9)$$

$$\mathbf{G} = \begin{bmatrix} \mathbf{0} & \cdots & \mathbf{0} & \mathbf{G}_0[0] & \cdots & \mathbf{G}_{M-1}[0] \\ \mathbf{0} & \cdots & \mathbf{0} & \mathbf{G}_0[1] & \cdots & \mathbf{G}_{M-1}[1] \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ \mathbf{0} & \cdots & \mathbf{0} & \mathbf{G}_0[N-1] & \cdots & \mathbf{G}_{M-1}[N-1] \end{bmatrix}. \quad (10)$$

In the case of the LZ method, while the precoder remains stable $\mathbf{F} = \mathbf{F}_0$, the first LR columns of the equalizer $\mathbf{G}$ are set to zero, consequently, the equalizer can be described shortly as following form

$$\mathbf{G} = [\mathbf{0}_{N \times LR} \quad \mathbf{G}_0], \quad (11)$$

here the equalizer $\mathbf{G}_0 \in \mathbb{C}^{N \times (P-L)R}$.

As a result, the term $\mathbf{GH}_1\mathbf{Fs}[i-1]$ equals to zero, it means that the ISI is eliminated completely. And then, the precoder $\mathbf{F}_0$ and the equalizer $\mathbf{G}_0$ are jointly designed by either the ZF or the MMSE criteria in order to improve the system performance. Therefore, the Eq. (6) can be rewritten as follows

$$\hat{\mathbf{s}}[i] = \mathbf{G}_0\mathbf{H}\mathbf{F}_0\mathbf{s}[i] + \mathbf{G}_0\mathbf{v}[i], \quad (12)$$

where $\mathbf{H}$ is the last M rows of $\mathbf{H}_0$

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}[\mathrm{L}] & \cdots & \mathbf{H}[0] & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \ddots & \ddots & \ddots & \cdots & \vdots \\ \vdots & \vdots & \ddots & \ddots & \ddots & \mathbf{0} \\ \mathbf{0} & \cdots & \mathbf{0} & \mathbf{H}[\mathrm{L}] & \cdots & \mathbf{H}[0] \end{bmatrix}. \quad (13)$$

Under the condition that the transmit power is constrained to P_0, the MMSE algorithm is used to optimize the precoder and equalizer [29], hence the matrices $\mathbf{F}_0$ and $\mathbf{G}_0$ are derived as

$$\mathbf{F}_0 = \mathbf{V}\boldsymbol{\Phi}\mathbf{U}^H, \quad (14)$$

$$\mathbf{G}_0 = \mathbf{R}_{ss}\mathbf{F}_0^H\mathbf{H}^H\left(\mathbf{R}_{vv} + \mathbf{H}\mathbf{F}_0\mathbf{R}_{ss}\mathbf{F}_0^H\mathbf{H}^H\right)^{-1}, \quad (15)$$

where $\mathbf{R}_{ss}$ is the input symbol covariance matrix and $\mathbf{R}_{vv}$ is the noise covariance matrix. In addition, $\mathbf{U}$ and $\mathbf{V}$ are the unitary matrices that can be determined by the eigenvalue decompositions (EVD) as in [29]

$$\mathbf{R}_{ss} = \mathbf{U}\boldsymbol{\Delta}\mathbf{U}^H, \quad (16)$$

$$\mathbf{H}^H\mathbf{R}_{vv}^{-1}\mathbf{H} = \mathbf{V}\boldsymbol{\Lambda}\mathbf{V}^H. \quad (17)$$

In the Eqs. (14), (16) and (17), the matrices $\boldsymbol{\Phi}$, $\boldsymbol{\Delta}$ and $\boldsymbol{\Lambda}$ are diagonal with nonnegative elements. Moreover, the main diagonal elements of $\boldsymbol{\Phi}$ are found out by the MMSE [29]

$$|\Phi_{jj}|^2 = \frac{P_0 + \sum_{i=1}^{k}\lambda_{ii}^{-1}}{\sum_{i=1}^{k}\lambda_{ii}^{-1/2}\delta_{ii}^{1/2}}\frac{1}{\sqrt{\lambda_{jj}\delta_{jj}}} - \lambda_{jj}^{-1}\delta_{jj}^{-1}, \quad (18)$$

with k is the number of Φ_{jj} satisfying $|\Phi_{jj}|^2 > 0$ and $\lambda_{jj}, \delta_{jj}$ are the jth main diagonal elements of $\boldsymbol{\Lambda}$ and $\boldsymbol{\Delta}$, respectively.

