Throughput Performance of SC-MIMO HARQ Using Training Sequence Aided Near ML Block Detection

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Abstract—**Very high-speed and high-quality packet transmissions in a limited bandwidth can be achieved by the joint use of multi-input multi-output (MIMO) multiplexing and hybrid automatic repeat request (HARQ). However, if single-carrier (SC) transmission is used, the transmission performance significantly degrades due not only to an inter-antenna interference (IAI) but also to an inter-symbol interference (ISI) resulting from a severe frequency-selective fading. Recently, we proposed a training sequence (TS) aided near maximum likelihood (ML) block detection using QR decomposition and M-algorithm (QRM-MLBD) to improve the transmission performance of SC-MIMO while significantly reducing computational complexity. In this paper, we evaluate, by computer simulation, the throughput performance of SC-MIMO HARQ using TS aided QRM-MLBD and show that the TS aided QRM-MLBD provides a better throughput performance than the conventional QRM-MLBD. We also compare the achievable throughput performance of TS aided QRM-MLBD to that of the frequency-domain detection based on the minimum mean square error criterion (MMSED).**

Keywords-component; Single-carrier, MIMO, maximum likehilood detection, QR decomposition, M-algorithm, training sequence

I. INTRODUCTION

In next generation mobile communication systems, high data rate wireless communication systems are demanded. Multi-input multi-output (MIMO) spatial multiplexing technique has been gaining much attention in a band-limited wireless channel [1]. However, since the mobile wireless channel is composed of many propagation paths with different time delays, the channel becomes severely frequency-selective as the transmission data rate increases [2]. MIMO spatial multiplexing with orthogonal frequency-division multiplexing (OFDM) [3] is attractive. However, OFDM suffers from high peak-to-average power ratio (PAPR). Recently, the single-carrier (SC) MIMO spatial multiplexing using frequency-domain block detection technique has been gaining an increasing popularity because of its lower PAPR property [4-5]. The use of cyclic prefix (CP) and frequency-domain signal detection such as the frequency-domain linear detection based on the minimum mean square error criterion (MMSED) can improve the transmission performance of SC-MIMO spatial multiplexing with a low complexity. However, a big performance gap from the maximum likelihood (ML) performance still exists due to the presence of residual inter-symbol interference (ISI) and inter-antenna interference (IAI).

The ML detection [6] is the optimal detection scheme in terms of transmission performance. However, its computational complexity becomes extremely high because the number of symbol candidate sequence is exponentially increased to $X^{N_tN_c}$ for X-QAM, where N_t is the number of transmit antennas and N_c is the block size. Recently, near ML-based reduced complexity signal detection was proposed [7] for CP inserted SC (CP-SC) MIMO spatial multiplexing (we call this detection scheme as the ML block detection using QR decomposition and M-algorithm (QRM-MLBD)). In QRM-MLBD, QR decomposition is applied to a concatenation of the space and frequency-domain channel and discrete Fourier transform (DFT). It was shown [7] that QRM-MLBD can significantly improve the packet error rate (PER) performance of CP-SC MIMO spatial multiplexing when compared to the MMSE detection. However, in order to achieve the sufficiently improved performance, the use of a fairly large number M of surviving paths in the M-algorithm is required, leading to high computational complexity. This is because if smaller M is used, the probability of removing the correct path at early stages increases.

Recently, in order to further reduce the required number M of surviving paths for achieving the sufficiently improved PER performance, we proposed a training sequence (TS) aided QRM-MLBD for SC-MIMO spatial multiplexing [8]. In TS aided QRM-MLBD, TS aided SC (TS-SC) block transmission [9] is used instead of CP-SC block transmission and its TS is utilized to reduce the probability of removing the correct path at early stages. We showed [8] that TS aided QRM-MLBD achieves the PER performance similar to the conventional QRM-MLBD while significantly reducing the computational complexity.

Packet access will be the core technology of the next generation mobile communication systems. Very high-speed and high-quality packet transmissions in a limited bandwidth can be achieved by the joint use of MIMO multiplexing and hybrid automatic repeat request (HARQ) [10]. In this paper, we evaluate, by computer simulation, the throughput performance of turbo coded SC-MIMO HARQ using TS aided QRM-MLBD.

