Adaptive Single-carrier MIMO Transmission Using Joint Tx/Rx MMSE Filtering

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***Abstract*—In this paper, we propose a new broadband single-carrier (SC) multiple-input multiple-output (MIMO) transmission method to reduce the residual inter-antenna interference (IAI) and inter-symbol interference (ISI) after the minimum mean square error based receive filtering (Rx MMSE filtering). The proposed method jointly performs transmit and receive MMSE filtering (joint Tx/Rx MMSE filtering) to transform the MIMO channel to the orthogonal eigenmodes and allocate the transmit power based on MMSE criterion. Also, as the received signal-to-interference plus noise power ratio (SINR) of each eigenmode is significantly different, rank adaptation and adaptive modulation are jointly introduced to narrow the received SINR gap between each eigenmode. The superiority of the proposed method (adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering) is confirmed by computer simulation.**

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

I. INTRODUCTION

Multiple-input multiple-output (MIMO) spatial multiplexing [1] is a powerful 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 it is robust 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 been attracting much attention as an alternative technique 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 linear receive filtering (Rx MMSE 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 after the Rx MMSE filtering. To further improve the transmission performance, non-linear signal detection techniques such as the iterative interference cancellation [4] and the maximum likelihood detection [5] have been studied. However, their computational complexity is extremely high.

Another interesting technique to improve the transmission performance is a 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 in a frequency-selective fading channel. Since joint Tx/Rx MMSE-FDE allocates the transmit power so as to sufficiently suppress the residual ISI, it outperforms the Rx MMSE-FDE [7] which uses the CSI at the receiver only.

In this paper, we propose adaptive SC-MIMO transmission using MMSE based joint transmit and receive filtering (joint Tx/Rx MMSE filtering) to suppress the residual IAI and ISI after the Rx MMSE filtering. The proposed joint Tx/Rx MMSE 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) to avoid the IAI. It also jointly performs MMSE based frequency-domain transmit power allocation and receive FDE on each eigenmode to suppress the ISI. However, the eigenmodes having low received signal-to-interference plus noise power ratio (SINR) limit the improvement of transmission performance. Therefore, rank adaptation [8] and adaptive modulation [9,10] are jointly optimized to narrow the received SINR gap between each eigenmode. The optimal combination of the number of data streams (i.e., rank) and modulation levels is found based on the minimum bit error rate (BER) criterion.

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

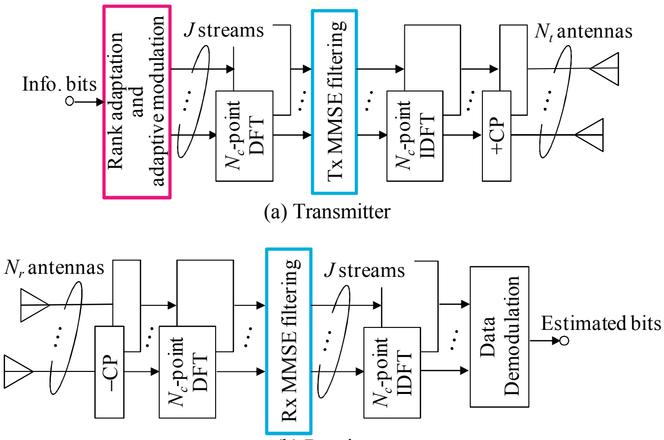
II. ADAPTIVE SC-MIMO TRANSMISSION USING JOINT TX/RX MMSE FILTERING

A. System model

System model of adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering is illustrated in Fig. 1. Transmitter and receiver have N_t and N_r antennas, respectively. The information bit sequence is data modulated based on the rank adaptation and 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.



(a) Transmitter

(b) Receiver

Figure 1. System model.

