Design of equalizers for frequency selective MIMO systems with low redundancy

Ta Chi Hieu*

Faculty of Radio-Electronic Engineering, Le Quy Don Technical University.
*Corresponding author: hieunda@gmail.com

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ABSTRACT

In this paper, we propose several designs of filter bank based decision feedback equalizers (DFE) for frequency selective multi-input multi-output (MIMO) systems. Transmission schemes over frequency selective channels are typically block based, and redundancy in form of guard intervals must be used to avoid inter-block interference (IBI). This paper will focus on the case where spectral efficiency demands low levels of redundancy. Then the remaining IBI will be eliminated by a separate feedback loop. The co-channel interference (CCI) inherited in MIMO channels will be eliminated by different DFE schemes. In another approach, a multi-tap MIMO DFE will be designed to deal with both IBI and CCI. The simulation results show that for quasi-static frequency selective MIMO channels, the multi-tap DFE provides the best performance while for Raleigh frequency selective MIMO channels, the dual loop DFE is a good solution for transmission with low redundancy.

Keywords: MIMO; IBI; Frequency selective; Broadband; Decision-feedback equalizer.

1. INTRODUCTION

Capacity enhancement, link reliability and coverage improvement are the key advantages of multi-input multi-output (MIMO) systems. These advantages have made MIMO systems widely adopted in many wireless communications systems [1, 2]. Simultaneously, block transmission techniques such as Orthogonal Frequency Division Multiplexing (OFDM) or Digital Multi-Tone (DMT) are considered to be very effective method to deal with inter-symbol interference (ISI) caused by frequency selectivity of the channels when data rate becomes higher [3]. Thanks to the above advantages, these techniques are widely applied in modern communication systems. The block transmission, however, always requires an amount of redundancy in the form of guard intervals whose length must be equal to or larger than the channel in order to eliminate inter-block interference (IBI). This makes the block transmission not applicable for channels with long impulse response since the spectrum efficiency is reduced.

An approach to deal with the long channel impulse response (CIR) is channel shortening [4-7]. In this approach, time or frequency domain equalizers or transceivers are designed such that the equalized channel length is shorter than that of the original. Channel shortening equalizers and/or transceivers can be designed under different criteria such as minimum mean square error (MMSE) [4, 5], maximum shortening signal-to-noise ratio (SNR) [6], maximum geometric SNR [7]. The shortened channel allows for the deployment of a maximum likelihood sequence estimator or multicarrier modulation scheme, although part of channel energy (and therefore capacity) is lost [4].

The problem of long CIRs has also been approached in [8] where a Wiener filter Tomlinson-Harashima precoder was proposed and optimized in terms of precoding order and latency time. The precoder, however, requires channel state information (CSI) at the transmitter. In [9] block decision-feedback equalizers (BDFE) for the case of single-input single output (SISO) frequency selective channels with IBI presented were proposed. These BDFEs are designed under zero-forcing (ZF) and MMSE criteria and are shown to be very effective even with small transmit redundancy, however, in [9] they have only been applied on SISO channels and therefore their applicability for MIMO channels still needs to be considered.

In this paper, we will focus on designing a decision-feedback equalizer (DFE) for frequency selective MIMO (FS-MIMO) channels that allows for the use of less redundancy than the conventional block transmission schemes. For this purpose, we will consider the expansion of the approaches in [9]. First, we propose a design that uses a small amount of redundancy in combination with a separate DFE loop to eliminate the IBI. The FS-MIMO channel will then become flat and another DFE will be applied to tackle co-channel interference (CCI) in resulted MIMO channel. Second, we still use the same transmit filter bank as the previous design, but for the receiver, we extend for the case of FS-MIMO channels the design of MMSE-BDFE that uses multi-tap feedforward and feedback filter banks in [9]. We will also point out under which conditions these proposed designs can be applied for dispersive MIMO channels.

The paper is organized as follows, Sec. 2 describes the system model and the proposed designs of DFEs. Simulation results and discussion are provided in Sec. 3 and conclusions are drawn in Sec. 4.

In our notation, lower- and uppercase bold face fonts are used for vector and matrix quantities, respectively. The operator $E\{\cdot\}$ denotes expectation, $(\cdot)^H$ the Hermitian transpose and $(\cdot)^T$ the transpose operation.