The remainder of this paper is organized as follows. Sect. II introduces the transmission system model of turbo-coded SC-

MIMO HARQ using TS aided QRM-MLBD. In Sect. III, we show the computer simulation results of the throughput performance. Sect. IV offers some concluding remarks.

II. SC-MIMO HARQ USING TS AIDED QRM-MLBD

A. *HARQ*

In general, HARQ can be classified into three types [11]: HARQ type I, type II and type III. In this paper, we consider the HARQ type II S-P4 [12] and turbo coding with rate R=1/3, as illustrated in Fig. 1. The turbo encoder outputs the systematic bit sequence and two parity bit sequences. These sequences are punctured into five sequences (including systematic bit sequence) by the puncturing matrices given by

$$\begin{bmatrix}1&1&1&1\\0&0&0&0\\0&0&0&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\1&0&0&0\\0&0&1&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&1&0&0\\0&0&0&1\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&0&1&0\\1&0&0&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&0&0&1\\0&1&0&0\end{bmatrix}. \tag{1}$$

For the first transmission, only the systematic bit sequence is transmitted. At the receiver, data decision and error detection are performed. If any error is detected in the received packet, second transmission is requested from the receiver by sending an NACK signal. When the NACK signal is received at the transmitter, the second packet (consisting of the punctured parity bit sequence) is transmitted. At the receiver, turbo decoding is carried out by using the first and second received packets. If any error is detected after turbo decoding, the NACK signal is transmitted again. One of the punctured parity bit sequences is transmitted each time the NACK signal is received at the transmitter until the 5th packet transmission. After the 5th packet transmission, the same packet is retransmitted.

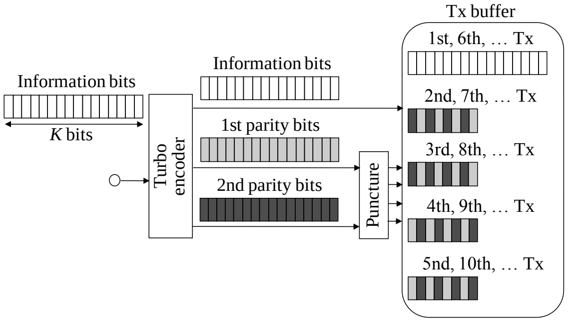


Figure 1. HARQ type II S-P4.

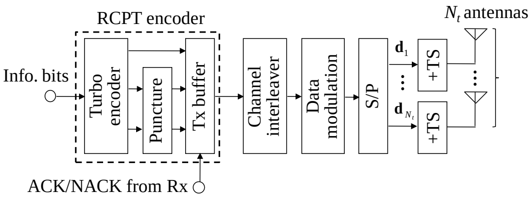
B. *Signal Transmission Model*

System model of SC-MIMO HARQ using TS aided QRM-MLBD is illustrated in Fig. 2. Throughout the paper, the symbol-spaced discrete time representation is used. At the transmitter, after turbo coding and puncturing, the coded bit sequence is stored in the transmitter buffer. The coded bit sequence is transformed into a data-modulated symbol sequence. Then, the data-modulated symbol sequence is serial-to-parallel (S/P) converted to N_t parallel symbol sequence, each to be transmitted from a different transmit antenna and each parallel symbol sequence is divided into a sequence of symbol blocks of N_c symbols each. The data symbol block of n_t-th transmit antenna can be expressed using the vector form as $\mathbf{d}_{n_t}=[d_{n_t}(0),\ldots,d_{n_t}(N_c-1)]^T$, where $(.)^T$ expresses the transposition. Before the transmission, the TS of length N_g symbols is appended at the end of each block. The block $\mathbf{s}_{n_t}=[s_{n_t}(0),\ldots,s_{n_t}(N_c+N_g-1)]^T$ to be transmitted is expressed using the vector form as

