Labeling Diversity for Media-Based Space-Time Block Coded Spatial Modulation

BABATUNDE S. ADEJUMOBI, (Member, IEEE), AND THOKOZANI SHONGWE, (Member, IEEE)

1Centre for Telecommunication, Department of Electrical and Electronic Engineering Science, University of Johannesburg, Johannesburg 2092, South Africa
2Department of Electrical and Electronic Engineering Technology, University of Johannesburg, Doornfontein Campus, Johannesburg 2028, South Africa

Corresponding author: Babatunde S. Adejumobi (bsaadejumobi@uj.ac.za)

This work was supported by the University of Johannesburg, Johannesburg, South Africa.

ABSTRACT Media-based space-time block coded spatial modulation (MBSTBC-SM) combines the advantages of media-based modulation (MBM) and STBC-SM. Meanwhile, labelling diversity (LD) has been applied to STBC-SM and improve the error performance. Hence, this paper proposes the application of LD to MBSTBC-SM in the form of MBSTBC-SM-LD over a fast, frequency-flat Rayleigh fading channel, without channel estimation. The proposed MBSTBC-SM-LD demonstrates an improved average bit error rate (BER) performance over MBSTBC-SM and STBC-SM. For example, a $4 \times 4, m_{rf} = 2$ MBSTBC-SM-LD demonstrates a 2 dB gain in signal-to-noise ratio (SNR) over MBSTBC-SM. Furthermore, the analytical framework for the union bound on the average BER of MBSTBC-SM-LD is formulated, and validates the Monte Carlo simulation results for the MBSTBC-SM system over a fast, frequency-flat Rayleigh fading channel. In addition, a low-complexity detector, which is able to achieve 73% reduction in computational complexity of MBSTBC-SM-LD, while maintaining a near-ML error performance is proposed.

INDEX TERMS Labeling diversity, media-based modulation, spatial modulation, space-time block codes, RF mirrors, index modulation.

I. INTRODUCTION

The 5G and future wireless communication systems require higher data rates, increased spectral efficiency and improved quality of service or link reliability for improved real-time multimedia services. Hence, there has been an upsurge of research focusing on improving pre-existing multiple-input-multiple-output (MIMO) and massive MIMO systems to meet up with these requirements. Some of the schemes that have improved MIMO include the Alamouti space-time block codes (ASTBC) [1]. ASTBC employs two time-slots to transmit two amplitude and/or phase modulation (APM) symbols. The ability for ASTBC to achieve full diversity and employ low-complexity detection has been of immense attraction to researchers [2].

Over the recent years, a new technique known as index modulation, which employs alternative ways than amplitude/frequency/phase to transmit information has gained immense attraction. For example, several schemes, which employ the indexes of the transmit antennas to transmit additional information in the form of index modulation, have been proposed in the literature. These schemes, in contrast with traditional MIMO systems, employ both index modulation and APM symbols to transmit information, hence, improving the spectral efficiency of traditional MIMO.

Spatial modulation (SM) [3], a novel MIMO scheme, which activates a single antenna to transmit an APM symbol in a single time-slot. However, the selected antenna also transmits additional information, thereby improving the spectral efficiency of traditional MIMO. Furthermore, SM is able to reduce power consumption because it employs a single RF chain. Since SM employs a single antenna, it is able to eliminate intersymbol interference and interchannel interference. However, SM is not able to achieve diversity. To remedy this disadvantage, several schemes which employ the principles in SM have been presented in the literature. For example, space-time block coded spatial modulation (STBC-SM) has been proposed in [4]. STBC-SM combines the advantages of ASTBC and SM to improve the error performance of the duo. STBC-SM transmits a pair of APM symbols using two time-slots, through a pair of transmit antennas, which are selected from a set. Furthermore, the computational complexity of the STBC-SM detector is significantly reduced because of the orthogonality of the STBC-SM codeword.

To further improve the error performance of SM, labelling diversity (LD) has been applied to STBC-SM in the form of
Hence, our contributions in this paper are as follows:

1) We propose the application of labelling diversity to MBSTBC-SM, which we have termed MBSTBC-SM-LD to improve the error performance of MBSTBC-SM.

2) We investigate the effect of labelling diversity to the different MAP schemes, which were used in [6], [9], [10], viz; Scheme 1 and 3.

3) A theoretical expression to evaluate the union bound on the average BER of an $N_{RF} \times N_T$ MBSTBC-SM-LD, having $m_{rf}$ RF mirrors, over a fast, frequency-flat, Rayleigh fading channel, where $N_T$, $N_R$ and $m_{rf}$ are the numbers of transmit antennas, receive antennas and RF mirrors, respectively.

4) The challenge of employing the maximum-likelihood detector is the computational complexity. Hence, we investigate a low complexity detector for MBSTBC-SM-LD, which employs orthogonal projection. One of the major advantage of orthogonal projection, is that it offers communication engineers a choice on the extent of trade-off between the BER performance and the computational complexity that is desired. Furthermore, it can be applied for fast or quasi-static fading channels.

The remainder of this paper is organised as follows: The background of MBSTBC-SM is presented in Section II. The system model of the proposed $N_R \times N_T$ MBSTBC-SM-LD over a fast, frequency-flat Rayleigh fading channel, having $m_{rf}$ RF mirrors is presented in Section III. In Section IV, the theoretical union bound on the average bit error probability for the ML detector of the proposed MBSTBC-SM-LD over an independent and identically distributed (i.i.d.) fast, frequency-flat Rayleigh fading channel is formulated. Section V presents the proposed low-complexity detector for MBSTBC-SM-LD, while in Section VI the analysis of the computational complexities for the different detectors, viz; the ML and low-complexity detectors are analyzed. The Numerical results of the proposed MBSTBC-SM-LD are presented and discussed in Section VII. Finally, this paper is concluded in Section VIII.

**Notation:** The following notations are employed throughout this paper; bold and capital letters represent matrices, while bold small letters denote column vectors of matrices. Other notations include $(\cdot)^T$ and $(\cdot)^H$ which represent transpose and Hermitian, respectively. $(\cdot)^*$ and $(\cdot)^{-1}$ represent the complex conjugate and inverse, respectively, while $|| \cdot ||_F$ and $Q(\cdot)$ represent Frobenius norm and Gaussian Q-function, respectively. Furthermore, $argmin(\cdot)$ represents the minimum of an argument with respect to $w$, and $(\cdot)$ represents the binomial coefficient, $|w|^2_{2p}$ represents the nearest power of two, less than or equal to the $w$.

**II. BACKGROUND OF LABELLING DIVERSITY**

This section presents a background of ASTBC and labelling diversity.

STBC employs two time-slots to transmit two symbols. During the first time-slot, the symbols $s_1$ and $s_2$, which are selected from an $M$-ary amplitude and/or phase modulation (APM) constellation $\Omega_1$ are transmitted, while in
the second time-slot, the conjugates of the symbols transmitted during the first time-slot are transmitted. The transmit codeword of ASTBC may be represented as [1]:

\[
\begin{bmatrix}
    s_p & -s_q^*
\end{bmatrix}
\]  

(1)

where \( s_p \) and \( s_q \), for \( p, q \in [1 : M] \), are the \( p \)-th and \( q \)-th symbol of the \( \Omega_1 \) constellation. \( M \) is the constellation size of \( \Omega_1 \). Each row in (1) corresponds to the individual transmit antenna and the columns correspond to the time-slots.

