Joint Tx/Rx MMSE Filtering and Joint Rank Adaptation/Adaptive Modulation for Single-carrier MIMO Packet Transmission

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Abstract—**In this paper, we propose a joint transmit and receive filtering based on minimum mean square error criterion (joint Tx/Rx MMSE filtering) and joint rank adaptation/adaptive modulation for single-carrier (SC) multiple-input multiple-output (MIMO) packet transmission. Joint Tx/Rx MMSE filtering transforms the MIMO channel to the orthogonal eigenmodes to avoid the inter-antenna interference (IAI) and allocates the transmit power based on MMSE criterion to sufficiently suppress the inter-symbol interference (ISI). Joint rank adaptation/adaptive modulation narrows the received signal-to-interference plus noise power ratio (SINR) gap between eigenmodes. The superiority of the SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation is confirmed by computer simulation.**

Keywords; Single-carrier transmission, MIMO, HARQ, MMSE filtering, Rank adaptation, Adaptive modulation

I. INTRODUCTION

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

Recently, single-carrier (SC) block transmission with MIMO has attracted much attention as an alternative technique because of its low PAPR property [4]. SC-MIMO 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 linear receive filtering (Rx MMSE filtering) [4] can achieve a good transmission performance with low-complexity. However, its performance improvement is limited due to the residual IAI and ISI after the Rx MMSE filtering. To improve the transmission performance, non-linear signal detection techniques such as the iterative interference cancellation [4] and the maximum likelihood detection [5] are proposed. However, their computational complexity is extremely high.

Another interesting technique to improve the transmission performance is joint transmit/receive linear signal detection. If the channel state information (CSI) is available at both the transmitter and receiver, the transmission performance can be improved while keeping the computational complexity low. In [6], joint transmit and receive MMSE based frequency-domain equalization (joint Tx/Rx MMSE-FDE) was proposed for the single-input single-output (SISO) SC block transmissions. By allocating the transmit power based on the MMSE criterion, joint Tx/Rx MMSE-FDE sufficiently suppresses the residual ISI and achieves a better performance than the Rx MMSE-FDE [7] which uses the CSI at the receiver only.

Recently, we proposed a joint Tx/Rx MMSE filtering for SC-MIMO packet transmission [8]. Unlike our previously proposed joint Tx/Rx MMSE-FDE (called the conventional joint Tx/Rx MMSE-FDE in this paper), joint Tx/Rx MMSE filtering transforms the MIMO channel to multiple orthogonal channels (i.e., eigenmodes) to avoid the IAI and jointly performs MMSE based frequency-domain transmit power allocation and receive FDE on each eigenmode to suppress the ISI. However, only the special case of full spatial multiplexing case is considered in [8], and therefore, the eigenmodes having low received signal-to-interference plus noise power ratio (SINR) limit the improvement of transmission performance.

In this paper, we propose a new joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation for SC-MIMO packet transmissions. Unlike conventional joint Tx/Rx MMSE filtering [8], our new joint Tx/Rx MMSE filtering allows an arbitrary number of data streams (i.e., rank) in every retransmission. A joint use of rank adaptation [9] and adaptive modulation [10] is introduced to narrow the received SINR gap between eigenmodes. The optimal combination of rank and modulation levels is found based on the minimum bit error rate (BER) criterion. Similar to conventional joint Tx/Rx MMSE filtering, MMSE packet combining can be applied to our proposed joint Tx/Rx MMSE filtering. However, rank and modulation levels cannot be changed every retransmission since MMSE packet combining can only be applied to packet retransmission using the same data-modulation. In this paper, to select the optimal rank and modulation levels every retransmission, bit log-likelihood ratio (LLR) packet combining [11] is applied to the joint Tx/Rx MMSE filtering.

The remainder of this paper is organized as follows. Sect. II presents the transmission system model and signal representation for SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation. Sect. III introduces joint rank adaptation/adaptive modulation. In Sect. IV, we evaluate the achievable throughput performance with the SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation by computer simulation. Section V gives the concluding remarks.