From the above analysis, we can conclude that the idea of [29] also can be applied for the ISI MIMO systems, consequently, the optimal precoder and equalizer can improve the system performance, and the SNR on every flat subchannel is proportional to eigenvalues of $\mathbf{H}^H\mathbf{R}_{vv}^{-1}\mathbf{H}$. It is clear that the sum of these eigenvalues is proportional to the square of the Frobenius norm of $\mathbf{H}$, and $\mathbf{H}$ is the last M rows of $\mathbf{H}_0$ (in the case of TZ, $\mathbf{H}$ is the first M columns of $\mathbf{H}_0$) in which $M = PT - LR$ is the number of the flat subchannels. Due to adding guard interval, LR, for cancelling the ISI, a part of channel energy is lost. To reduce the loss, the guard interval should be decreased, however, decreasing the guard interval let the ISI remain, and hence the performance of system is extremely deteriorated. Consequently, in order to improve the performance of system with the guard interval, we propose the sharing redundancy method in the following section.

3.2 The sharing redundancy method

In this paper, we combine the ideas reported [29, 31, 38] and propose a method in order to optimize jointly designing of the precoding and equalization in the ISI MIMO channel model. The proposed method can reduce the loss in matrix $\mathbf{H}_0$, in other words, reduce the loss in channel energy. Besides, the proposed method also guarantees that some subchannels with too low eigenvalues will be discarded so that they do not affect the BER. The key idea about sharing redundancy in our approach is that instead of setting the last LT rows of the precoder matrix to zero, we only set the last $KT = \lfloor \frac{LT}{2} \rfloor$ rows to zero at the transmitter and the first $(L-K)R$ columns of $\mathbf{G}$ are also set to zero at the receiver. Therefore, the precoder and equalizer can be expressed mathematically as the following forms

$$\mathbf{F} = \begin{bmatrix} \mathbf{F}_0 \\ \mathbf{0}_{KT\times N} \end{bmatrix}, \tag{19}$$

$$\mathbf{G} = \begin{bmatrix} \mathbf{0}_{N\times(L-K)R} & \mathbf{G}_0 \end{bmatrix} \tag{20}$$

where $\mathbf{F}_0 \in \mathbb{C}^{(P-K)T\times N}$ and $\mathbf{G}_0 \in \mathbb{C}^{N\times(P-L+K)R}$ are designed by the MMSE criterion [31]. In fact, the MMSE criterion can minimize the sum of estimated symbol errors, however this criterion does not ensure that the MSEs on each subchannel are minimized separately. Moreover, when the *water-pouring* algorithm [31] is applied to the power distribution strategy of the MMSE design, the weakest eigenmodes will be dropped, consequently, the power is then redistributed according to remaining eigenvalues. As a result, more power is distributed to the higher eigenvalues, and the optimal precoder and equalizer are given by [31]

$$\mathbf{F}_0 = \mathbf{V}\boldsymbol{\Phi}_\mathbf{f}, \tag{21}$$

$$\mathbf{G}_0 = \boldsymbol{\Phi}_\mathbf{g}\mathbf{V}^H\mathbf{H}^H\mathbf{R}_{vv}^{-1}, \tag{22}$$

where $\mathbf{V}$ is unitary matrices derived from following EVD algorithm

$$\mathbf{H}^H\mathbf{R}_{vv}^{-1}\mathbf{H} = \mathbf{V}\boldsymbol{\Lambda}\mathbf{V}^H \tag{23}$$

