Abstract—The use of optical spatial modulation (OSM), which has been recently emerged as a powerful and bandwidth-efficient pulsed modulation technique for indoor optical wireless communication, is proposed as a simple, low-complexity means of achieving spatial diversity in coherent free space optical (FSO) communication systems. In doing so, this paper makes several novel contributions as follows. It presents a generic analytical framework for obtaining the Average Bit Error Probability (ABEP) of uncoded OSM with coherent detection in the presence of turbulence-induced fading. Although the framework is general enough to accommodate any type of models based on turbulence scattering, the focus in this paper is the H-K distribution. Although this distribution represents a very general scattering model valid over a wide range of atmospheric conditions, it has not been considered in the past in conjunction with FSO systems possibly because of its mathematical complexity. The proposed analytical framework yields exact performance evaluation results for MIMO systems with two transmit and an arbitrary number of receive apertures. In addition, tight upper bounds are derived for the performance of OSM systems with an arbitrary number of transmit apertures as well as for convolutionally encoded signals. The performance of OSM is compared to that of well-established coherent FSO schemes, employing spatial diversity at the transmitter or the receiver only. Specifically, it is shown that OSM can offer comparable performance with conventional coherent FSO schemes while outperforming the latter in terms of spectral efficiency and hardware complexity. Various numerical performance evaluation results are also presented and compared with equivalent results obtained by Monte Carlo simulations which verify the accuracy of the derived analytical expressions.

Index Terms—average bit error probability, atmospheric turbulence, coherent detection, free space optical communication systems, H-K distribution, multiple-input multiple-output (MIMO) systems, optical spatial modulation.

I. INTRODUCTION

Free-space optical (FSO) communication systems have recently attracted great attention within the research community as well as for commercial use. FSO systems can provide ultra-high data rates (at the order of multiple gigabits per second), immunity to electromagnetic interference, excellent security and large unlicensed bandwidth i.e. hundred and thousand times higher than radio-frequency (RF) systems, along with low installation and operational cost [2].

K. P. Peppas is the Department of Informatics and Telecommunications, University of Peloponnese, 22100 Tripoli, Greece. He is also with with the Institute of Informatics and Telecommunications, National Centre for Scientific Research—“Demokritos,” Patriarhou Grigorou and Neapoleos, 15310 Agia Paraskevi, Athens, Greece (e-mail: kpeppas@iit.demokritos.gr).

P. T. Mathiopoulos is with the Department of Informatics and Telecommunications, National and Kapodistrian University of Athens, 15784 Zografou, Athens, Greece (e-mail: mathio@di.uoa.gr).

The challenge in employing such systems is that FSO links are highly vulnerable due to the detrimental effects of attenuation under adverse weather conditions (e.g., fog), pointing errors and atmospheric turbulence. One method to improve the reliability of the FSO link is to employ spatial diversity, i.e., multiple-lasers and multiple-apertures to create a multiple-input multiple-output (MIMO) optical channel. Because of its low complexity, spatial diversity is a particularly attractive fading mitigation technique and performance enhancements have been extensively studied in many past research works in the field of FSO communications [3]–[6].

In order to evaluate the impact of atmospheric turbulence on the performance of OSM, accurate models for the fading distribution are necessary. For example the lognormal distribution is often used to model weak turbulence conditions whereas the negative exponential and the K-distribution are used to model strong turbulence conditions [7]. Other more general statistical models have also been proposed to model scintillation over all turbulence conditions, including the Gamma-Gamma [8], the lognormal-Rice (or Beckmann) [9] the homodyned K distribution (H-K) [10] and the I-K [11]–[13] distributions. All these three models are based on the argument that scintillation is a doubly stochastic random process modeling both small and large scale turbulence effects. Besides, they agree well with measurement data and simulations for a wide range of turbulence conditions.

In this paper, the H-K distribution is adopted to model turbulence-induced fading. The main reason for this choice is the fact that this distribution is based on a very general scattering model which is valid for a wide range of atmospheric conditions. It is also noted that the H-K distribution generalizes existing models such as the K-distribution. The H-K distribution models the field of the optical wave as the sum of a deterministic component and a random component, the intensity of which follows the Rice (Nakagami-\(n\)) distribution. The average intensity of the random portion of the field is treated as a fluctuating quantity [14]. It is important to underline that, to the best of our knowledge, in the open technical literature there have been no papers published analyzing and evaluating the performance of FSO systems over such channels, because of the complicated mathematical form of their respective probability density functions (PDF).

Depending on their detection, FSO systems can be classified into two main categories, namely coherent (heterodyne detection) and non-coherent (direct detection) systems. Coherent FSO systems have the information bits encoded directly onto
the electric field of the optical beam. At the receiver, a local oscillator (LO) is employed to extract the information encoded on the optical carrier electric field. On the one hand, coherent FSO systems can provide significant performance enhancements due to spatial temporal selectivity and heterodyne gain in comparison to direct detection systems. Moreover, they are more versatile as any kind of amplitude, frequency, or phase modulation can be employed. On the other hand, coherent receivers are more difficult to implement as the LO field should be spatially and temporally coherent with the received field.

Recently, the so-called optical spatial modulation (OSM) has emerged as a power- and bandwidth-efficient single-carrier transmission technique for optical wireless communication systems [14]–[16]. This spatial diversity scheme, initially proposed in [17] and further investigated in [18], [19], employs a simple modulation mechanism that foresees to activate just one out of several MIMO transmitters at any time instant and to use the index of the activated transmitter as an additional dimension for conveying implicit information. It has been shown that OSM can increase the data rate by base two logarithm of the number of transmit units [14]. Also, OSM can increase the data rate by by a factor of 2 and 4, respectively, as compared to on-off keying (OOK) and pulse position modulation (PPM) [14], [15]. It is underlined that such performance gains are obtained with a significant reduction in receiver complexity and system design.

