Joint Tx/Rx Filtering for Single-carrier MIMO Spatial Multiplexing and HARQ Packet Combining

Shinya KUMAGAI† Tatsunori OBARA† Tetsuya YAMAMOTO† and Fumiyuki ADACHI‡

Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University
6-6-05, Aza-Aoba, Aramaki, Aoba-ku, Sendai, 980-8579, JAPAN
E-mail: †{kumagai, obara, yamamoto}@mobile.ecei.tohoku.ac.jp ‡adachi@ecei.tohoku.ac.jp

Abstract—**Recently, we proposed a joint transmit/receive frequency-domain minimum mean square error based filtering (joint Tx/Rx MMSE-FD filtering) to improve the transmission performance of broadband single-carrier (SC) multiple-input multiple-output (MIMO) spatial multiplexing. For wireless packet communications, hybrid automatic repeat-request (HARQ) plays an important role as the error control technique to achieve a higher throughput performance. In this paper, we modify and optimize the joint Tx/Rx MMSE-FD filtering for its application to SC-MIMO spatial multiplexing and HARQ packet combining. At each retransmission of the same packet, joint Tx/Rx MMSE-FD filtering is carried out so as to minimize the MSE after packet combining. Computer simulation results show that joint Tx/Rx MMSE-FD filtering offers an improved throughput performance compared to the conventional Rx MMSE-FD filtering.**

Keywords; Single-carrier transmission, MIMO spatial multiplexing, MMSE filtering, HARQ

I. INTRODUCTION

Multiple-input multiple-output (MIMO) spatial multiplexing [1] is a promising technique to increase the transmission data rate without increasing the signal bandwidth. MIMO spatial multiplexing with orthogonal frequency-division multiplexing (OFDM) [2] has been attracting much attention because of its robustness against the frequency-selective fading [3]. However, the OFDM signal has a disadvantage of its high peak-to-average power ratio (PAPR) property.

Recently, single-carrier (SC) block transmission with MIMO spatial multiplexing has gained an increasing popularity because of its lower PAPR property [4]. SC-MIMO spatial multiplexing suffers not only from the inter-antenna interference (IAI) but also from the inter-symbol interference (ISI) caused by severe frequency-selectivity of the channel. The minimum mean square error based frequency-domain receive filtering (Rx MMSE-FD filtering) [4] for SC-MIMO spatial multiplexing can achieve a good transmission performance with low-complexity. However, its performance improvement is limited due to the residual IAI and ISI.

To improve the transmission performance of SC-MIMO spatial multiplexing, we recently proposed a joint transmit/receive MMSE-FD filtering (joint Tx/Rx MMSE-FD filtering) [5]. Joint Tx/Rx MMSE-FD filtering requires the channel state information (CSI) at both transmitter and receiver. This filtering transforms the MIMO channel to multiple orthogonal channels (i.e., eigenmodes) so as to avoid the IAI. It also jointly performs MMSE based transmit power allocation and receive frequency-domain equalization (FDE) on each eigenmode so as to suppress the ISI.

Hybrid automatic repeat-request (HARQ) [6,7] is an important error control technique to realize a high-speed wireless packet access. In HARQ using the Chase combining (CC) strategy (or Type I), if any error is detected in a received packet after decoding, the same packet is retransmitted [6]. The retransmitted packets are combined to achieve the time diversity gain. In [8], MMSE based transmit power allocation and receive FDE are applied to single-input single-output (SISO) SC-HARQ packet combining. At each retransmission of the same packet, the transmit power allocation and receive FDE weight are updated to minimize the MSE after the packet combining.

In this paper, we modify and optimize the joint Tx/Rx MMSE-FD filtering for its application to SC-MIMO spatial multiplexing and HARQ packet combining. At each retransmission of the same packet, the transmitter applies the Tx MMSE-FD filtering to the packet to be retransmitted and the receiver applies the Rx MMSE-FD filtering to past and present received packets to combine them. It is shown by computer simulation that joint Tx/Rx MMSE-FD filtering offers an improved throughput performance compared to the Rx MMSE-FD filtering.