B. Transmit and received signals

At the transmitter, the $N_t\times 1$ transmit symbol vector $\mathbf{S}(k)$ at the kth frequency is obtained by applying the Tx MMSE filtering to the $J\times 1$ 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

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

where $(.)^T$ is the transpose operation and $\mathbf{W}_t(k)$ is the $N_t\times J$ 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 $N_r\times 1$ frequency-domain received signal vector $\mathbf{R}(k)$ at the kth frequency after N_c-point DFT is expressed as

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

where E_s and T_s are the average transmit symbol energy and symbol duration, respectively. $\mathbf{H}(k)$ is the $N_r\times N_t$ 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 $J\times 1$ frequency-domain soft-output vector $\hat{\mathbf{D}}(k)$ is obtained by performing the Rx MMSE filtering on $\mathbf{R}(k)$ as

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

where $\mathbf{W}_r(k)$ is the $J\times N_r$ 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.

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

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

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

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

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

$$\begin{aligned}&\min_{\{\mathbf{W}_t(k),\mathbf{W}_r(k);k=0\sim N_c-1\}}\ \varepsilon\\&\text{s.t.}\ \sum_{k=0}^{N_c-1}\mathrm{tr}\{\mathbf{W}_t(k)\mathbf{W}_t^H(k)\}=N_c\end{aligned}.\qquad(6)$$

The transmit and receive filters which satisfy Eq. (6) are the optimal solution of MMSE filters. However, it is quite difficult to derive a set of MMSE matrices $\{\mathbf{W}_t(k), \mathbf{W}_r(k)\}$ 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 Rx filter matrix $\mathbf{W}_r(k)$ considering (the Tx filter + channel) as the equivalent channel $\overline{\mathbf{H}}(k)=\mathbf{H}(k)\mathbf{W}_t(k)$. Then, we derive the Tx filter matrix $\mathbf{W}_t(k)$ by solving the optimization problem of Eq. (6) for the given $\mathbf{W}_r(k)$.

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 optimal $\mathbf{W}_r(k)$ is given as

$$\mathbf{W}_r(k)=\overline{\mathbf{H}}^H(k)\{\overline{\mathbf{H}}(k)\overline{\mathbf{H}}^H(k)+\gamma^{-1}\cdot\mathbf{I}_{N_r}\}^{-1}.\qquad(7)$$

The objective function is expressed as the function only of the Tx filter matrix $\mathbf{W}_t(k)$ by substituting the optimal $\mathbf{W}_r(k)$ into the objective function. The optimization problem is rewritten by substituting Eq. (7) into Eq. (6) and using the matrix inversion lemma [11] as

$$\min_{\{\mathbf{W}_t(k);k=0\sim N_c-1\}} \varepsilon = \sum_{k=0}^{N_c-1} \mathrm{tr}\left\{\gamma \cdot \mathbf{H}(k)\mathbf{W}_t(k)\mathbf{W}_t^H(k)\mathbf{H}^H(k) + \mathbf{I}_{N_r}\right\}^{-1}$$
$$\text{s.t.} \sum_{k=0}^{N_c-1} \mathrm{tr}\left\{\mathbf{W}_t(k)\mathbf{W}_t^H(k)\right\} = N_c \quad . (8)$$

$\mathbf{H}(k)$ and $\mathbf{W}_t(k)$ can be transformed by singular value decomposition [11] as

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

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[**AB**]=tr[**BA**], where **A** and **B** are respectively a $A \times B$ and $B \times A$ matrices, Eq. (8) can be rewritten by substituting Eq. (9) as

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

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 $\mathbf{V}_t(k) = \mathbf{I}_J$ for the sake of brevity. In general, tr$[\mathbf{A}^{-1}]$ is minimized when **A** is a diagonal matrix [11]. Therefore, the objective function expressed as Eq. (10) is minimized when $\mathbf{U}_t(k)$=$\mathbf{V}_h(k)$. From the above, $\mathbf{W}_t(k)$ is expressed as

$$\mathbf{W}_t(k) = \mathbf{V}_h(k)\sqrt{\mathbf{P}(k)} . \quad (11)$$

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

$$\min_{\{P_j(k);j=0\sim J-1,k=0\sim N_c-1\}} \varepsilon = \sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1} \frac{1}{\gamma P_j(k)\Lambda_j(k)+1}$$
$$\text{s.t.} \sum_{k=0}^{N_c-1}\sum_{j=0}^{J-1} P_j(k) = N_c \quad , (12)$$

where $P_j(k)$ and $\Lambda_j(k)$ are respectively the jth diagonal elements of $\mathbf{P}(k)$ and $\mathbf{\Lambda}(k)$. Following [12], the optimal solution is given as (for the sake of brevity, the derivation is omitted)