Because of the above mentioned advantages of OSM over other more conventional transmission schemes and given the wide applicability of FSO, it is of interest to investigate the potential performance enhancements obtained by incorporating OSM in FSO systems. However, in general this research topic has not been dealt within our research community. Only recently, there have been papers published in the open technical literature dealing with performance analysis studies of FSO systems employing spatial modulation and operating in the presence of atmospheric turbulence, e.g. see [20] and [21]. Specifically, in [20], the combination of subcarrier intensity modulation and spatial modulation with receiver diversity was proposed to enhance the performance of intensity modulated direct detection (IM/DD) FSO systems. In [21], another IM/DD based system FSO system which combines antenna shift keying with joint pulse position and amplitude modulations was considered. For this system, which was denoted as spatial pulse position and amplitude modulation (SPPAM), the atmospheric turbulence channel was modeled as log-normal or Gamma-Gamma distributions and was evaluated, in terms of bounds, for uncoded and coded signals. ABEP performance evaluation results have shown that SPPAM offers a compromise between spectral and power efficiencies as well as a certain degree of robustness against atmospheric turbulence. Despite these two papers which deal with non-coherent detection schemes, the potential enhancements of OSM on the performance of FSO systems with coherent detection still remains an open research topic which, to the best of our knowledge, has not been addressed so far in the open technical literature.

Motivated by the above, in this paper we present for the first time a generic analytical framework which can be used to accurately obtain the performance of outdoor OSM with coherent detection in the presence of turbulence-induced fading. More specifically and within this novel analytical framework, the main novel research contributions of the paper are as follows:

| TABLE I |
| --- |
| **LIST OF MATHEMATICAL NOTATIONS** |

\[ f^2 = -1 \] denotes the imaginary unit
\[ |z| \] denotes the magnitude of the complex number \( z \)
\[ \Re\{z\} \] denotes the real part of the complex number \( z \)
\[ \Im\{z\} \] denotes the imaginary part of the complex number \( z \)
\[ f(x) = o(g(x)) \quad \text{as} \quad x \rightarrow x_0 \quad \text{if} \quad \lim_{x \rightarrow x_0} \frac{f(x)}{g(x)} = 0 \]
\[ \| \cdot \|_F^2 \] denotes the square Frobenius norm
\[ (\cdot)^T \] denotes the matrix transpose
\[ * \] denotes convolution
\[ \mathbb{E}\{\cdot\} \] denotes expectation
\[ f_X(\cdot) \] denotes the Probability Density Function (PDF) of the random variable \( X \)
\[ F_X(\cdot) \] denotes the Cumulative Distribution Function (CDF) of the random variable \( X \)
\[ M_X(\cdot) \] denotes the Moment Generating Function (MGF) of the random variable \( X \)
\[ I_\alpha(\cdot) \] is the modified Bessel function of the first kind and order \( \alpha \) \[ (8.431) \]
\[ K_\alpha(\cdot) \] is the modified Bessel function of the second kind and order \( \alpha \) \[ (8.432) \]
\[ \Gamma(x) = \int_0^\infty \exp(-t)t^{x-1}dt \] is the Gamma function \[ (8.310/1) \]
\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp(-t^2/2)dt \] is the Gauss Q-function
\[ W_{P,Q}(\cdot) \] is the Whittaker function \[ (9.220) \]
\[ Pr\{\cdot\} \] denotes the probability operator
\[ \hat{\cdot} \] denotes estimated value at the receiver side
New analytical expressions for the ABEP of coherent OSM under turbulence conditions modeled by the H-K distribution are derived. When the transmitter is equipped with two apertures the resulting analytical expressions are exact, whereas for an arbitrary number of transmit apertures tight upperbounds can be obtained.

Error probability performance bounds for coded OSM systems are derived and the performance enhancements when channel coding is employed are presented and analyzed.

The error probability performance of OSM is also compared to that of conventional FSO schemes with transmit or receive diversity only, i.e. when Maximal Ratio Combining (MRC), Selection Combining (SC) or Alamouti-type Space-Time Block Codes (STBC) are employed. It is noted that the theoretical analysis is substantiated by comparing the theoretical and equivalent simulated performance evaluation results obtained by means of Monte Carlo techniques.

The paper is organized as follows. After this introduction, Section II outlines the system and channel models. In Section III analytical expressions for the ABEP of uncoded OSM systems are presented. Asymptotic ABEP expressions are also derived, wherefrom the diversity gain of coherent OSM can be readily deduced. The performance of coded OSM systems is discussed in Section IV. In Section V the various performance evaluation results and their interpretations as well as comparisons are presented. Finally, concluding remarks can be found in Section VI.

II. SYSTEM AND CHANNEL MODEL

In this section, a detailed description of the OSM FSO system model, i.e. transmitter, channel and receiver is provided. Moreover, the H-K distribution is introduced and analytical expressions for its parameters in terms of equivalent physical parameters of the turbulence phenomenon, such as the refractive-index structure parameter, optical wave number, and propagation path length, are derived.

A. Preliminaries

Let us consider a $M \times N$ MIMO FSO system with $M$ transmit units (lasers) and $N$ coherent receivers. It is assumed that the receiving apertures are separated by more than a coherence wavelength to ensure the independency of fading channels. The basic principle of OSM modulation is as follows:

i) The transmitter encodes blocks of $\log_2(M)$ data bits into the index of a single transmit unit. Such a block of bits is hereafter referred to as “message” and is denoted by $b_m$, $\forall m = 1,2,\ldots,M$. It is assumed that the $M$ messages are transmitted with equal probability by the encoder and that the related transmitted signal is denoted by $E_m = E_m \exp(j\phi_{b_m})$. During each time slot, only one transmitter $\ell$, where $\ell = 1,2,\ldots,M$ is active for data transmission. The information bits are modulated on the electric field of an optical signal beam through an external modulator. During this particular time slot, the remaining transmit lasers are kept silent, i.e. they do not transmit.

ii) At the receiver, the incoming optical field is mixed with a local oscillator (LO) field and the combined wave is first converted by the photodetector to an electrical one. A bandpass filter is then employed to extract the intermediate frequency (IF) component of the total output current. Finally, a $N$-hypothesis detection problem is solved to retrieve the active transmit unit index, which results in the estimation of the unique sequence of bits emitted by the transmitter.

B. Receiver Structure

The received electric field at the aperture plane of the $n$-th receiver after mixing with a LO beam, can be expressed as

$$e_n(t) = \sqrt{2P_0 Z_0 E_m h_{m,n} \cos(\omega_0 t + \phi_{m,n} + \phi_{b_m})} + \sqrt{2P_{LO} Z_0 \cos(\omega_{LO} t)}.$$  (1)

In the above equation, $P_t$ is the transmit laser power, $Z_0$ is the free space impedance, $h_{m,n}$ and $\phi_{m,n}$ denote the magnitude and the phase of the complex channel coefficient between the $m$-th transmit and the $n$-th receive aperture, respectively. Furthermore, $P_{LO}$ denotes the power of the local oscillator, $\omega_{LO} = \omega_0 + \omega_{IF}$ where $\omega_0$ and $\omega_{IF}$ are the carrier and the intermediate radian frequencies, respectively.

