**Offset Spatial Modulation With Multiple Receive Antennas**

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**ABSTRACT** Recently, a class of novel spatial modulation (SM) schemes, termed offset spatial modulation (OSM), were developed to reduce the radio frequency (RF) switching frequency, in the context of multiple-input single-output (MISO) transmission. In this paper, the structure of original OSM is extended to adapt multiple-input multiple-output (MIMO) systems by introducing multiple receive antennas (MRA). The detailed system design is presented and further developed for offset space shift keying (OSSK). Furthermore, the theoretical bounds of bit-error rate (BER) performance for the proposed OSM-MRA and OSSK-MRA systems are both quantified. It is demonstrated that the proposed OSM-MRA and OSSK-MRA systems outperform their counterparts including conventional OSM and OSSK with improved spectral efficiency.

**INDEX TERMS** Offset spatial modulation, offset space shift keying, multiple receive antennas, spectral efficiency.

**I. INTRODUCTION**

Multiple-input multiple-output (MIMO) technology has attracted considerable attentions in recent years owing to its enhanced performance on data rate and spatial diversity [1]. For example, spatial multiplexing [2] has been widely developed for supporting high data rate transmission, especially in the context of massive MIMO [3], [4]. However, conventional MIMO schemes will consume multiple radio frequency (RF) chains, which limits their applications in practical massive MIMO implementation.

For alleviating the above-mentioned problem, spatial modulation (SM) [5]–[9] was conceived by employing only one single RF chain, toward a low-cost solution for the massive MIMO implementation [10], [11]. Specifically, it conveys the information bits by not only traditional digital modulation but also the spatial degrees of freedom where only one antenna is activated during each transmission slot. On the other hand, a simplified case of SM, namely space shift keying (SSK) modulation, has been developed in [12] by only exploiting the indices of antennas to convey information.

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On the other hand, precoding-aided spatial modulation (PSM) [13]–[16] has been proposed toward an extension of conventional SM. Specifically, a class of PSM schemes, termed as receive antenna modulation (RSM) [17], [18] has been emerged by exploiting the receive antenna (instead of the transmit antenna) as an extra source of information. Moreover, by attaining the feedback of the channel state information (CSI) at the transmit side, the performance could be further enhanced with the aid of transmit antenna selection (TAS) [19]–[21] and receive antenna selection (RAS) [22] techniques. More specifically, by adopting TAS techniques, such as Euclidean distance optimized antenna selection (EDAS) [23], [24], conventional SM schemes are capable of achieving a significant bit-error rate (BER) performance enhancement at the cost of extra transmit antennas as well as the computational complexity. Furthermore, general link adaptation techniques for SM-MIMO and SSK-MIMO techniques can be detailed in [13], [14], [25], [26].

However, toward future high-rate wireless communications [27], despite the above-mentioned advantages provided by the single RF chain, SM-MIMO and SSK-MIMO have the limitation in frequent switching between the RF chain and transmit antennas. That is, the transmission rate of the
SM-MIMO and SSK-MIMO systems is limited by the RF switching speed, which may become a bottleneck for their future implementations, toward extremely high transmission rate for next generation mobile systems.

For alleviating this problem, offset spatial modulation (OSM) and offset space shift keying (OSSK) have been recently developed in [28] for the sake of simplifying the processing of RF switching, by exploiting the CSI at the transmit side. More specifically, for OSM and OSSK schemes [28], the RF chain can be offset to a fixed antenna in order to completely resolve the RF switching problem by precoding at the transmitter. In addition, OSM and OSSK also provide an extra transmit performance enhancement compared to SM at the cost of moderately increased computational complexity. However, it is noted that the aforementioned OSM-MIMO and OSSK-MIMO schemes are only designed for multiple-input single-output (MISO) systems, which limits their deployment in the high-spectral efficiency required MIMO systems. Hence, it is advantageous to extend traditional OSM and OSSK by adopting multiple receive antennas.

Based on the above facts, the contribution of this paper is twofold. Firstly, we integrate multiple receive antennas (MRA) into OSM and OSSK systems, toward a flexibility configuration in the context of MIMO channels.

