Diversity Combining for Mobile FSO Nodes in SIMO Setup

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***Abstract*— The term *mobile FSO* refers to wireless optical nodes with mobility capabilities. Mobile FSO has emerged as a way to combat some of limitations facing FSO links, i.e. incompatibility of transmitters and receivers relative to movement and misalignment. In this paper we utilize an angular diversity model for characterizing the performance of mobile FSO over atmospheric turbulence. Switch-and-examine combining (SEC) technique is our suggestion for the diversity combining solution in such unbalanced multi-transceiving configuration. The channel fading is modeled as a lognormal distribution with spatially correlated samples. Analytical/ statistical discussion on the resultant output is presented and bit error performance and processing load are numerically evaluated and compared to a selection combining (SC) diversity.**

***Index Terms*—** Mobile FSO, Channel state information (CSI), Spatial diversity, Selection combining (SC), Switched diversity, Switch-and-examine combining (SEC).

I. Introduction

Wireless optical communication, also known as free space optics (FSO), offers advantages over Radio Frequency (RF) including unlicensed frequency spectrum and high data rate transmission. However, line of sight (LoS) connection and the directional reception of narrow FSO beams are major deterrents for practical development. Classified as single-input multiple-output (SIMO), mobile FSO offers mobility to an optical network; asymmetrically unbalanced transceiver apertures with intentional misalignment are installed to establish node mobility while maintaining connection.

FSO-based mobile nodes can potentially be used either in a mobile ad hoc network (MANET) where infrastructure is unavailable [1] or in a wireless sensor network (WSN) where security of communication, including freedom from susceptibility to jamming, is important [2]. Previous works on mobile FSO focus mainly on alignment and tracking issues [3] and the design of multi-element structure [4], [5] which are characterized to ensure uninterrupted data flow by auto-aligning transmitter and receiver modules. These concentrate on design and/or experimental views based on a simple deployed angular diversity model. Unbalanced multireceiving structure of optical power distribution in a nonrandom model is characterized in [6]. The work presented in this paper introduces random atmospheric turbulence into the unbalanced diversity model of mobile FSO, where branches have unequal average SNRs.

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Since mobile FSO can be modeled as multi-receiving spatial diversity, an efficient diversity combining scheme is required to improve performance. When compared to all diversity techniques, Maximum Ratio Combining (MRC)—also known as Optimal Combining (OC)—provides superior combining performance. But implementation complexities are inherent, and the system is extremely sensitive to channel estimation error. This is especially true for low SNR signals. Thus, MRC is an inefficient choice for mobile FSO. Equal Gain Combining (EGC) is inefficient for systems with branches having acutely low SNR conditions. Selection Combining (SC) adheres to a selection strategy based on the highest received SNR. Although the process may seem simple, the high processing load and repetitive switching characterized in SC diversity causes increased implementation and does not fully exploit diversity offered by the branches.

This paper investigates the feasibility of deploying a switch-and-examine combining (SEC) diversity technique for mobile FSO. Receiver switching is initiated when SNR reading is low. This takes into account the SNR of a new branch. Provided the SNR level is above a predefined threshold level, it remains on this branch; otherwise, switching continues to alternate branches until an acceptable SNR is observed.

We assume a circular configuration of transceiver apertures, where all are placed in the same plane. Intensity modulation techniques are used for transmitting user binary information through a transmission medium. Modulation format selection is outside the scope of this paper, as it is independent of Intensity Modulation/Direct Detection (IM/DD) scheme.

The balance of this paper is organized as follows. Section II offers basic definitions and channel model assumptions. Section III presents node structure analysis of SIMO-based setup in terms of probability distribution. A combining approach for diversity in mobile FSO can be found in Section IV. Simulation results are presented in Section V, and the Section VI concludes the paper.

II. Channel Model

The fading channel coefficient, i.e., gain, which models the channel intensity gain from transmit aperture to receive aperture, is given by $h = e^{2\chi}$ where log-amplitude χ is a normal random variable (RV) with mean μ_χ and standard deviation σ_χ, i.e., *fading strength* or "Rytov variance/parameter" in the literature. The selection of $E[h] =$

1 leads us to $\mu_\chi = -\sigma_\chi^2$. Instantaneous electrical SNR will be defined by [7]

$$\gamma = \bar{\gamma}\alpha^2 h^2 \tag{1}$$

where, in our analysis, α is misalignment coefficient and $\bar{\gamma}$ is defined as the average, excluding fading effect

$$\bar{\gamma} \triangleq \frac{R^2 P_t^2}{\sigma_v^2} \tag{2}$$

for On-Off Keying (OOK) modulation. In (2), σ_v is the noise standard deviation assumed equal for both symbols '1' and '0' [8], R is the receiver's responsivity, i.e. optical-to-electrical conversion coefficient, and P_t is the average optical power. For simplicity, we include $2RP_t = 1$ to normalize the power.