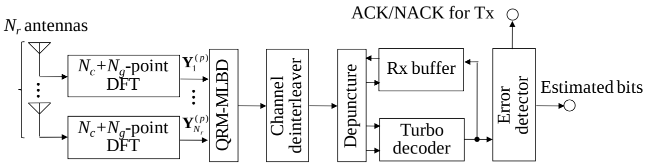
$$\mathbf{s}_{n_t}=\begin{bmatrix}\mathbf{d}_{n_t}\\ \mathbf{u}_{n_t}\end{bmatrix}, \tag{2}$$

where $\mathbf{u}_{n_t}=[u_{n_t}(0),\ldots u_{n_t}(N_g-1)]^T$ denotes the TS vector which is identical for all blocks. The TS-SC block structure is illustrated and compared to CP-SC transmission in Fig. 3. The difference from CP-SC transmission is that CP is replaced by TS. In order to let TS to play the role of CP, DFT size at the receiver must be the sum of number of useful data symbols and the TS length. In this paper, to keep the same data rate as CP-SC, the data symbol block length and the TS length need to be set to N_c and N_g, respectively. Therefore, the DFT size to be used at the receiver is N_c+N_g symbols for TS-SC while it is N_c symbols for CP-SC.

The signal block is transmitted over a frequency-selective fading channel. The received signal is transformed by N_c+N_g-point DFT into the frequency-domain signal. Then, QRM-MLBD is carried out.

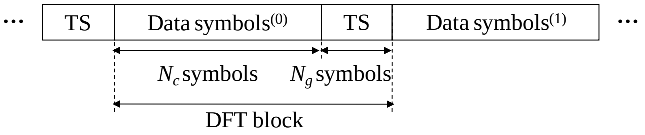


(a) Transmitter

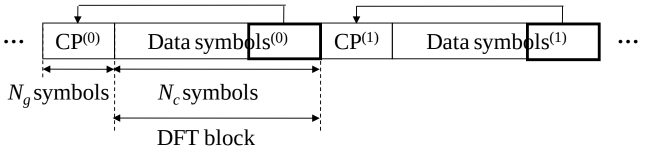


(b) Receiver

Figure 2. System model of SC- MIMO HARQ using TS aided QRM-MLBD.



(a) TS-SC



(b) CP-SC

Figure 3. Block structure.

C. *QRM-MLBD*

Consider that the *p*th retransmitted packet has been received. The frequency-domain received signal vector at the n_r-th receive antenna $\mathbf{Y}_{n_r}^{(p)}=[Y_{n_r}^{(p)}(0),\ldots,Y_{n_r}^{(p)}(N_c+N_g-1)]^T$ after N_c+N_g-point DFT is expressed as

$$\mathbf{Y}_{n_r}^{(p)}=\sqrt{\frac{2E_s}{T_s}}\sum_{n_t=1}^{N_t}\mathbf{H}_{n_r,n_t}^{(p)}\mathbf{F}\mathbf{s}_{n_t}+\mathbf{N}_{n_r}^{(p)}, \qquad (3)$$

where E_s and T_s are respectively the symbol energy and duration, **F** is the DFT matrix of size $(N_c+N_g)\times(N_c+N_g)$, $\mathbf{H}_{n_r,n_t}^{(p)}$ is the frequency-domain channel matrix between the n_tth transmit antenna and n_rth receive antenna in the *p*th retransmission, and $\mathbf{N}_{n_r}^{(p)}=[N_{n_r}^{(p)}(0),\ldots,N_{n_r}^{(p)}(N_c+N_g-1)]^T$ is the frequency-domain noise vector.

From Eq. (3), the $N_r(N_c+N_g)\times 1$ overall frequency-domain received signal $\mathbf{Y}^{(p)}=[\{\mathbf{Y}_1^{(p)}\}^T,\ldots,\{\mathbf{Y}_{N_r}^{(p)}\}^T]^T$ is given by