Second, by exploiting the indices of the receive antennas, the spectral efficiency is also enhanced. Furthermore, we also quantify the performance bound of BER through theoretical analysis for the two developed systems.

In general, both analysis and simulation results show that the proposed systems outperform traditional OSM and OSSK in terms of BER under the same spectral efficiency.

The remainder of this paper is organized as follows: Section II elaborates on the system model of OSM-MRA and OSSK-MRA toward a general configuration for MIMO transmission. In Section III, the theoretical analyses of the BER bound of OSM-MRA and OSSK-MRA are derived. Our complexity comparisons and numerical results are outlined in Section IV, and finally, our conclusions are offered in Section V.

**Notations:** We use $| \cdot |$ to denote the absolute value of the element. $\binom{\cdot}{\cdot}$ denotes the binomial coefficient. $\| \cdot \|_F^2$ denotes the Frobenius norm of a matrix or vector. $E(\cdot)$ represents the statistical expectation of a random variable. $(\cdot)^H$ denotes the transposition while $(\cdot)^T$ denotes the conjugate transposition. $Q(\cdot)$ represents the tail probability of standard normal distribution given by $\frac{1}{\sqrt{2\pi}} \int_{\infty}^{\infty} \exp \left(-\frac{t^2}{2}\right) dt$. We use $\exp_0(\cdot)$ to denote the $i-th$ column of identity matrix with $m$ order. $\Gamma(\cdot)$ is the gamma function given by $\Gamma(x) = \int_0^{\infty} t^{x-1} e^{-t} \, dt$, $\delta(x)$ denotes the unit impulse function.

**II. SYSTEM MODEL**

In this Section, the system models of original OSM/OSSK as well as the developed OSM/OSSK-MRA are described. We show that the advantages of OSM/OSSK-MRA systems are achieved at the cost of the introduction of one extra transmit antenna.

**A. SYSTEM MODEL OF OSM/OSSK**

We consider a static OSM system [28] with $N_t$ transmit antennas and a single receive antenna. For convenience, we assume that $N_t$ and $M$ are powers of 2 in order to achieve the spectral efficiency of $m = \log_2(MN_t)$ bits/s/Hz, where $M$ is the size of conventional amplitude and phase modulation (APM) constellations. Moreover, both noise vector $n \in \mathbb{C}^{N_t \times 1}$ and the channel matrix $H \in \mathbb{C}^{N_r \times N_t}$ are independent and identically distributed (i.i.d) complex Gaussian random variables obeying $CN(0,1)$. Let $h_i, i \in \mathcal{T} = \{1,2,\ldots,N_t\}$ denote the channel gain corresponding to the channel from $i-th$ transmit antenna to the receive antenna and $s_k, k \in \mathcal{S} = \{1,2,\ldots,M\}$ be the $k-th$ symbol selected from $M$ APM symbols. More especially, when $M = 1$ and $s_k = 1$, the system could be reduced to an OSSK system [28]. Considering there are $MN_t$ different transmit modes, let $x_{k,i}$ denote the transmit signal, which can be expressed as

$$x_{k,i} = \frac{h_i}{h_{\text{tran}}} s_k e_{\text{tran}}^*, \quad (1)$$

where $\text{tran}$ is the label of the fixed transmit antenna in the concept of static OSM. Furthermore, in order to normalize the transmit power, we define $\beta_i = \frac{h_{\text{tran}}}{h_i}$ as the normalized transmit power factor. After passing through the Rayleigh fading channels and the additive white Gaussian noise (AWGN), the received signal can be denoted as

$$y = \rho \beta_i h_i x_{k,i} + n, \quad (2)$$

where $\rho$ is the transmit power and $h$ is the channel gain vector due to the single receive antenna. By substituting (1) into (2), the received signal will be transformed into

$$y = \rho \beta_i h_i s_k + n. \quad (3)$$

At the receiver side, the objective functions of the maximum likelihood (ML) detector can be formulated as

$$k, i = \arg \min_{k \in \mathcal{S}, i \in \mathcal{T}} \|y - \sqrt{\rho} \beta_i h_i x_{k,i}\|_F^2 \quad (4)$$

for OSM schemes and

$$i = \arg \min_{i \in \mathcal{T}} \|y - \sqrt{\rho} \beta_i h_i x_{k,i}\|_F^2 \quad (5)$$

for OSSK schemes.