The output current of the $n$-th photodetector can be mathematically expressed as

$$i_n(t) = \frac{R}{Z_0} |e_n(t)|^2$$  (2)

where $R = \eta q_e / (h \nu_0)$ is the responsivity of the photodetector with $q_e = 1.6 \times 10^{-19}$Cb is the charge of an electron, $h = 6.6 \times 10^{-34}$J·s is the Planck constant, $\eta$ is the photodetector efficiency, and $\nu_0 = \omega_0 / (2\pi)$ is the optical center frequency. Expanding (2) and ignoring the double-frequency terms that are filtered out by the bandpass filter, the resulting photocurrent can be expressed as

$$i_n(t) = R P_t E_m^2 h_{m,n}^2 + R P_{LO}$$
$$+ 2R \sqrt{P_t P_{LO} E_m h_{m,n}} \cos(\omega_{IF} t - \phi_{m,n} - \phi_{b_m})$$
$$\triangleq i_{DC}(t) + i_{AC}(t).$$  (3)

In (3), $i_{DC}(t) \triangleq R P_t E_m^2 h_{m,n}^2 + R P_{LO}$ is the DC component generated by the signal and local oscillator fields, respectively, $i_{AC}(t) \triangleq 2R \sqrt{P_t P_{LO}} \cos(\omega_{IF} t - \phi_{m,n} - \phi_{b_m})$ is the AC component in the received photocurrent which, unlike for direct detection, contains information about the frequency and phase of the received signal. It is assumed that for coherent detection the intermediate frequency $\omega_{IF}$ is nonzero, so that the signal power can be expressed as $P_s = 2R^2 P_t P_{LO} E_m^2 h_{m,n}^2$.

As in [22]–[25], we also consider that $P_{LO} \gg P_t$ and thus, the DC photocurrent can be approximated as $i_{DC}(t) \approx R P_{LO}$. The photodetection process is impaired by shot noise with variance $\sigma_{shot,L}^2 = 2q_e R P_{LO} B_e$ where $B_e$ is the electrical bandwidth of the photodetector. It is also noted that because of the large value of $R P_{LO}$ the photocurrent due to thermal noise and the dark current can be ignored [22].
Following [23] and [24], the sufficient statistics at the $n$-th coherent receiver can be expressed as
\[
y_n = \sqrt{P} h_{m,n} E_m \exp(j[\phi_{m,n} + \phi_{b_m}]) + z_n
\]  
(4)
where $\mu = RP/(q_i B_e)$ is the average signal-to-noise ratio (SNR) and $z_n$ is the noise at the $n$-th receiver. Assuming that the LO power is large and the receiver noise is dominated by LO related noise terms, the Additive White Gaussian Noise (AWGN) model can be employed as an accurate approximation of the Poisson photon-counting detection model [22], [23]. Thus, $z_n$ can be modeled as a zero-mean unit variance complex Gaussian random variable [23].

Similar to [26], it is assumed that the receiver has knowledge of the actual fading gains and that the total fading remains constant over one bit interval and changes from one interval to another in an independent manner. At the receiver, the optimal spatial modulation detector estimates the active transmitter constant over one bit interval and changes from one interval to another. Thus, $P$ is a Gaussian random variable [23].

\[
\begin{align*}
\mathbb{E}\{y_n\} &= \mathbb{E}\{h_{m,n} E_m \exp(j[\phi_{m,n} + \phi_{b_m}]) + z_n\} \\
&= \sqrt{P} h_{m,n} E_m \mathbb{E}\{\exp(j[\phi_{m,n} + \phi_{b_m}])\} + \mathbb{E}\{z_n\}
\end{align*}
\]  
(6)

\[
\begin{align*}
\mathbb{E}\{y_n^2\} &= \mathbb{E}\{h_{m,n}^2 E_m^2 \exp(2j[\phi_{m,n} + \phi_{b_m}]) + 2z_n z_n^* + 2z_n h_{m,n} E_m \exp(j[\phi_{m,n} + \phi_{b_m}]) + z_n^2\} \\
&= \mathbb{E}\{h_{m,n}^2 E_m^2 \exp(2j[\phi_{m,n} + \phi_{b_m}])\} + \mathbb{E}\{z_n^2\}
\end{align*}
\]  
(7)

\[
\begin{align*}
\mathbb{E}\{y_n^4\} &= \mathbb{E}\{h_{m,n}^4 E_m^4 \exp(4j[\phi_{m,n} + \phi_{b_m}]) + 4z_n z_n^* z_n^2 + 4z_n^2 h_{m,n} E_m \exp(2j[\phi_{m,n} + \phi_{b_m}]) + 2z_n^4\} \\
&= \mathbb{E}\{h_{m,n}^4 E_m^4 \exp(4j[\phi_{m,n} + \phi_{b_m}])\} + 4 \mathbb{E}\{z_n^2\}^2 + 2 \mathbb{E}\{z_n^4\}
\end{align*}
\]  
(8)

\[
\begin{align*}
\mathbb{E}\{y_n^6\} &= \mathbb{E}\{h_{m,n}^6 E_m^6 \exp(6j[\phi_{m,n} + \phi_{b_m}]) + 6z_n z_n^* z_n^2 z_n^3 + 6z_n^2 h_{m,n} E_m \exp(3j[\phi_{m,n} + \phi_{b_m}]) + 3z_n^4 h_{m,n} E_m \exp(j[\phi_{m,n} + \phi_{b_m}]) + z_n^6\} \\
&= \mathbb{E}\{h_{m,n}^6 E_m^6 \exp(6j[\phi_{m,n} + \phi_{b_m}])\} + 6 \mathbb{E}\{z_n^2\}^3 + 6 \mathbb{E}\{z_n^4\} \mathbb{E}\{z_n^2\} + 3 \mathbb{E}\{z_n^6\}
\end{align*}
\]  
(9)

\[
\mathbb{E}\{y_n^8\} = \mathbb{E}\{h_{m,n}^8 E_m^8 \exp(8j[\phi_{m,n} + \phi_{b_m}]) + 8z_n z_n^* z_n^2 z_n^4 + 8z_n^2 h_{m,n} E_m \exp(4j[\phi_{m,n} + \phi_{b_m}]) + 4z_n^4 h_{m,n} E_m \exp(2j[\phi_{m,n} + \phi_{b_m}]) + z_n^8\} \\
\]  
(10)