Lognormal distribution is considered for representing atmospheric turbulence statistics (i.e. scintillation). It has a probability distribution function (PDF) in the form of [7]

$$f_h(h) = \frac{1}{\sqrt{8\pi}h\sigma_\chi}\exp\left[-\frac{\left(\ln(h) + 2\sigma_\chi^2\right)^2}{8\sigma_\chi^2}\right] \tag{3}$$

for the channel coefficient.

Channel coefficients are correlated in time and space domains as a result of atmospheric eddies movement. Temporal correlativeness may affect optimal detection performance when a single-input single-output (SISO) system is investigated [9], [10]. Spatial correlativeness should be considered when a spatial-based diversity system is used to prove improved detection performance. Two important parameters—coherence time and coherence length—in particular, represent the variation of the time-varying fading channel in time and space domain, respectively. In this work, we consider a multi-receiving system with only spatially correlated links.

Without loss of generality, the probability distribution of N is identically distributed; however, correlated fading coefficients are a N-dimensional joint PDF. The joint PDF of the channel coefficient is presented in the form of $f_{\boldsymbol{H}}(h_1, h_2, \ldots, h_N)$; $\boldsymbol{H} = [h_1\ h_2 \ldots h_N]_{1\times N}$, which is identically but not independently distributed (not i.i.d). Also we assume $E[h_n] = 1$ for any $n = 1,2,\ldots,N$ and $\bar{\boldsymbol{\Psi}}$ is defined as the average vector by $\bar{\boldsymbol{\Psi}} \triangleq \{E\{\chi_n\}\}_{n=1}^N = [-\sigma_x^2\ -\sigma_x^2\ \ldots\ -\sigma_x^2]_{1\times N}$, as the paths are identically faded. Due to symmetry, $\bar{\gamma}_i = \bar{\gamma}$ for any $i = 1,2,\ldots,N$. Also, Σ_χ is the defined as the covariance matrix of the channel coefficients

$$\Sigma_\chi = \begin{bmatrix} \sigma_x^2 & C_{1,2} & \cdots & C_{1,N} \\ C_{2,1} & \sigma_x^2 & \cdots & C_{2,N} \\ \vdots & \vdots & \ddots & \vdots \\ C_{N,1} & C_{N,2} & \cdots & \sigma_x^2 \end{bmatrix}_{N\times N} \tag{4}$$

where $C_{i,j}$ is the mutual covariance function between the log-amplitudes χ_i and χ_j of any two fading coefficients h_i and h_j associated with any two receivers (or transmitters) i and j (Fig. 1.). Note that $C_{i,j} = C_{j,i}$ result from symmetry property of correlation. The correlation coefficient is defined as $\rho_{i,j} \triangleq C_{i,j}/\sigma_x^2$, $0 \le \rho_{i,j} \le 1$.

III. Node Structure Analysis

Various configurations can be considered for mobile FSO in an ad hoc network. We assume nodes have a circular structure with two nodes and all receiver apertures placed in the same plane. The resultant configuration denotes a receive diversity in SIMO. As Fig. 1 shows, any two adjacent receiver apertures are separated by angles $\{\theta_n\}_{n=1}^N$ on a framework. In this part, the dynamics of such configuration is addressed to mathematically characterize the jointly distribution of the SNRs of the branches.

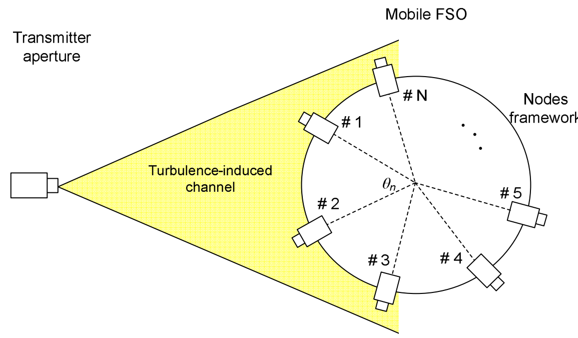

Fig. 1: Illustration of mobile FSO in a SIMO setup.