A drawback of the system model of OSM/OSSK lies in that it only adapts the MISO transmission with one single receive antenna if all channel gains convey bits. In the extreme case, when conventional OSM and OSSK schemes adopt $N_t$ ($N_r > 1$) receive antennas with the indices $\rho \in \mathcal{R} = \{1,2,\ldots,N_t\}$ and $i \in \mathcal{T} = \{1,2,\ldots,N_t\}$ to convey the information bits, the transmit signal will be formulated as $x_{k,p,i} = s_k e_{\text{tran}}^*$ if $\text{tran} = i$, which is difficult to conduct the channel information, since this configuration will result in detection errors at the receiver side.
For alleviating this problem, and hence to extend the application of OSM/OSSK into a general MIMO configuration, we introduce an extra transmit antenna to OSM/OSSK toward MRA transmission with spectral efficiency enhancement. Meanwhile, we also exhibit that a careful design of selecting the transmit antenna can also bring additional BER performance gain to the original system.

**B. SYSTEM MODEL OF OSM-MRA/OSSK-MRA**

The schematic of the proposed OSM-MRA and OSSK-MRA systems is portrayed in Fig. 1 with $N_t = N_r + 1$ transmit antennas and $N_r$ ($N_r > 1$) receive antennas. Moreover, the system model of Fig. 1 could be regarded as an OSM-MRA system when $s_k$ is confirmed as a conventional APM symbol with the constellation size of $M$ while an OSSK-MRA system in case of $s_k = 1$ and $M = 1$.

Explicitly, as shown in Fig. 1, the input binary bits are modulated by the spatial modulation block as an APM symbol $s_k$ with spatially modulated indices $p, i$. Moreover, the index of the receive antenna also conveys information so that the spectral efficiency of $log_2(MN_rN_t)$ bps/Hz and $log_2(N_rN_t)$ bps/Hz can be achieved for OSM-MRA and OSSK-MRA, respectively. Additionally, the throughput equations for OSSK-MRA and OSSK-MRA are still expressed as the frequency of the switching multiplied by the spectral efficiency with data rate gains of $log_2(N_r)$ bps/Hz.

Meanwhile, since the transmitter knows the CSI, after the antenna selection block sends the index of the offset antenna, which is recognized as $\text{tran}$, the channel gain $h_{p,i}$ and $h_{p,\text{tran}}$ will be pre-coded into the transmit signal $x_{k,p,i}$ along with the APM symbol $s_k$. Specifically, considering $\text{tran} = N_t + 1$ and Rayleigh fading channels, the transmit signal can be denoted as

$$x_{k,p,i} = \frac{h_{p,i}}{h_{p,N_r+1}}s_k e^{j\frac{2\pi}{N_r+1}},$$

where $h_{p,i}$ denotes the channel gain corresponding to the channel from the $i - th$ transmit antenna to the $p - th$ receive one. Therefore, the signal at the receiver side can be denoted as

$$y = \sqrt{\rho} \beta_{p,i} H x_{k,p,i} + n, \tag{7}$$

where $\beta_{p,i} = \left| \frac{h_{p,N_r+1}}{h_{p,i}} \right|$. By substituting (6) into (7), we can obtain

$$y = \sqrt{\rho} \beta_{p,i} \times [h_{1,N_r+1} h_{p,i}, \ldots, h_{N_r,N_r+1} h_{p,i}]^T s_k + n.$$ \tag{8}

At the receiver side, the objective function of the detector adopting the ML criterion for OSM-MRA and OSSK-MRA can be formulated as

$$\kappa, p, i = \arg \min_{\kappa \in S, p \in R, i \in T} \left| ||y - \sqrt{\rho} \beta H x_{k,p,i}||^2_F \right. \tag{9}$$

and

$$p, i = \arg \min_{p \in R, i \in T} \left| ||y - \sqrt{\rho} \beta H x_{k,p,i}||^2_F \right., \tag{10}$$

respectively, where $x_{k,p,i}$ has $N_rN_tM$ possible signal modes and $\beta$ is a normalized transmit power factor which could be also replaced by the long-term average transmit power in practical applications.