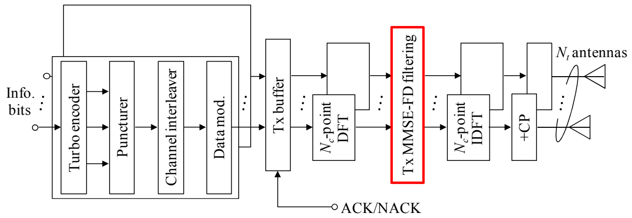
The remainder of this paper is organized as follows. Sect. II presents the system model and signal expressions for SC-MIMO spatial multiplexing and HARQ packet combining with joint Tx/Rx MMSE-FD filtering. In Sect. III, we derive the Tx and Rx MMSE-FD filters and discuss their behavior. In Sect. IV, we evaluate by computer simulation the throughput performance achievable with the proposed method. Section V gives some concluding remarks.

II. SYSTEM MODEL

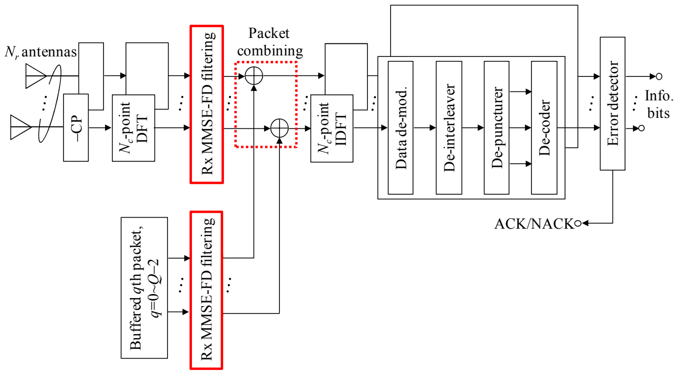
In this paper, the CC strategy is considered for HARQ. The proposed packet combining method can be easily extended to the incremental redundancy (IR) strategy [7].

System model of SC-MIMO spatial multiplexing and HARQ packet combining with joint Tx/Rx MMSE-FD filtering is illustrated in Fig. 1. In this paper, turbo coding is used for HARQ. At the transmitter, the information bit sequence is turbo encoded with the coding rate R. A packet is generated by data-modulating the resultant codeword. The data sequence of the packet is grouped in a sequence of N_c-symbol blocks, where N_c is the size of discrete Fourier transform (DFT) and inverse DFT (IDFT). Each symbol block is transformed into a frequency-domain symbol block by N_c-point DFT. After the Tx MMSE-FD filtering is applied to these N_t frequency-domain symbol blocks, where N_t is the number of transmit antennas, each block is transformed back to time-domain symbol block by N_c-point IDFT. Finally, the last N_g symbols of each transmit block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI) at the beginning of each transmit block and then transmitted from N_t antennas.

At the receiver, each CP is removed from the signal blocks received by N_r antennas and then, each block is transformed into the frequency-domain signal block by N_c-point DFT. After the Rx MMSE-FD filtering and packet combining are applied to all retransmitted N_r frequency-domain signal blocks, each block is transformed back to time-domain soft-output block by N_c-point IDFT. Error detection is performed after turbo decoding. If any errors are detected, the Negative acknowledgement (NACK) signal is sent to the transmitter to request the retransmission of the same packet. On the other hand, if no error is detected, the ACK signal is sent to the transmitter to request the transmission of a new packet.



(a) Transmitter



(b) Receiver

Figure 1. System model.

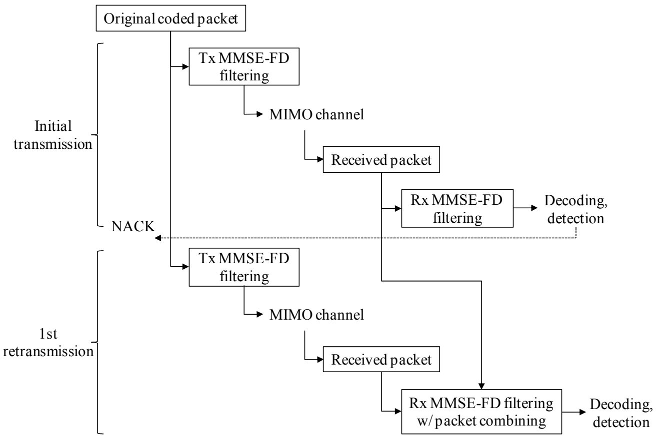


Figure 2. Diagram of packet combining and retransmission.

A conceptual diagram of SC-MIMO spatial multiplexing and HARQ packet combining with joint Tx/Rx MMSE-FD filtering is illustrated in Fig. 2. The transmitter applies the Tx FD filtering to the packet to be retransmitted and the receiver applies the Rx FD filtering to combine the past and present received packets.