The electrical signal of the n-th receiver in discrete time samples would be expressed by

$$r_n = 2RPh_n\alpha_n s + v_n\ ; \quad n = 1,2,\ldots,N \tag{5}$$

where $h_n > 0$ is the normalized time-varying channel fading coefficient associated to the n-th receiver, and v_n is total additive Gaussian noise associated with the n-th receiver, having mean I_0 and variance σ_v^2, equal for all receivers. For such a configuration, CSI is defined as

$$\{\gamma_n\}_{n=1}^N \tag{6}$$

where γ_n is the instantaneous SNR of the branch n given by (1). Possessing knowledge on this equivalently leads to the knowledge on $\{\alpha_n h_n\}_{n=1}^N$. Note that the notation of discrete-time sampling is neglected in the model in (5). In (5), α_n, $0 \le \alpha_n \le 1$, is the attenuation factor of the n-th receiver due to misalignment or angular reception. Specifically, the misalignment coefficient α_n for any receiver aperture n directly relates to its azimuth angle in the transmission plane,

$$\alpha_n = \begin{cases} \cos\varphi_n\,, & -\frac{\pi}{2} < \varphi_n < \frac{\pi}{2} \\ 0, & \frac{\pi}{2} \le \varphi_n \le \frac{3\pi}{2} \end{cases}\ ; \quad n = 1,2,\ldots,N \tag{7}$$

where $\{\varphi_n\}_{n=1}^N$ are the azimuth angles of receivers in counterclockwise direction, as shown in Fig. 2. Although these angles are random variables, they are dependent on each other based on the separation angles $\{\theta_n\}_{n=1}^N$. For any given receiver n, the azimuth angle is given by

$$\varphi_n = \varphi_q + \sum_{i=n}^{q-1}\theta_i\ ; \quad n = 1,2,\ldots,N \tag{8}$$

where q is a given aperture with angle φ_q. Note that $\{\theta_n\}_{n=1}^N$ are deterministic and fixed; therefore, (8) indicates that all the random variables $\Phi = \{\varphi_n\}_{n=1}^N$ can be expressed in only one given angle.

Some symmetry can be included into the mobile node structure to characterize the received signal power within the receivers. If it is assumed that the nodes are uniformly placed on the framework, then $\theta_1 = \theta_2 = \cdots = \theta_N = \theta$, which would be given by $\theta = 2\pi/N$. We can assume $q = 1$, thus

$$\varphi_n = \varphi_1 + \frac{2(n-1)\pi}{N}\ ; \quad n = 1,2,\ldots,N \tag{9}$$

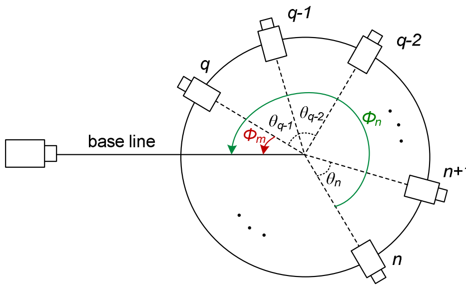

Fig. 2: Angle analysis of nodes in a SIMO setup.

Equation (9) indicates all RVs $\{\varphi_n\}_{n=1}^{N}$ are known if one is known. Thus, the RVs due to misalignment are totally correlated and can be reduced to only one variable. If the RV defined as $\varphi \triangleq \varphi_1$; $-\frac{\pi}{2} < \varphi < \frac{3\pi}{2}$, has a PDF expressed by $f_\varphi(\varphi)$, then the PDF of all other RVs are expressed by

$$f_{\varphi_n}(\varphi_n) = f_\varphi\left(\varphi - \frac{2(n-1)\pi}{N}\right); \quad n = 1, 2, \ldots, N \tag{10}$$

where $-\frac{\pi}{2} < \varphi < \frac{3\pi}{2}$. Consequently, the PDF of α_n can be calculated by

$$f_{\alpha_n}(\alpha_n) = \begin{cases} \frac{1}{\sqrt{1-\alpha_n^2}} f_{\varphi_n}(\arccos \alpha_n), & -\frac{\pi}{2} \le \arccos \alpha_n < \frac{\pi}{2} \\ 0, & \frac{\pi}{2} \le \arccos \alpha_n \le \frac{3\pi}{2} \end{cases} \tag{11}$$

where $n = 1, 2, \ldots, N$ and $0 \le \alpha_n \le 1$. The vector of joint PDF of misalignment angle is then given by

$$f_\Phi(\varphi) = \left\{ f_\varphi\left(\varphi - \frac{2(n-1)\pi}{N}\right)\right\}_{n=1}^{N} \tag{12}$$

The vector of misalignment coefficient is a function of α

$$f_{\mathrm{A}}(\alpha) = \{f_{\alpha_n}(\alpha)\}_{n=1}^{N} \tag{13}$$

where $\mathrm{A} = \{\alpha_n\}_{n=1}^{N}$. Mathematically, the instantaneous SNR is defined by

$$\gamma_n = \frac{R^2 P^2 h_n^2 \alpha_n^2}{\sigma_v^2} = h_n^2 \alpha_n^2 \bar{\gamma} \quad ; \; n = 1, 2 \ldots, N \tag{14}$$

where $\bar{\gamma}_n$ is the average received signal with no fading defined in (2) and $h_n^2 = e^{4\chi_n}$. In unbalanced applications, e.g., angular diversity, noise power σ_v^2 can be assumed constant at different apertures; however, α_n varies for different receivers due to position fluctuation.