$$\begin{aligned}\mathbf{Y}^{(p)}&=\sqrt{\frac{2E_s}{T_s}}\begin{bmatrix}\mathbf{H}_{1,1}^{(p)}\mathbf{F} & \mathbf{H}_{1,2}^{(p)}\mathbf{F} & \cdots & \mathbf{H}_{1,N_t}^{(p)}\mathbf{F}\\ \mathbf{H}_{2,1}^{(p)}\mathbf{F} & \mathbf{H}_{2,2}^{(p)}\mathbf{F} & \cdots & \mathbf{H}_{2,N_t}^{(p)}\mathbf{F}\\ \vdots & \vdots & \ddots & \vdots\\ \mathbf{H}_{N_r,1}^{(p)}\mathbf{F} & \mathbf{H}_{N_r,2}^{(p)}\mathbf{F} & \cdots & \mathbf{H}_{N_r,N_t}^{(p)}\mathbf{F}\end{bmatrix}\begin{bmatrix}\mathbf{s}_1\\ \vdots\\ \mathbf{s}_{N_t}\end{bmatrix}+\begin{bmatrix}\mathbf{N}_1^{(p)}\\ \vdots\\ \mathbf{N}_{N_r}^{(p)}\end{bmatrix},\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}^{(p)}\mathbf{s}+\mathbf{N}^{(p)}\end{aligned} \qquad (4)$$

where $\mathbf{H}^{(p)}$ is an equivalent channel matrix of size $N_r(N_c+N_g)\times N_t(N_c+N_g)$, which is a concatenation of the space and frequency-domain channel and DFT, **s** is the $N_t(N_c+N_g)\times 1$ overall transmit symbol vector, and $\mathbf{N}^{(p)}$ is the $N_r(N_c+N_g)\times 1$ overall noise vector.

Until the 5th packet is received ($p\le 5$), the following TS aided QRM-MLBD [8] is carried out. QRM-MLBD consists of three steps; ordering, QR decomposition, and M-algorithm. The ordered overall transmit symbol vector $\mathbf{s}^{order}$ can be expressed as

$$\begin{aligned}\mathbf{s}^{order}&=[s_1(0),\ldots,s_{N_t}(0),\ldots,s_1(N_c+N_g-1),\ldots,s_{N_t}(N_c+N_g-1)]^T\\ &=[\mathbf{d}^T(0),\mathbf{d}^T(1),\ldots,\mathbf{d}^T(N_c-1),\mathbf{u}^T(0),\ldots,\mathbf{u}^T(N_g-1)]^T\end{aligned}, \qquad (5)$$

where $\mathbf{d}^T(t)$ and $\mathbf{u}^T(t)$ denote the data symbol vector and TS vector at *t*th symbol of size $N_t\times 1$, respectively. After ordering, the QR decomposition is applied to the ordered equivalent channel matrix $\mathbf{H}^{(p)}$ to obtain $\mathbf{H}^{(p)}=\mathbf{Q}^{(p)}\mathbf{R}^{(p)}$, where $\mathbf{Q}^{(p)}$ is an $N_r(N_c+N_g)\times N_t(N_c+N_g)$ unitary matrix and $\mathbf{R}^{(p)}$ is an $N_t(N_c+N_g)\times N_t(N_c+N_g)$ upper triangular matrix. The transformed frequency-domain received signal $\hat{\mathbf{Y}}^{(p)}=[\hat{Y}^{(p)}(0),\ldots,\hat{Y}^{(p)}(N_t(N_c+N_g)-1)]^T$ is obtained as

$$\begin{aligned}\hat{\mathbf{Y}}^{(p)}&=\{\mathbf{Q}^{(p)}\}^H\mathbf{Y}^{(p)}\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{R}^{(p)}\mathbf{s}^{order}+\{\mathbf{Q}^{(p)}\}^H\mathbf{N}^{(p)}\end{aligned}. \qquad (6)$$

From Eq. (6), the ML solution can be obtained by searching for the best path having the minimum Euclidean distance in the tree diagram composed of $N_t(N_c+N_g)$ stages. M-algorithm [13] can be applied to reduce the computational complexity. In each stage, the best *M* paths are selected as surviving paths by comparing the path metrics based on the squared Euclidean distance for all surviving paths and are passed to the next stage. It can be understood from Eqs. (5) and (6) that in TS-SC MIMO multiplexing with ordering, the TSs of all transmit antennas are moved to the bottom of overall transmit signal vector. This property can be of considerable help in the MLD using M-algorithm. The symbols to be detected at early stages in M-algorithm belong to the known TSs, and therefore, the probability of removing the correct path can be significantly reduced.