and $\boldsymbol{\Phi}_\mathbf{f}$, $\boldsymbol{\Phi}_\mathbf{g}$ are diagonal matrices with their main diagonal elements are designed by the MMSE and then given by

$$\left|\phi_{\mathrm{f},jj}\right|^2 = \frac{P_0 + \sum_{i=1}^{k}\lambda_{ii}^{-1}}{\sum_{i=1}^{k}\lambda_{ii}^{-1/2}}\frac{1}{\sqrt{\lambda_{jj}}} - \lambda_{jj}^{-1}, \tag{24}$$

$$\left|\phi_{\mathrm{g},jj}\right|^2 = \left[\frac{\sum_{i=1}^{k}\lambda_{ii}^{-1/2}}{P_0 + \sum_{i=1}^{k}\lambda_{ii}^{-1}}\lambda_{jj}^{-1/2} - \left(\frac{\sum_{i=1}^{k}\lambda_{ii}^{-1/2}}{P_0 + \sum_{i=1}^{k}\lambda_{ii}^{-1}}\lambda_{jj}^{-1/2}\right)^2\lambda_{jj}^{-1}\right]\frac{1}{\lambda_{jj}}, \tag{25}$$

where λ_{jj} is the main diagonal elements of $\boldsymbol{\Lambda}$, and the number of $\phi_{\mathrm{f},jj}$ satisfying $\left|\phi_{\mathrm{f},jj}\right|^2 > 0$ is the same with $\phi_{\mathrm{g},jj}$ satisfying $\left|\phi_{\mathrm{g},jj}\right|^2 > 0$ and equals to k. As a result, the subchannel SNRs can be given by following [30]

$$\mathrm{SNR}_{jj} = \left|\phi_{\mathrm{f},jj}\right|^2\lambda_{jj}. \tag{26}$$

For the optimal linear precoder and equalizer in (21) and (22), since the ISI is completely eliminated, Eq. (6) can be rewritten as

$$\hat{\mathbf{s}}[i] = \mathbf{G}_0\hat{\mathbf{H}}\mathbf{F}_0\mathbf{s}[i] + \mathbf{G}_0\mathbf{v}'[i], \tag{27}$$

where $\hat{\mathbf{H}} \in \mathbb{C}^{(P-L+K)R\times(P-K)T}$ is illustrated as

$$\hat{\mathbf{H}} = \begin{bmatrix} \mathbf{H}[L-K] & \cdots & \mathbf{H}[0] & \mathbf{0} & \cdots & \mathbf{0} \\ \vdots & & & \ddots & \ddots & \vdots \\ \mathbf{H}[L] & & & & \ddots & \mathbf{0} \\ \mathbf{0} & \ddots & & & & \mathbf{H}[0] \\ \vdots & \ddots & \ddots & & & \vdots \\ \mathbf{0} & \cdots & \mathbf{0} & \mathbf{H}[L] & \cdots & \mathbf{H}[K] \end{bmatrix} \tag{28}$$

and $\mathbf{v}'[i]$ is the block length of noise samples and equals to $(P-L+K)R$.

In this paper, the proposed method based on shared redundancy will be named shortly as Sharing method. In order to prove produced improvement, the Sharing method will be compared with the LZ method logically as in Fig. 2 and quantitatively as in following section. In this section, the loss in $\mathbf{H}_0$ of the Sharing method and the loss in the case of the LZ or TZ method is compared logically and expressed obviously. From Eq. (13), in the LZ case the first LR rows of $\mathbf{H}_0$ are discarded by the $\mathbf{G}$ equalizer, therefore, the elements in the shaded triangle in Fig. 2a will be lost, resulting in a reduction in channel energy.