Space-time labelling diversity (STLD) is similar to STBC, however, whereas the symbols \( s_1 \) and \( s_2 \) are taken from the same \( M \)-ary APM symbol constellation mapper \( \Omega_1 \), the two symbols for STLD are taken from two different optimized mapper sets \( \Omega_1 \) and \( \Omega_2 \) of an \( M \)-ary APM constellation. A \( 2 \times N_R \) STLD system is as shown in Figure 1, while a pictorial example of 16 QAM labelled mappers, as presented in [12] is given in Figures 2 and 3. The codeword for STLD, may be formulated as:

\[
\begin{bmatrix}
    s_{p_1} & s_{q_1} \\
    s_{p_2} & s_{q_2}
\end{bmatrix}
\]  

(2)

where \( s_{p_1} \) and \( s_{q_1} \), for \( p_1, q_1 \in [1 : M] \), are the \( p \)-th and \( q \)-th symbols of \( \Omega_1 \) APM symbol mapper, while \( s_{q_2} \) and \( s_{p_2} \), for \( p_2, q_2 \in [1 : M] \), are the \( q \)-th and \( p \)-th symbols of \( \Omega_2 \) APM symbol mapper, respectively.

### III. SYSTEM MODEL OF THE PROPOSED MBSTBC-SM-LD

This section presents the system model of the proposed MBSTBC-SM-LD system, having \( N_R \) receive antennas and \( N_T \) transmit antennas, over a fast frequency-flat Rayleigh fading channel. The channel gains of a fast fading channel are assumed to be constant during each time-slots and have independent values for different time slots [5]. Each transmit antenna of MBSTBC-SM-LD is equipped with \( m_{rf} \) RF mirrors as depicted in Figure 4. Furthermore, the transmission of the MBSTBC-SM-LD symbols employ two time-slots, which shall be referred to as Time-slot A and Time-slot B, for the first and second time-slots, respectively.

A group of \( d \) bits which is fed into the input of the MBSTBC-SM-LD system is split into three groups, such that \( 2 \log_2 M \) bits are employed to select two symbols from two different \( M \)-ary APM symbol mappers. The symbols \( x_{p_1} \) and \( x_{q_1} \), where \( x_{p_1}, x_{q_1} \in \Omega_1 \), for, for \( p_1, q_1 \in [1 : M] \), are selected from the first APM symbol mapper \( \Omega_1 \), such that \( x_{p_1} \) and \( x_{q_1} \) are the \( p \)-th and \( q \)-th symbol of \( \Omega_1 \). In the same manner, the same bits are employed to select the symbols \( x_{p_2} \) and \( x_{q_2} \), where \( x_{p_2}, x_{q_2} \in \Omega_2 \), for \( p_2, q_2 \in [1 : M] \), which are the \( p \)-th and \( q \)-th symbol of the second APM symbol mapper \( \Omega_2 \). We assume that the normalised signal power \( E\{ |x_i|^2 \} = E\{ |x_{q_i}|^2 \} = 1 \) where \( i \in [1 : 2] \) and \( E\{ \} \) represents the expectation operator. These symbols are transmitted during Time-Slot A, while the conjugates are transmitted during Time-slot B. The second group of bits are the \( \log_2 c \) bits, \( c = 0.5(N_T(N_T - 1))_{2g} \) bits, which are employed to select the transmit antenna pair \( tr_1 \) and \( tr_2 \), where \( tr_1, tr_2 \in [1 : N_T] \).

The last (third) group of \( \log_2 N_m^2 \) bits are the \( \psi = \gamma m_{rf} \) bits, which are employed to select a MAP of the available \( N_m = 2^\nu m_{rf} \) MAPS corresponding to each transmit antenna, where \( \gamma \) is a scalar multiplier, which is determined by the scheme being used, details of this are given in [9]. For example, in Scheme 2 of [9], where \( \gamma = 1 \), the \( \psi \) bits are used to select a single MAP index, which is employed...
by all mirrors, while in Scheme 1 and 3, where \( \gamma = 2 \),
the \( \psi \) bits are further subdivided into two subgroups, \( \psi_1 \)
and \( \psi_2 \) which are employed to select two different MAPs
corresponding to the transmit antenna pair \( tr_1 \) and \( tr_2 \). The
scheme employed determines the MAP arrangement for the
first and second time-slot. Hence, the spectral efficiency
scheme employed during the second time-slot. Throughout this paper, \( x_{p1} \) and
\( x_{q1} \) shall represent the \( p \)-th and \( q \)-th symbol of the \( M \)-QAM
Mapper 1, while \( x_{p2} \) and \( x_{q2} \) shall represent \( p \)-th and \( q \)-th
symbols of the \( M \)-QAM Mapper 2.

The second group of the \( d \) bits \( b_x \) is employed to transmit the selected symbols. For \( N_T = 4 \), the \( c \)
transmit antenna pairs \((tr_1, tr_2)\) are \((1, 2), (3, 4), (1, 4) \) and
\((2, 3) \). Hence, the \( \tau \)-th transmit pair that will be selected by the \( b_x \) bits is \((3, 4) \). Finally, the third group of \( \log_2 N_m^2 \) bits, are
the \( \psi = 1000 \) bits, which is further subdivided into \( \psi_1 = 10 \)
and \( \psi_2 = 00 \) bits, which are employed to select the MAP to
be activated during transmission. The \( \psi_1 \) and \( \psi_2 \) bits are
employed to activate the \( \omega_k = 3 \)-rd and \( \omega_l = 1 \)-st MAPs of
the RF mirrors on the 3-rd and 4-th transmit antennas, respectively.

The bit assignment for Scheme 3 follows the same method
as Scheme 1 during the first time-slot, however, in the
second time-slot, there are changes made to the MAP indices.
 Whereas \( \omega_k = \omega_m \) and \( \omega_l = \omega_n \) in Scheme 1, \( \omega_k = \omega_n \) and
\( \omega_l = \omega_m \) in Scheme 3. Hence, using the same input bits as
that in Table 1, the bit assignment for Scheme 3 is as shown in Table 2