\[
\mathbb{E}\{y_n^{10}\} = \mathbb{E}\{h_{m,n}^{10} E_m^{10} \exp(10j[\phi_{m,n} + \phi_{b_m}]) + 10z_n z_n^* z_n^3 z_n^5 + 10z_n^2 h_{m,n} E_m \exp(5j[\phi_{m,n} + \phi_{b_m}]) + 5z_n^4 h_{m,n} E_m \exp(3j[\phi_{m,n} + \phi_{b_m}]) + z_n^{10}\} \\
\]  
(11)

\[
\mathbb{E}\{y_n^{12}\} = \mathbb{E}\{h_{m,n}^{12} E_m^{12} \exp(12j[\phi_{m,n} + \phi_{b_m}]) + 12z_n z_n^* z_n^4 z_n^7 + 12z_n^2 h_{m,n} E_m \exp(6j[\phi_{m,n} + \phi_{b_m}]) + 6z_n^4 h_{m,n} E_m \exp(3j[\phi_{m,n} + \phi_{b_m}]) + z_n^{12}\} \\
\]  
(12)

The $\mathbb{E}$-th normalized moment of $y_n^m$ is given by [10] as
\[
\frac{\mathbb{E}\{y_n^m\}}{(\mathbb{E}\{y_n^2\})^{m/2}} = \frac{\alpha_{m/2}^m}{\alpha_{m/2}} \prod_{k=0}^{m-1} \left( \frac{\nu}{\nu - k} \right) \frac{\Gamma(\alpha_{m/2} + \nu - k)}{\Gamma(\alpha_{m/2})} \frac{\mathbb{E}\{y_n^2\}}{\mathbb{E}\{y_n^2\}}.
\]  
(13)

The $\nu$-th normalized moment of $y_n^m$ is given by [10] as
\[
\mathbb{E}\{y_n^m\} = \frac{\alpha_{m/2}^m}{\alpha_{m/2}} \prod_{k=0}^{m-1} \left( \frac{\nu}{\nu - k} \right) \frac{\Gamma(\alpha_{m/2} + \nu - k)}{\Gamma(\alpha_{m/2})} \frac{\mathbb{E}\{y_n^2\}}{\mathbb{E}\{y_n^2\}}.
\]  
(14)

Using (10), the scintillation index can be readily calculated as
\[
\sigma_{b_i}^2 = \frac{\mathbb{E}\{y_n^2\}}{(\mathbb{E}\{y_n^2\})^2} - 1 = \frac{\alpha_{m/2} + 2\alpha_{m/2}\rho_{b_i} + 2}{\alpha_{m/2} + 1 + \rho_{b_i}^2}. \quad \sigma_{b_i} \ll 1
\]  
(15)

Under the assumption of spherical wave propagation, $\sigma_{b_i}^2$ can be directly related to atmospheric conditions as [12] Eq. (7), Eq. (9)]
\[
\sigma_{b_i}^2 \approx \frac{0.41\alpha_{m/2}^2(1 + 0.5\alpha_{m/2}^2)}{1 + 2.8/\sigma_1^{4/5}}, \quad \sigma_1 \gg 1
\]  
(16)

where $\sigma_1 = 1.23C_{n_i}^2 k^{7/6} L_{ij}^{11/6}$ is the Rytov variance, $k = 2\pi/\lambda$ is the optical wave number with $\lambda$ being the wavelength, $L_{ij}$ is the link distance and $C_{n_i}$ denotes the index of refraction structure parameter. For FSO links near the ground, $C_{n_i}^2 \approx 1.7 \times 10^{-14}$m$^{-2/3}$ and $8.4 \times 10^{-15}$m$^{-2/3}$ for the daytime and night, respectively [23]. Moreover, $\sigma_1 \ll 1$ and $\sigma_1 \gg 1$ correspond to weak and strong turbulence conditions, respectively.

Using (12), the parameters of the H-K distribution, $\alpha$ and $\rho$, can be directly related to physical parameters of the turbulence by following a similar line of arguments as in [12], where similar results were derived for the I-K distribution. In particular, on the one hand, weak turbulence conditions are characterized in the H-K distribution by large values of
\( \rho_{ij} \). In this case the scintillation index given by (11) can be approximated as

\[
\sigma_{ij}^2 \approx \frac{2}{\rho_{ij}}, \quad \text{with } \rho_{ij} \gg 1.
\]  

(13)

On the other hand, assuming strong turbulence conditions where \( \rho_{ij} \) tends to zero, (11) can be approximated as

\[
\sigma_{ij}^2 \approx 1 + \frac{2}{\rho_{ij}}, \quad \text{with } \rho_{ij} \ll 1.
\]  

(14)

By comparing (13) and (14) with the first and second branches of (12), respectively, \( \alpha_{ij} \) and \( \rho_{ij} \) can be obtained as

\[
\alpha_{ij} = 0.71\sigma_{ij}^{1/5}
\]  

(15)

\[
\rho_{ij} = \frac{4.88}{\sigma_{ij}^2(1 + 0.2\sigma_{ij}^1)}.
\]  

(16)

To the best of our knowledge, the relationship of \( \alpha_{ij} \) and \( \rho_{ij} \) with \( \sigma_{ij} \) given by (15) and (16) is a novel result.

### III. PERFORMANCE ANALYSIS OF UNCoded OSM

In this section, by employing the well-known MGF-based approach for the performance analysis of digital communications over fading channels [29], analytical expressions for the ABEP of uncoded OSM systems will be derived. Expressions for the diversity and coding gains of OSM systems are also presented, thus providing useful insight as to how these parameters affect the overall system performance.