In order to statistically derive the marginal PDF of γ_n, the PDFs of both h_n^2 and α_n^2 are needed. We can easily find the expression of h_n^2 by using (1) and (3)

$$f_{h_n^2}(h) = \frac{1}{\sqrt{32\pi h^2 \sigma_\chi}} \exp\left(-\frac{\left(\ln(h) + 4\sigma_\chi^2\right)^2}{32\sigma_\chi^2}\right) \tag{15}$$

and for the joint distribution,

$$f_{h^2}(h_1, h_2, \ldots, h_N) = \frac{\exp\left(-\frac{1}{32}(\ln[\boldsymbol{H}] - 4\overline{\boldsymbol{\Psi}})\Sigma_\chi^{-1}(\ln[\boldsymbol{H}] - 4\overline{\boldsymbol{\Psi}})^T\right)}{4^N (2\pi)^{\frac{N}{2}} \left(\det\left[\Sigma_\chi\right]\right)^{\frac{1}{2}} \mathrm{P}[\boldsymbol{H}]} \tag{16}$$

where T is transpose operator, and $\mathrm{P}[\boldsymbol{H}] \triangleq \prod_{n=1}^{N} h_n$ is the product function. Also, $f_{\alpha_n^2}(\alpha)$ is considered as the PDF of α_n^2, which becomes

$$f_{\alpha_n^2}(\alpha_n) = \frac{1}{2\sqrt{\alpha_n}} f_{\alpha_n}\left(\sqrt{\alpha_n}\right) \tag{17}$$

The corresponding vector can be accordingly defined by

$$f_{\mathrm{A}^2}(\alpha) \triangleq \{f_{\alpha_n^2}(\alpha)\}_{n=1}^{N} \tag{18}$$

where $\mathrm{A}^2 = \{\alpha_n^2\}_{n=1}^{N}$. Since h_n and α_n can be confidently assumed uncorrelated, using the help of [11, Ex. 7] as the PDF of the product of two RVs,

$$f_{\gamma_n}(\gamma_n) = \frac{1}{\bar{\gamma}} \int_0^1 \frac{1}{u} f_{h_n^2}\left(\frac{\gamma_n}{\bar{\gamma} u}\right) f_{\alpha_n^2}(u)\, du \tag{19}$$

where $f_{h_n^2}(.)$ is the marginal PDF of h_n^2 for the n-th receiver given by (15). We extend this to joint representation

$$f_\Gamma(\gamma_1, \gamma_2, \ldots, \gamma_N) = \frac{1}{\bar{\gamma}^N} \times \int_0^1 \frac{1}{u} f_{h^2}\left(\frac{\gamma_1}{\bar{\gamma} u}, \frac{\gamma_2}{\bar{\gamma} u}, \ldots, \frac{\gamma_N}{\bar{\gamma} u}\right) \circ f_{\mathrm{A}^2}(u)\, du \tag{20}$$

where $\circ$ is the Hadamard product operator.

IV. The SEC Combining Approach

The signals from receiver apertures require combining before detection occurs, as shown in Fig. 3. In this section, the authors statistically depict SEC switched diversity technique for mobile FSO.

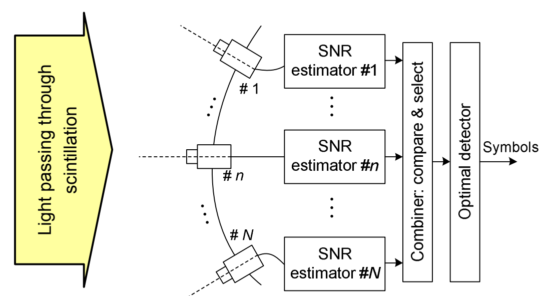

Fig. 3: Structure of an unbalanced multireceiver.

A. Switching Strategy

Basically, a few potential diversity approaches are available for switched diversity. Based on a switch-and-stay combining (SSC) scheme [12], [13], the combiner switches to the next branch subsequent to the existing received SNR dropping below a threshold γ_T, regardless of the new branch's SNR. This remains true even if the new branch is inferior to the original branch. A major deficiency with SSC exists, as the probability of a non-illuminated receiver is very high. Therefore, SSC is not an acceptable choice for mobile FSO.

The SEC diversity scheme is similar to SSC with minor modification. Although a low SNR reading initiates receiver switching, the new branch SNR is first taken into consideration. If the SNR level is above the threshold level, the original branch is maintained. Switching continues to alternate branches until an acceptable SNR is observed, at which time switching ceases. SEC is designed on a switching threshold basis and is basically proposed to reduce the volume of processing load and, thus, implementation complexity induced to receiver [12], [14].