The received signal matrix \( Y \) at the receiver of the
MBSTBC-SM-LD system may be represented as:

\[
Y = \begin{bmatrix} y_A & y_B \end{bmatrix} = \sqrt{\frac{\rho}{2}} \begin{bmatrix} H^A u_A & H^B u_B \end{bmatrix} + N
\]

where \( \rho \) is the average signal-to-noise ratio (SNR) at each
receive antenna. \( Y = \begin{bmatrix} y_A & y_B \end{bmatrix} \) is an \( N_R \times 2 \) matrix having
each column representing the signal vector for each time-slot, viz: Time-slot A and B. \( u_A \) is an \( N_T N_m \times 1 \) transmit vec-
tor for Time-slot A, having the symbols \( x_{p1} \) and \( x_{q1} \) as the
only non-zero entry on the \( t_{p1} \)-th, \( t_{q1} = N_m(t_{r1} - 1) + \omega_k \)}
TABLE 1. Bit assignments for Scheme 1

| Row | d Bits | v1 | v2 | q | bory | τ | ψ1 | ωk | ω2 | ωm | ωn |
|-----|--------|----|----|---|------|---|----|----|----|----|----|
| 1   | 01 11 01 10 00 | 01 | 2 | 11 | 4 | 0 | 2 | 10 | 3 | 00 | 1 |
| 2   | 11 01 10 01 11 | 11 | 4 | 01 | 2 | 01 | 2 | 10 | 3 | 01 | 2 |
| 3   | 01 10 00 01 10 | 01 | 2 | 10 | 3 | 00 | 1 | 01 | 2 | 10 | 3 |

TABLE 2. Bit assignments for Scheme 3

| Row | d Bits | v1 | v2 | q | bory | τ | ψ1 | ωk | ω2 | ωm | ωn |
|-----|--------|----|----|---|------|---|----|----|----|----|----|
| 1   | 01 11 01 10 00 | 01 | 2 | 11 | 4 | 0 | 2 | 10 | 3 | 00 | 1 |
| 2   | 11 00 11 01 11 | 11 | 4 | 00 | 1 | 01 | 2 | 11 | 4 | 01 | 2 |
| 3   | 01 10 00 10 10 | 01 | 2 | 10 | 3 | 00 | 1 | 10 | 3 | 01 | 2 |

and $t_{q_k}$-th, $t_{q_n} = N_m(t_{r_2} - 1) + \omega_l$ position, respectively. Conversely, $u_B$ is an $N_T N_m \times 1$ transmit vector for Time-slot B, having the symbols $x_{q_2}$ and $x_{p_2}$ as the only non-zero entry on the $t_{q_k}$-th, $t_{q_n} = N_m(t_{r_1} - 1) + \omega_m$ and $t_{p_k}$-th, $t_{p_n} = N_m(t_{r_1} - 1) + \omega_m$ position, respectively. The transmit codeword matrix $U = [u_A u_B]$ for the proposed MBSTBC-SM-LD system is similar to the codeword matrix of STBC-SM given in [4]. However, whereas the transmit codeword matrix for STBC-SM is an $N_T \times 2$ matrix, MBSTBC-SM-LD employs an $N_T N_m \times 2$ codeword matrix.

The fast frequency flat Rayleigh fading channel matrices $H^A$ and $H^B$ are each defined as an $N_R \times N_T N_m$ channel matrix, which follows a zero mean and unit variance $CN(0, 1)$ distribution, where $H^A = \left[H^A_{1:H^A_{N_R}}\right]$ and $H^B = \left[H^B_{1:H^B_{N_T N_m}}\right]$. However, for a fast fading channel, it is assumed that $H^A \neq H^B$ [5], [9]. The noise $N$ at the receiver is defined as an $N_R \times 2$ additive white Gaussian noise (AWGN) matrix, whose entries are independent and identically distributed (i.i.d.) according to $CN(0, 1)$ distribution.

A simplified form of (3), for the received signal vectors $y_A$ and $y_B$ of MBSTBC-SM-LD over a fast frequency-flat Rayleigh fading channel for two time-slots; viz, times-slots A and B, may be written as [9]:

$$y_A = \sqrt{\frac{\rho}{2}} \left( H^A_{1:t_{r_1}} x_{p_2} e_{q_2} + H^A_{2:t_{r_2}} x_{q_2} e_{o_2} \right) + n_A$$  (4)

$$y_B = \sqrt{\frac{\rho}{2}} \left( H^B_{1:t_{r_1}} x_{q_2} e_{o_2} + H^B_{2:t_{r_2}} x_{o_2} e_{o_2} \right) + n_B$$  (5)

where $H^A_{1:t_{r_1}}$ and $H^B_{1:t_{r_1}}$, for $t_{r_1}, t_{r_2} \in [1 : N_T]$, are each the $N_R \times N_m$ channel matrix for the transmit antenna pair $t_{r_1}$ and $t_{r_2}$, employed during time-slot A, respectively, while $H^B_{2:t_{r_2}}$ and $H^B_{2:t_{r_2}}$ are each the $N_R \times N_m$ channel matrix for the transmit antenna pair $t_{r_1}$ and $t_{r_2}$, employed in time-slot B, respectively. $n_A$ and $n_B$ are each an $N_R \times 1$ AWGN vector, whose entries are i.i.d. with $CN(0, 1)$ distribution for time-slot A and time-slot B, respectively, such that $N = [n_A n_B]$, where $n_B = \{n_B_{1} n_B_{2} \cdots n_B_{N_T N_m}\}$, for $\beta \in \{A, B\}$, $H_{t_{r_1}} = \{h_{t_{r_1},1} h_{t_{r_1},2} \cdots h_{t_{r_1},N_m}\}$, for $a \in [1, 2]$ and $i \in \{A, B\}$. Each vector of $H_{t_{r_2}}$ can be defined by $h_{t_{r_2}} = \{h_{t_{r_2},1} h_{t_{r_2},2} \cdots h_{t_{r_2},N_m}\}$, for $\psi \in [1 : N_m]$. The vectors $e_{o_2}$ and $e_{o_2}$ are each an $N_m \times 1$ vector having the $\omega_m$-th and $\omega_m$-th element, respectively, as unity during the first time-slot while $e_{o_2}$ and $e_{o_2}$ are each an $N_m \times 1$ vector having the $\omega_m$-th and $\omega_m$-th element, respectively, as unity during the second time-slot. Hence, a simplified form for the received signal vector for time-slots A and B given in (4) and (5) respectively, of the proposed MBSTBC-LD is formulated as [9]:

$$y_A = \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^A x_{p_2} + h_{t_{r_2},o_2}^A x_{q_2} \right) + n_A$$  (6)

$$y_B = \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^B x_{o_2} + h_{t_{r_2},o_2}^B x_{o_2} \right) + n_B$$  (7)

From (4) and (5), the joint maximum-likelihood (ML) detector for the received signal vector of MBSTBC-SM-LD may be formulated as:

$$\begin{aligned}
\hat{r}, & \hat{\omega}_k, \hat{\omega}_l, \hat{p}_1, \hat{q}_2 \\
= & \arg \min \limits_{r \in \{1,2\}, \omega_k, \omega_l, p_1, q_2 \in \{1,M\}} \left\{ \left\| y_A - \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^A x_{p_2} + h_{t_{r_2},o_2}^A x_{q_2} \right) \right\|_F^2 \\
+ & \left\| y_B - \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^B x_{o_2} + h_{t_{r_2},o_2}^B x_{o_2} \right) \right\|_F^2 \right\}_F \\
\end{aligned}$$  (8)

while the reduced form may be represented as:

$$\begin{aligned}
\hat{r}, & \hat{\omega}_k, \hat{\omega}_l, \hat{p}_1, \hat{q}_2 \\
= & \arg \min \limits_{r \in \{1,2\}, \omega_k, \omega_l, p_1, q_2 \in \{1,M\}} \left\{ \left\| y_A - \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^A x_{p_2} + h_{t_{r_2},o_2}^A x_{q_2} \right) \right\|_F^2 \\
+ & \left\| y_B - \sqrt{\frac{\rho}{2}} \left( h_{t_{r_1},o_2}^B x_{o_2} + h_{t_{r_2},o_2}^B x_{o_2} \right) \right\|_F^2 \right\}_F \\
\end{aligned}$$  (9)

where $\hat{r}, \hat{\omega}_k, \hat{\omega}_l, \hat{p}_1$ and $\hat{q}_2$ are estimates of $r, \omega_k, \omega_l, p_1$ and $q_2$, respectively.