Different from the LZ method, the guard interval is shared between transmitter and receiver in the Sharing method, in other words, the first $(L-K)R$ rows of $\mathbf{H}_0$ are discarded by the equalizer and the last KT columns of $\mathbf{H}_0$ are discarded by the precoder. As a result, the loss in $\mathbf{H}_0$ can be described by the two shaded triangles as illustrated in Fig. 2b in which the triangle at the top-left corner and the triangle at bottom-right corner correspond to the loss caused by the equalizer and the loss caused by the

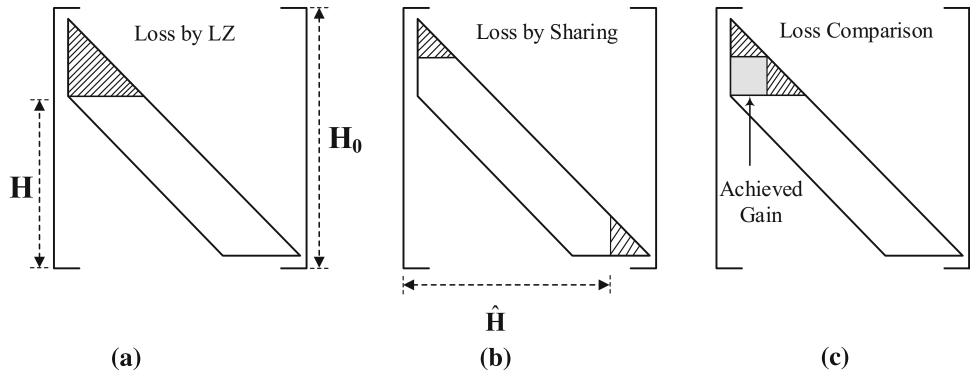

Fig. 2 Comparison of the loss in $\mathbf{H}_0$ in two approaches

precoder, respectively. We can shift the triangle at the bottom-right corner to the top-left corner area, consequently, a comparison of the losses in two cases can be illustrated obviously as in Fig. 2c. It is clear that the loss of $\mathbf{H}_0$ in the Sharing approach is smaller than the loss in LZ or TZ cases, leading to archived gain between two approaches. The elements of $\mathbf{H}_0$ lying in the dotted gray quadrilateral area are retained in $\hat{\mathbf{H}}$ and they will contribute significantly for an increase in the SNR due to the EVD algorithm usually distributes higher channel energy to the larger eigenvalues. The precoder and equalizer matrices $\mathbf{F}_0$ and $\mathbf{G}_0$ are designed by the MMSE criterion and are calculated as in (21), (22) with matrix $\mathbf{H}$ is replaced by $\hat{\mathbf{H}}$ while $\mathbf{R}_{vv}$ is resized accordingly.

4 Simulation results

In order to evaluate the proposed LZ and Sharing redundancy designs, a design without adding the guard interval named as Non LZ design is also simulated. The system performance is evaluated by the subchannel SNR, BER, and throughput according to three parameters, such as the transmission block size, FIR order and modulation order. Let the number of transmit and receive antennas be the same as two, and the CIR is generated from the Saleh–Valenzuela indoor channel model as in [41]. In addition, the total transmit power across two transmit antennas is normalized to unity ($P_0 = 1$), and the Monte Carlo simulation is utilized to calculate the BER and throughput of system.

Firstly, the simulation program is developed to analyze the SNR of the subchannels in three cases that are the total SNR equals to 0, 10 and 20 dB. The total SNR in this paper is defined as the received SNR in the case one symbol is transmitted from transmitter to receiver via single input single output channel. The setting parameters are the FIR order $L = 15$, the transmission block size $P = 24$.

As can be seen from Fig. 3, the subchannel SNR of the Non LZ design is much lower than that of the LZ and Sharing designs. The reason is that in the Non LZ design the received signal energy is slightly higher due to no loss in channel energy, however, without LZ or TZ there is an existence of the ISI and then the total noise defined in this paper consists of the AWGN and ISI, consequently, it is also too high. Moreover, the subchannel SNR of the Sharing design is always slightly higher than that of the LZ design and both the subchannel SNR of the two methods reduces gradually when the subchannel index increases in all three cases. For more detail, when the total SNR increases from 0 to 20 dB, the gap between the Sharing and LZ designs rises significantly. The reason of this issue is that the archived gain is likely to be stable, consequently, the gap rises slightly when the total SNR increases. Moreover, with the same total SNR, this gap is larger with the higher subchannels index. As a result, the system BER of the Non LZ design is much higher than that of the LZ and Sharing designs as shown in Fig. 4, therefore, in the rest of this paper, we will concentrate on comparison of the LZ design and Sharing design.