In the case of first transmission (p=1), the data demodulation is carried out by tracing back the path having the smallest path metric at the last stage. On the other hand, when p<1, the log likelihood ratio (LLR) is used as the soft-input in the decoder. When QRM-MLBD is used, however, the LLR values cannot be directly computed since surviving paths at the last stage do not necessarily contain both 1 and 0 for every coded bit. In this paper, we apply the LLR calculation method proposed in [14]. The approximate LLR values are computed at every stage by using path metric and are updated successively as tree search progresses. If the LLR value cannot be computed at the last stage, the recently updated approximate LLR value at the upper stage is used.

After the 5th packet transmission (p>5), the same packet is retransmitted. When the same packet is retransmitted, joint QRM-MLBD and packet combining (called recursive QRM-MLBD) [15] is carried out. Recursive QRM-MLBD is formed as

$$\begin{bmatrix}\tilde{\mathbf{R}}^{(p-5)}\\ \mathbf{H}^{(p)}\end{bmatrix}=\tilde{\mathbf{Q}}^{(p)}\tilde{\mathbf{R}}^{(p)} \qquad (7)$$

and

$$\hat{\mathbf{Y}}^{(p)}=\{\tilde{\mathbf{Q}}^{(p)}\}^H\begin{bmatrix}\hat{\mathbf{Y}}^{(p-5)}\\ \mathbf{Y}^{(p)}\end{bmatrix}, \qquad (8)$$

where $\tilde{\mathbf{Q}}^{(p)}$ is an $(N_t(N_c+N_g)+N_r(N_c+N_g))\times N_t(N_c+N_g)$ unitary matrix and $\tilde{\mathbf{R}}^{(p)}$ is an $N_t(N_c+N_g)\times N_t(N_c+N_g)$ upper triangular matrix. Eq (7) implies that previously obtained upper triangular matrix $\tilde{\mathbf{R}}^{(p-5)}$ and the equivalent channel matrix $\mathbf{H}^{(p)}$ associated with the present received packet are required to carry out the QR decomposition. Eq (8) implies that the transformed signal vector to be used in M-algorithm $\hat{\mathbf{Y}}^{(p)}$ can be obtained recursively using the previously obtained transformed vector $\hat{\mathbf{Y}}^{(p-5)}$ and the present overall frequency-domain received signal vector $\mathbf{Y}^{(p)}$. Recursive QRM-MLBD does not need to store all of the received retransmitted packets and equivalent channel matrixes.

III. COMPUTER SIMULATION

The throughput performance of SC-MIMO HARQ using TS aided QRM-MLBD is evaluated by computer simulation. The simulation condition is summarized in Table I. We assume N_t=N_r=4, N_c=64, N_g=16, and L=16-path frequency-selective block Rayleigh fading channel with uniform power delay profile. Independent channel is assumed for each retransmission. The ideal channel estimation, ideal error detection, and no transmission error in ACK/NACK are assumed. We employ a rate 1/3 turbo encoder using two (13, 15) recursive systematic convolutional (RSC) component encoders. Log-MAP decoding with 6 iterations is assumed. The packet size is set to K=3072.

TABLE I. COMPUTER SIMULATION CONDITION

Channel coding	R=1/3 (13, 15) RSC encoder Log-MAP decoding with 6 iterations	
ARQ	HARQ type II S-P4	
Transmitter	Data modulation	QPSK, 16QAM, 64QAM
	Number of transmit antennas	N_t=4
	Data symbol block length	N_c=64
	TS or CP lengths	N_g=16
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	L=16 path uniform power delay profile
Receiver	Number of receive antennas	N_r=4
	Channel estimation	Ideal

A. *Throughput Performance*

The throughput performance of SC-MIMO HARQ using TS aided QRM-MLBD is plotted in Fig. 4 as a function of average received symbol energy-to-noise power spectrum density ratio E_s/N_0 for M=1, 4, 16, 64, and 256. For comparison, the throughput performance of SC-MIMO HARQ using the conventional QRM-MLBD is also plotted. The modulation level which gives the best throughput is selected for each E_s/N_0. It can be seen form Fig. 4 that TS aided QRM-MLBD provides better throughput performance than the conventional QRM-MLBD when the same M is used. This is because TS aided QRM-MLBD significantly reduces the probability of removing the correct path at early stages in the M-algorithm even if small M is used. When M=16 is used, the E_s/N_0 reduction is about 3.6dB, 4.6dB, and 5.5dB for the throughput=5.0bps/Hz, 10bps/Hz, and 15bps/Hz, respectively.