A reduced form of (8), which will be employed to calculate the computational complexity of MBSTBC-SM-LD may be represented as [6], [9]:

$$\begin{aligned}
\hat{r}, & \hat{\omega}_k, \hat{\omega}_l, \hat{p}_1, \hat{q}_2 \\
= & \arg \min \limits_{r \in \{1,2\}, \omega_k, \omega_l, p_1, q_2 \in \{1,M\}} \left\{ \left\| \rho^\frac{1}{2} g_{p_1}^H \right\|_F^2 + \left\| 2\rho^\frac{1}{2} g_{q_2}^H \right\|_F^2 \\
+ & \left\| h_{t_{r_1},o_2}^A x_{p_2} + h_{t_{r_2},o_2}^B x_{o_2} \right\|_F^2 \\
- & 2\rho \left\| y_A^H g_{p_1} \right\|_F^2 \right\}_F \\
\end{aligned}$$  (10)
where $g_{p_1}^i = h_{r_1, o_1}^i x_{p_1}, g_{q_1}^i = h_{r_2, o_1}^i x_{q_1}, g_{q_2}^m = h_{r_1, o_2}^m x_{q_2}$ and $g_{p_2}^m = h_{r_2, o_2}^m x_{p_2}$.

IV. ANALYTICAL ABEP OF MBSTBC-SM-LD

In this section, the analytical union bound on the ABEP of MBSTBC-SM-LD is formulated for a fast, frequency-flat Rayleigh fading channel.

The union bound on the ABEP of MBSTBC-SM-LD is defined as [9]:

$$ABEP \leq \frac{1}{cN^2M^2} \sum_u \sum_{\hat{U}} N_{U, \hat{U}} P(U \rightarrow \hat{U})$$

(11)

where $P(U \rightarrow \hat{U})$ is the pair error wise probability (PEP) event, given that $U$ is the transmit codeword matrix of the form defined in (3), which is erroneously decided by the receiver as $\hat{U}$. $U$ and $\hat{U}$ are each an $N_T N_m \times 2$ codeword matrix. $N_{U, \hat{U}}$ is the number of bits received in error, given that the PEP event $P(U \rightarrow \hat{U})$ has occurred.

Considering $H_t^A = [H_{r_1}^A H_{r_2}^A], H_t^B = [H_{r_1}^B H_{r_2}^B] \in H$, the conditional probability $P(U \rightarrow \hat{U} \mid H)$ may be formulated as [4]:

$$P(U \rightarrow \hat{U} \mid H) = \left| \begin{array}{l}
|y_A - \frac{\rho}{\sqrt{2}} (H_{r_1}^A x_{q_1}, e_{o_1})|_F^2 \\
|y_B - \frac{\rho}{\sqrt{2}} (H_{r_1}^B x_{q_2}, e_{o_2})|_F^2 \\
|y_M - \frac{\rho}{\sqrt{2}} (H_{r_1}^B x_{q_2}, e_{o_2})|_F^2 \\
|y_B - \frac{\rho}{\sqrt{2}} (H_{r_2}^B x_{q_2}, e_{o_2})|_F^2 \\
\end{array} \right|$$

(12)

Similar to [6], [12], the reduced form of the conditional probability in (12) may be formulated as:

$$P(U \rightarrow \hat{U} \mid H) = Q\left(\frac{\rho}{\sqrt{8}} |H_t^A|_F^2 |u_{A, A}|^2_F + \frac{\rho}{\sqrt{8}} |H_t^B|_F^2 |u_{B, B}|^2_F\right)$$

(13)

where $u_{A, A} = u_A - \hat{u}_A, u_A$ and $\hat{u}_A$ are the A-th $N_T N_m \times 1$ column vectors, representing the first time-slots of $U$ and $\hat{U}$, respectively. Furthermore, $u_{B, B} = u_B - \hat{u}_B, u_B$ and $\hat{u}_B$ are the B-th $N_T N_m \times 1$ column vectors, representing the second time-slots of $U$ and $\hat{U}$, respectively. Therefore, $U = [u_A u_B]$ and $\hat{U} = [\hat{u}_A \hat{u}_B]$ are each an $N_T N_m \times 2$ codeword matrix.

The unconditional PEP can be obtained by averaging the conditional PEP $P(U \rightarrow \hat{U} \mid H)$, which may be formulated as [12]:

$$P(U \rightarrow \hat{U}) = \int_0^\infty \int_0^\infty Q(\sqrt{k_A} + k_B) f_{k_A}(k_A) f_{k_B}(k_B) \times dk_A dk_B$$

(14)

where $f_{k_A}(k_A)$ and $f_{k_B}(k_B)$ are the probability density functions of $k_A$ and $k_B$, respectively, which follows a Rayleigh distribution given as [12]:

$$f_{k_i}(k_i) = \frac{k_i^{N_R-1}}{(\frac{\sigma_i^2}{2})^{N_R}} e^{-k_i/(2\sigma_i^2)}$$

(15)

To simplify the expression in (14), the exponential expression of $Q(\sqrt{x})$ is applied to (13). The exponential expression of (13) may be represented as:

$$Q(\sqrt{k_A} + k_B) = \frac{1}{2a} \left[ \frac{1}{2} e^{-\frac{k_A}{N_R}} e^{-\frac{k_B}{N_R}} + \sum_{g=1}^{a-1} e^{-\frac{k_A}{N_R}} e^{-\frac{k_B}{N_R}} \right]$$

(16)

where $\theta_R = \frac{\pi}{2N_R}, a$ is the number of iterations needed for convergence of the Q-function.