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II. SC-MIMO PACKET TRANSMISSION USING JOINT TX/RX MMSE FILTERING AND JOINT RANK ADAPTATION/ADAPTIVE MODULATION

A. System model

System model of SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation is illustrated in Fig. 1. Transmitter and receiver have N_t and N_r antennas, respectively. In this paper, turbo coding is used for hybrid automatic repeat-request (HARQ) [12]. At the transmitter, the information bit sequence is turbo encoded. A packet is generated by data-modulating the resultant codeword. The turbo coded bit sequence is data modulated based on joint rank adaptation/adaptive modulation expressed in Sect. III to J (is less than or equal to $\min(N_t,N_r)$) sequences 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 filtering is applied to these J frequency-domain symbol blocks, the output of N_t blocks are 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 filtering is applied to N_r frequency-domain signal blocks, the output of J blocks are transformed back to time-domain soft-output blocks by N_c-point IDFT. The bit log-likelihood ratio (LLR) is calculated from the time-domain soft-output, and the bit LLR packet combining (i.e., addition of bit LLR) is done if the received packet is the retransmitted packet. 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.

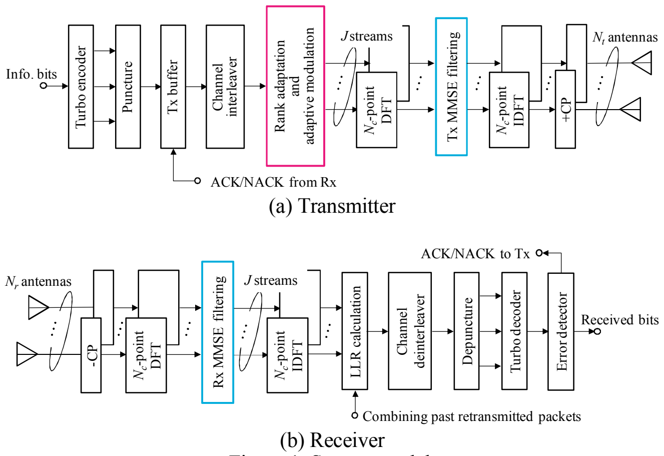


Figure 1. System model.

B. Transmit and received signals

At the transmitter, the N_t×1 transmit symbol vector $\mathbf{S}(k)$ at the kth frequency is obtained by applying the Tx MMSE filtering to the J×1 frequency-domain symbol vector $\mathbf{D}(k)=[D_0(k),\ldots,D_j(k),\ldots,D_{J-1}(k)]^T$ at the kth frequency, which is expressed as

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

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

The N_r×1 frequency-domain received signal vector $\mathbf{R}(k)$ at the kth frequency after N_c-point DFT is expressed as

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

where E_s and T_s are the average transmit symbol energy and symbol duration, respectively. $\mathbf{H}(k)$ is the N_r×N_t MIMO channel matrix and $\mathbf{Z}(k)=[Z_0(k),\ldots,Z_m(k),\ldots,Z_{N_r-1}(k)]^T$ 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 J×1 frequency-domain soft-output vector $\hat{\mathbf{D}}(k)$ is obtained by performing the Rx MMSE filtering on $\mathbf{R}(k)$ as

$$\begin{aligned}\hat{\mathbf{D}}(k) &= [\hat{D}_0(k),\ldots,\hat{D}_j(k),\ldots,\hat{D}_{J-1}(k)]^T \\ &= \mathbf{W}_r(k)\mathbf{R}(k) \\ &= \sqrt{2E_s/T_s}\,\mathbf{W}_r(k)\mathbf{H}(k)\mathbf{W}_t(k)\mathbf{D}(k)+\mathbf{W}_r(k)\mathbf{Z}(k),\end{aligned} \tag{3}$$

where $\mathbf{W}_r(k)$ is the J×N_r Rx filter matrix. N_c-point IDFT is applied to each frequency-domain soft-output block $\{\hat{D}_j(k);\ k=0\sim N_c-1\}$, $j=0\sim J-1$, and then the time-domain soft-output block is obtained. The bit LLR is calculated from the time-domain soft-output, and the bit LLR packet combining (i.e., addition of bit LLR) is done if the received packet is the retransmitted packet.