Secondly, the BER performance of the Sharing and LZ designs is compared in three cases that are different in the

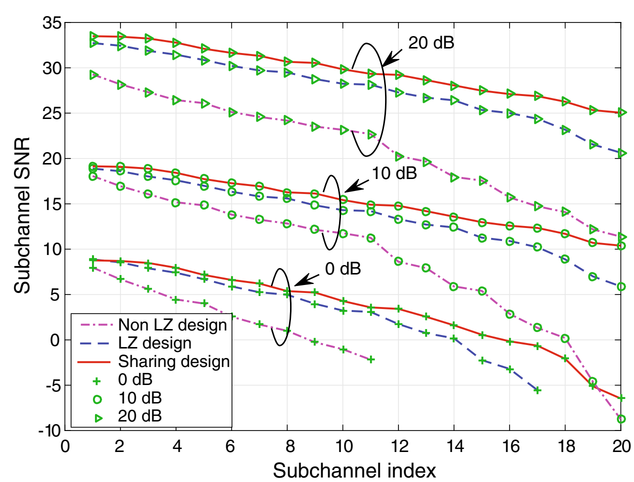

Fig. 3 The subchannel SNR of the non LZ, LZ and sharing designs

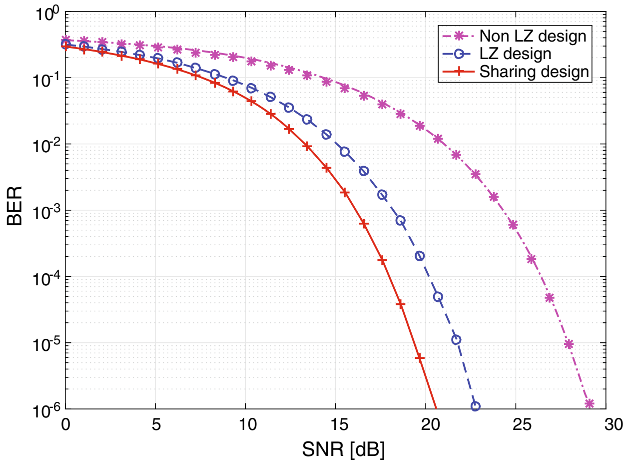

Fig. 4 The BER performance of the non LZ, LZ and sharing designs

transmission block size P, the FIR order L and modulation scheme, as shown in Figs. 5, 6 and 7, respectively.

In the Fig. 5, the BER performance of two proposed methods is evaluated with the differences in the transmission block size $P = 24, 36$ and 48. It is clear that the BER of the Sharing design is significantly lower than that of the LZ design with all three P values. For example, at the total SNR $= 25$ dB, the BER of the Sharing design is much smaller than that of the LZ design and the gap between two approaches decreases when the P increases. It is reasonable that when the P increases, the number of subchannels rises and since the achieved gain is unchanged, the subchannel SNRs of two methods become to close, leading to the curves of two methods becoming closer. Furthermore, when the transmission block size P increases, the BER rises in the both methods.

In the Fig. 6, the Sharing design also states its higher performance when the FIR order is changed. For instance,