Figure 5 compares the throughput performances of TS-SC MIMO HARQ using TS aided QRM-MLBD, MMSED, and frequency-domain iterative ISI+IAI cancellation (FDI^{2}C) [16]. With MMSED and FDI^{2}C, when the same packet is retransmitted, MMSE packet combing [17] is used as the packet combing scheme. For FDI^{2}C, the number of iteration set to five (i.e., I=5). It can be seen from Fig. 5 that the throughput performance is significantly improved in a high E_s/N_0 region by using TS aided QRM-MLBD. When M=16 is used, the E_s/N_0 reduction from MMSED (FDI^{2}C) is about 5.0(1.0) dB and 13.2 (3.8) dB for the throughput=10bps/Hz and 15bps/Hz, respectively. On the other hand, the throughput of TS aided QRM-MLBD is inferior to that of FDI^{2}C in the region of E_s/N_0>7dB and E_s/N_0<12dB.

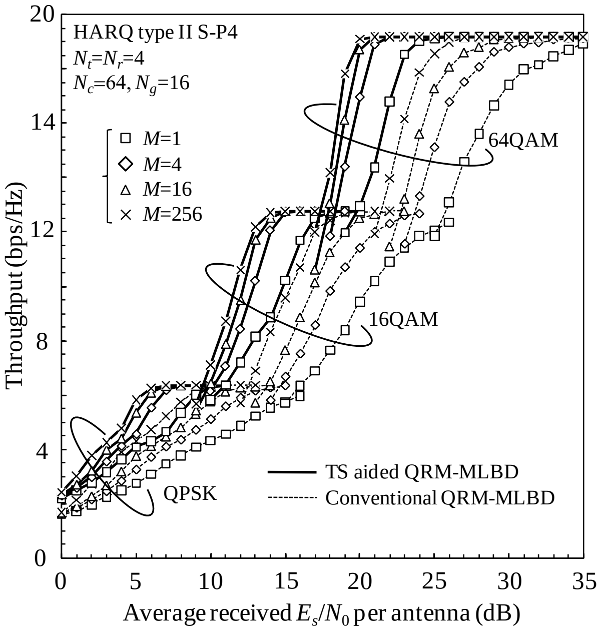


Figure 4. Throughput performance of SC-MIMO HARQ using TS aided QRM-MLBD and the conventional QRM-MLBD.

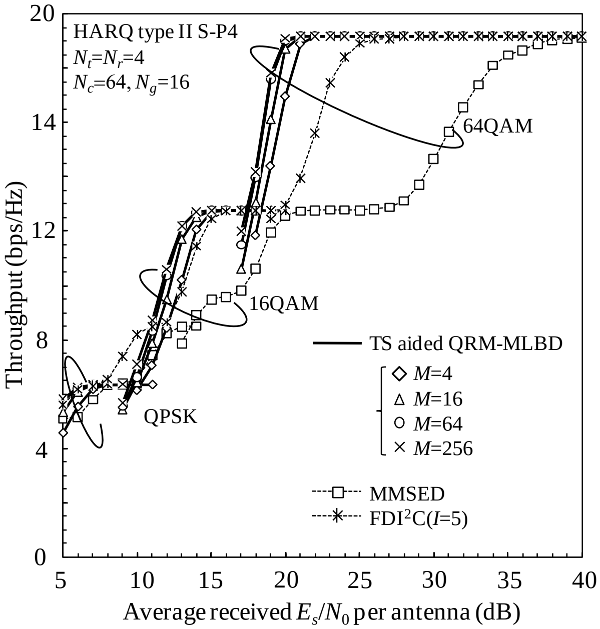


Figure 5. Throughput performance comparison between TS aided QRM-MLBD, MMSED, and FDI^{2}C.