Substituting (15) and (16) to (14) becomes [13]:

$$P(U \rightarrow \hat{U}) = \frac{1}{2a} \left[ M_A(\frac{1}{2}) M_B(\frac{1}{2}) + \sum_{g=1}^{a-1} M_A(\frac{1}{2}) M_B(\frac{1}{2}) \left( \frac{1}{2} \right)^{\frac{1}{2}} \right]$$

(17)

where $M_A(w) = (1 + 2\sigma_A^2 w)^{-N_R}$ and $M_B(w) = (1 + 2\sigma_B^2 w)^{-N_R}$

V. LOW-COMPLEXITY NEAR-ML DETECTOR FOR MBSTBC-SM-LD

The computational complexity for MBSTBC-SM-LD is very large, because the computation/search algorithm parses large amount of data as would be shown in the next section. Hence, in this section, we propose an orthogonal projection-based low-complexity detector for the proposed MBSTBC-SM-LD system, over a fast frequency-flat Rayleigh fading channel, having $N_R$ receive antennas, $N_T$ transmit antennas and $m_r$ RF mirrors associated with each transmit antenna. The proposed low-complexity near-ML detector for MBSTBC-SM-LD firstly determines the $f_1$ and $f_2$ nearest estimates $u_1 = [u_{1,1}, u_{1,2}, \ldots, u_{1,f_1}]$ and $u_2 = [u_{2,1}, u_{2,2}, \ldots, u_{2,f_2}]$, which are the indexes of the nearest estimates of the $p$-th and $q$-th, $p, q \in [1 : M]$ transmitted symbols, respectively, for the APM mappers $\Omega_1$ and $\Omega_2$, for all the $c$ antenna pairs $(r_1, r_2), (r_1, r_2) \in [1 : N_T]$, such that $r_1 \neq r_2$ and $f_1, f_2 \ll M^2$. 
To determine the indexes of the most likely candidate set of the transmitted symbols, the orthogonal projection matrices $P^A_{tr_a,\omega_0}$ and $P^B_{tr_a,\omega_0}$ where $a \in [1 : 2]$ and $\omega_0 \in [1 : N_0]$ corresponding to the channel subspace $h^A_{tr_a,\omega_0}$ and $h^B_{tr_a,\omega_0}$ respectively, are computed, such that $P^A_{tr_a,\omega_0} h^A_{tr_a,\omega_0} = h^A_{tr_a,\omega_0} P^A_{tr_a,\omega_0}$ and $P^B_{tr_a,\omega_0} h^B_{tr_a,\omega_0} = 0$. $P^A_{tr_a,\omega_0}$ and $P^B_{tr_a,\omega_0}$ are defined in (18) and (19) as [5], [9]:

$$P^A_{tr_a,\omega_0} = I_{NR} - h^A_{tr_a,\omega_0} \left( h^A_{tr_a,\omega_0} H h^A_{tr_a,\omega_0} \right)^{-1} \left( h^A_{tr_a,\omega_0} H \right)^H$$

(18)

$$P^B_{tr_a,\omega_0} = I_{NR} - h^B_{tr_a,\omega_0} \left( h^B_{tr_a,\omega_0} H h^B_{tr_a,\omega_0} \right)^{-1} \left( h^B_{tr_a,\omega_0} H \right)^H$$

(19)

where $P^A_{tr_a,\omega_0}$ and $P^B_{tr_a,\omega_0}$ are the projection matrices for the time-slots $A$ and $B$, respectively, and given that the $tr_a$-th, $a \in [1 : 2]$ transmit antenna has been employed, while the $\omega_0$-th MAP has been activated.

If $z^A_{q_{1}} = x_{q_1}$ and $z^B_{q_{2}} = x_{q_2}$, where $j \in [1 : f_1]$, $q_1, q_2 \in [1 : M]$, $z^A_{q_1} \subseteq \Omega_1$ and $z^B_{q_2} \subseteq \Omega_2$ then, the sum of the projections can be formulated as [5], [9], [14]:

$$P^A_{tr_1,\omega_0} r^A_{tr_2,\omega_0} \in \Omega_1 \quad P^B_{tr_2,\omega_0} r^B_{tr_2,\omega_0} \in \Omega_2$$

(20)

where $r^A_{tr_2,\omega_0}$ and $r^B_{tr_2,\omega_0}$ are the projection spaces corresponding to the projection matrices in $P^A_{tr_1,\omega_0}$ and $P^B_{tr_2,\omega_0}$, respectively, and are defined in (21) and (22), respectively as:

$$r^A_{tr_2,\omega_0} = y_A - \sqrt{\frac{\rho}{2}} h^A_{tr_2,\omega_0} z^A_{q_1}$$

(21)

$$r^B_{tr_2,\omega_0} = y_B - \sqrt{\frac{\rho}{2}} h^B_{tr_2,\omega_0} z^A_{q_2}$$

(22)

however, if $z^A_{q_1} \neq x_{q_1}$ and $z^B_{q_2} \neq x_{q_2}$, (20) yields [5], [9], [14]:

$$\sqrt{\frac{\rho}{2}} P^A_{tr_1,\omega_0} h^A_{tr_2,\omega_0} \left( x_{q_1} - z^A_{q_1} \right)$$

(23)

In the same manner as (20), if $z^A_{q_1} \neq x_{q_1}$ and $z^B_{q_2} \neq x_{q_2}$, where $k \in [1 : f_2]$, $p_1, p_2 \in [1 : M]$, $z^A_{q_1} \subseteq \Omega_1$ and $z^B_{q_2} \subseteq \Omega_2$ then, the sum of the projections can be formulated as [5], [9], [14]:

$$P^A_{tr_2,\omega_0} r^A_{tr_1,\omega_0} \in \Omega_1 \quad P^B_{tr_1,\omega_0} r^B_{tr_2,\omega_0} \in \Omega_2$$

(24)

where $r^A_{tr_1,\omega_0}$ and $r^B_{tr_1,\omega_0}$ are the projection spaces corresponding to the projection matrices in $P^A_{tr_2,\omega_0}$ and $P^B_{tr_1,\omega_0}$, respectively, and are defined in (25) and (26), respectively as:

$$r^A_{tr_1,\omega_0} = y_A - \sqrt{\frac{\rho}{2}} h^A_{tr_1,\omega_0} z^A_{q_1}$$

(25)

$$r^B_{tr_1,\omega_0} = y_B - \sqrt{\frac{\rho}{2}} h^B_{tr_1,\omega_0} z^A_{q_2}$$

(26)

However, if $z^A_{q_1} \neq x_{p_1}$ and $z^B_{q_2} \neq x_{p_2}$, (24) yields [5], [9], [14]:

$$\sqrt{\frac{\rho}{2}} P^A_{tr_2,\omega_0} h^A_{tr_1,\omega_0} \left( x_{p_1} - z^A_{q_1} \right)$$

$$+ \sqrt{\frac{\rho}{2}} P^B_{tr_2,\omega_0} h^B_{tr_1,\omega_0} \left( x_{p_2} - z^B_{q_2} \right)$$

$$+ P^A_{tr_1,\omega_0} n_A + P^B_{tr_1,\omega_0} n_B$$

(27)

As can be seen from (20), (23), (24) and (27), the Frobenius norms of (20) and (24) is less than (23) and (27), respectively. Hence, the most likely candidates of the transmitted symbols $x_{p_1}, x_{p_2}, x_{q_1}$ and $x_{q_2}$ are chosen by selecting the $f_1$ and $f_2$ most likely symbols, which offer the least Frobenius norms. Then, the ML rule is employed to perform an exhaustive search over the $f_1$ and $f_2$ selected most likely candidates, in order to estimate the transmitted APM symbols.