#### A. Preliminaries

For \( M = 2 \), the conditional bit error probability (BEP) of OSM systems when no turbulence induced fading is considered can be obtained in closed form as [26]

\[
P_{E}(h_1, h_2) = Q \left( \sqrt{\frac{\mu}{4}} \| h_1 - h_2 \|_F^2 \right).
\]  

(17)

The squared Frobenius norm in (17) can be expressed as

\[
\| h_1 - h_2 \|_F^2 = \sum_{n=0}^{N} |h_{1,n} - h_{2,n}|^2
\]  

(18)

where \( h_{1,n} \) is the \( n \)-th element of \( h_i \), \( i \in \{1, 2\} \). When \( M > 2 \) transmitters are considered, a tight upper bound for the conditional BEP of the above system can be obtained as [14, Eq. (7)]

\[
P_{E}(H) \leq \frac{M^{-1}}{\log_2(M)} \times \sum_{m_1=1}^{M} \sum_{m_2=m_1+1}^{M} N_0(m_1, m_2) \text{PEP}(m_1 \to m_2)
\]  

(19)

where \( \text{PEP}(m_1 \to m_2) \) denotes the pairwise error probability (PEP) related to the pair of transmitters \( m_1 \) and \( m_2 \), where \( m_1 \) and \( m_2 \in \{1, 2, \ldots, M\} \), and \( N_0(m_1, m_2) \) is the number of bit which have occurred when the receiver decides incorrectly that \( m_2 \) instead of \( m_1 \) has been active. The \( \text{PEP}(m_1 \to m_2) \) can be evaluated as [14, Eq. (8)]

\[
\text{PEP}(m_1 \to m_2) = Q \left( \sqrt{\frac{\mu}{4}} \| h_{m_1} - h_{m_2} \|_F^2 \right).
\]  

(20)

#### B. MGF-Based Approach

When atmospheric turbulence is taken into account, the conditional error probabilities in (17) and (19) need to be averaged over the elements of the channel matrix \( H \) in order to evaluate the ABEP. Without loss of generality, let us consider the case of a \( 2 \times N \) MIMO system. Since \( h_{1,n} \) are complex Gaussian random variables, the difference \( \Delta_n \equiv h_{1,n} - h_{2,n} \) is a complex Gaussian random variable having mean equal to the difference of the means of \( h_{1,n} \) and variance equal to the sum of variances of \( h_{1,n} \). In order to deduce a closed form expression for the ABEP, it is further assumed that \( h_{1,n} \) have uncorrelated real and imaginary components with the same variance \( \sigma_n^2 = b_n/2 \). It is noted that such an assumption is justified for link distances of the order of km and for aperture separation distances of the order of cm [30, 31]. For example, in [31] it was reported that for a link distance of 1.5 km, a wavelength of 1550 nm, an aperture diameter of 1 mm and photodetectors separated by as little as 35 mm, which validates the independence assumption.

Consequently, \( \Delta_n \) has uncorrelated components too and its squared envelope, \( |\Delta_n|^2 \), is characterized by a non-central chi-square PDF as follows

\[
f_{|\Delta_n|^2}(x|b_n) = \frac{1}{2b_n} \exp \left( -\frac{x + A_n^2}{2b_n} \right) I_0 \left( \frac{A_n \sqrt{x}}{b_n} \right)
\]  

(21)

where \( A_n = |A_{2,n}e^{j\theta_{2,n}} - A_{1,n}e^{j\theta_{1,n}}| \). Assuming that \( b_n \) follows a gamma distribution with parameters \( \alpha_n \) and \( b_{0,n} \), the unconditional PDF of \( |\Delta_n|^2 \) is obtained by averaging (21) with respect to \( b_n \), i.e.

\[
f_{|\Delta_n|^2}(x) = \frac{(\alpha_n/b_{0,n})^{\alpha_n}}{2^\alpha_n \Gamma(\alpha_n)}
\]  

\[
\times \int_0^\infty b_n^{\alpha_n-2} \exp \left( -\frac{\alpha_n b_n}{b_{0,n}} - \frac{x + A_n^2}{2b_n}\right) I_0 \left( \frac{A_n \sqrt{x}}{b_n} \right) db_n.
\]  

(22)

As was pointed out in [11], the integral in (22) cannot be solved in closed form. Nevertheless, for the special case of \( \alpha_n = 1 \), i.e. when one scatterer per branch is considered, and by employing [11, Eq. (10)], this integral can be evaluated in closed form as

\[
f_{|\Delta_n|^2}(x) = \begin{cases} 
1 & x < A_n^2, \\
\frac{1}{b_{0,n}} K_0 \left( \sqrt{2A_n/b_{0,n}} \right) I_0 \left( \sqrt{2x/b_{0,n}} \right), \quad x < A_n^2, \\
\frac{1}{b_{0,n}} K_0 \left( \sqrt{2A_n/b_{0,n}} \right) K_0 \left( \sqrt{2x/b_{0,n}} \right), \quad x > A_n^2.
\end{cases}
\]  

(23)

Moreover, for the special case where \( h_{1,n} \) and \( h_{2,n} \) have identical mean value, i.e. when \( A_n = 0 \), (22) yields the well known K-distribution with PDF given by

\[
f_{|\Delta_n|^2}(x) = 2^{(1-\alpha_n)/2} \Gamma(\alpha_n) \left( \frac{\alpha_n x}{b_{0,n}} \right)^{(\alpha_n-1)/2}
\]  

\[
\times K_{\alpha_n-1} \left( \sqrt{2\alpha_n x/b_{0,n}} \right).
\]  

(24)

By employing the MGF-based approach for the performance analysis of digital communications over fading channels, the
average PEP (APEP) can be obtained as

$$\text{APEP} = \frac{1}{\pi} \int_0^{\pi/2} \prod_{n=1}^{N} \left[ |M|_{\Delta_n}|^2 \left( \frac{\mu}{8 \sin^2 \theta} \right) \right] d\theta.$$  

(25)

Moreover, using the tight approximation for the Gaussian Q-function presented in [32 Eq. (14)] (i.e., $Q(x) \approx 1/12 \exp(-x^2) + 1/4 \exp(-x^2/3)$), an expression accurately approximating APEP can be deduced as

$$\text{APEP} \approx \frac{1}{12} \prod_{n=1}^{N} \left[ |M|_{\Delta_n}|^2 \left( \frac{\mu}{8} \right) \right] + \frac{1}{4} \prod_{n=1}^{N} \left[ |M|_{\Delta_n}|^2 \left( \frac{\mu}{2} \right) \right].$$  

(26)

In the following analysis, analytical expressions for the MGF of $|\Delta_n|^2$ will be deduced. Specifically, the following result holds:

**Proposition 1.** An integral representation for the MGF of $|\Delta_n|^2$ can be deduced as

$$M_{|\Delta_n|^2}(s) = \frac{(\alpha_n/b_{0,n})^{\alpha_n}}{\Gamma(\alpha_n)} \times \int_0^{\infty} \frac{b^{\alpha_n-1}}{2 \pi s^{2}} \exp \left( -\frac{\hat{A}_n s}{2 b s + 1} - \alpha_n b_{0,n} \right) db.$$  