A. Transmit signal

It is assumed that the same packet has been retransmitted Q times (including the initial transmission). In the (Q−1)th retransmission, at the transmitter, the N_t×1 transmit symbol vector $\mathbf{S}^{(Q-1)}(k)$ at the kth frequency is obtained by applying the Tx MMSE-FD filtering to the N_t×1 frequency-domain symbol vector $\mathbf{D}(k)=[D_0(k),\ldots, D_n(k),\ldots, D_{N_t-1}(k)]^T$ at the kth frequency, which is expressed as

$$\begin{aligned}\mathbf{S}^{(Q-1)}(k)&=[S_0^{(Q-1)}(k),...,S_n^{(Q-1)}(k),...,S_{N_t-1}^{(Q-1)}(k)]^T\\&=\mathbf{W}_t^{(Q-1)}(k)\mathbf{D}(k)\end{aligned}, \qquad (1)$$

where $\mathbf{W}_t^{(Q-1)}(k)$ is an N_t×N_t Tx MMSE-FD filter matrix. $(.)^T$ represents the transpose operation. N_c-point IDFT is applied to each transmit symbol block $\{S_n^{(Q-1)}(k);\, k=0\sim N_c-1\}$, $n=0\sim N_t-1$, and each block is transmitted after CP-insertion.

B. Received signal

After the (Q−1)th retransmitted packet is received, the receiver combines Q received packets. The N_r×1 frequency-domain received signal vector $\mathbf{R}^{(q)}(k)$, $q=0\sim Q-1$, at the kth frequency after N_c-point DFT is expressed as

$$\begin{aligned}\mathbf{R}^{(q)}(k)&=[R_0^{(q)}(k),...,R_m^{(q)}(k),...,R_{N_r-1}^{(q)}(k)]^T\\&=\sqrt{2E_s/T_s}\,\mathbf{H}^{(q)}(k)\mathbf{S}^{(q)}(k)+\mathbf{Z}^{(q)}(k)\end{aligned}, \qquad (2)$$

where E_s and T_s are respectively the transmit symbol energy and the symbol duration, $\mathbf{H}^{(q)}(k)$ is the N_r×N_t MIMO channel matrix and $\mathbf{Z}^{(q)}(k)=[Z_0^{(q)}(k),...,Z_m^{(q)}(k),...,Z_{N_r-1}^{(q)}(k)]^T$ is the noise vector whose elements are zero-mean complex-valued random variables having variance $2N_0/T_s$ with N_0 being the one-sided power spectrum density of additive white Gaussian noise (AWGN).

The N_t×1 frequency-domain soft-output vector $\hat{\mathbf{D}}(k)$ is obtained by performing the Rx MMSE-FD filtering and packet combining as

$$\begin{aligned}\hat{\mathbf{D}}(k)&=[\hat{D}_0(k),...,\hat{D}_n(k),...,\hat{D}_{N_t-1}(k)]^T\\&=\sum_{q=0}^{Q-1}\mathbf{W}_r^{(q)}(k)\mathbf{R}^{(q)}(k)\\&=\sqrt{\frac{2E_s}{T_s}}\sum_{q=0}^{Q-1}\mathbf{W}_r^{(q)}(k)\mathbf{H}^{(q)}(k)\mathbf{W}_t^{(q)}(k)\mathbf{D}(k)+\sum_{q=0}^{Q-1}\mathbf{W}_r^{(q)}(k)\mathbf{Z}^{(q)}(k)\end{aligned}, \qquad (3)$$

where $\mathbf{W}_r^{(q)}(k)$ is the N_t×N_r Rx MMSE-FD filter matrix for the qth received packet. N_c-point IDFT is applied to each frequency-domain soft-output block $\{\hat{D}_n(k);\, k=0\sim N_c-1\}$, $n=0\sim N_t-1$, and then the time-domain soft-output block is obtained.