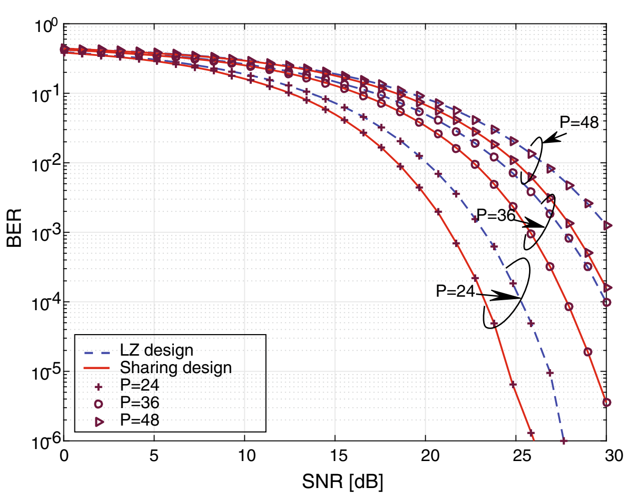

Fig. 5 The BER performance of the sharing and LZ designs ($P = 24, 36$ and 48)

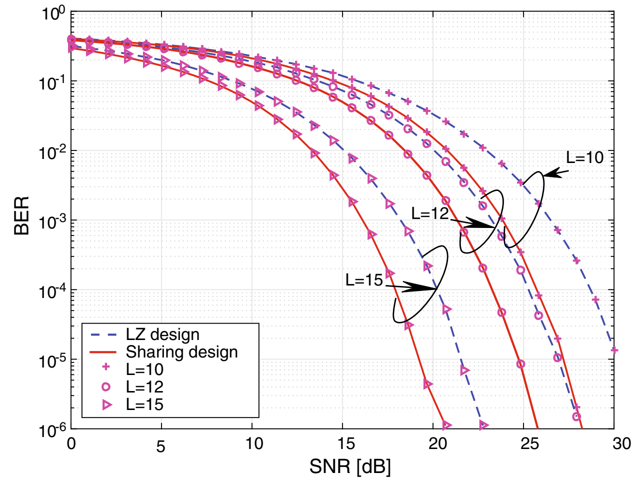

Fig. 6 The BER performance of the sharing and LZ designs ($L = 10, 12$ and 15)

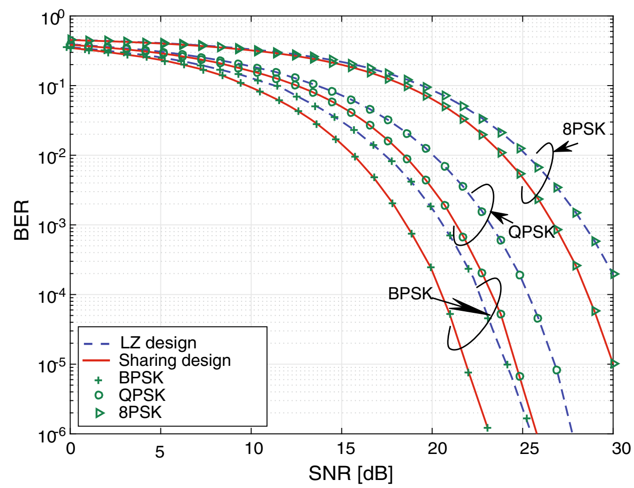

Fig. 7 The BER performance of the sharing and LZ designs (BPSK, QPSK and 8PSK)

at the total SNR $= 20$ dB, the system BER of the Sharing design is also much smaller than that of the LZ design. Moreover, when the FIR order L decreases from 15 to 10, the BER will increase and the gaps between curves of two methods will reduce. This can be explained as when the L rises, the number of transmission paths increases, consequently, the detection of received signal at receiver becomes more effective and then the BER will go down. Different from Fig. 5, in the Fig. 6 when the FIR order decreases, the BER rises in both methods. In the Fig. 7, the system performance is also enhanced remarkably by employing the Sharing design with different modulation orders.

Finally, the system performance evaluated by the throughput of two designs is compared in the cases of the different transmission block size P, FIR order L and modulation order as shown in Figs. 8, 9 and 10, respectively.