(27)

**Proof:** By employing the definition of the MGF, $M_{|\Delta_n|^2}(s)$ can be obtained as

$$M_{|\Delta_n|^2}(s) = \int_0^{\infty} \exp(-s x) f_{|\Delta_n|^2}(x) dx$$

$$= \frac{(\alpha_n/b_{0,n})^{\alpha_n}}{2 \Gamma(\alpha_n)} \int_0^{\infty} \exp \left( -\frac{\alpha_n b_{0,n} x}{2 b} - x + \hat{A}_n^2 \right) \times I_0 \left( \frac{\hat{A}_n \sqrt{x}}{b} \right) db.$$

(28)

By changing the order of integration, the above equation can be expressed as

$$M_{|\Delta_n|^2}(s) = \frac{(\alpha_n/b_{0,n})^{\alpha_n}}{2 \Gamma(\alpha_n)} \int_0^{\infty} b^{\alpha_n-2} \exp \left( -\frac{\alpha_n b_{0,n}}{b} \right) \times I_0 \left( \frac{\hat{A}_n \sqrt{x}}{b} \right) db \times dx.$$  

(29)

The inner integral, i.e. with respect to $x$, can be evaluated by employing [33 Eq. (3.15.2.2)] as

$$\int_0^{\infty} \exp \left( -\frac{s x - x + \hat{A}_n^2}{2 b} \right) I_0 \left( \frac{\hat{A}_n \sqrt{x}}{b} \right) dx = \frac{2 b}{2 s b + 1} \exp \left[ \frac{1}{2 A_n b(2 s b + 1)} \right].$$  

(30)

Substituting (30) into (29) and after some straightforward manipulations, (27) is readily deduced thus completing the mathematical proof.

The integral in (27) can be accurately approximated by employing a Gauss-Chebyshev Quadrature (GCQ) technique as [34]

$$M_{|\Delta_n|^2}(s) \approx (\alpha_n/b_{0,n})^{\alpha_n} \times \sum_{j=0}^{J} w_j \frac{\alpha_n-1}{2 t_j s + 1} \exp \left( -\frac{\hat{A}_n s}{2 t_j s + 1} - \alpha_n t_j \right)$$  

(31)

where $J$ is the number of integration points, $t_j$ are the abscissas and $w_j$ the corresponding weights. In [35 eqs. (22) and (23)], $t_j$ and $w_j$ are defined as

$$t_j = \tan \left[ \frac{\pi}{4} \cos \left( \frac{2 j - 1}{2 J} \pi \right) + \frac{\pi}{4} \right]$$

(32a)

$$w_j = \frac{\pi^2}{4 J \cos^2 \left( \frac{2 j - 1}{2 J} \pi \right) + \frac{\pi}{4}}.$$  

(32b)

For the special case of $\hat{A}_n = 0$, it can be shown that (27) can be evaluated in closed form. Specifically, the following result holds:

**Corollary 1.** For the special case of $\hat{A} = 0$ the MGF of $|\Delta_n|^2$ can be deduced in closed form as

$$M_{|\Delta_n|^2}(s) = \left( \frac{\alpha_n}{2 s b_{0,n}} \right)^{\alpha_n} \exp \left( \frac{\alpha_n}{4 s b_{0,n}} \right)$$

$$\times W_{-\frac{\alpha_n}{2 s b_{0,n}}} \frac{(\alpha_n/2 s b_{0,n})^{\alpha_n-1}}{\Gamma(\alpha_n)}.$$  

(33)

This result can be readily deduced by employing the integral representation of the Whittaker W-function given in [1 Eq. (9.222)]. Moreover it is worth pointing out that (33) is in agreement with a previously known result, namely the analytical expression for the MGF of the K-distribution. [36 Eq. (4)].

C. Analysis of the Diversity Gain

The diversity gain of the considered OSM MIMO system can be obtained by using the approach presented in [37]. In particular, a generic analytical expression, which becomes asymptotically tight at high SNR values, will be derived for the APEP appearing in (25), as follows:

**Proposition 2.** For high SNR values, [45] can be approximated by

$$\text{APEP} \approx 2^{N-1} \prod_{n=1}^{N} c_n \left( \frac{N}{4} \right)^{-N}$$  

(34)

where

$$c_n = \left( \frac{\hat{A}_n}{2} \right)^{\alpha_n-1} \frac{(\alpha_n/b_{0,n})^{\alpha_n}}{\Gamma(\alpha_n)} K_{\alpha_n-1} \left( \frac{2 A_n \alpha_n}{b_{0,n}} \right).$$  

(35)

**Proof:** According to [37 Proposition 3], the asymptotic error performance of the OSM system depends on the behavior of $M_{|\Delta_n|^2}(s)$, as $s \to \infty$. To determine an analytical asymptotic expression for APEP a Taylor series expansion is employed to approximate $M_{|\Delta_n|^2}(s)$ as

$$|M_{|\Delta_n|^2}(s)| = c_n |s|^{-\alpha_n} + o(|s|^{-\alpha_n}), \ s \to \infty$$  

(36)
where \(c_n\) and \(d_n\) are parameters that determine the diversity and coding gains of the \(n\)-th diversity branch, respectively. Observe that since \(\frac{b_s}{(2s+1)} \overset{s \to \infty}{\approx} \frac{1}{(2s+1)} \approx \frac{1}{2s}\), (27) yields
\[
\mathcal{M}_{|\Delta_n|^2}(s) \approx \left(\frac{\alpha_n/b_{0,n}}{2s\Gamma(\alpha_n)}\right)^{\alpha_n} \times \int_0^\infty b_n^{\alpha_n-2} \exp\left(-\frac{\bar{A}_n}{2b_n} - \frac{\alpha_n b_n}{b_{0,n}}\right) db_n.
\]
By employing \([33]\) Eq. (2.2.2.1), (37) can be solved in closed form yielding
\[
\mathcal{M}_{|\Delta_n|^2}(s) \approx \left(\frac{\hat{A}_n}{2}\right)^{\alpha_n-1} \left(\frac{\alpha_n/b_{0,n}}{s\Gamma(\alpha_n)}\right)^{\alpha_n} \times K_{\alpha_n-1} \left(\sqrt{\frac{2\hat{A}_n}{b_{0,n}}\alpha_n}\right).
\]
By comparing \([38]\) and \([36]\) it is readily deduced that \(d_n = 1\) and \(c_n\) is given by \([33]\). Thus, by substituting \([36]\) into \([25]\), the asymptotic PEP expression can be obtained as in \([34]\) which concludes the proof.