III. DERIVATION OF TX AND RX MMSE-FD FILTERS

A. Formulation of objective function

We define the QN_r×1 expanded received signal vector $\mathbf{R}(k)$ as

$$\mathbf{R}(k) \equiv \begin{bmatrix} \mathbf{R}^{(0)}(k) \\ \vdots \\ \mathbf{R}^{(Q-1)}(k) \end{bmatrix} = \sqrt{2E_s/T_s}\,\overline{\mathbf{H}}(k)\mathbf{D}(k) + \overline{\mathbf{Z}}(k), \tag{4}$$

where

$$\overline{\mathbf{H}}(k) = \begin{bmatrix} \mathbf{H}^{(0)}(k)\mathbf{W}_t^{(0)}(k) \\ \vdots \\ \mathbf{H}^{(Q-1)}(k)\mathbf{W}_t^{(Q-1)}(k) \end{bmatrix},\ \overline{\mathbf{Z}}(k) = \begin{bmatrix} \mathbf{Z}^{(0)}(k) \\ \vdots \\ \mathbf{Z}^{(Q-1)}(k) \end{bmatrix}. \tag{5}$$

Defining the expanded Rx MMSE-FD filter matrix as $\overline{\mathbf{W}}_r(k) \equiv [\mathbf{W}_r^{(0)}(k),\ldots,\mathbf{W}_r^{(Q-1)}(k)]$, Eq. (3) can be rewritten as

$$\hat{\mathbf{D}}(k) = \sqrt{2E_s/T_s}\,\overline{\mathbf{W}}_r(k)\overline{\mathbf{H}}(k)\mathbf{D}(k) + \overline{\mathbf{W}}_r(k)\overline{\mathbf{Z}}(k). \tag{6}$$

The total MSE of the blocks between the transmit symbol vector $\mathbf{D}(k)$ and the soft-output vector $\hat{\mathbf{D}}(k)$ is defined as

$$\varepsilon \equiv E\left[\sum_{k=0}^{N_c-1} tr\left\{\left(\mathbf{D}(k) - \hat{\mathbf{D}}(k)\Big/\sqrt{\frac{2E_s}{T_s}}\right)\left(\mathbf{D}(k) - \hat{\mathbf{D}}(k)\Big/\sqrt{\frac{2E_s}{T_s}}\right)^H\right\}\right]. \tag{7}$$

where $(.)^H$ is the Hermitian transpose operation. From Eq. (6), the total MSE can be rewritten as

$$\begin{aligned}\varepsilon &= \sum_{k=0}^{N_c-1} tr\left[\left\{\mathbf{I}_{N_t} - \overline{\mathbf{W}}_r(k)\overline{\mathbf{H}}(k)\right\}\left\{\mathbf{I}_{N_t} - \overline{\mathbf{W}}_r(k)\overline{\mathbf{H}}(k)\right\}^H\right] \\ &\quad + \gamma^{-1}\sum_{k=0}^{N_c-1} tr\left\{\overline{\mathbf{W}}_r(k)\overline{\mathbf{W}}_r^H(k)\right\}\end{aligned}, \tag{8}$$

where $\mathbf{I}_X$ is the X×X identity matrix and $\gamma=E_s/N_0$. At the receiver, the Rx MMSE-FD filter matrix $\overline{\mathbf{W}}_r(k)$ is updated at every retransmission. On the other hand, at the transmitter, only the Tx MMSE-FD filter matrix $\mathbf{W}_t^{(Q-1)}(k)$ for the currently retransmitting packet can be optimized. Therefore, the minimization of the total MSE given by Eq. (8) under the total transmit power constraint is rewritten by the optimization problem as

$$\begin{aligned}&\min.\ \varepsilon \\ &\text{s.t.} \sum_{k=0}^{N_c-1} tr\left[\mathbf{W}_t^{(Q-1)}(k)\left\{\mathbf{W}_t^{(Q-1)}(k)\right\}^H\right] = N_t N_c.\end{aligned} \tag{9}$$

We want to find the Tx and Rx MMSE-FD filters which satisfy Eq. (9). In this paper, as the case in [8], we first derive the Rx MMSE-FD filter matrix $\overline{\mathbf{W}}_r(k)$ conditioned on the Tx MMSE-FD filter. Then, we derive the Tx MMSE-FD filter matrix $\mathbf{W}_t^{(Q-1)}(k)$ by solving the optimization problem for the given $\overline{\mathbf{W}}_r(k)$.