C. Derivation of Tx/Rx filters

In this section, we derive the optimal transmit and receive filters based on MMSE criterion. 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

$$\varepsilon \equiv E\left[\sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\left(\mathbf{D}(k)-\frac{\hat{\mathbf{D}}(k)}{\sqrt{2E_s/T_s}}\right)\left(\mathbf{D}(k)-\frac{\hat{\mathbf{D}}(k)}{\sqrt{2E_s/T_s}}\right)^H\right\}\right], \tag{4}$$

where tr(.) and $(.)^H$ are the trace operation and Hermitian transpose operation, respectively. From Eq. (3) and (4), the total MSE can be rewritten as

$$\varepsilon = \sum_{k=0}^{N_c-1} \mathrm{tr}\left[\left\{\mathbf{I}_J - \mathbf{W}_r(k)\mathbf{H}(k)\mathbf{W}_t(k)\right\}\left\{\mathbf{I}_J - \mathbf{W}_r(k)\mathbf{H}(k)\mathbf{W}_t(k)\right\}^H\right] + \gamma^{-1}\sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\mathbf{W}_r(k)\mathbf{W}_r^H(k)\right\}, \quad (5)$$

where $\mathbf{I}_A$ is the $A\times A$ identity matrix and $\gamma=E_s/N_0$. The minimization of the total MSE given by Eq. (5) under the total transmit power constraint is rewritten by the optimization problem as

$$\begin{aligned}&\left\{\mathbf{W}_{t,opt}(k), \mathbf{W}_{r,opt}(k)\right\} = \underset{\{\mathbf{W}_t(k),\mathbf{W}_r(k);k=0\sim N_c-1\}}{\arg\min.}\ \varepsilon \\ &\text{s.t.} \sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\mathbf{W}_t(k)\mathbf{W}_t^H(k)\right\} = N_c.\end{aligned} \quad (6)$$

The Tx and Rx filters which satisfy Eq. (6) are the optimal solution of MMSE filters. However, it is quite difficult to derive a set of MMSE matrices {$\mathbf{W}_{t,opt}(k)$, $\mathbf{W}_{r,opt}(k)$} at the same time since $\mathbf{W}_t(k)$ (or $\mathbf{W}_r(k)$) is the function of $\mathbf{W}_r(k)$ (or $\mathbf{W}_t(k)$). Therefore, in this paper, as is the case in [6], we first derive the optimal Rx filter matrix $\mathbf{W}_{r,opt}(k)$ considering (the Tx filter + channel) as the equivalent channel $\overline{\mathbf{H}}(k)=\mathbf{H}(k)\mathbf{W}_t(k)$. Then, we derive the optimal Tx filter matrix $\mathbf{W}_{t,opt}(k)$ by solving the optimization problem of Eq. (6) for the given $\mathbf{W}_{r,opt}(k)$.

By considering $\overline{\mathbf{H}}(k)=\mathbf{H}(k)\mathbf{W}_t(k)$ as the equivalent channel transfer function, the objective function becomes a concave function so that it is minimized when $\partial\varepsilon/\partial\mathbf{W}_r(k)=0$. Therefore, the $\mathbf{W}_{r,opt}(k)$ is given as

$$\mathbf{W}_{r,opt}(k) = \overline{\mathbf{H}}^H(k)\left\{\overline{\mathbf{H}}(k)\overline{\mathbf{H}}^H(k)+\gamma^{-1}\mathbf{I}_{N_r}\right\}^{-1}. \quad (7)$$

The objective function is expressed as only the function of the Tx filter matrix $\mathbf{W}_t(k)$ by substituting the $\mathbf{W}_{r,opt}(k)$ into the objective function. The optimization problem is rewritten by substituting Eq. (7) into Eq. (6) and using the matrix inversion lemma [13] as

$$\begin{aligned}&\mathbf{W}_{t,opt}(k) = \underset{\{\mathbf{W}_t(k);k=0\sim N_c-1\}}{\arg\min.} \sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\gamma\mathbf{H}(k)\mathbf{W}_t(k)\mathbf{W}_t^H(k)\mathbf{H}^H(k)+\mathbf{I}_{N_r}\right\}^{-1} \\ &\text{s.t.} \sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\mathbf{W}_t(k)\mathbf{W}_t^H(k)\right\} = N_c.\end{aligned} \quad (8)$$