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

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

$$\begin{aligned} \hat{\mathbf{H}}(k) &= \mathbf{W}_r(k)\mathbf{H}(k)\mathbf{W}_t(k) \\ &= \mathrm{diag}\left[\frac{P_0(k)\Lambda_0(k)}{P_0(k)\Lambda_0(k)+\gamma^{-1}}, \ldots, \frac{P_{J-1}(k)\Lambda_{J-1}(k)}{P_{J-1}(k)\Lambda_{J-1}(k)+\gamma^{-1}}\right] . \\ &\equiv \mathrm{diag}\left[\hat{H}_0(k), \ldots, \hat{H}_{J-1}(k)\right] \end{aligned} \quad (14)$$

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 theory both across eigenmodes and frequencies (2D-WF) expressed as

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

However, unlike the conventional 2D-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 2D-WF power allocation is also plotted for comparison. The 2D-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 1st 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.

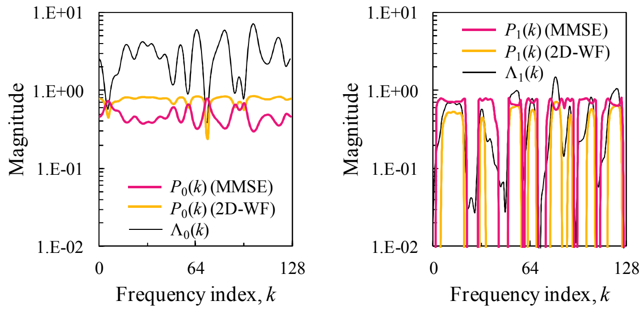


Figure 2. One shot observation of the power allocation.

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 2D-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.

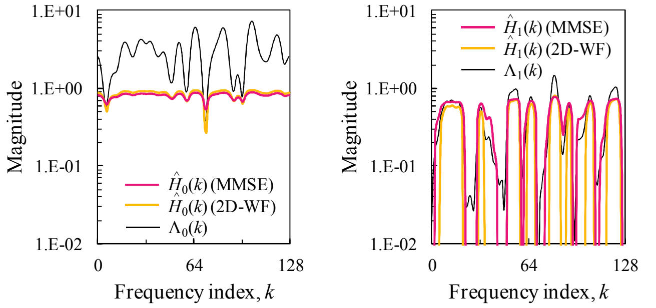


Figure 3. One shot observation of the equivalent channel.

III. Rank Adaptation and 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 [9,10] to the SC-MIMO transmission. The number J of data streams (rank) and modulation level for each eigenmode are jointly determined based on the minimum BER criterion. 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 adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering is given as

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

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]

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

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

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

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 Eq. (19) for all possible combinations of J and modulation levels, and then, the optimal combination which minimizes the average conditional BER is found. For example, when $N_t=N_r=2$ and η=4(bps/Hz), possible combinations of J and modulation levels are $(J;M_0,M_1)$=(1;4,0), (2;3,1), and (2;2,2), because $\Lambda_0(k)$ is always larger than $\Lambda_1(k)$ for $k=0\sim N_c-1$. The average conditional BER is computed for the above 3 combinations to find the optimal one.

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. 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.