The study of SEC has been characterized by Yang and Alouini in [15], and Alexandropoulos *et al.* in [14] while the result for two uncorrelated paths has been rewritten by Simon and Alouini in [12, Eq. (9.340)]. We use a similar but more basic and general method for characterizing the probability distribution of the resultant SNR for FSO links, as shown below.

B. Analytical Analysis

Let's include the discrete sampling time k in the analytical discussion. Without loss of generality, suppose the combiner is currently connected to branch n of a N-branch receiver. To avoid confusion of notation, we recall several definitions:

$\gamma_n[k]$: instantaneous received SNR at branch n at time k
$\gamma[k]$: combined (resultant) received SNR at time k
γ: combined (resultant) received SNR variable
n: active branch at time k
γ_T: switching threshold

We emphasize that γ_n is the total instantaneous SNR at receiver n; thus, the analysis is applicable to SIMO configuration. Since the events of sequence $\{\gamma[k]=\gamma_i[k]\}_{i=1}^{N}$ are mutually exclusive, the CDF of the resultant SNR can be generally rewritten as [15], [16]

$$F_{SEC}(\gamma)=\sum_{i=1}^{N} P(\gamma[k]=\gamma_i[k] \ \text{ and } \ \gamma_i[k]\le\gamma) \quad (21)$$

Again, by considering the mutually exclusive property of sequence $\{\gamma[k]=\gamma_i[k]\}_{i=1}^{N}$, the CDF can be deduced from (21):

$$F_{SEC}(\gamma)=N\,P(\gamma[k]=\gamma_n[k] \ \text{ and } \ \gamma_n[k]\le\gamma) \quad (22)$$

In the case of SEC switching strategy, possible cases exist in which equality $\gamma[k]=\gamma_n[k]$ may occur, where

$$\gamma[k]=\gamma_n[k] \text{ iff} \begin{cases} \gamma[k-1]=\gamma_{n-1}[k-1] \\ \text{and } \gamma_{n-1}[k]<\gamma_T \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{or} \\ \gamma[k-1]=\gamma_{n-2}[k-1] \\ \text{and } \bigcap_{i=n-2}^{n-1}\gamma_i[k]<\gamma_T \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{or} \\ \vdots \\ \gamma[k-1]=\gamma_1[k-1] \\ \text{and } \bigcap_{i=1}^{n-1}\gamma_i[k]<\gamma_T \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{or} \\ \gamma[k-1]=\gamma_N[k-1] \\ \text{and } \bigcap_{i=N}^{n-1}\gamma_i[k]<\gamma_T \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{or} \\ \vdots \\ \gamma[k-1]=\gamma_{n+1}[k-1] \\ \text{and } \bigcap_{i=n+1}^{n-1}\gamma_i[k]<\gamma_T \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{or} \\ \gamma[k-1]=\gamma_n[k-1] \text{ and } \gamma_n[k]\ge\gamma_T \end{cases} \quad (23)$$

Eq. (23) lists all possible cases in which the n-th receiver is the current receiver at time k. Since the receiver apertures are symmetrically placed on the platform, $\bigcap_{i=t}^{s}\gamma_i[k]<\gamma_T$ for $t>s$ becomes

$$\bigcap_{i=t}^{s}\gamma_i[k] \triangleq \left(\bigcap_{i=t}^{N}\gamma_i[k]\right)\bigcap\left(\bigcap_{i=1}^{s}\gamma_i[k]\right) \quad (24)$$

recalling that events $\gamma[k-1]=\gamma_i[k-1]\,, i=1,2,\ldots,N$ are mutually exclusive and, consequently, any "and" combination of them will be also exclusive. Branches are equally likely chosen due to the symmetry of receiver apertures on the platform. Thus, applying the occurrence probability of each case as $1/(N-1)$ in (23) and by substituting the simplified expression in the CDF in (21), it yields

$$F_{SEC}(\gamma) \quad (25)$$
$$=\frac{N}{N-1}\sum_{j=1}^{N-1} P\left\{\begin{matrix}\gamma[k-1]=\gamma_q[k-1] \text{ and } \gamma_n[k]\ge\gamma_T \\ \text{and } \bigcap_{i=n-1-j}^{n-1}\gamma_i[k]\le\gamma_T \text{ and } \gamma_n[k]\le\gamma\end{matrix}\right\}$$
$$+NP\{\gamma[k-1]=\gamma_n[k-1] \text{ and } \gamma_n[k]\ge\gamma_T \text{ and } \gamma_n[k]\le\gamma\}$$