2. SYSTEM MODEL

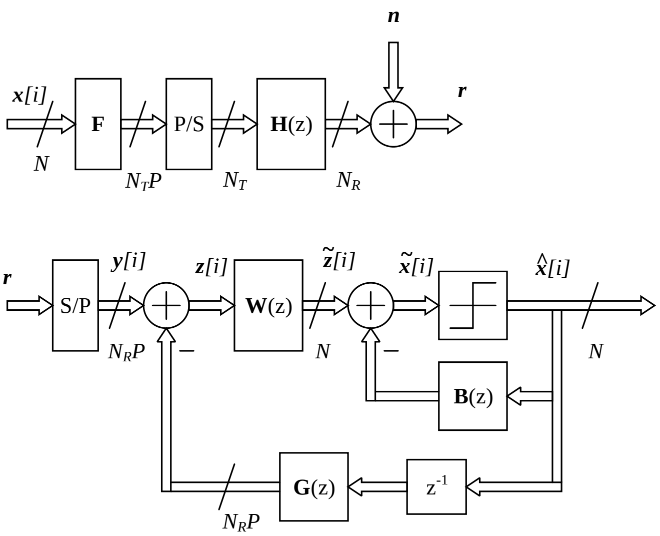

Figure 1. *System model for MIMO transmission scheme using DFEs.*

The system generic model is illustrated in Fig. 1. The MIMO channel with N_T inputs and N_R outputs (N_T transmit and N_R receive antennas) is assumed to be frequency selective with finite impulse response (FIR) of order L and CIR taps $[\mathbf{H}[0],...,\mathbf{H}[L]]$ where $\mathbf{H}[0],...,\mathbf{H}[L]\in\mathbb{C}^{N_R\times N_T}$.

The symbol blocks $\mathbf{x}[i], \mathbf{y}[i], \tilde{\mathbf{x}}[i], \hat{\mathbf{x}}[i]$ are defines as

$$\mathbf{x}[i]=\left[x[iN],...,x[iN+N-1]\right]^T,\ \mathbf{y}[i]=\left[y[iN_RP],...,y[iN_RP+N_RP-1]\right]^T,$$

$$\tilde{\mathbf{x}}[i]=\left[\tilde{x}[iN],...,\tilde{x}[iN+N-1]\right]^T,\ \hat{\mathbf{x}}[i]=\left[\hat{x}[iN],...,\hat{x}[iN+N-1]\right]^T,$$

respectively, and the blocks of noise samples are correspondingly defined as

$$\mathbf{n}[i]=\left[n[iN],...,n[iN+N_RP-1]\right]^T.$$

The transmit filter bank $\mathbf{F}\in\mathbb{C}^{N_TP\times N}$ has a simple structure as follows:

$$\mathbf{F}=\begin{bmatrix}\mathbf{I}_{N\times N}\\ \mathbf{0}_{(N_TP-N)\times N}\end{bmatrix}. \tag{1}$$

This structure allows to add $L_r=N_TP-N$ zeroes into each input symbol block $\mathbf{x}[i]$, making it length to be N_TP. Since we use low redundancy, L_r/N_T must be less than L. The transmit symbol blocks then are de-multiplexed into P vectors, each of length N_T by the parallel-to-serial converter. At the output of FS-MIMO channel, the received symbol vectors of length N_R mixed with noise are stacked up into symbol blocks of length N_RP by the serial-to-parallel converter. With $P>L$, the symbol blocks $\mathbf{y}[i]$ are given by [11]

$$\mathbf{y}[i]=\mathbf{H_0}\mathbf{F}\mathbf{x}[i]+\mathbf{H_1}\mathbf{F}\mathbf{x}[i-1]+\mathbf{n}[i], \tag{2}$$

where $\mathbf{H}_0$, $\mathbf{H}_1\in\mathbb{C}^{N_RP\times N_TP}$ represent the channel operation and are written as

$$\mathbf{H_0}=\begin{bmatrix}\mathbf{H}[0] & \mathbf{0} & \cdots & \mathbf{0} & \mathbf{0}\\ \vdots & \mathbf{H}[0] & \mathbf{0} & \vdots & \mathbf{0}\\ \mathbf{H}[L] & \vdots & \ddots & \ddots & \vdots\\ \mathbf{0} & \ddots & \vdots & \ddots & \mathbf{0}\\ \mathbf{0} & \cdots & \mathbf{H}[L] & \cdots & \mathbf{H}[0]\end{bmatrix},\ \mathbf{H}_1=\begin{bmatrix}\mathbf{0} & \mathbf{0} & \mathbf{H}[L] & \cdots & \mathbf{H}[1]\\ \vdots & \ddots & \mathbf{0} & \vdots & \mathbf{0}\\ \mathbf{0} & \cdots & \ddots & \ddots & \mathbf{H}[L]\\ \vdots & \vdots & \vdots & \ddots & \vdots\\ \mathbf{0} & \cdots & \mathbf{0} & \cdots & \mathbf{0}\end{bmatrix}. \tag{3}$$