The steps following outlines the algorithm for the proposed orthogonal projection-based low-complexity detector of MBSTBC-SM-LD.

Firstly, the projection matrices for all $\tau$-th, $\tau \in [1 : c]$ antenna pair and for the time-slots $A$ and $B$, given as $P^A_{tr_1,\omega_0}$, $P^A_{tr_1,\omega_0}$, $P^A_{tr_2,\omega_0}$ and $P^B_{tr_2,\omega_0}$ are obtained. Furthermore, the projection spaces $r^A_{tr_1,\omega_0}$, $r^A_{tr_2,\omega_0}$, $r^B_{tr_2,\omega_0}$ and $r^B_{tr_1,\omega_0}$ are also determined for all values of $p_1, p_2, q_1$ and $q_2$ where $p_1, p_2, q_1, q_2 \in [1 : M]$ and for all MAP combinations $\omega_A, \omega_B, \omega_1, \omega_2$, where $\omega_A, \omega_B, \omega_1, \omega_2 \in [1 : N_0]$. The projection matrices $P^A_{tr_1,\omega_0}$ and $P^A_{tr_2,\omega_0}$ of the proposed MBSTBC-SM-LD, for the $\tau$-th antenna, during time-slot $A$, may be defined as [5], [14]:

$$P^A_{tr_1,\omega_0} = I_{NR} - h^A_{tr_1,\omega_0} \left( h^A_{tr_1,\omega_0} H h^A_{tr_1,\omega_0} \right)^{-1} \left( h^A_{tr_1,\omega_0} H \right)^H$$

(28)

$$P^B_{tr_2,\omega_0} = I_{NR} - h^B_{tr_2,\omega_0} \left( h^B_{tr_2,\omega_0} H h^B_{tr_2,\omega_0} \right)^{-1} \left( h^B_{tr_2,\omega_0} H \right)^H$$

(29)

while the projection matrices for time-slot $B$, $P^B_{tr_1,\omega_0}$ and $P^B_{tr_2,\omega_0}$ may be represented as:

$$P^B_{tr_1,\omega_0} = I_{NR} - h^B_{tr_1,\omega_0} \left( h^B_{tr_1,\omega_0} H h^B_{tr_1,\omega_0} \right)^{-1} \left( h^B_{tr_1,\omega_0} H \right)^H$$

(30)

$$P^B_{tr_2,\omega_0} = I_{NR} - h^B_{tr_2,\omega_0} \left( h^B_{tr_2,\omega_0} H h^B_{tr_2,\omega_0} \right)^{-1} \left( h^B_{tr_2,\omega_0} H \right)^H$$

(31)

The projection spaces for the $\tau$-th transmit antenna pair and the $N_0^2$ MAP combination of the proposed MBSTBC-SM-LD may be formulated as [5], [14]:

$$r^A_{tr_2,\omega_0} = y_A - \sqrt{\frac{\rho}{2}} h^A_{tr_2,\omega_0} \left( x_{p_1} - z^A_{q_1} \right)$$

(32)

$$r^B_{tr_1,\omega_0} = y_B - \sqrt{\frac{\rho}{2}} h^B_{tr_1,\omega_0} z^A_{q_1}$$

(33)
\[ r_{\tau_{1}, \omega_{1}; p, q_{2}} = y_{A} - \sqrt{\frac{\rho}{2}} h_{\tau_{1}, \omega_{1}; p, q_{2}} z_{p, q_{2}}; \quad \forall z_{p, q_{2}} \in \Omega_{2} \]  

where \( p_{1}, p_{2}, q_{1}, q_{2} = [1 : M] \), \( h_{\tau_{1}, \omega_{1}}^{A} \) and \( h_{\tau_{2}, \omega_{2}}^{A} \) are the \( \omega_{k} \)-th and \( \omega_{r} \)-th, column vectors of the \( \tau \)-th antenna pair channel matrices \( H_{\tau_{1}}^{A} \) and \( H_{\tau_{2}}^{A} \), respectively, which are employed during time-slot \( A \). The channel subspace \( h_{\tau_{1}, \omega_{1} \omega_{m}} \) and \( h_{\tau_{2}, \omega_{2} \omega_{m}} \) are the \( \omega_{m} \)-th and \( \omega_{m} \)-th column vectors of the channel matrices \( H_{\tau_{1}}^{B} \) and \( H_{\tau_{2}}^{B} \), respectively, which are the channel matrices for the \( \tau \)-th antenna pair employed during time-slot \( B \).

Secondly, the \( f_{1} \) and \( f_{2} \) nearest estimates of \( x_{p_{1}} \) and \( x_{q_{1}} \),

\[ z_{u_{1}}^{q_{1}} = \frac{z_{u_{1}}^{q_{1}}}{z_{u_{1}1}} \frac{x_{q_{1}}}{z_{u_{1}1}} \cdot \cdot \cdot \frac{z_{u_{1}}^{q_{1}}}{z_{u_{1}1}} \subseteq \Omega_{1}, \quad z_{u_{2}}^{q_{2}} = \frac{z_{u_{2}}^{q_{2}}}{z_{u_{2}1}} \frac{x_{q_{2}}}{z_{u_{2}1}} \cdot \cdot \cdot \frac{z_{u_{2}}^{q_{2}}}{z_{u_{2}1}} \subseteq \Omega_{1} \]

and

\[ z_{u_{2}1}^{q_{2}} = \frac{z_{u_{2}1}}{z_{u_{2}1}} \frac{x_{q_{2}}}{z_{u_{2}1}} \cdot \cdot \cdot \frac{z_{u_{2}1}}{z_{u_{2}1}} \subseteq \Omega_{2} \]

for the \( \tau \)-th, \( \tau \in [1 : c] \) transmit antenna pair and the \( N_{m}^{2} \) MAP combinations are obtained. These symbols are obtained by determining the \( f_{1} \) and \( f_{2} \) symbols, which offer the minimum metric in (36) and (37) formulated as [5], [9]:

\[ z_{u_{1}}^{q_{1}}, z_{u_{2}}^{q_{2}} = \arg \min_{r_{A}, r_{B}} \| p_{r_{1}, o_{1}}^{A}, r_{B_{1}, o_{1}}^{B} \|^2 \]  

where \( \tau \in [1 : c] \), \( \omega_{k}, \omega_{r}, \omega_{m}, \omega_{k} = [1 : N_{m}], p, q \in [1 : M] \).