B. Rx MMSE-FD filter conditioned on Tx MMSE-FD filter

In this section, we derive the Rx MMSE-FD filter matrix $\overline{\mathbf{W}}_r(k)$ by considering $\overline{\mathbf{H}}(k)$ as the equivalent channel transfer function. In this case, it is minimized when $\partial\varepsilon/\partial\,\overline{\mathbf{W}}_r(k)=0$ since the objective function is a concave function. Therefore, the optimal $\overline{\mathbf{W}}_r(k)$ is given as

$$\overline{\mathbf{W}}_r(k) = \overline{\mathbf{H}}^H(k)\left\{\gamma^{-1}\cdot\mathbf{I}_{QN_r} + \overline{\mathbf{H}}(k)\overline{\mathbf{H}}^H(k)\right\}^{-1}. \tag{10}$$

The Rx MMSE-FD filter $\mathbf{W}_r^{(q)}(k)$ for each received packet can be derived by using the matrix inversion lemma [9] as

$$\begin{aligned}\mathbf{W}_r^{(q)}(k) &= \left[\gamma^{-1}\cdot\mathbf{I}_{N_t} + \sum_{q'=0}^{Q-1}\left\{\mathbf{H}^{(q')}(k)\mathbf{W}_t^{(q')}(k)\right\}^H\mathbf{H}^{(q')}(k)\mathbf{W}_t^{(q')}(k)\right]^{-1} \\ &\quad\times\left\{\mathbf{H}^{(q)}(k)\mathbf{W}_t^{(q)}(k)\right\}^H\end{aligned}$$

$$, q=0\text{~}Q-1. \tag{11}$$

C. Tx MMSE-FD filter

The objective function is expressed as the function of only the Tx MMSE-FD filter matrix $\mathbf{W}_t^{(Q-1)}(k)$ by substituting the optimal $\overline{\mathbf{W}}_r(k)$ which is derived in the foregoing section into the objective function. The optimization problem is rewritten by substituting Eq. (10) into Eq. (8) and using the matrix inversion lemma as

$$\begin{aligned}&\min.\ \varepsilon = \sum_{k=0}^{N_c-1} tr\left[\mathbf{I}_{N_t} + \gamma\cdot\sum_{q'=0}^{Q-1}\left\{\mathbf{H}^{(q')}(k)\mathbf{W}_t^{(q')}(k)\right\}^H\mathbf{H}^{(q')}(k)\mathbf{W}_t^{(q')}(k)\right]^{-1} \\ &\text{s.t.} \sum_{k=0}^{N_c-1} tr\left[\mathbf{W}_t^{(Q-1)}(k)\left\{\mathbf{W}_t^{(Q-1)}(k)\right\}^H\right] = N_t N_c.\end{aligned} \tag{12}$$

$\mathbf{H}^{(q)}(k)$ and $\mathbf{W}_t^{(q)}(k)$ can be transformed by singular value decomposition (SVD) [9] as

$$\begin{aligned}\mathbf{H}^{(q)}(k) &= \mathbf{U}_h^{(q)}(k)\sqrt{\mathbf{\Lambda}_h^{(q)}(k)}\left\{\mathbf{V}_h^{(q)}(k)\right\}^H \\ \mathbf{W}_t^{(q)}(k) &= \mathbf{U}_t^{(q)}(k)\sqrt{\mathbf{P}_t^{(q)}(k)}\left\{\mathbf{V}_t^{(q)}(k)\right\}^H\end{aligned}, \tag{13}$$

where $\mathbf{V}_h^{(q)}(k)$, $\mathbf{U}_t^{(q)}(k)$, and $\mathbf{V}_t^{(q)}(k)$ are respectively the N_t×N_t unitary matrices. $\mathbf{U}_h^{(q)}(k)$ is the N_r×N_r unitary matrix. $\mathbf{\Lambda}_h^{(q)}(k)$ is the N_r×N_t matrix whose (j,j)th element equals to the jth eigenvalue of $\left\{\mathbf{H}^{(q)}(k)\right\}^H\mathbf{H}^{(q)}(k)$; j=0~rank[$\left\{\mathbf{H}^{(q)}(k)\right\}^H\mathbf{H}^{(q)}(k)$]=$J$, and any other elements are zero. $\mathbf{P}_t^{(q)}(k)$ is the N_t×N_t diagonal matrix whose diagonal elements equal to the eigenvalues of $\left\{\mathbf{W}_t^{(q)}(k)\right\}^H\mathbf{W}_t^{(q)}(k)$. Since tr[**AB**]=tr[**BA**], where **A** and **B** are respectively N_t×N_t matrices, Eq. (12) can be rewritten by substituting Eq. (13) as