$\mathbf{H}(k)$ and $\mathbf{W}_t(k)$ can be transformed by singular value decomposition [13] as

$$\begin{cases}\mathbf{H}(k) = \mathbf{U}_h(k)\sqrt{\mathbf{\Lambda}(k)}\mathbf{V}_h^H(k) \\ \mathbf{W}_t(k) = \mathbf{U}_t(k)\sqrt{\mathbf{P}(k)}\mathbf{V}_t^H(k)\end{cases}, \quad (9)$$

where $\mathbf{V}_h(k)$ and $\mathbf{U}_t(k)$ are respectively the $N_t\times N_t$ unitary matrices. $\mathbf{U}_h(k)$ is the $N_r\times N_r$ unitary matrix and $\mathbf{V}_t(k)$ is the $J\times J$ unitary matrix. $\mathbf{\Lambda}(k)$ is the $N_r\times N_t$ matrix whose (i,i) element has the ith eigenvalue of $\mathbf{H}(k)\mathbf{H}^H(k)$; i=0~rank[$\mathbf{H}(k)\mathbf{H}^H(k)$], and any other elements are zero. $\mathbf{P}(k)$ is the $N_t\times J$ matrix whose (j,j) element has the jth eigenvalue of $\mathbf{W}_t(k)\mathbf{W}_t^H(k)$. Since tr[$\mathbf{AB}$]=tr[$\mathbf{BA}$], where $\mathbf{A}$ and $\mathbf{B}$ are respectively a $A\times B$ and $B\times A$ matrices, Eq. (8) can be rewritten by substituting Eq. (9) as

$$\begin{aligned}&\left\{\mathbf{P}_{opt}(k), \mathbf{U}_{t,opt}(k)\right\} \\ &= \underset{\{\mathbf{P}(k),\mathbf{U}_t(k);k=0\sim N_c-1\}}{\arg\min.} \sum_{k=0}^{N_c-1}\mathrm{tr}\begin{Bmatrix}\gamma\cdot\sqrt{\mathbf{\Lambda}(k)}\mathbf{V}_h^H(k)\mathbf{U}_t(k)\sqrt{\mathbf{P}(k)} \\ \times\sqrt{\mathbf{P}^T(k)}\mathbf{U}_t^H(k)\mathbf{V}_h(k)\sqrt{\mathbf{\Lambda}^T(k)}+\mathbf{I}_{N_r}\end{Bmatrix}^{-1} \\ &\text{s.t.} \sum_{k=0}^{N_c-1}\mathrm{tr}\left\{\sqrt{\mathbf{P}(k)}\sqrt{\mathbf{P}^T(k)}\right\} = N_c.\end{aligned} \quad (10)$$

It can be seen from Eq. (10) that the optimization problem does not depend on $\mathbf{V}_t(k)$ (i.e., $\mathbf{V}_t(k)$ can be set to arbitrary $J\times J$ unitary matrix). In this paper, we set $\mathbf{V}_{t,opt}(k)=\mathbf{I}_J$ for the sake of brevity. In general, tr[$\mathbf{A}^{-1}$] is minimized when $\mathbf{A}$ is a diagonal matrix [13]. Therefore, the objective function expressed as Eq. (10) is minimized when $\mathbf{U}_{t,opt}(k)=\mathbf{V}_h(k)$. From the above, $\mathbf{W}_{t,opt}(k)$ is expressed as

$$\mathbf{W}_{t,opt}(k) = \mathbf{V}_h(k)\sqrt{\mathbf{P}_{opt}(k)}. \quad (11)$$

The optimization problem is rewritten by substituting Eq. (11) into Eq. (10) as

$$\begin{aligned}&P_{j,opt}(k) = \underset{\{P_j(k);j=0\sim J-1,k=0\sim N_c-1\}}{\arg\min.} \sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1}\frac{1}{\gamma P_j(k)\Lambda_j(k)+1} \\ &\text{s.t.} \sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1}P_j(k) = N_c,\end{aligned} \quad (12)$$

where $P_j(k)$ and $\Lambda_j(k)$ are respectively the jth diagonal elements of $\mathbf{P}(k)$ and $\mathbf{\Lambda}(k)$. Following [14], the optimal solution is given as (for the sake of brevity, the derivation is omitted)

$$P_{j,opt}(k) = \max\left\{\frac{1}{\sqrt{\mu}}\frac{1}{\sqrt{\gamma\Lambda_j(k)}}-\frac{1}{\gamma\Lambda_j(k)}, 0\right\}, \quad (13)$$

where μ is chosen to satisfy the constraint condition.