The combiner monitors the instantaneous SNRs at a combining sampling period time ξ, which is amply large when compared to coherence time. Consequently, the correlated pair $\gamma_n[k]$ and $\gamma_q[k]$ are independent of their corresponding values at time $k-1$. By understanding that $P\{\gamma[k]=\gamma_n[k]\}=1/N$ and similarly, $P\{\gamma[k-1]=\gamma_n[k-1]\}=1/N$ for any value of n, the CDF is evaluated in the form

$$F_{SEC}(\gamma)=\frac{1}{N-1}\sum_{j=1}^{N-1} P\left\{\bigcap_{i=n-j}^{n-1}\gamma_i[k]\le\gamma_T \cap \gamma_T\le\gamma_n[k]\le\gamma\right\}$$
$$+P\{\gamma_n[k]\ge\gamma_T \text{ and } \gamma_n[k]\le\gamma\} \quad (26)$$

Depending on the value of γ_T, $F_{SEC}(\gamma)$ converts to

$$F_{SEC}(\gamma)=\begin{cases} G_\gamma(\gamma) & \gamma<\gamma_T \\ F_\gamma(\gamma)-F_\gamma(\gamma_T)+G_\gamma(\gamma) & \gamma\ge\gamma_T \end{cases} \quad (27)$$

where $F_\gamma(.)$ is the marginal CDF of the resultant SNR, γ. Recalling that n can be any given receiver $n\in\{1,2,\ldots,N\}$, thus, simply choosing $n=N$, $G_\gamma(\gamma)$ is calculated by

$$G_\gamma(\gamma)=\frac{1}{N-1}\sum_{j=1}^{N-1} P\left\{\bigcap_{i=N-j}^{N-1}\gamma_i[k]\le\gamma_T \text{ and } \gamma_N[k]\le\gamma\right\} \quad (28)$$

which extends to

$$G_\gamma(\gamma)=\frac{1}{N-1}\{P\{\gamma_{N-1}[k]\le\gamma_T \text{ and } \gamma_N[k]\le\gamma\}$$
$$+P\{\gamma_{N-2}[k]\le\gamma_T \text{ and } \gamma_{N-1}[k]\le\gamma_T \text{ and } \gamma_N[k]\le\gamma\}+\cdots$$
$$+P\{\gamma_1[k]\le\gamma_T \ \ldots \text{and } \gamma_{N-1}[k]\le\gamma_T \text{ and } \gamma_N[k]\le\gamma\}\} \quad (29)$$

To provide a clear understanding of Eq. (29), we recall that for an N-dimensional jointly variable $\mathbf{\Gamma}$,

$$P\{\gamma_1[k]\le\gamma_T \ \ldots \text{and } \gamma_{K-1}[k]\le\gamma_T \text{ and } \gamma_K[k]\le\gamma\}=$$
$$=\underbrace{\int_0^{\gamma_T}\ldots}_{K-1}\underbrace{\int_0^{\infty}\ldots}_{N-K}\int_0^{\gamma} f_{\mathbf{\Gamma}}(\gamma_1,\ldots,\gamma_N)\ \underbrace{d\gamma_N\ldots d\gamma_1}_{N} \quad (30)$$

for any integer $2\le K\le N$. The PDF of the resultant γ can be mathematically expressed by

$$f_{SEC}(\gamma)=\begin{cases} g_\gamma(\gamma) & \gamma<\gamma_T \\ f_\gamma(\gamma)+g_\gamma(\gamma) & \gamma\ge\gamma_T \end{cases}, \quad (31)$$

where $f_\gamma(\gamma)$ is the marginal PDF derivable from (19), and $g_\gamma(\gamma) \triangleq \frac{d}{d\gamma} G_\gamma(\gamma)$. By applying an optimal detection for a very large number of symbols, the average BER is achievable by

$$P_{SEC}^{e} = \int_0^\infty f_{SEC}(\gamma)\, Q(\gamma) d\gamma \qquad (32)$$

where $Q(.)$ is one dimensional Q-function.

V. Simulation Analysis

We numerically measure the BER probability by sending intensity modulated symbols through a fading channel and detecting the received symbols using an optimal metric. Additional processing load due to diversity is defined as the number of branch monitoring occurrences (exclusive of the current branch) to the maximum possible cases. Subsequently, the processing load, $Pload$, for SC is $Pload_{SC} = N - 1$, and for SEC is $Pload_{SEC} = \epsilon\xi/\tau$, where ϵ is the number of additional branch monitoring occurrences in a time period τ and ξ is the combining sampling period, as previously defined. Based on the switching strategy of SEC, $Pload_{SEC} \leq N - 1$.