$$\begin{aligned}&\min.\ \varepsilon \\ &= \sum_{k=0}^{N_c-1} tr\left[\left\{\mathbf{I}_{N_t} + \gamma\cdot\sum_{q=0}^{Q-1}\begin{bmatrix}\sqrt{\mathbf{P}_t^{(q)}(k)}\left\{\mathbf{U}_t^{(q)}(k)\right\}^H\mathbf{V}_h^{(q)}(k)\mathbf{\Lambda}_h^{(q)}(k) \\ \times\left\{\mathbf{V}_h^{(q)}(k)\right\}^H\mathbf{U}_t^{(q)}(k)\sqrt{\mathbf{P}_t^{(q)}(k)}\end{bmatrix}\right\}^{-1}\right] \\ &\text{s.t.} \sum_{k=0}^{N_c-1} tr\left\{\mathbf{P}_t^{(Q-1)}(k)\right\} = N_t N_c.\end{aligned} \tag{14}$$

It can be seen from Eq. (14) that the optimization problem does not depend on $\mathbf{V}_t^{(q)}(k)$ (i.e., $\mathbf{V}_t^{(q)}(k)$ can be set to arbitrary N_t×N_t unitary matrix). In this paper, we set $\mathbf{V}_t^{(q)}(k) = \mathbf{I}_{N_t}$ for the sake of brevity. In general, $tr[\mathbf{A}^{-1}]$ is minimized when **A** is a diagonal matrix [9]. Therefore, the objective function in Eq. (14) is minimized when $\mathbf{U}_t^{(q)}(k) = \mathbf{V}_h^{(q)}(k)$. From the above explanation, $\mathbf{W}_t^{(q)}(k)$ is expressed as

$$\mathbf{W}_t^{(q)}(k) = \mathbf{V}_h^{(q)}(k)\sqrt{\mathbf{P}_t^{(q)}(k)}. \tag{15}$$

The optimization problem is rewritten by substituting Eq. (15) into Eq. (12) as

$$\begin{aligned}&\min.\ \varepsilon=\sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1}\left\{1+\gamma\cdot\sum_{q=0}^{Q-1}P_j^{(q)}(k)\Lambda_j^{(q)}(k)\right\}^{-1}\\&\text{s.t.}\ \sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1}P_j^{(Q-1)}(k)=N_tN_c\end{aligned}, \tag{16}$$

where $P_j^{(q)}(k)$ and $\Lambda_j^{(q)}(k)$ are respectively the jth diagonal elements of $\mathbf{P}_t^{(q)}(k)$ and $\mathbf{\Lambda}_h^{(q)}(k)$. Following [10], the optimal solution is given as (for the sake of brevity, the derivation is omitted)

$$P_j^{(Q-1)}(k)=\max\left\{\frac{1}{\sqrt{\mu}}\frac{1}{\sqrt{\gamma\Lambda_j^{(Q-1)}(k)}}-\frac{\gamma^{-1}+\sum_{q=0}^{Q-2}P_j^{(q)}(k)\Lambda_j^{(q)}(k)}{\Lambda_j^{(Q-1)}(k)},0\right\}, \tag{17}$$

where μ is chosen to satisfy the constraint condition.

D. *Discussion of joint Tx/Rx MMSE-FD filtering*

In this section, we discuss the behavior of joint Tx/Rx MMSE-FD filtering derived in Sect. III-*B* and III-*C*. The equivalent channel matrix $\hat{\mathbf{H}}^{(q)}(k)$ for the qth retransmitted packet is expressed as

$$\begin{aligned}\hat{\mathbf{H}}^{(q)}(k)&=\mathbf{W}_r^{(q)}(k)\mathbf{H}^{(q)}(k)\mathbf{W}_t^{(q)}(k)\\&=diag\left[\frac{P_0^{(q)}(k)\Lambda_0^{(q)}(k)}{\gamma^{-1}+\sum_{q'=0}^{Q-1}P_0^{(q')}(k)\Lambda_0^{(q')}(k)},\ldots,\frac{P_{J-1}^{(q)}(k)\Lambda_{J-1}^{(q)}(k)}{\gamma^{-1}+\sum_{q'=0}^{Q-1}P_{J-1}^{(q')}(k)\Lambda_{J-1}^{(q')}(k)},\mathbf{0}\right].\end{aligned} \tag{18}$$