D. Discussion of joint Tx/Rx MMSE filtering

In this section, we discuss the behavior of joint Tx/Rx MMSE filtering. The equivalent channel matrix $\hat{\mathbf{H}}(k)$ after the Rx MMSE filtering is expressed as

$$\begin{aligned}\hat{\mathbf{H}}(k) &= \mathbf{W}_{r,opt}(k)\mathbf{H}(k)\mathbf{W}_{t,opt}(k) \\ &= \mathrm{diag}\left[\frac{P_{0,opt}(k)\Lambda_0(k)}{P_{0,opt}(k)\Lambda_0(k)+\gamma^{-1}},\ldots,\frac{P_{J-1,opt}(k)\Lambda_{J-1}(k)}{P_{J-1,opt}(k)\Lambda_{J-1}(k)+\gamma^{-1}}\right] \\ &\equiv \mathrm{diag}\left[\hat{H}_0(k),\ldots,\hat{H}_{J-1}(k)\right]\end{aligned} \quad (14)$$

It can be seen from Eq. (14) that the MIMO channel matrix $\mathbf{H}(k)$ is diagonalized (i.e., the IAI is avoided) by joint Tx/Rx MMSE filtering. In addition, the ISI can be significantly suppressed by applying the MMSE based power allocation to each eigenmode following Eq. (13).

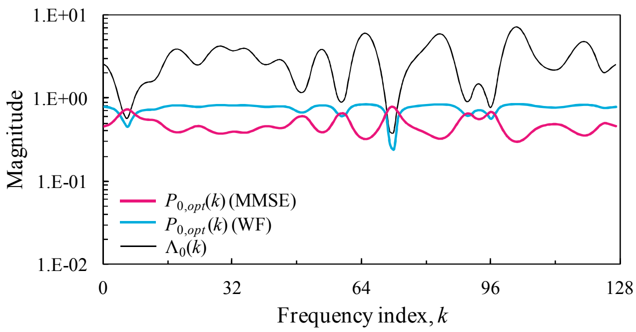
Fig. 2 shows one shot observation of the proposed MMSE power allocation. $N_t=N_r=J=2$, $N_c=128$, E_s/N_0=6dB and a 16-path frequency-selective block Rayleigh fading having uniform power delay profile is assumed. It can be seen from Eq. (13) that the proposed MMSE power allocation is similar to the power allocation based on the well-known water-filling (WF) theory expressed as

$$P_{j,opt}(k) = \max\left\{\frac{1}{\sqrt{\mu}} - \frac{1}{\gamma\Lambda_j(k)}, 0\right\}. \quad (15)$$

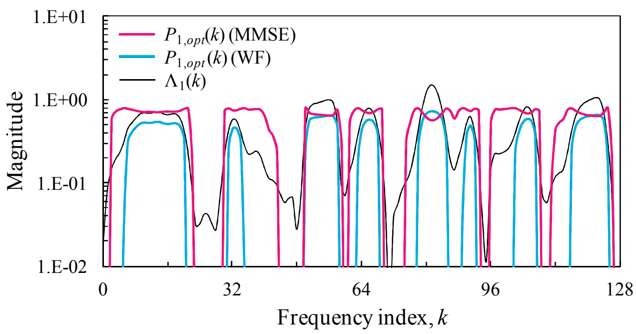
However, unlike the conventional WF power allocation, in the proposed MMSE power allocation, each eigenmode and frequency has different threshold (first term of right side of Eq. (13)) because it depends on $\Lambda_j(k)$. Therefore, the MMSE power allocation avoids the ISI increase by allocating power to the eigenmodes and frequencies which have comparatively low $\Lambda_j(k)$.