In general, for the system model described, the spatial modulation index will be pre-coded into the transmit signal in advance so that the RF chain will be offset to the $N_t + 1 - th$. For instance, we consider an OSSK-MRA system with $N_t = 5$ and $N_r = 2$. The channel matrix can be denoted as

$$H = \begin{pmatrix} h_{1,1} & h_{1,2} & h_{1,3} & h_{1,4} & h_{1,5} \\ h_{2,1} & h_{2,2} & h_{2,3} & h_{2,4} & h_{2,5} \end{pmatrix}.$$ \tag{11}

Since the offset antenna is the $5-th$ one, part of $H$, denoted as $H_S = \begin{pmatrix} h_{1,1} & h_{1,2} & h_{1,3} & h_{1,4} \\ h_{2,1} & h_{2,2} & h_{2,3} & h_{2,4} \end{pmatrix}$, is exploited for modulation. Meanwhile, the mapping table and the precoding method of transmitted signal are shown in TABLE 1.

It is shown that OSM-MRA and OSSK-MRA can combine the benefits of conventional OSM and OSSK - a fixed RF chain to completely solve the RF switching problem toward high-rate wireless transmission - while achieving a higher spectral efficiency.

On the other hand, since $N_t + 1$ transmit antennas are employed in the proposed system, with the increase of the...
freedom on the spatial domain, the transmitter requires to select \( N_t \) antennas to convey the information bits for the sake of a transmission performance enhancement.

### III. BER ANALYSIS FOR OSM-MRA/OSSK-MRA

In this section, we aim at providing an analysis of the BER bounds with ML detection for OSM-MRA and OSSK-MRA systems. The objective of the detector is to find the optimal \( \kappa, p \) and \( i \) condition on \( \mathcal{S} \) for OSM-MRA and \( p \in \mathcal{R} \) and \( i \in \mathcal{T} \) for OSSK-MRA by maximizing the maximum-likelihood probabilities as

\[
\kappa, p, i = \arg \max_{\kappa \in \mathcal{S}, p \in \mathcal{R}, i \in \mathcal{T}} p_y(y|x_{k,p,i}, H) \tag{11}
\]

and

\[
p, i = \arg \max_{p \in \mathcal{R}, i \in \mathcal{T}} p_y(y|x_{k,p,i}, H). \tag{12}
\]

Note that \( p_y(y|x_{k,p,i}, H) \) denotes the probability density function (PDF) of \( y \) conditioned on \( H \) and \( x_{k,p,i} \), which can be given by

\[
p_y(y|x_{k,p,i}, H) = \frac{1}{\pi} \exp \left( -\|y - \sqrt{\rho} H x_{k,p,i}\|_F^2 \right). \tag{13}
\]

Therefore, the objective functions (11) and (12) are simplified to (9) and (10), respectively.

For the ML detector, the bit error probability can be derived using the well-known union bounding technique [29] in (14), as shown at the bottom of this page, where \( N(\kappa, p, i \rightarrow \lambda, q, j) \) represents the number of bits in error of deciding \( x_{\lambda,q,j} \) given that \( x_{k,p,i} \) is transmitted and \( p(x_{k,p,i} \rightarrow x_{\lambda,q,j}) \) is the pairwise error probability (PEP) of deciding \( x_{\lambda,q,j} \) given that \( x_{k,p,i} \) is transmitted. Coefficient \( \gamma = N_t N_r M \) is the total number of transmit modes and \( n_{neig} \) denotes the number of neighbors for \( x_{k,p,i} \) that may cause detection errors. For example, when \( n_{neig} = \gamma - 1 \), it represents the case that all the possible neighbors of \( x_{k,p,i} \) should be considered. Therefore, when \( n_{neig} = \gamma - 1 \), the bound in Eq.(14) may be loose since \( P_{e,biu} \) only can be calculated if the Euclidean distance between \( x_{\lambda,q,j} \) and \( x_{k,p,i} \) is small. And the bound can be tightened by considering the appropriate \( n_{neig} \).