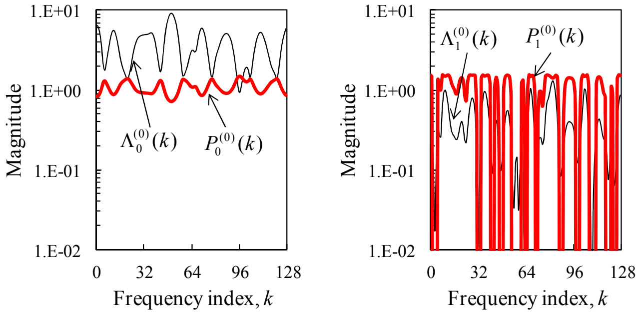
It can be seen from Eq. (18) that the MIMO channel matrix $\mathbf{H}^{(q)}(k)$ is diagonalized (i.e., the IAI is avoided) by joint Tx/Rx MMSE-FD filtering. $P_j(k)$ given by Eq. (17) is the transmit power allocated to the jth eigenmode at the kth frequency. Therefore, joint Tx/Rx MMSE-FD filtering achieves the eigenbeam-space division multiplexing (E-SDM) [1].

In the case of SISO SC-HARQ, the transmit power $P_t^{(Q-1)}(k)$ allocated to the kth frequency is given as [6]

$$\begin{aligned}&P_t^{(Q-1)}(k)\\&=\max\left\{\frac{1}{\sqrt{\mu}}\frac{1}{\sqrt{\gamma\left|H^{(Q-1)}(k)\right|^2}}-\frac{\gamma^{-1}+\sum_{q=0}^{Q-2}P_t^{(Q-1)}(k)\left|H^{(q)}(k)\right|^2}{\left|H^{(Q-1)}(k)\right|^2},0\right\},\end{aligned} \tag{19}$$

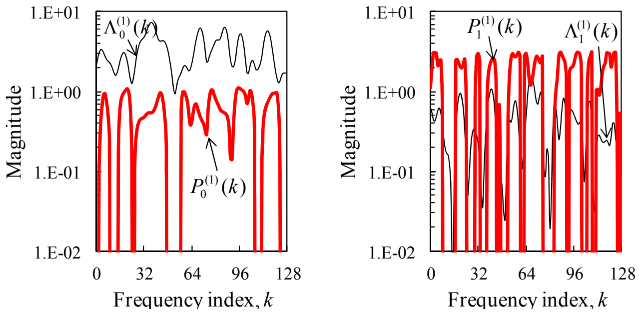
where $H^{(q)}(k)$ is the channel transfer function for the qth retransmitted packet. Comparing Eq. (17) to Eq. (19), it can be understood that in the case of MIMO transmission, the MMSE based transmit power allocation and receive FDE are jointly carried out on each eigenmode.

From the above, joint Tx/Rx MMSE-FD filtering can improve the transmission performance by avoiding the IAI by E-SDM and suppressing the ISI by the MMSE based transmit power allocation and receive FDE.



i) 0th eigenmode ii) 1st eigenmode

(a) Initial transmission (q=0)



i) 0th eigenmode ii) 1st eigenmode

(a) 1st retransmission (q=1)

Figure 3. One-shot observation of the power allocation.

Fig. 3 shows one-shot observation of the power allocation using the proposed method. N_t=N_r=2, N_c=128, average transmit E_s/N_0=0dB and an L=16-path frequency-selective block Rayleigh fading having uniform power delay profile are assumed.

Fig. 3 (a) shows the power allocation in the initial transmission. In this case, the power allocation is the same as that without taking packet combining into account [5]. At the 0th eigenmode which has a large $\Lambda_j(k)$, the impact of ISI is dominant because the received signal-to-noise ratio (SNR) after the Rx filtering is comparatively high. Therefore, the power is allocated like the inverse function of $\Lambda_j(k)$ to suppress the ISI. On the other hand, at the 1st eigenmode which has a small $\Lambda_j(k)$, the impact of noise is dominant because the received SNR after the Rx filtering is low. Therefore, the proposed method allocates no power to the frequencies which have small $\Lambda_j(k)$ but allocates much power to the other frequencies to improve the received SNR.

As shown in Fig. 3 (b), the behavior of the power allocation in the retransmission is different from the initial transmission. For the power allocation of the retransmission, the CSI and power allocation in the initial transmission, the present CSI, and the MMSE packet combing at the receiver are taken into account. Therefore, the proposed method allocates no power to the frequencies which were allocated much power in the initial transmission but allocates much power to the frequencies which were allocated no power in the initial transmission to make the equivalent channel flat after the packet combining. In addition, the proposed method allocates more power to the 1st eigenmode than the 0th eigenmode to reduce the received SNR gap between eigenmodes. Thus, the proposed method achieves high packet combining gain.