From \([34]\) it is clear that the diversity gain achieved by the considered system is equal to \(N\). It is also evident that the diversity gain depends only on the number of the receive apertures and is independent of the fading severity. This finding is in agreement with relevant findings reported in \([26]\) and \([38]\), for the case of radio-frequency MIMO wireless systems.

It is noted that for the special case \(\hat{A}_n = 0\), i.e. when \(|\Delta_n|^2\) follows the K-distribution, by employing the asymptotic result \(K_2(x) \overset{x \to 0}{\approx} (\Gamma(t)/2)(2/x)^t\) \([34]\), \(c_n\) can be further simplified as
\[
c_n = \frac{\alpha_n}{2b_{0,n}(\alpha_n - 1)}.
\]

**IV. Performance Analysis of Coded OSM over Turbulence Channels**

When coded OSM is employed, the input signal \(s(t)\) is first encoded by a convolutional encoder. The encoded data are interleaved by a random block interleaver and transmitted through the optical wireless channels using spatial modulation. It is also assumed that perfect interleaving at the transmitter and de-interleaving at the receiver is used. Assuming maximum likelihood soft decision decoding, the log likelihood ratios (LLRs) for the \(i\)-th constellation branch when the \(\ell\)-th transmitting antenna is active are computed as \([14]\) Eq. (6)]
\[
\text{LLR} = \log\frac{\Pr[\ell = 1 | \mathbf{y}]}{\Pr[\ell = 0 | \mathbf{y}]} = \log\frac{\sum_{\mathbf{h}_i \in \mathcal{L}_1} \exp\left(-\frac{1}{2N_0} \| \mathbf{y} - \mathbf{h}_i s_\ell \|^2\right)}{\sum_{\mathbf{h}_i \in \mathcal{L}_0} \exp\left(-\frac{1}{2N_0} \| \mathbf{y} - \mathbf{h}_i s_\ell \|^2\right)}
\]
where \(\mathcal{L} \in \{1 : M\}\) is the set of spatial constellation points, \(\mathcal{L}_1\) and \(\mathcal{L}_0\) are subsets from \(\mathcal{L}\) containing the transmitter indices having “1” and “0” at the \(i\)-th bit, respectively. The resulting data are finally decoded by a Viterbi decoder.

A union bound on the ABEP of a coded communication system can be evaluated as \([29]\]
\[
\tilde{P}_{ub} \leq \frac{1}{N} \sum_X P(X) \sum_{X' \neq X} q(X, X') \text{PEP}(X, X')
\]
where \(P(X)\) is the probability that the coded sequence \(X\) is transmitted, \(q(X, X')\) is the number of information bit errors in choosing another coded sequence \(X'\) instead of \(X\) \(n\) is the number of information bits per transmission and \(\text{PEP}(X, X')\) is the pairwise error probability, i.e. the probability of selecting \(X'\) when \(X\) was actually transmitted.

By employing \([29]\) p. 510, (41) can be efficiently evaluated as
\[
P_{ub} \leq \frac{1}{N} \sum_X P(X) \int_0^{\pi/2} \left|\frac{\partial}{\partial N} T[D(\theta), N]\right|_{N=1} d\theta
\]
where \(T[D(\theta), N]\) is the transfer function of the employed convolutional code, \(N\) is an indicator variable taking into account the number of the erroneous bits and \(D(\theta)\) depends on the underlying PEP expression. Furthermore, assuming that uniform error probability (UEP) codes are considered and taking into account the symmetry property this code family exhibits, thus making the distance structure of a UEP code independent of the transmitted sequence, \([42]\) can be further simplified as \([29]\)
\[
\tilde{P}_{ub} \leq \frac{1}{\pi} \int_0^{\pi/2} \left|\frac{1}{n} \frac{\partial}{\partial N} T[D(\theta), N]\right|_{N=1} d\theta.
\]
For \(M = 2\), using \([17]\) and Craig’s formula for the Gauss Q-function, i.e. \(Q(x) = 1/\pi \int_0^{\pi/2} \exp(-x^2/2\sin^2\theta) d\theta\), \(D(\theta)\) can be expressed as
\[
D(\theta) = \prod_{n=1}^N \mathcal{M}_{|\Delta_n|^2}\left(\frac{\mu}{8\sin^2\theta}\right)
\]
where \(\mathcal{M}_{|\Delta_n|^2}\) can be obtained from \([27]\). When \(M > 2\), by employing \([14]\) Eq. (13)), and using a similar line of arguments as in the case of \(M = 2\), \(D(\theta)\) can be written as
\[
\prod_{m_1=1}^M \prod_{m_2 \neq m_1=1}^M \mathcal{M}_{|\Delta_{m_1, m_2}|^2}\left(\frac{\mu}{8\sin^2\theta}\right)
\]
where \(|\Delta_{m_1, m_2}|^2 = || \mathbf{h}_{m_1} - \mathbf{h}_{m_2} ||^2\). The last MGF can be easily computed analytically with the help of \([27]\).

**V. Performance Evaluation Results and Discussion**

In this section the various performance evaluation results which have been obtained by numerically evaluating the mathematical expressions presented in Sections \(\text{III}\) and \(\text{IV}\) for uncoded and coded OSM systems operating over H-K turbulent channels will be presented. In particular, for uncoded OSM systems the following performance evaluation results have been obtained: i) ABEP vs. SNR for \(2 \times N_r\) OSM systems (obtained using \([26]\) with \([27]\), and \([34]\) - see Figs. \(\text{I}\) 2 and 3); ii) ABEP vs. SNR for \(2 \times N\) MIMO OSM systems, \(2 \times N\) MIMO (obtained using \([26]\) with \([27]\). For the uncoded schemes, in order to validate the accuracy
of the previously mentioned expressions, comparisons with complementary Monte Carlo simulated performance results are also included in these figures. As far as the performance of coded OSM systems is concerned, ABEP upper bounds vs. SNR have been obtained using (43) with (27) (see Fig. 3).