It can be seen from equation (2) that the term $\mathbf{H}_1\mathbf{F}\mathbf{x}[i-1]$ represents the IBI. The transmit filter in (1) helps to remove the impact of the last L_r columns of the matrix $\mathbf{H}_1$ but since $L_r<LN_T$, due to low redundancy, the product $\mathbf{H}_1\mathbf{F}\neq\mathbf{0}$ and the IBI still remain. To tackle the remaining IBI and detect transmitted symbols we will consider two approaches as stated in Sec. 1. The subsections below will present details about those designs.

2.1. Design with dual loop DFE

As stated above, in this design two feedback loops will be used, one for remaining IBI removal and the other for CCI cancellation and symbol detection. Thus feedback filter bank $\mathbf{G}(z)$ will be set as $\mathbf{G}(z)=\mathbf{H}_1\mathbf{F}$. With the assumption that the detected symbols are correct, the remaining IBI will be removed completely and symbol blocks $\mathbf{z}[i]$ at the input of the second DFE can be written as

$$\mathbf{z}[i] = \mathbf{H_0Fx}[i] + \mathbf{H_1Fx}[i-1] - \mathbf{H_1F\hat{x}}[i-1] + \mathbf{n}[i] \approx \mathbf{H_0Fx}[i] + \mathbf{n}[i]. \quad (4)$$

Thus, the FS-MIMO channel is now rendered into a flat one. In the next step, we will deal with CCI inherited in MIMO channels. For doing this, we choose to apply the MMSE-DFE with an optimal order scheme in [10] since it has been shown to provide a very good bit error ratio (BER). From (3) and figure 1, the error vector can be written as

$$\boldsymbol{\varepsilon} = \mathbf{Px}[i] - \mathbf{\hat{x}}[i] = (\mathbf{B}+\mathbf{I})\mathbf{Px}[i] - \mathbf{Wz}[i] \quad (5)$$

where $\mathbf{P} \in \mathbb{R}^{N\times N}$ is the permutation matrix, $\mathbf{W} \in \mathbb{C}^{N\times N_R P}$ the feedforward filter bank and $\mathbf{B} \in \mathbb{C}^{N\times N}$ the feedback filter bank matrix. With the covariance matrices of the input signal and the noise defined as $\mathbf{R}_{xx} = E\{\mathbf{xx}^H\}$ and $\mathbf{R}_{nn} = E\{\mathbf{nn}^H\}$, one can write the error covariance matrix as

$$\mathbf{R}_{\varepsilon\varepsilon} = E\{\boldsymbol{\varepsilon\varepsilon}^H\} = \left(\mathbf{R}_{xx}^{-1} + \mathbf{H}_F^H\mathbf{R}_{nn}^{-1}\mathbf{H}_F\right)^{-1}, \quad (6)$$

where $\mathbf{H}_F = \mathbf{H}_0\mathbf{F}$.

From following Cholesky factorization with symmetric permutation

$$\mathbf{PR}_{\varepsilon\varepsilon}\mathbf{P}^T = \mathbf{LDL}^H \quad (7)$$

where $\mathbf{L}$ is unit lower triangular matrix and $\mathbf{D}$ is diagonal matrix, feedforward and feedback filter banks that minimize $\mathbf{R}_{\varepsilon\varepsilon}$ can be expressed as

$$\mathbf{W} = \mathbf{DL}^H\mathbf{PH}_F^H\mathbf{R}_{nn}^{-1}, \quad \mathbf{B} = \mathbf{L}^{-1} - \mathbf{I}. \quad (8)$$