According to (13), the PEP conditioned on channel matrix \( H \) is given by in (15), as shown at the bottom of this page. Then, by substituting (7) into (15), we arrive at (16), as shown at the bottom of this page.

Since \( n^H (H x_{k,p,i} - H x_{\lambda,q,j}) \) is a complex Gaussian variable obeying \( CN(0, \|H x_{k,p,i} - H x_{\lambda,q,j}\|^2_F) \), we have

\[
p(x_{k,p,i} \rightarrow x_{\lambda,q,j}|H) = Q \left( \sqrt{\frac{\rho}{4} \|H x_{k,p,i} - H x_{\lambda,q,j}\|^2_F} \right). \tag{17}
\]

For convenience, let \( d = x_{k,p,i} - x_{\lambda,q,j} \) and \( H d \) be complex Gaussian vectors with \( N_r \) dimension obeying \( CN(0, d^H d) \). After normalization, vector \( \frac{H d}{\|H d\|} \) obeys a distribution of \( CN(0, 1) \). Moreover, it can be readily shown that \( \|H d\|^2 \) becomes chi-square distributed with \( 2N_r \) degrees of freedom, and its probability density function can be further expressed as

\[
p_{Hd^2}(x) = \begin{cases} \frac{1}{2^{N_r} \Gamma(N_r)} x^{N_r-1} e^{-x/2}, & x \geq 0 \\ 0, & x < 0 \end{cases} \tag{18}
\]

Thus, the PEP conditioned on \( \beta \) and \( d \) can be expressed as

\[
p(x_{k,p,i} \rightarrow x_{\lambda,q,j}|\beta_{p,i}, d) = \int_0^{+\infty} Q \left( \sqrt{\frac{\rho}{4} \beta_{p,i}^2 d^H d} \right) p_{Hd^2}(x) dx. \tag{19}
\]

Let

\[
a = \frac{\rho}{4} \beta_{p,i}^2 d^H d = \frac{\rho}{4} \left( |s_k - h_{p,i} h_{p,N_r+1} s_k| \right)^2
\]

\[
= \frac{\rho}{4} |s_k - u \cdot s_j|^2
\]

\[
= \frac{\rho}{4} |s_k|^2 - \frac{\rho}{4} |u|^2 |s_j|^2 - \frac{\rho}{2} |s_k||s_j| \cos \theta, \tag{20}
\]

where \( u = \frac{h_{p,i} h_{p,N_r+1}}{h_{p,i} h_{p,N_r+1}} \) and \( \theta = \varphi(s_k) - \varphi(u) - \varphi(s_j) \). Thus the variable \( a \) could be seen as a binary function of \( |u|^2 \) and \( \cos \theta \).
Let $Z = |u|^2 = \frac{h_{h, j}^2}{h_{p, q}^2} \frac{N_t}{N_r} + 1 |^2$ and $T = \cos \theta$, then Eq.(19) can be further formulated as

$$p (z, t) = \int_{0}^{\infty} Q \left( \sqrt{a} (Z, T) x \right) p_{\frac{2}{2N_r}} (x) dx.$$  \hfill (21)

Eq.(21) has a closed form expression given in [30], which can be expressed as

$$p (z, t) = \frac{1}{2N_r} \left( 1 - \frac{a (Z, T)}{1 + a (Z, T)} \right)^{N_r} \sum_{k=0}^{N_r-1} \left( N_r - 1 + k \right)$$

$$\times \left[ 1 - \frac{1}{2} \left( 1 - \sqrt{a (Z, T)} \right) \right]^k.$$  \hfill (22)

Since $Z$ and $T$ have different distributions in different cases, we will discuss each case separately.