IV. PERFORMANCE EVALUATION

A. *Computer simulation condition*

TABLE I. COMPUTER SIMULATION CONDITION.

Channel coding	R=1/2 (13,15) RSC encoder Log-MAP decoding with 8 iterations	
HARQ	HARQ Type I (Chase combining)	
Transmitter & Receiver	Data modulation	QPSK, 16QAM
	No. of information bits	K=3072bits
	No. of DFT points	N_c=128
	Guard interval length	N_g=16
	Channel estimation	Ideal
	No. of transmit antennas	N_t=2
	No. of receive antennas	N_r=2
Channel	Fading	Frequency-selective block Rayleigh
	Path model	L=16-path with uniform power delay profile
	Time delay difference	1 Symbol

Computer simulation conditions are summarized in Table I. A turbo encoder with the original coding rate 1/3 using two (13,15) recursive systematic convolutional (RSC) encoders, a block interleaver/deinterleaver, and log-MAP turbo decoding with 8 iterations are used. Information sequence length is K=3072 bits. The two parity sequences from the turbo encoder are punctured to obtain rate-1/2 turbo codes. Ideal ACK/NACK transmissions are assumed. The channel is assumed to be an L=16-path frequency-selective block Rayleigh fading having uniform power delay profile. No antenna correlation and ideal channel estimation at both the transmitter and receiver are assumed. Independent channel is assumed for each retransmission.

B. *HARQ throughput performance*

Fig. 4 shows the throughput performance achievable with the proposed method. The throughput performance achievable with Rx MMSE-FD filtering is also plotted for comparison. It can be seen from Fig. 4 that the proposed method achieves a better throughput performance than Rx MMSE-FD filtering especially in the low average transmit E_s/N_0 region (about −10~10dB). This is because, as noted in Section III-*D*, the proposed method avoids the IAI by E-SDM, and suppresses the ISI and improves the received SNR (i.e., improves the received signal-to-interference plus noise ratio (SINR)) by applying the MMSE based transmit power allocation and receive FDE.

On the other hand, in the high average transmit E_s/N_0 region (about 11~14dB), the throughput performance of the proposed method is worse than that of Rx MMSE-FD filtering. When average transmit E_s/N_0 is high, the packet transmitted through the 0th eigenmode which has large eigenvalue is received without error in initial transmission. On the other hand, the packet transmitted through the 1st eigenmode which has small eigenvalue is not received correctly in initial transmission and hence, it limits the throughput improvement. Therefore, performance improvement of the minimum eigenmode by using adaptive modulation [11] is important to further increase the throughput.

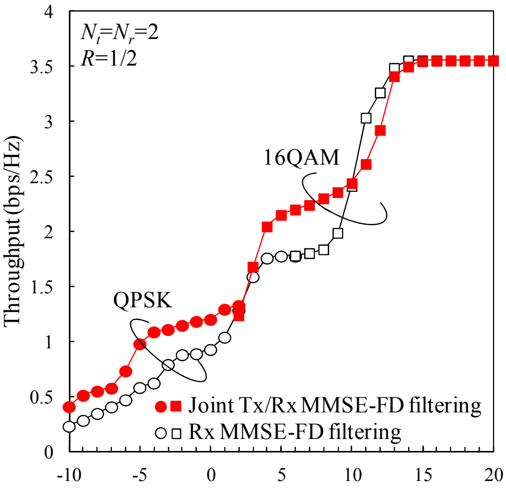


Figure 4. Throughput performance.

V. CONCLUSION

In this paper, we presented a joint Tx/Rx MMSE-FD filtering for HARQ with SC-MIMO spatial multiplexing. Joint Tx/Rx MMSE-FD filtering transforms the MIMO channel to eigenmodes so as to avoid the IAI and suppresses the ISI. In addition, it achieves high packet combining gain. It was shown by computer simulation that joint Tx/Rx MMSE-FD filtering can achieve better throughput performance than the Rx MMSE-FD filtering. A modification of joint Tx/Rx MMSE-FD filtering with SC-MIMO spatial multiplexing and HARQ packet combining using adaptive modulation [11] is left as an interesting future study.

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