For comparison, we also extend the MMSE-DFE in [9] for MIMO channels and propose an alternative design for the second DFE as follows:

$$\mathbf{W} = \mathbf{UR}_{xx}\mathbf{H}_F^H\left(\mathbf{H}_F\mathbf{R}_{xx}\mathbf{H}_F^H + \mathbf{R}_{nn}\right)^{-1} \text{ and } \mathbf{B} = \mathbf{U} - \mathbf{I}, \quad (9)$$

where $\mathbf{U}$ is the upper triangular matrix with unit diagonal, derived from Cholesky factorization

$$\left(\mathbf{R}_{xx}^{-1} + \mathbf{H}_F^H\mathbf{R}_{nn}^{-1}\mathbf{H}_F\right)^{-1} = \mathbf{U}^H\boldsymbol{\Delta}\mathbf{U}. \quad (10)$$

2.2. Design of multi-tap MIMO DFE

In this design, instead of using a separate DFE to eliminate IBI, we will use a DFE with multi-tap feedforward and feedback filter banks to simultaneously deal with IBI and CCI. Thus, we set $\mathbf{G}(z) = \mathbf{0}$ and assume that the feedforward filter bank has three taps: $\mathbf{W}_{-1}, \mathbf{W}_0$, $\mathbf{W}_1$ and feedback filter bank has two taps: $\mathbf{B}_0$ and $\mathbf{B}_1$. We can re-write the input signal of the feedforward filter bank as

$$\mathbf{z}[i] = \mathbf{H_0Fx}[i] + \mathbf{H_1Fx}[i-1] + \mathbf{n}[i], \quad (11)$$

and the signal at the input of the decision device as

$$\mathbf{\tilde{x}}[i] = \mathbf{W}_{-1}\mathbf{z}[i+1] + \mathbf{W_0z}[i] + \mathbf{W_1z}[i-1] - \mathbf{B}_0\mathbf{\hat{x}}[i] - \mathbf{B}_1\mathbf{\hat{x}}[i-1]. \quad (12)$$

Defining following matrices

$$\mathbf{S}=\begin{bmatrix}\mathbf{H}_0\mathbf{F} & \mathbf{H}_1\mathbf{F} & \mathbf{0}\\ \mathbf{0} & \mathbf{H}_0\mathbf{F} & \mathbf{H}_1\mathbf{F}\\ \mathbf{0} & \mathbf{0} & \mathbf{H}_0\mathbf{F}\end{bmatrix},\qquad \mathbf{R}_{\overline{nn}}=\begin{bmatrix}\mathbf{R}_{nn} & \mathbf{0} & \mathbf{0}\\ \mathbf{0} & \mathbf{R}_{nn} & \mathbf{0}\\ \mathbf{0} & \mathbf{0} & \mathbf{R}_{nn}+\mathbf{H}_1\mathbf{F}(\mathbf{H}_1\mathbf{F})^H\end{bmatrix} \tag{13}$$

and $\mathbf{R}_{\overline{yy}}=\mathbf{S}\mathbf{S}^H+\mathbf{R}_{\overline{nn}}$, we can now write the tap weights of the feedforward and feedback filter banks that minimize the mean square error as [9]

$$\begin{aligned}\left[\mathbf{W}_{-1}\quad \mathbf{W}_0\quad \mathbf{W}_1\right]&=\left[\mathbf{0}\quad \mathbf{Q}_{22}\quad \mathbf{Q}_{23}\right]\mathbf{S}^H\mathbf{R}_{\overline{yy}}^{-1}\\ \mathbf{B}_0&=\mathbf{Q}_{22}-\mathbf{I}\\ \mathbf{B}_1&=\mathbf{Q}_{23}\end{aligned} \tag{14}$$

where $\mathbf{Q}_{22}$, $\mathbf{Q}_{23}\in\mathbb{C}^{N\times N}$ are submatrices of the matrix $\mathbf{Q}\in\mathbb{C}^{3N\times 3N}$ derived from the following Cholesky factorization

$$\mathbf{I}+\mathbf{S}^H\mathbf{R}_{\overline{nn}}^{-1}\mathbf{S}=\mathbf{Q}^H\mathbf{\Sigma}\mathbf{Q} \tag{15}$$

$$\mathbf{Q}=\begin{bmatrix}\mathbf{Q}_{11} & \mathbf{Q}_{12} & \mathbf{Q}_{13}\\ \mathbf{0} & \mathbf{Q}_{22} & \mathbf{Q}_{23}\\ \mathbf{0} & \mathbf{0} & \mathbf{Q}_{33}\end{bmatrix}.$$

3. SIMULATION RESULTS AND DISCUSSION

In order to assess the performance of the proposed equalizers, we will perform a simulation and compare their BER performance. The channel is assumed to be FIR MIMO with the length of $L+1=15$, channel coefficients drawn from complex Gaussian distribution with zero mean and unit variance. Number of transmit and receive antenna $N_T=N_R=4$, the length of transmit symbol block $PN_T=184$ and the length of input symbol block N is chosen to be either 152 or 164, corresponding to cases where $L_r/N_T=8$ and $L_r/N_T=5$, respectively. We use 4-QAM modulation for all designs.