1) FOR $p \neq q$

In this case, $Z = \frac{h_{h, j}^2}{h_{p, q}^2} \frac{N_t}{N_r} |^2$. Since these four channel gain factors are independent complex Gaussian variables and the square of the mode of each channel gain factor obeys chi-distribution with 2 degrees of freedom, $Z$ can be seen as the product of two independent $F$ distributions, whose first and second degrees of freedom are 2. And the PDF of $F(2, 2)$ distribution is expressed as

$$f_{F(2,2)} (x) = \left\{ \begin{array}{ll}
(1 + x)^{-2}, & x > 0 \\
0, & x \leq 0.
\end{array} \right. \hfill (23)$$

Since $Z$ is the product of two independent variables, the PDF of $Z$ can be expressed as [31]

$$f_z (z) = \int_{0}^{\infty} f_{F(2,2)} (z/x) f_{F(2,2)} (x) \cdot \frac{1}{|x|} dx.$$  \hfill (24)

Substituting (23) into (24) yields

$$f_z (z) = \int_{0}^{\infty} \left( 1 + \frac{z}{x} \right)^{-2} (1 + t)^{-2} \cdot \frac{1}{t} dt$$

$$= \left\{ \begin{array}{ll}
\frac{z \log z + \log z - 2z + 2}{z^3 - 3z^2 + 3z - 1}, & z > 0 \text{ and } z \neq 1 \\
\frac{1}{6}, & z = 1 \\
0, & z \leq 0.
\end{array} \right. \hfill (25)$$

In order to better analyze the distribution of $\varphi (u)$, Theorem 3.1 is proposed as follows.

**Theorem 3.1:** For any two independent phase random variables obeying $[0, 2\pi]$ uniform distribution, the random variable of their sum or difference still obeys $[0, 2\pi]$ uniform distribution.

**Proof:** The detailed proof is provided in Appendix A.

By Theorem 3.1, $\varphi (u)$ still obeys $[0, 2\pi]$ uniform distribution. Therefore the PDF of $T$ can be expressed as

$$f_T (t) = \left\{ \begin{array}{ll}
\frac{1}{\sqrt{1-t^2}}, & t \in [-1, 1] \\
0, & \text{others}.
\end{array} \right. \hfill (26)$$

2) FOR $p = q$ AND $i \neq j$

In this case, $Z = \frac{h_{h, i}^2}{h_{p, q}^2}$, which obeys $F(2, 2)$. Thus, the PDF of $Z$ is expressed as

$$f_z (z) = \left\{ \begin{array}{ll}
(1 + z)^{-2}, & z \geq 0 \\
0, & z < 0.
\end{array} \right. \hfill (27)$$

Similarly, $\varphi (u)$ obeys $[0, 2\pi]$ uniform distribution and the PDF of $T$ is same as (26).

3) FOR $p = q$ AND $i = j$

In this case, $Z = 1$ and $T = \cos (\varphi (s_a) - \varphi (s_i))$ are determined variables. So the PDF of $Z$ and $T$ can be expressed by the impulse function as

$$f_z (z) = \delta (z - 1) \hfill (28)$$

and

$$f_T (t) = \delta [t - \cos (\varphi (s_a) - \varphi (s_i))]. \hfill (29)$$

Since the mode and the phase of the channel gain factor are independent, the joint PDF of $Z$ and $T$ can be expressed as

$$f_{z, T} (z, t) = f_z (z) f_T (t). \hfill (30)$$

Similarly, in the context of OSSK-MRA, Eq.(20) can be simplified to

$$a = \frac{p}{4} + \frac{p}{4} - \frac{p}{2} \sqrt{Z T}, \hfill (31)$$

and the PDF of variables $Z$ and $T$ are the same as that of OSM-MRA. As results, the PEP of both OSM-MRA and OSSK-MRA can be expressed as

$$p (x, p, i \rightarrow x, q, j) = \int_{-1}^{1} \int_{0}^{\infty} \rho (x, p, i \rightarrow x, q, j | z, t) f_{z, T} (z, t) dz dt. \hfill (32)$$

Finally, the BER bounds of OSM-MRA and OSSK-MRA can be evaluated by substituting (32), (31), (30) and (21), and the other related parameters into (14).