In Fig. 2, the conventional WF power allocation is also plotted for comparison. The WF power allocation is done by allocating much power to the eigenmodes and frequencies which have high $\Lambda_j(k)$, and no power to those which have low $\Lambda_j(k)$. Therefore, the ISI is enhanced and the transmission performance degrades. However, the proposed MMSE power allocation is quite different. At the first eigenmode which has low $\Lambda_j(k)$, the impact of noise is much larger than the ISI because the received signal-to-noise ratio (SNR) is low. Therefore, the proposed method allocates no power to the frequencies which has low $\Lambda_j(k)$ but allocates much power to the other frequencies to improve the received SNR. On the other hand, at the 0th eigenmode which has comparatively high $\Lambda_j(k)$, the impact of the ISI is dominant. Therefore, the power is allocated like the inverse function of $\Lambda_j(k)$ to suppress the ISI.

Fig. 3 shows one shot observation of the equivalent channel $\hat{\mathbf{H}}(k)$ when the power allocation shown in Fig. 2 has been done. As noted above, the MMSE power allocation is done to suppress the ISI. Therefore, the ISI is suppressed especially at the 0th eigenmode. On the other hand, the WF power allocation does not consider the impact of ISI. Therefore, the ISI is not much suppressed in spite of using the Rx MMSE filtering.

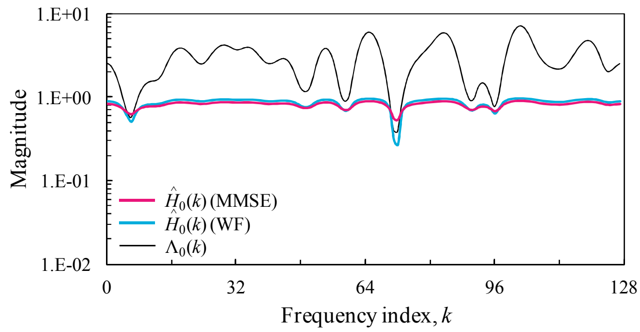


(a) 0th eigenmode

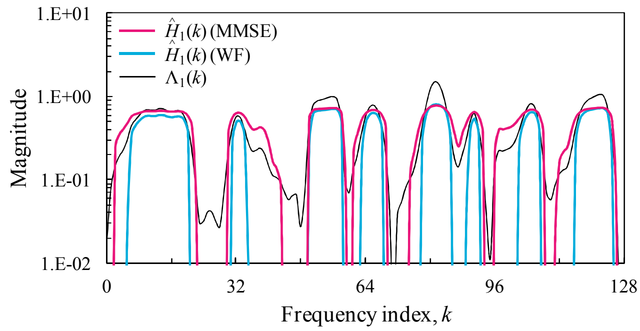


(b) 1st eigenmode

Figure 2. One shot observation of the power allocation.



(a) 0th eigenmode



(b) 1st eigenmode

Figure 3. One shot observation of the equivalent channel.

III. JOINT RANK ADAPTATION/ADAPTIVE MODULATION

In this paper, to relieve the degradation of transmission quality due to the received SINR gap among eigenmodes, we introduce rank adaptation [8] and adaptive modulation [9] to the SC-MIMO packet transmission. The number J of data streams (rank) and modulation level for each eigenmode are jointly determined based on the minimum BER criterion every retransmission. The transmission performance gap among eigenmodes can be narrowed by allocating no or a few bits (i.e., reducing the rank J or applying low level modulation) to the eigenmodes which have low received SINR and allocating many bits (i.e., applying high level modulation) to the eigenmodes which have high received SINR.

The received SINR Γ_j of the jth eigenmode in the proposed SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation is given as

$$\Gamma_j = \frac{\left|\tilde{H}_j\right|^2}{\left\{\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\bar{H}_j(k)\right|^2 - \left|\tilde{H}_j\right|^2\right\} + \frac{\gamma^{-1}}{N_c}\sum_{k=0}^{N_c-1}\sum_{m=0}^{N_r-1}\left|W_{j,m}^{(r)}(k)\right|^2}, \quad (16)$$

where

$$\tilde{H}_j = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}_j(k), \quad (17)$$

and $W_{j,m}^{(r)}(k)$ is the (j,m)th element of $\mathbf{W}_r(k)$.