We use symbol-by-symbol (S-by-S) detection method assuming perfect CSI in Eq. (1) or (6) is available; OOK modulation is employed in the simulation. When referred, definition $\bar{\gamma}$ in (2) is used as the *average received SNR* or *average SNR* in the performance plots. To normalize the power, we include $2RP_t = 1$. Lognormal distribution is applied for the channel fading distribution, and correlation coefficient values are assumed as $\{\rho_{i,j}\}_{i,j=1,i\neq j}^{N} = \rho$, where $\rho_{i,j} = 1$ for $i = j$. Cholesky decomposition is employed to generate correlated samples. To proceed we assume uniform but independent distributions for misalignment angle φ_1 through PDFs $f_\varphi(\varphi)$, changing from $-\pi/2$ and $3\pi/2$. Using (9), all other variables $\{\varphi_n\}_{n=2}^{N}$ can be determined. Any misalignment gain factors of $\{\alpha_n\}_{n=1}^{N}$ have a value between 0 and 1, given in Eq. (7).

An optimum threshold for SEC which minimizes BER performance may exist, as shown in Figs. 4 and 5. Optimal points will be applied in forthcoming simulations. For high values of γ_T, SEC works as an unbalanced single receiver. However, for low values of γ_T, situation is superior. Note that a complete correlation of $\rho = 1$ does not present the worst performance, but rather, $\rho = 0.8$ is, as shown in Fig 4. This result is an interesting, as it is not expected to be observed in regular balanced configurations.

Fig. 6 is presented to highlight the effect of the number of branches on BER performance of SEC. In this figure, branches are considered uncorrelated, i.e., $\rho = 0$, and the fading strength is $\sigma_\chi = 0.2$. As demonstrated, $N = 2$ results in poor mobile FSO performance even when using the SC method. Fig. 7 compares SC and SEC BER performances. SC performance is superior, e.g., the BER performance of SC having $N = 6$ is superior than that of SEC with $N = 8$, which is about 2.2dB at $\bar{\gamma} = 17$dB. Additionally, correlation can degrade the combining performance of SEC and SC, as plotted in Fig. 8 for SEC. The figure shows combining performance with correlation between branches, i.e., between channels. SEC performance using $N = 6$ and uncorrelated branches is superior to $N = 8$ and $\rho = 0.2$. Thus, the correlation between branches greatly affects BER performance when using SEC or SC.

Processing load due to combining as a performance criterion is presented in Fig. 9. As shown, SEC decreases the processing load in the combiner when compared to SC performance. Additionally, setup model, i.e., SIMO, does not affect the processing load for SC. Increasing the number of receivers contributes to a higher load for both SC and SEC.

VI. Conclusions

Deployment of mobility-based FSO links is complicated, as highly sensitive FSO links are subject to misalignment. In this paper, we propose and analyze a circular structure for the apertures of mobile FSO providing mobility to FSO nodes in WSNs, MANETs. In this paper, mobile FSO was evaluated in SIMO setup was comprised of a small number of receiver apertures, which were intentionally misaligned to provide angular diversity. In this event, channels are considered unbalanced and correlated. Accordingly, an SEC diversity technique is suggested for combining purposes. SEC helps achieve a low processing load; however it requires an optimal switching threshold for performance optimization.

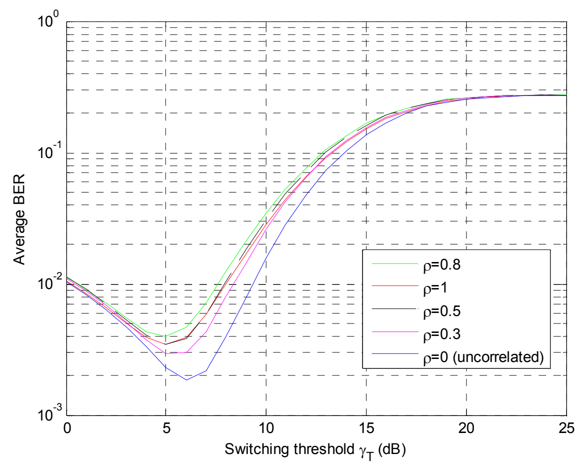

Fig. 4: BER of SEC vs. switching threshold γ_T, for several different values of correlation coefficients ρ. N=4, $\sigma_\chi = 0.2$ and $\bar{\gamma} = 17$dB.

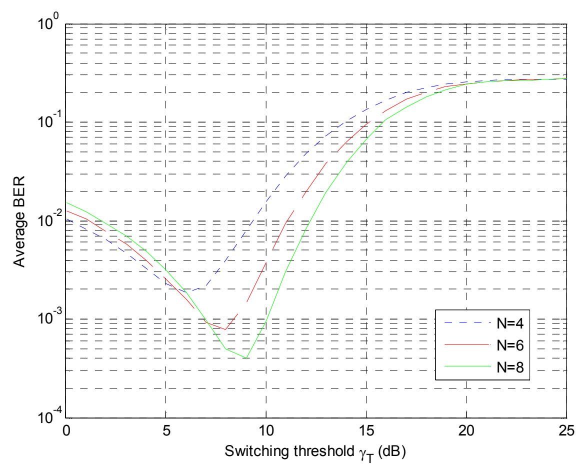

Fig. 5: Bit error rate vs. switching threshold γ_T in SIMO, for a N-branch SEC (N=4, 6, 8). $\rho = 0$, $\sigma_\chi = 0.2$ and $\bar{\gamma} = 17$dB.