Transmitter & Receiver	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. Average BER performance

Fig. 4 shows the average BER performance (left figure) and the distribution of selected rank and modulation levels (right figure) of the proposed adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering (MMSE) for $N_t=N_r=4$ and η=16(bps/Hz). For comparison, the average BER performance of SC-MIMO transmission using Rx MMSE filtering only and the conventional eigenmode SC-MIMO transmissions using the minimum BER based power allocation (Min. BER) [9], the water-filling power allocation across eigenmode only (1D-WF) [10], and 2D-WF power allocation, are also plotted. 16QAM modulation is applied to N_t data streams when Rx MMSE filtering is used. It can be seen from Fig. 4 that the proposed adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering achieves better average BER performance than SC-MIMO 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 suppress the ISI by performing the MMSE power allocation shown in Fig. 2. In addition, a joint use of rank adaptation and adaptive modulation narrows the SINR gap between eigenmodes. The detailed discussion is given below.

It can be seen from Fig. 4 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 improved 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 improved by allocating the transmitted bits and applying low level modulation to all of the eigenmodes (i.e., increasing the rank J) because all of the eigenmodes have high received SINR.

It can also be seen from Fig. 4 that the proposed MMSE power allocation achieves the best average BER performance among {MMSE, Min. BER, 1D-WF, 2D-WF}. This is because the proposed MMSE can allocate the transmit power across frequencies to suppress the ISI. Min. BER and 1D-WF cannot suppress the ISI sufficiently because they perform equal power allocation across frequencies. 2D-WF does not suppress the ISI because it performs power allocation across frequencies without considering the impact of ISI.

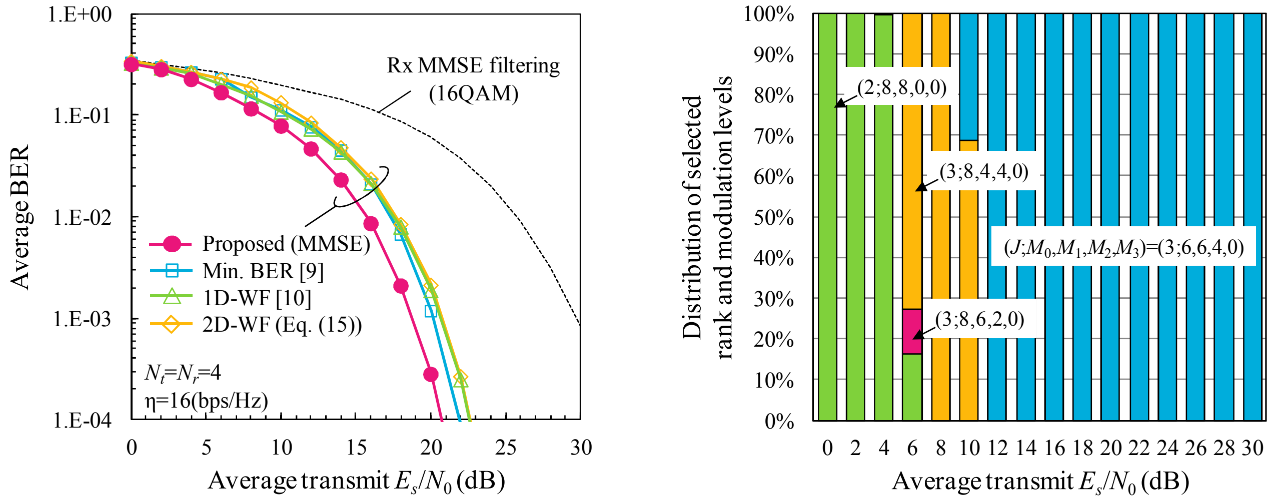


Figure 4. Average BER performance (left figure) and distribution of selected rank and modulation levels on the proposed method (right figure).

V. Conclusion

In this paper, we proposed an adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering. In adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering, the IAI is avoided by transforming the MIMO channel to eigenmodes and the ISI is suppressed by performing the MMSE power allocation. In addition, a joint use of rank adaptation and adaptive modulation narrows the SINR gap between eigenmodes. Computer simulation confirmed that the proposed adaptive SC-MIMO transmission using joint Tx/Rx MMSE filtering significantly improves the average BER performance and achieves better BER performance than the conventional eigenmode SC-MIMO transmissions.

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