Fig. 1 illustrates the ABEP performance as a function of the average SNR, $\mu$, of $2 \times N$ MIMO H-K turbulent channels as a function of the average SNR, $\mu$, for various values of link distances, $L$. Simulation Parameters: $\lambda = 1550\text{nm}$, $C_n^2 = 1.7 \times 10^{-14}m^{-2/3}$, $\theta_{1,n} = \pi/3$, $\theta_{2,n} = \pi/4$.

Fig. 2 presents the ABEP performance as a function of the average SNR, $\mu$, of $2 \times N$ MIMO OSM systems with $N \in \{1, 2, 3, 4\}$. Independent and identically distributed branches are considered with $A_{1,n} = 2$, $A_{2,n} = 1$, $\theta_{1,n} = \pi/3$, $\theta_{2,n} = \pi/4$, $\alpha_n = 2$, $b_{0,n} = 2$. The obtained results clearly indicate that the ABEP curves, obtained using (26), are in close agreement with those obtained via simulations, verifying the correctness of the proposed analysis. Moreover, it is evident that the asymptotic ABEP curves correctly predict the diversity gain of the considered system for all tested cases.

In Fig. 2 the dependence on the link distance $L$ of the ABEP of a $2 \times N$ MIMO OSM system is illustrated. The considered system is again equipped with either $N = 2$ or $N = 4$ receiving apertures and identically distributed branches are assumed. The parameters of the H-K distribution are calculated from (15) and (16) assuming spherical wave propagation. Following (39), it is further assumed that the operating
For MRC case, the ABEP can be deduced as [29]

\[ \gamma_{\text{MRC}} = \mu \sum_{n=1}^{N} I_n \]  

(46)

whereas for SC is

\[ \gamma_{\text{SC}} = \max\{\mu I_1, \mu I_2\}. \]  

(47)

For MRC case, the ABEP can be deduced as [29]

\[ P_E = \frac{1}{2} \prod_{n=1}^{N} \mathcal{M}_{I_n}(\mu). \]  

(48)

For SC case, an analytical expression for the ABEP is more difficult to be deduced and, therefore, ABEP will be evaluated by means of Monte Carlo simulation only.

As far as the Alamouti scheme is concerned, the instantaneous SNR at the input of the demodulator of the optical receiver has a similar form as [40] [22]. For this scheme, the ABEP of BPSK can be evaluated as

\[ P_E = \frac{1}{\pi} \int_{0}^{\pi/2} \prod_{n=1}^{N} \mathcal{M}_{I_n} \left( \frac{\mu}{\sin^2 \theta} \right) d\theta. \]  

(49)

In order to simplify the underlying mathematical analysis, it is assumed that the PDF of \( I_n \) is given by [29] with the parameters \( A_n \) being all zero, i.e. the PDF is the K-distribution. Thus, \( \mathcal{M}_{I_n}(\mu) \) can be readily obtained in closed form from [33] by replacing \( b_{0,n} \) with \( b_{0,n}/2 \). In Fig. 3 the ABEP of \( 2 \times 2 \) MIMO OSM links is compared with the ABEP of \( 1 \times 2 \) coherent FSO systems with DPSK, and identically distributed links are considered. In order to compare these systems under the same propagation conditions, it is assumed that \( \alpha_n = 1, 5, b_{0,n} = 1.5, A_{2,n} = 0 \) and \( A_{1,n} = \{1, 2, 3\} \). As it can be observed, when either MRC or SC are employed, although coherent DPSK performs worse than OSM for values of \( A_{1,n} \) up to approximately 1, it outperforms OSM at lower values of \( A_{1,n} \). Moreover, although the OSM outperforms the Alamouti scheme for \( A_{1,n} = 2 \) and 3, it performs similarly for high SNR values when \( A_{1,n} = 1 \). It is noted that for \( A_{1,n} = 1 \) and lower values of \( A_{1,n} \), the Alamouti scheme yields the best performance of the considered OSM schemes. When more transmit apertures are employed, however, this advantage is compensated by the superior spectral efficiency of OSM and its lower hardware complexity as compared to coherent MRC. Specifically, as pointed out in [14], OSM offers increased spectral efficiency by a factor \( \log_2(M) \). Moreover, as only one transmitting aperture is activated at any symbol duration, OSM has a lower decoding complexity as compared to conventional MRC and Alamouti schemes.

In Fig. 4 upper bounds on the ABEP of convolutional coded \( 2 \times 1 \) and \( 2 \times 1 \) OSM systems are depicted, assuming similar propagation conditions to those considered in Fig. 2. Considering a convolutional code with rate 1/3 and constraint length of 3, its transfer function is given as [40] Eq. (8.2.6)

\[ T[D(\theta), N] = \frac{D(\theta)^6 N}{1 - 2ND(\theta)^2}. \]  

(50)

Substituting [50] to [43], a union bound on the ABEP can be obtained as

\[ P_{\text{ub}} \leq \frac{1}{\pi \log_2(M)} \int_{0}^{\pi/2} D(\theta)^6 \left( \frac{1}{1 - 2D(\theta)^2} \right) d\theta. \]  

(51)

The performance results of Fig. 4 clearly show that, as expected, the incorporation of convolutional coding significantly enhances the performance of OSM systems, even when a small number of receive apertures is employed, even for \( N = 1 \).

VI. CONCLUSION

In this paper, the use of spatial modulation technique for coherent FSO communication systems has been proposed. We have provided a comprehensive analytical framework for error performance analysis in the presence of atmospheric turbulence scattering channel models which include the H-K distribution. The proposed framework reveals important information about the performance of OSM over such turbulent channels, including the effect of fading severity and the achievable diversity gain. It also provides valuable insight into the impact of channel parameters on performance of OSM. Upper bounds for the ABEP performance of coded OSM systems have also been derived, demonstrating that coding techniques can greatly enhance the performance of OSM. Extensive
computer simulation performance evaluation results have been also obtained which have verified the accuracy of the analytical approach. Important trends about the performance of OSM for a variety of atmospheric turbulent scenarios and MIMO setups have also been identified. For example, it was shown that OSM can provide significant performance enhancements in the presence of atmospheric turbulence. The improvements are comparable to the ones offered by conventional coherent systems with spatial diversity, while outperforming the latter in terms of spectral efficiency and hardware complexity. Besides, under specific propagation conditions, OSM can yield better performance than conventional SIMO systems employing MRC or SC. We believe that the proposed framework is a useful tool for understanding the performance trend, important properties and tradeoffs of outdoor OSM operating in the presence of atmospheric turbulence.

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