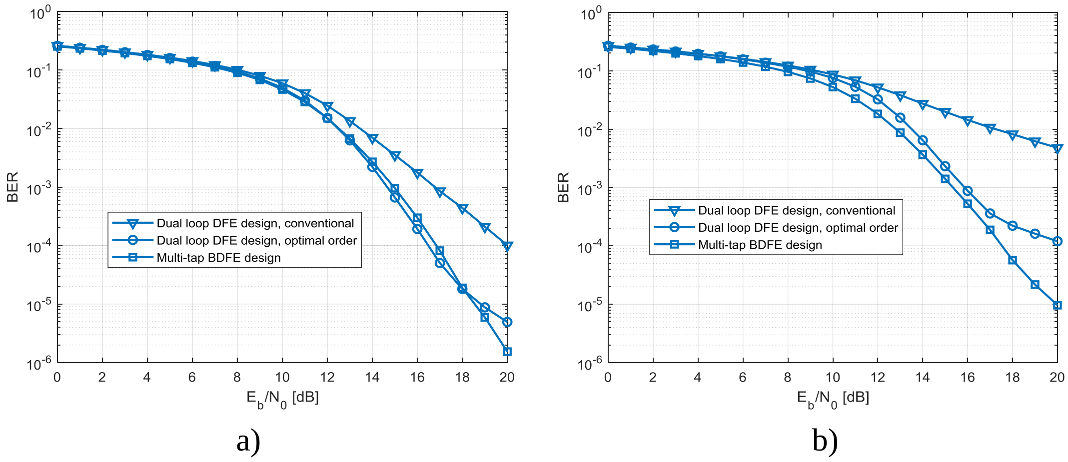

Figure 2. *BER performance of the proposed designs in case of quasi-static channel.*
a) $L_r/N_T=8$, *b)* $L_r/N_T=5$

Figure 2 shows the simulated BER performance of proposed designs for the case where the channel is assumed to be quasi-static. Figure 2.a corresponds to the case of $L_r / N_T = 8$ and figure 2.b corresponds to the case of $L_r / N_T = 5$. Note that with a channel length of 15, it can be considered that $L_r / N_T = 8$ equivalent to half redundancy and $L_r / N_T = 5$ equivalent to one-third redundancy. It can be seen from the figure that the multi-tap MIMO DFE works effectively with quasi-static channels and provides a steady BER performance. Especially, when $L_r / N_T = 5$ or the redundancy is low, the multi-tap MIMO DFE design can effectively deal with both IBI and CCI. The dual loop DFE designs, on the contrary, are not very effective in eliminating the remaining IBI. The design that uses DFE with optimal order does outperform the design that uses conventional MMSE-DFE and its BER performance can be comparable with that of multi-tap DFE design when the redundancy is at the medium level (Fig. 2.a). However, when the redundancy gets lower, the dual loop DFE design with optimal order cannot outperform the multi-tap DFE. This design also tends to have an error-floor at high SNR. In our view the reason is that the error in (5) assumes that the output symbols are correctly detected. Therefore, when the SNR is high, the error due to the remaining IBI will be dominated and the second DFE has no knowledge about this interference.

Figure 3 shows the simulated BER performance of dual loop DFEs when the channel is experienced Rayleigh fading and $L_r / N_T = 8$. It can be seen from the figure that the two DFEs perform well under Raleigh fading, although BER performance is higher than that of the case when the channel is quasi-static. The dual loop DFE with optimal order still maintains a better performance than that of the conventional DFE.