When Gray code mapping is used and if ISI + noise can be approximated as a complex-valued random variable, the conditional BER $p_b^{(j)}$ of the jth eigenmode for the given set of modulation level and received SINR Γ_j is given as [3]

$$p_b^{(j)} = a_j \mathrm{erfc}\left(\sqrt{\frac{\Gamma_j}{b_j}}\right), \tag{18}$$

where erfc(.) denotes the complementary error function and a_j and b_j are shown in Table I for various modulation levels. When M_j bits are allocated to the symbol of the jth eigenmode, the conditional BER averaged over eigenmodes is given as

$$\overline{P}_b = \frac{\sum_{j=0}^{J-1} M_j p_b^{(j)}}{\sum_{j=0}^{J-1} M_j} = \frac{1}{\eta}\sum_{j=0}^{J-1} M_j a_j \mathrm{erfc}\left(\sqrt{\frac{\Gamma_j}{b_j}}\right), \tag{19}$$

where $\eta = \sum_{j=0}^{J-1} M_j$ is the spectral efficiency in bps/Hz.

The rank J and modulation levels for J streams are jointly determined as follows. For the given spectral efficiency η, the average conditional BER is computed by using Eqs. (18) and (19) for all possible combinations of J and modulation levels, and then, the optimal combination which minimizes the average conditional BER is found.

TABLE I. a_j AND b_j .

Data modulation	a_j	b_j
BPSK	1/2	1
QPSK	1/2	2
8PSK	1/3	$1/\sin^2(\pi/8)$
16QAM	3/8	10
64QAM	7/24	42
256QAM	15/64	170

IV. COMPUTER SIMULATION

A. Computer simulation condition

Computer simulation condition is summarized in Table II. 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 6 iterations are used. HARQ using type II S-P4 [12] is considered. Information sequence length is K=4096 bits. Ideal ACK/NACK transmissions are assumed. The channel is assumed to be a 16-path frequency-selective block Rayleigh fading having uniform power delay profile. Uncorrelated fading and ideal channel estimation at both the transmitter and receiver are also assumed.

TABLE II. COMPUTER SIMULATION CONDITION.

Channel coding	Turbo encoding with (13,15) RSC encoder Log-MAP decoding with 6 iterations	
HARQ	Type II S-P4	
Transmitter & Receiver	No. of information bits	K=4096
	No. of DFT points	N_c=128
	Guard interval length	N_g=16
	No. of Tx/Rx antennas	(N_t,N_r)=(4,4)
	Channel estimation	Ideal
	Antenna correlation	Uncorrelated
Channel	Fading	Frequency-selective block Rayleigh
	Power delay profile	16-path uniform

B. HARQ throughput performance

Fig. 4 shows the HARQ throughput performance (left figure) and the distribution of selected rank and modulation levels (right figure) of the proposed SC-MIMO packet transmission using joint Tx/Rx MMSE filtering (MMSE) and joint rank adaptation/adaptive modulation for N_t=N_r=4. For comparison, the HARQ throughput performance of SC-MIMO packet transmission using Rx MMSE filtering only, the conventional joint Tx/Rx MMSE filtering without joint rank adaptation/adaptive modulation, and the conventional eigenmode SC-MIMO packet transmission using the WF power allocation and joint rank adaptation/adaptive modulation are also plotted. QPSK (16QAM) modulation is applied to N_t data streams for η=8 (16) and MMSE packet combining is done when Rx MMSE filtering or conventional joint Tx/Rx MMSE filtering are used. It can be seen from Fig. 4 that the proposed method achieves better HARQ throughput performance than SC-MIMO packet transmission using Rx MMSE filtering only. This is because the proposed joint Tx/Rx filtering avoids the IAI by transforming the MIMO channel to eigenmodes and suppresses the ISI by performing the MMSE power allocation shown in Fig. 2. It is also seen from Fig. 4 that the proposed method achieves better HARQ throughput performance than SC-MIMO packet transmission using conventional joint Tx/Rx MMSE filtering since joint rank adaptation/adaptive modulation narrows the SINR gap between eigenmodes. The detailed discussion is given below.