Finally, the joint ML rule is applied over all the symbols of \( z_{u_{1}}^{q_{1}}, z_{u_{2}}^{q_{2}} \), \( z_{u_{2}1}^{q_{2}} \) and \( z_{u_{2}1}^{q_{2}} \) obtained in (36) and (37), for all the \( c \) transmit antenna pair combinations and all \( N_{m}^{2} \) MAP combinations. The joint ML detector for Scheme 1 and 3 may be represented as:

\[ \hat{\tau}, \hat{\omega}_{k}, \hat{\omega}_{r}, \hat{p}, \hat{q} = \arg \min_{z_{u_{1}}^{q_{1}}, z_{u_{2}}^{q_{2}}} \| y_{A} - \sqrt{\frac{\rho}{2}} h_{\tau_{1}, \omega_{1} \omega_{m}} \|^{2} \]  

where \( \tau \in [1 : c], j \in [1 : f_{1}] \) and \( k \in [1 : f_{2}] \). Furthermore, \( \hat{\tau}, \hat{\omega}_{k}, \hat{\omega}_{r}, \hat{p} \) and \( \hat{q} \) are the estimates of \( \tau, \omega_{k}, \omega_{r}, p \) and \( q \). The values of \( \hat{p} \) and \( \hat{q} \) is known since \( \hat{p}_{1} = \hat{p}_{2} = \hat{p} \) and \( \hat{q}_{1} = \hat{q}_{2} = \hat{q} \) based on our system model. \( \hat{p} \) and \( \hat{q} \) are the estimates of \( p \) and \( q \).

## VI. COMPUTATIONAL COMPLEXITY ANALYSIS

This section analyses the computational complexity in terms of complex operations performed by the proposed MBSTBC-SM-LD. Furthermore, we compare the MBSTBC-SM-LD with MBSTBC-SM.

In (10), each term contains ten \( N_{R} \) complex operations. Since the ML search is performed over \( N_{m}^{2} \) MAP combinations, \( c \) transmit antenna pairs and \( M^{2} \) symbol combination pairs, ignoring the real operations performed by the proposed system, the computational complexity may be given as:

\[ 10cN_{R}N_{m}^{2}M^{2} \]  

The computational complexity of the proposed MBSTBC-SM-LD is calculated in three phases.

Firstly, the computational complexity employed in calculating the projection matrix is similar to [5], hence, the computational complexity for the first phase of detection \( \Delta_{\text{phase1}} \), may be given as:

\[ \Delta_{\text{phase1}} = 8N_{R}^{2} + 12N_{R} - 4 \]  

The second phase of detection \( \Delta_{\text{phase2}} \), involves determining the most-likely symbol set \( z_{u_{1}}^{q_{1}} \) and \( z_{u_{2}}^{q_{2}} \) that has been transmitted, as given in (36) and (37). The computational complexity of \( \Delta_{\text{phase2}} \), is given as:

\[ \Delta_{\text{phase2}} = 4(2MN_{R}^{2} + 4MN_{R} + N_{R} - M) \]  

It must be noted that the operations in (40) and (41) are performed across \( c \) transmit antenna pairs and \( N_{m}^{2} \) MAP combination.

Thirdly, the computational complexity employed to perform an exhaustive ML search across the \( f_{1} \) and \( f_{2} \) most-likely candidate sets is similar to the computational complexity given in (39), however, instead of the \( M^{2} \) symbol search, it reduces to \( f_{1}f_{2} \) symbol search. Hence, the computational complexity involved in searching the \( f_{1} \) and \( f_{2} \) symbol sets becomes:

\[ \Delta_{\text{phase3}} = 10cN_{f_{1}}f_{2}N_{m}^{2} \]  

The total computational complexity for the low-complexity detector is the sum of the computational complexity for the three phases, viz; \( \Delta_{\text{phase1}}, \Delta_{\text{phase2}} \) and \( \Delta_{\text{phase3}} \). The computational complexity is given as:

\[ 2cN_{m}^{2} \left( 4MN_{R}^{2} + 8MN_{R} + 2N_{R} - 2M + 4N_{R}^{2} + 6N_{R} + 5N_{f_{1}}f_{2} - 2 \right) \]  

A summary of the computational complexities of the ML detector (MLD) and the low-complexity detector (LCD) is presented in Table 3. The computational complexities are calculated for \( M = 16 \) and 64, while in both cases, arbitrary values are chosen for \( f_{1} \) and \( f_{2} \), which offer very close error performance to the MLD. Furthermore, the percentage reduction in computational complexity, in terms of the number of complex multiplication, of the LC detector is compared with the ML detector, for a specified spectral efficiency (SE).
TABLE 3. Comparison of MBSTBC-SM-LD computational complexity for ML with low-complexity detector for $N_T = 4$ and $c = 4$

| CONFIGURATION | SE | ML | LCD | Reduction |
|---------------|----|----|-----|-----------|
| 16-QAM        |    | 7  | 655,360 | 296,704 | 54.72% |
| $N_R = 4$, $m_{rf} = 2$, $f_1 = 6$, $f_2 = 6$ |    | 8  | 1,310,720 | 350,400 | 73.27% |

VII. NUMERICAL RESULTS

In this section, the numerical results of the Monte-Carlo simulations for MBSTBC-SM and the proposed MBSTBC-SM-LD are compared for $M = 16$ and 64, employing Scheme 1 and 3. Furthermore, the numerical results of the reduced computational complexity detector which offer very close error performance with the optimal-ML detector over a fast, frequency-flat Rayleigh fading channel are presented. For our simulation, we assume that the channel is fully known by the receiver.

In Figure 5, the average bit error rate (BER) performances of the different schemes of MBSTBC-SM-LD are compared with the BER performances of Scheme 1 and 3 of MBSTBC-SM employing $m_{rf} = 2$ RF mirrors, $M = 16$-QAM for Scheme 1 and 3. The notation (Scheme, $N_T, N_R, m_{rf}, M$) has been employed for Figure 5 and 6, where $N_T$ is the number of transmit antennas, $N_R$ is the number of receive antennas, $m_{rf}$ is the number of RF mirrors and $M$ is the constellation size of the graycoded $M$-ary QAM constellation. The legend for the numerical values of the union-bound on the ABEP of the MBSTBC-SM-LD system is given directly below the respective MBSTBC-SM-LD as “Theory” for Figures 5 and 6.

Considering Figure 5, the theoretical expression, on the union-bound of the ABEP of MBSTBC-SM-LD given in (17) is evaluated and is compared with the simulated results. As expected, the theoretical values of the proposed MBSTBC-SM-LD demonstrates a tight match with the simulated ABEP at higher SNR. Furthermore, the proposed MBSTBC-SM-LD demonstrates a superior error performance over MBSTBC-SM. For example, the Scheme 1 of MBSTBC-SM-LD outperforms Scheme 1 and Scheme 3 of MBSTBC-SM by 1.5 and 2 dB gain, respectively, when the BER is $10^{-5}$. Furthermore, Scheme 3 of the proposed MBSTBC-SM-LD outperforms Scheme 1 and Scheme 3 of MBSTBC-SM by 1 and 1.5 dB, respectively.

Comparing the different MBSTBC-SM-LD schemes, Scheme 1 of MBSTBC-SM-LD demonstrates a better performance than Scheme 3. For example, Scheme 1 shows a 0.5 dB gain over Scheme 3 of MBSTBC-SM-LD.