**IV. PERFORMANCE EVALUATION**

In this section, the BER performance and computational complexity of OSM-MRA, OSSK-MRA and their counterparts are evaluated and compared by theoretical analysis and computer simulations over i.i.d Rayleigh flat fading channels with AWGN.
A. COMPLEXITY COMPARISON

In this subsection, the complexity for calculating the best transmission mode is considered in the context of CSI-aided schemes including OSM-MRA, OSSK-MRA, OSM, OSSK, phase alignment assisted SM (SM-P) [32] and SM with the EDAS algorithm. We present the computational requirements at the transmit side as well as the receiver side in terms of the number of floating point operations (FLOPs) for complex multiplication based on [33]. Specifically, Table 2 shows the complexity required in terms of FLOPs, where $N$ is the total antenna number for EDAS, $p$ is the number of the offset antennas in OSM and OSSK. Moreover, the detection complexity of each scheme with the ML detector can be expressed as

$$C_1 = \frac{N_r N_t (N_r/\tau + 3N_r)}{\log_2(N_t N_r)}, \quad (33)$$
$$C_2 = \frac{N_t N_r M (N_t/\tau + 3N_r)}{\log_2(M N_t N_r)}, \quad (34)$$
$$C_3 = \frac{N_t/\tau + 3N_t}{\log_2(N_t)}, \quad (35)$$
$$C_4 = \frac{N_t M/\tau + 3N_t M}{\log_2(N_t M)}, \quad (36)$$

and

$$C_5 = \frac{N_r N_t M/\tau + N_t N_r M}{\log_2(N_t M)}, \quad (37)$$

where $\tau$ is the coherence interval [34].

It is clear from Table 2 that at the transmit side, the FLOPs of proposed OSSK-MRA and OSM-MRA are the same as those of OSSK ($p = 1$) and OSM ($p = 1$), achieving the lowest computational complexity by only calculating the normalization factor $\beta$. On the other hand, the FLOPs required by OSM-MRA and OSSK-MRA at the receiver are $N_r$ times compared to conventional OSM and OSSK due to the increase in the spectrum efficiency.

Specifically, Fig. 2 shows the detailed complexity of OSM-MRA, OSSK-MRA, OSM, OSSK, SM-P, and EDAS in terms of FLOPs, in case of $\tau = 1$, $N_t = 4$, $M = 4$, $N_r = 2$ and $N = 10$.

As shown in Fig. 2, the proposed OSM-MRA and OSSK-MRA schemes achieve the complexity as fewer than 100 FLOPs in this configuration, while exhibiting twice spectrum efficiency than that of OSM, OSSK, respectively. On the other hand, since the detection complexity of OSM-MRA and OSSK-MRA increases with $N_r$, some low complexity detection criteria could be applied to OSM-MRA and OSSK-MRA in future works.

B. SIMULATION RESULTS

In this subsection, we demonstrate the performances of OSM-MRA, OSSK-MRA and their counterparts by Monte Carlo simulations.

Figs. 3 and 4 give the BER performance comparison of OSM-MRA, SM, RSM with zero-forcing (ZF) based precoder [18] and OSM schemes with $p = 1$ and the tight bound of OSM-MRA. Specifically, Fig. 3 shows the BER performances of the aforementioned schemes at a spectrum efficiency of 4bps/Hz with $N_t = 5$, $N_r = 2$ and BPSK for OSM-MRA, $N_t = 8$, $N_r = 1$ and BPSK for OSM and SM, and $N_r = N_t = 4$ and QPSK for RSM. On the other hand, for non-RF switching schemes, it is shown that the proposed OSM-MRA scheme is capable of attaining a SNR
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FIGURE 4. BER comparison of OSM-MRA, OSM, SM, RSM and the tight bound of OSM-MRA.

The proposed OSM-MRA scheme outperforms conventional SM at a high SNR while solving the RF switching problem. For non-RF switching schemes, our proposed OSM-MRA scheme provides approximately 10dB SNR gains than that of conventional OSM at a BER of 0.1 with less antenna usage. Moreover, with the similar performance to RSM, the OSM-MRA scheme exhibits a SNR gain of 5dB in comparison to conventional SM while with only 13 antennas usage and without RF switching. Meanwhile, as expected, the tight bounds of OSM-MRA shown in Figs. 3 and 4 are almost overlapped with the simulation results at high SNRs, which proves the reliability of the theoretical analysis in Section III.

Similarly, Figs. 5 and 6 illustrate the performance comparison among OSSK-MRA, OSSK ($p = 1$), SSK and RSSK [18] schemes when achieving the spectrum efficiency of 5 bit/s/Hz and 6 bit/s/Hz, respectively.