From Fig. 4, we observe that the probability that small J is selected and many bits are allocated to the eigenmodes which have high eigenvalues is high when average transmit E_s/N_0 is low. In the low average transmit E_s/N_0 region, the overall BER is minimized by allocating the transmitted bits and applying high level modulation to the eigenmodes which have high eigenvalues (i.e., decreasing the rank J) because the received SINR of the eigenmode which have high eigenvalues are especially improved due to the high diversity gain. On the other hand, in the high average transmit E_s/N_0 region, the overall BER is minimized by allocating the transmitted bits and applying low level modulation to many eigenmodes (i.e., increasing the rank J) because almost all of the eigenmodes have high received SINR.

It can also be seen from Fig. 4 that the proposed MMSE power allocation achieves better HARQ throughput performance than the WF power allocation. This is because the proposed MMSE can allocate the transmit power across frequencies to suppress the ISI. The WF does not suppress the ISI because it performs power allocation across frequencies without considering the impact of ISI.

V. CONCLUSION

In this paper, we proposed a SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation. Joint Tx/Rx MMSE filtering transforms the MIMO channel to eigenmodes and performs the MMSE-based frequency-domain power allocation on each eigenmode in order to avoid the IAI and sufficiently suppress the ISI, respectively. Joint rank adaptation/adaptive modulation narrows the SINR gap between eigenmodes. Computer simulation confirmed that the proposed SC-MIMO packet transmission using joint Tx/Rx MMSE filtering and joint rank adaptation/adaptive modulation significantly improves the throughput performance and achieves better throughput performance than the conventional eigenmode SC-MIMO packet transmission.

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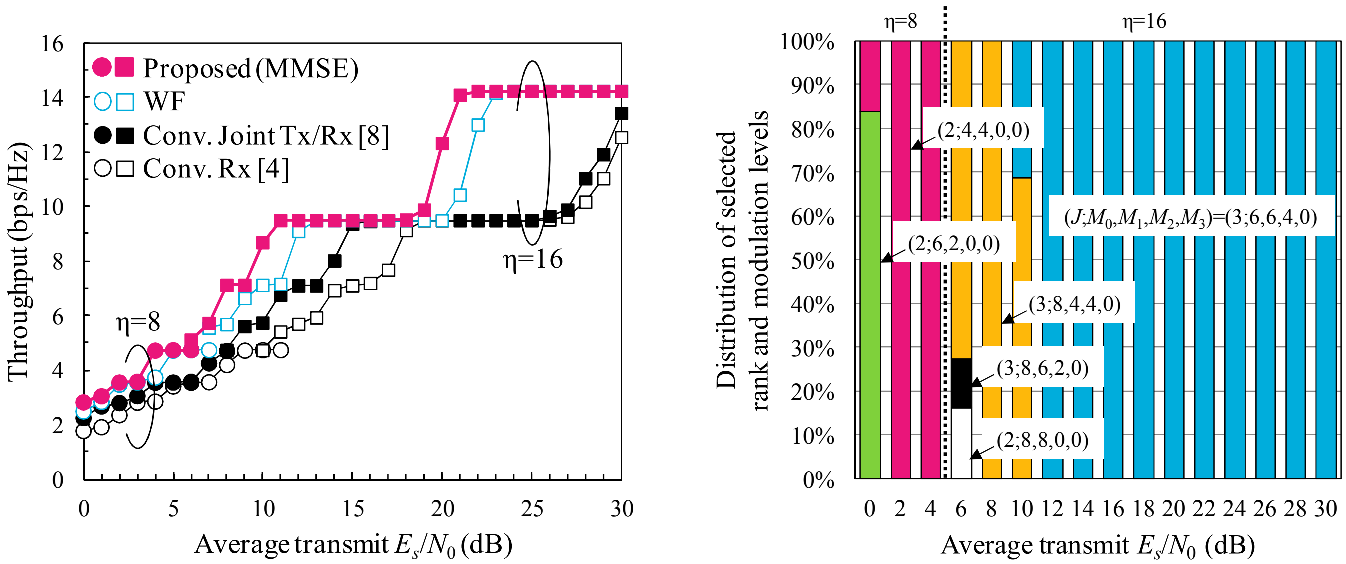

Figure 4. HARQ throughput performance (left figure) and distribution of selected rank and modulation levels on the proposed method (right figure).