B. *Computational Complexity*

The computational complexity of TS aided QRM-MLBD is discussed. The complexity here is defined as the number of complex multiplications. The required number of multiplications is shown in Table II. First, we discuss the complexity comparison between TS aided QRM-MLBD and the conventional QRM-MLBD. TS aided QRM-MLBD achieves better throughput performance even if small M is used. Hence, TS aided QRM-MLBD reduces significantly the computational complexity required for the squared Euclidean distance calculations. It can be seen from Fig. 4 that TS aided QRM-MLBD with M=16 achieves better throughput performance than the conventional QRM-MLBD with M=256 in all E_s/N_0 region. When $N_t=N_r=4$, the overall computational complexity of TS aided QRM-MLBD with M=16 is about 69%, 27%, and 12% of that of the conventional QRM-MLBD with M=256 for QPSK, 16QAM, and 64QAM, respectively (N_c=64 and N_g=16).

Next, we discuss the complexity comparison between TS aided QRM-MLBD, MMSED, and FDI^{2}C. TS aided QRM-MLBD provides significantly higher throughput performance than MMSED. However, when $N_t=N_r=4$, TS aided QRM-MLBD with M=4 requires about 429, 341, and 204 times higher computational complexity than MMSED for QPSK, 16QAM, and 64QAM, respectively (N_c=64 and N_g=16). Compared to FDI^{2}C, TS aided QRM-MLBD with M=16 requires about 73, 80, and 100 times higher computational complexity for QPSK, 16QAM, and 64QAM, respectively. TS aided QRM-MLD improves the throughput performance at the cost of increased complexity.

TABLE II. NUMBER OF MULTIPLICATIONS

TS aided QRM-MLBD	DFT	$N_r(N_c+N_g)^2$
	QR decomposition	$N_rN_t^2(N_c+N_g)^3+N_rN_t(N_c+N_g)^2$
	Multiplication of $\mathbf{Q}^H$	$N_rN_t(N_c+N_g)^2$
	Squared Euclidian distance calculations	$X\{2+(M/2)(N_tN_c+4)(N_tN_c-1)\}+N_t^2N_g(N_c+N_g)$
Conventional QRM-MLBD	DFT	$N_rN_c^2$
	QR decomposition	$N_rN_t^2N_c^3+N_rN_tN_c^2$
	Multiplication of $\mathbf{Q}^H$	$N_rN_tN_c^2$
	Squared Euclidian distance calculations	$X\{2+(M/2)(N_tN_c+4)(N_tN_c-1)\}$
MMSED	DFT/IDFT	$N_r(N_c+N_g)^2+N_t(N_c+N_g)^2$
	Weight generation	$(N_r^3+2N_r^2N_t)(N_c+N_g)$
	Weight multiplication	$N_tN_r(N_c+N_g)$
	LLR calculation	$N_t\{N_t(N_r+1)+N_r+2\}(N_c+N_g)+2XN_tN_c$
FDI^{2}C(I=5)	DFT/IDFT	$N_r(N_c+N_g)^2+11N_t(N_c+N_g)^2$
	Weight generation	$6(N_r^3+2N_r^2N_t)(N_c+N_g)$
	Weight multiplication	$6N_tN_r(N_c+N_g)$
	ISI+IAI cancellation	$5(N_t^2+N_tN_r)(N_c+N_g)$
	LLR calculation	$6N_t\{N_t(N_r+1)+N_r+2\}(N_c+N_g)+12XN_tN_c$

IV. CONCLUSION

In this paper, we evaluated the throughput performance of turbo coded SC-MIMO HARQ using TS aided QRM-MLBD and compared with the conventional QRM-MLBD, MMSED, and FDI^{2}C. We showed that TS aided QRM-MLBD significantly improves throughput performance while reducing the number of surviving paths in the M-algorithm. Therefore, TS aided QRM-MLBD provides better throughput performance than the conventional QRM-MLBD while significantly reducing the computational complexity. We also showed that TS aided QRM-MLD provides significantly higher throughput performance than MMSED and FDI^{2}C especially in a high E_s/N_0 region but at the cost of increased complexity.

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