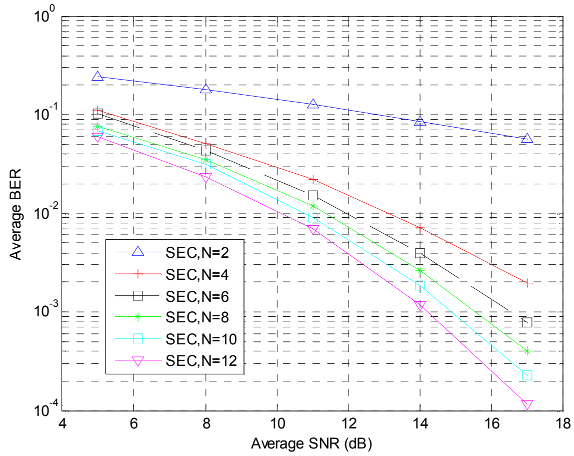

Fig. 6: Optimal performance of SEC vs. $\bar{\gamma}$ in SIMO, for $N = 2 - 12$ with uncorrelated branches, $\rho = 0$, and at $\sigma_\chi = 0.2$.

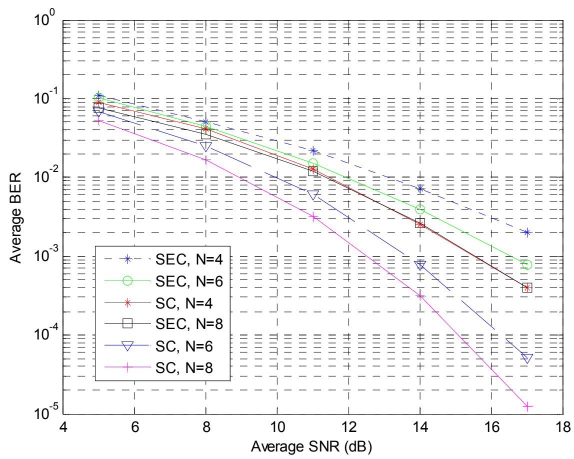

Fig. 7: Performance of SEC vs. $\bar{\gamma}$ in SIMO, compared to SC for $N = 4, 6, 8$ with uncorrelated branches, $\rho = 0$, and at $\sigma_\chi = 0.2$.

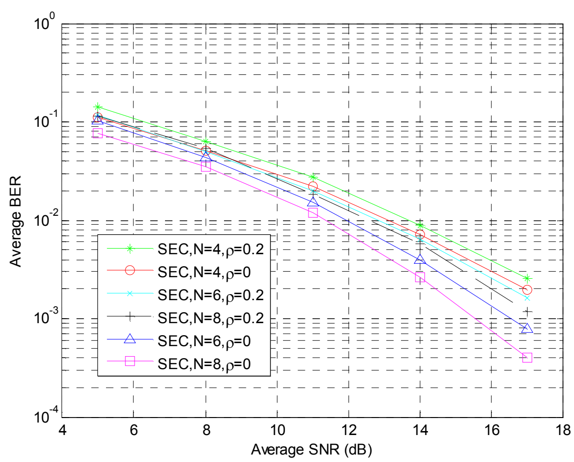

Fig. 8: The effect of correlation and number of receivers on the performance of FSO node in SIMO using SEC for $N = 4, 6, 8$ at $\sigma_\chi = 0.2$.

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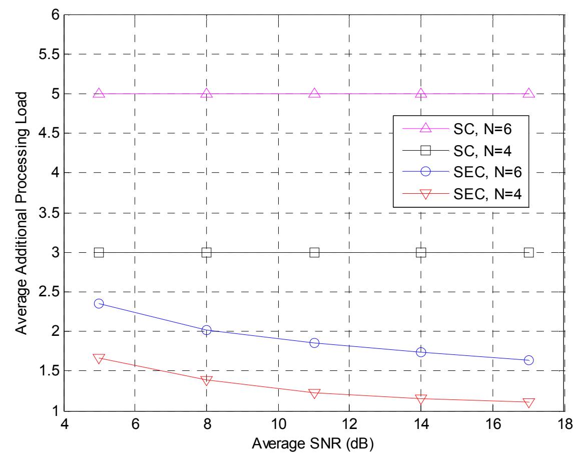

Fig. 9: Additional processing load due to diversity of SEC vs. $\bar{\gamma}$, compared with SC for $N = 4, 6$ with uncorrelated branches, $\rho = 0$, and at $\sigma_\chi = 0.2$.

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