As can be seen from Fig. 8, the system throughput of the Sharing design is higher than that of the LZ design with three different values of the transmission block size $P = 24, 36$ and 48. Moreover, when the transmission block size increases, the gaps between two design curves decrease. In addition, when the SNR is high enough at about 25 dB, the system performance of the two methods is almost the same and the system throughputs achieve the maximum value. In this case, the maximum throughput is 2 bit/s/Hz because we use quadrature phase shift keying (QPSK) modulation scheme and normalized transmission rate is set to be 1 symbol/s/Hz. In the Fig. 9, the Sharing design produces significantly larger throughput than that of the LZ design with the different FIR orders $L = 10, 12$ and 15. Different from Fig. 8, when the FIR order increases, the gap between the Sharing and LZ designs rises. Besides, if the SNR increases to 25 dB or more, two methods have the same performance. In the Fig. 10, the Sharing design also produces slightly better throughput with the different modulation orders in which the gap between the Sharing and LZ design curves increases when the modulation order rises.

5 Conclusion

Research on the block transmission systems via MIMO channels is the continuous work in order to meet the requirements of the state of the art communication systems. In fact, there have been a number of researches on this issue that concentrate basically on canceling the ISI and optimizing the parameters and designs of the systems. Basing on developing the conventional works, a combining design of precoder and equalizer for the ISI MIMO channels is developed and proposed in this paper. The key idea of this paper is to share redundancy for both precoder and equalizer. Thanks to shared redundancy and applied EVD algorithm, the Sharing design can take advantage of channel energy and produce a significant improvement in the system performance, reducing the BER, decreasing channel energy loss and increasing the system throughput in comparison with the LZ design as shown in simulation results.

It is fairly clear that the Sharing design has almost the same implementation complexity as the LZ design. However, a deep discussion and analysis about complexity of two proposed designs is essential. Moreover, a study on the proposed system in the imperfect CSI scenario and other weighted MMSE criteria such as maximum information

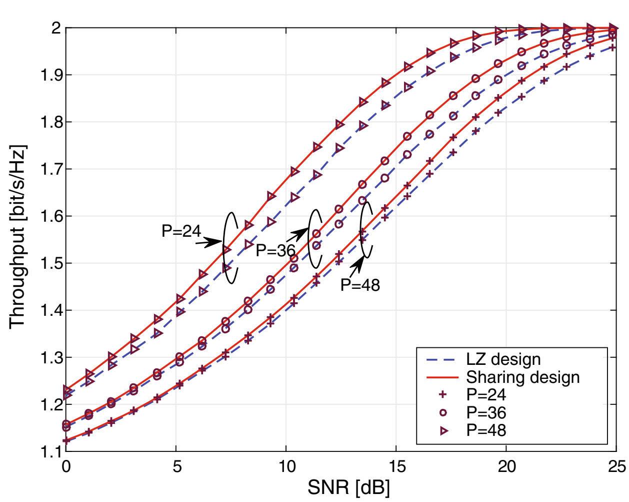

Fig. 8 The system throughput of the sharing and LZ designs ($P = 24, 36$ and 48)

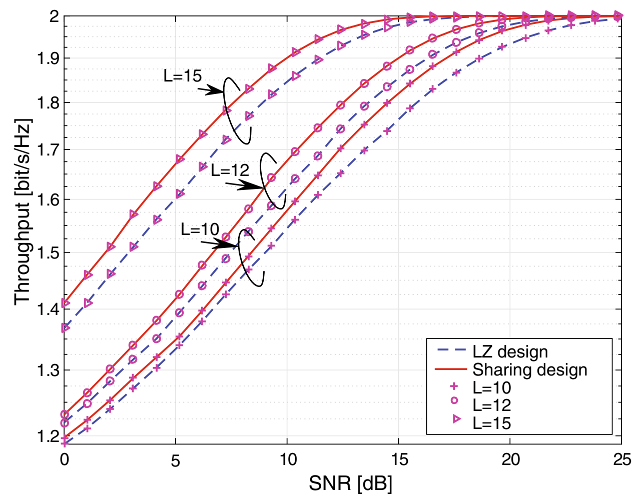

Fig. 9 The system throughput of the sharing and LZ designs ($L = 10, 12$ and 15)