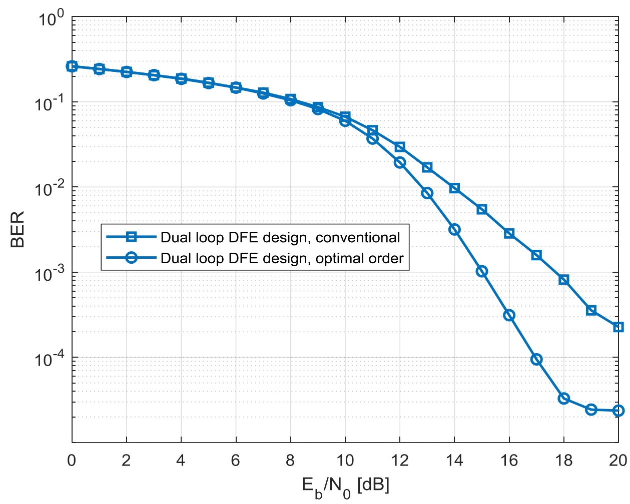

Figure 3. *BER performance of the dual loop DFE designs in case of Rayleigh fading channel,* $L_r / N_T = 8$.

The multi-tap MIMO DFE, unfortunately, cannot work under Rayleigh fading. The structure of matrix $\mathbf{S}$ in (13) reflects that this DFE is designed under the assumption that the channel is unchanged. The implementation of multi-tap DFE that works in the condition of Rayleigh fading is an interesting research topic and will be mentioned in a future work.

Although the usage of permutation matrix or multi-tap filter banks makes the DFEs proposed in Sec. 2 more complex, but with the gain achieved over conventional DFE this added complexity can be considered acceptable.

4. CONCLUSIONS

In our paper, several designs of DFE for FS-MIMO channels have been proposed. The transmit filter bank introduces an amount of redundancy, which is much smaller than that of conventional block transmission systems. In the dual loop DFE designs, we use the first feedback loop for remaining IBI elimination, the CCI inherited in MIMO channels is cancelled by the main DFE which can either be the MMSE-DFE or MMSE-DFE with optimal detection order. In the multi-tap MIMO DFE design, we use a DFE with multi-tap feedback and feedforward filter banks which we extended from a design for single-input single-output channels.

Simulation results show that the proposed designs can work well with low redundancy. With quasi-static channels, the multi-tap MIMO DFE design provides the lowest BER, the dual loop DFE with optimal order also has comparable BER with that of multi-tap DFE but tends to have error floor. As redundancy gets lower, the multi-tap MIMO DFE still maintains the best BER performance. When the channel experiences Rayleigh fading, the dual loop DFE with optimal order can provide good BER performance while the proposed multi-tap MIMO DFE is not suitable for Rayleigh fading.

Future works would concern the implementation of multi-tap DFE that can work with Rayleigh fading channels, the improvement of dual loop DFE with optimal order so that it can take remaining IBI into account at high SNRs.

The proposed designs would be suitable for the uplink in wireless communication systems where, due to their complexity, the DFEs can reside at the base station. These designs also are applicable for indoor transmission schemes where the channel is stable and a high data rate are required.

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TÓM TẮT

Thiết kế các bộ san bằng cho các hệ thống MIMO chọn lọc theo tần số với độ dư thấp

Trong bài này, chúng tôi đề xuất phương án thiết kế các bộ san bằng hồi tiếp quyết định (DFE) dựa trên các bộ lọc dãy cho các hệ thống MIMO có pha đinh chọn lọc theo tần số. Các hệ thống truyền dẫn trên các kênh pha đinh chọn lọc tần số thông thường là các hệ thống truyền dẫn theo khối và phải dùng một độ dư dưới dạng các khoảng bảo vệ để tránh nhiễu giữa các khối (IBI). Bài báo này sẽ tập trung vào trường hợp phải dùng một độ dư thấp để đảm bảo yêu cầu về hiệu quả phổ. Khi đó, phần nhiễu IBI còn lại sẽ được loại bỏ nhờ vào một vòng hồi tiếp riêng, sau đó, nhiễu đồng kênh tồn tại trong hệ thống MIMO sẽ được loại bỏ bằng một số bộ san bằng DFE khác nhau. Một phương án khác là sử dụng bộ san bằng DFE nhiều khâu để vừa loại bỏ nhiễu IBI, vừa loại bỏ nhiễu CCI và tách các symbol tín hiệu.

Kết quả mô phỏng cho thấy, với kênh MIMO chọn lọc theo tần số gần tĩnh, bộ san bằng DFE nhiều khâu cho phẩm chất tốt nhất trong khi với kênh có pha đinh Rayleigh bộ san bằng DFE với vòng hồi tiếp kép là giải pháp tốt cho trường hợp truyền dẫn với độ dư thấp.

Từ khoá: MIMO; IBI; Chọn lọc theo tần số; Băng rộng; Bộ san bằng hồi tiếp quyết định.