As shown in Fig. 5, OSSK-MRA can efficiently improve the performances of SSK and OSSK in the whole SNR region although it is slightly inferior to RSM in terms of BER. More specifically, OSSK-MRA is 5dB and 10dB superior in terms of SNR compared to SSK and OSSK at a BER of $5 \times 10^{-2}$, respectively, while adopting only 40% and 20% amount of antennas compared to OSSK and RSSK.

As depicted in Fig. 6, RSSK outperforms other three schemes at the cost of 128 antenna usage, which is 2 times and 6 times that of OSSK and OSSK-MRA, respectively. On the other hand, except for RSSK, the curves of other three mentioned schemes shown in Fig. 6 are similar to those in Fig. 5, which illustrates that OSSK-MRA offers significant BER improvement compared to OSSK and SSK.

Based on the above facts, we conclude that both OSM-MRA and OSSK-MRA can totally resolve the RF-switching problem of SM and SSK while facilitating a better BER performance as well as significant antenna usage reduction. It is clear that OSM-MRA and OSSK-MRA constitute beneficial choices compared to their counterparts in terms of their performances versus spectrum efficiency characteristics in the whole SNR region.

V. CONCLUSION

In this paper, OSM-MRA and OSSK-MRA schemes were proposed toward high-spectral efficiency MIMO systems with better BER performances compared to conventional OSM and OSSK schemes, and for more balanced...
tradeoffs among the RF switching, the BER performance and the spectral efficiency. On the one hand, with the aid of the CSI, the OSM-MRA and OSSK-MRA could completely solve the RF switching problem while providing considerable BER performance gain compared to OSM, OSSK and conventional SM schemes. Meanwhile, the BER bounds of OSM-MRA and OSSK-MRA were also derived by theoretical analysis and validated by simulation results. On the other hand, from the simulation results we conclude that, the proposed OSM-MRA and OSSK-MRA schemes outperform traditional OSM, OSSK and SM at the same spectrum efficiency while achieving almost the same computational complexity. In general, we demonstrated that the proposed OSM-MRA and OSSK-MRA schemes substantially improve the attainable performances compared to OSM, OSSK, SM and SSK schemes and are particularly suitable for single-RF transmission scenarios.

Furthermore, since an extra transmit antenna is employed at the transmit side, some transmit antenna selection techniques could be used for OSM-MRA and OSSK-MRA in order to achieve better BER performances. Therefore, we will focus on the OSM-MRA and OSSK-MRA schemes combined with TAS and the derivation of their BER bounds in our future work.

APPENDIX A PROOF OF THEOREM 3.1

Assume that \( x \) and \( y \) are two independent phase variables obeying \([0, 2\pi]\) uniform distribution. Let \( z_1 = x + y \) and \( z_2 = x - y \). According to [35], the PDF of \( z_1 \) and \( z_2 \) can be expressed as

\[
 f_{z_1}(z) = \int f_X(z-u)f_Y(u)\,du = \begin{cases} 
 \frac{z}{4\pi^2}, & z \in [0, 2\pi) \\
 \frac{1}{4\pi^2}, & z \in [2\pi, 4\pi] \\
 0, & \text{otherwise} 
\end{cases}
\]  

(38)

and

\[
 f_{z_2}(z) = \int f_X(z+u)f_Y(u)\,du = \begin{cases} 
 \frac{2\pi + z}{4\pi^2}, & z \in [0, 2\pi) \\
 \frac{2\pi - z}{4\pi^2}, & z \in [-2\pi, 0) \\
 0, & \text{otherwise} 
\end{cases}
\]  

(39)

Since \( z_1 \) and \( z_2 \) are both phase variables, the ranges of \( z_1 \) and \( z_2 \) could be change into \([0, 2\pi]\). Thus (38) and (39) could be reformulated as

\[
 f_{z_1}(z) = f_{z_2}(z) = \begin{cases} 
 1, & z \in [0, 2\pi) \\
 0, & \text{otherwise} 
\end{cases}
\]  

(40)

The Theorem 3.1 has been proved.

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