In Figures 7 and 8, the BER performances of the optimal ML detector for MBSTBC-SM-LD are compared with the performance of the low-complexity detector for 16-QAM. For example, at a BER of $10^{-5}$, the difference in BER between Scheme 1 and Scheme 3 is negligible as there is a tight match between both schemes. Furthermore, the MBSTBC-SM-LD scheme demonstrate significant improvement over MBSTBC-SM, for example, the MBSTBC-SM-LD schemes outperforms the MBSTBC-SM schemes by $\approx 2.5$ dB gain in SNR, when the BER is $10^{-5}$.

In Figures 7 and 8, the BER performances of the optimal ML detector for MBSTBC-SM-LD are compared with the
BER performances of the low-complexity near-ML detector for MBSTBC-SM-LD. Similar to Figure 5 and 6, the notation (Scheme, \(N_T, N_R, m_{sf}, M\)) has been employed for Figure 7 and 8 also, where \(N_T\) is the number of transmit antennas, \(N_R\) is the number of receive antennas, \(m_{sf}\) is the number of RF mirrors and \(M\) is the constellation size of the gray-coded \(M\)-ary QAM constellation. Furthermore, the corresponding low-complexity near-ML detector of MBSTBC-SM-LD, which employs \(f_1\) and \(f_2\) most likely candidates, is given directly below the individual optimal ML detector of MBSTBC-SM-LD. From the different graphs in Figures 7 and 8, the low-complexity detector shows a tight match with its counterpart optimal near-ML detector with high reduction in computational complexity under the same channel condition.

**VIII. CONCLUSION**

In this paper, we have proposed the application of labelling diversity to MBSTBC-SM. Two schemes were proposed for MBSTBC-SM-LD, viz., Scheme 1 and Scheme 3. Both schemes of MBSTBC-SM-LD have demonstrated significant improvement over MBSTBC-SM. Furthermore, a theoretical expression for the union-bound on the ABEP of MBSTBC-SM-LD was formulated and agrees well with the numerical values of the Monte Carlo simulations for MBSTBC-SM-LD. Finally, a low-complexity near-ML detector which is based on orthogonal projection was proposed. A 73% reduction in computational complexity is achieved by the low-complexity detector when compared to the optimal ML detector of MBSTBC-SM-LD, especially at high SNR for \(M = 64\)-QAM.

**REFERENCES**

[1] S. M. Alamouti, “A simple transmit diversity technique for wireless communications,” *IEEE J. Sel. Areas Commun.*, vol. 16, no. 8, pp. 1451–1458, Oct. 1998.

[2] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, “Space-time block coding for wireless communications: Performance results,” *IEEE J. Sel. Areas Commun.*, vol. 17, no. 3, pp. 451–460, Mar. 1999.

[3] R. Y. Mesleh, H. Haas, S. Sinanovic, C. W. Ahn, and S. Yun, “Spatial modulation,” *IEEE Trans. Veh. Technol.*, vol. 57, no. 4, pp. 2228–2241, Jul. 2008.

[4] E. Basar, U. Aygolu, E. Panayirci, and H. V. Poor, “Space-time block coded spatial modulation,” *IEEE Trans. Commun.*, vol. 59, no. 3, pp. 823–832, Mar. 2011.

[5] K. Govindasamy, H. Xu, and N. Pillay, “Space-time block coded spatial modulation with labeling diversity,” *Int. J. Commun. Syst.*, vol. 31, no. 1, p. e395, Jan. 2018.

[6] N. Pillay and H. Xu, “Uncoded space-time labeling diversity—Application of media-based modulation with RF mirrors,” *IEEE Commun. Lett.*, vol. 22, no. 2, pp. 272–275, Feb. 2018.

[7] E. Seifi, M. Atamanesh, and A. K. Khandani, “Media-based MIMO: A new frontier in wireless communications,” 2015, arXiv:1507.07516. [Online]. Available: http://arxiv.org/abs/1507.07516

[8] B. S. Adejumobi, N. Pillay, and S. H. Mneney, “A study of spatial media-based modulation using RF mirrors,” in *Proc. IEEE AFRICON*, Sep. 2017, pp. 336–341.

[9] B. S. Adejumobi and N. Pillay, “RF mirror media-based space-time block coded spatial modulation techniques for two time-slots,” *IET Commun.*, vol. 13, no. 15, pp. 2313–2321, Sep. 2019.

[10] Y. Naresh and A. Chockalingam, “On media-based modulation using RF mirrors,” *IEEE Trans. Veh. Technol.*, vol. 66, no. 6, pp. 4967–4983, Jun. 2017.

[11] E. Basar and I. Altunbas, “Space-time channel modulation,” *IEEE Trans. Veh. Technol.*, vol. 66, no. 8, pp. 7609–7614, Aug. 2017.

[12] H. Xu, K. Govindasamy, and N. Pillay, “Uncoded space-time labeling diversity,” *IEEE Commun. Lett.*, vol. 20, no. 8, pp. 1511–1514, Aug. 2016.

[13] I. Al-Shahrani, “Performance of M-QAM over generalized mobile fading channels using MRC diversity,” M.S. thesis, Dept. Elect. Eng. College Eng., King Saud Univ., Riyadh, Saudi Arabia, 2007.

[14] S. Bahg, S. Shin, and Y.-O. Park, “ML approaching MIMO detection based on orthogonal projection,” *IEEE Commun. Lett.*, vol. 11, no. 6, pp. 474–476, Jun. 2007.

**BABATUNDE S. ADEJUMOBI** (Member, IEEE) received the B.Sc. and M.Sc. degrees in electronic and computer engineering from Lagos State University, Lagos, Nigeria, in 2007 and 2012, respectively, and the Ph.D. degree in electronic engineering (wireless communication) from the Department of Electrical, Electronic, and Computer Engineering, University of KwaZulu-Natal, Durban, South Africa. His current research includes spatial modulation, space-time block coded modulation and orthogonal frequency division multiplexing. However, he also has interest in optical communication, wireless communications, and image processing.

**THOKOZANI SHONGWE** (Member, IEEE) received the B.Eng. degree in electronic engineering from the University of Swaziland, Swaziland, in 2004, and the M.Eng. degree in telecommunications engineering from the University of the Witwatersrand, South Africa, in 2006, and the D.Eng. degree from the University of Johannesburg, South Africa, in 2014. He is currently an Associate Professor with the Department of Electrical and Electronic Engineering Technology, University of Johannesburg. He was a recipient of the 2014 University of Johannesburg Global Excellence Stature (GES) Award, which was awarded to him to carry out his postdoctoral research at the University of Johannesburg. In 2016, he was a recipient of the TWAS-DFG Cooperation Visits Program funding to do research in Germany. Other awards that he has received in the past are: the Post-Graduate Merit Award scholarship to pursue his master’s degree at the University of the Witwatersrand, in 2005, which is awarded on a merit basis. In the year 2012, he (and his co-authors) received an Award of the Best Student Paper at the IEEE ISPLC 2012 (power line communications conference) in Beijing, China. His research fields are in digital communications and error correcting coding. His research interests include power-line communications, cognitive radio, smart grid, and visible light communications.