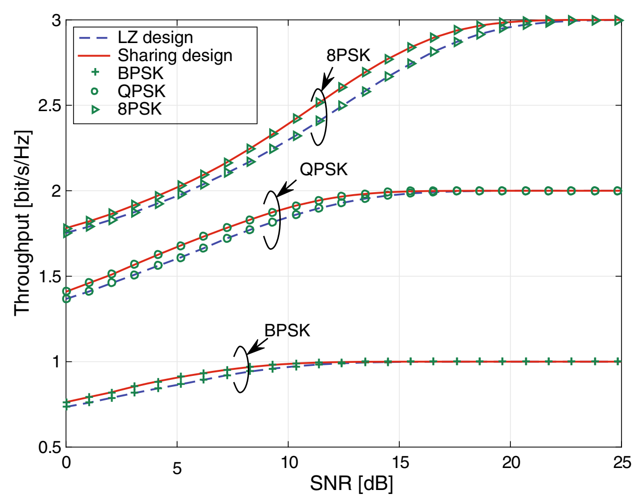

Fig. 10 The system throughput of the sharing and LZ designs (BPSK, QPSK and 8PSK)

rate, QoS based or equal error are interesting ideas. Above mentioned issues are future works of this paper.

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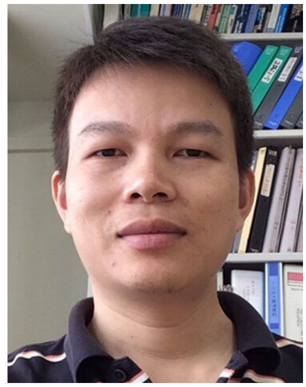
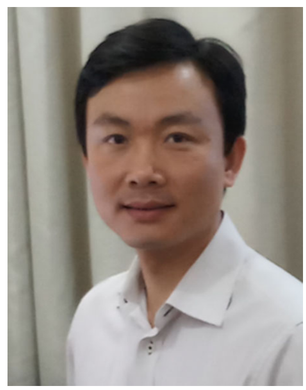
Bui Quoc Doanh received the B.S. degree in Communication Command at Telecommunication University, Ministry of Defense, Vietnam, in 2007, and the M.E. degree in Electronics Engineering from Posts and Telecommunications Institute of Technology (PTIT), Vietnam, in 2011. He is currently working towards the Ph.D. degree in electronic engineering at Le Quy Don Technical University, Hanoi, Vietnam. His research intersests include precoding and equalization for MIMO systems and wireless information.



Do Thanh Quan received the B.E. and M.E. degrees of electronics engineering at Le Quy Don Technical University in 2006 and 2008, respectively. He also got the first rank in entrance exam for bachelor course of this university. After graduation with bachelor's degree, the university offered him for a supporting researcher and lecturer position because of his great achievement. In 2013, he went to Japan for Ph.D. degree at Kohno lab, Graduate School of Engineering, Yokohama National University and received the degree in 2017. At the moment, he is working as a lecturer in Faculty of Radio and Electronical Engineering, Le Quy Don Technical University, Hanoi, Vietnam. His research interests are medical information communication technology, signal processing and information theory.



Pham Thanh Hiep received the B.E. degree in Communications Engineering from National Defence Academy, Japan, in 2005; received the M.E. and Ph.D. degree in Physics, Electrical and Computer Engineering from Yokohama National University, Japan, in 2009 and 2012, respectively. He was working as associate researcher at Yokohama National University, Yokohama, Japan from 2012 to 2015. Now, he is a lecturer at Le Quy Don Technical University, Ha Noi, Viet Nam. His reseach interests lie in the area of wireless information and communications technologies.

Ta Chi Hieu was born in Vinh Phuc in 1970. He graduated the Military Technical Academy in 1994 with hornor. He got his M.Sc. degree in electronic engineering in the National Defense Academy of Japan in 2002 and his Ph.D. degree in signal processing in the University of Strathclyde, United Kingdom in 2008. He is currently working at the Faculty of Radio Electronics, Military Technical Academy. His research interests include precoding and equalization for MIMO systems, microwave engineering and computational electromagnetics.