

Polarization-Dependent Loss Mitigation in MISO Coherent Optical Ground-to-GEO Feeder Links

Youssef Minyari^{*a,b}, Raphaël Le Bidan^a, Charly Poulliat^c, and Aubin Lecointre^d

^aIMT Atlantique, Brest, France

^bTéSA, Toulouse, France

^cToulouse INP-ENSEEIHT, Toulouse, France

^dThales Alenia Space, Toulouse, France

ABSTRACT

In this paper, we propose integrating the Polarization Pairwise Coding technique into Multiple-Input Single-Output (MISO) architectures to mitigate Polarization-Dependent Loss (PDL) in coherent free-space optical (FSO) ground-to-GEO feeder links. We demonstrate that this simple linear-processing strategy efficiently mitigates the system link budget penalty associated with worst-case PDL, bringing performance close to the PDL-free limit, without increasing receiver complexity. The precoded MISO architectures maintain performance nearly identical to that of PDL-free conditions. Our findings demonstrate that this simple integration preserves the advantage of spatial diversity while mitigating PDL.

Keywords: Coherent optical feeder links, ground-to-GEO, atmospheric turbulence, MISO, diversity, PDL, pairwise coding

1. INTRODUCTION

Coherent free-space optical (FSO) communications for ground-to-GEO satellite feeder links have emerged as a critical technology to meet the growing demand for terabit-per-second capacities,¹ overcoming the limitations of traditional radio frequency (RF)^{2,3} and direct detection (IM/DD) systems.^{4,5} However, these optical links suffer from severe signal fading induced by atmospheric turbulence.⁶ Although deep temporal interleaving, complex adaptive optics, large telescope apertures, and High-Power Optical Amplifiers (HPOA) can mitigate this fading, they introduce excessive latency and increase the overall system complexity. To avoid this, a Multiple-Input Single-Output (MISO) 2×1 architecture using Alamouti⁷ Space-Time Coding (STC) is being investigated to provide the necessary spatial diversity, a technique highly effective for latency-constrained optical feeder links.⁸ As demonstrated in our previous work,⁹ this architecture successfully stabilizes the link against atmospheric effects while relaxing hardware implementation constraints in terms of power amplification, interleaving, telescope size, and adaptive optics.

Despite these system-level advantages, the overall link budget remains highly vulnerable to specific optical component imperfections, most notably the Polarization-Dependent Loss (PDL), which introduces a differential gain between the two orthogonal polarizations of the studied Dual-Polarization (DP) system. Although analog optical components, such as polarization scramblers,^{10,11} can alleviate this effect, Digital Signal Processing (DSP) solutions are preferred to avoid adding hardware complexity to the optical payload. To mitigate the loss of capacity entirely in the digital domain, several DSP-based Polarization-Time Coding (PTC) techniques have been developed and studied for long-haul optical fiber networks. These include the Golden and Silver codes,^{12,13} the Number Theory (NT) code,¹⁴ and the capacity-achieving scheme proposed by Shehadeh and Kschischang.¹⁵ This scheme was expanded in a study demonstrating its effectiveness in a practical 16-QAM setting.¹⁶ More recently, to avoid the delay induced by coding over multiple consecutive time-slots, single time-slot DSP solutions have been introduced, such as Spatially Balanced (SB) signaling¹⁷ and New Spatially Balanced (NSB) signaling.¹⁸ In this context, recent contributions have also introduced the IQ-code for multi-carrier systems.^{19,20}

* youssef.minyari@imt-atlantique.fr

However, integrating these PDL mitigation strategies from optical fibers into an FSO context is not straightforward. In a ground-to-GEO link, the optical signal is affected by PDL and severe atmospheric turbulence, which requires spatial diversity techniques at the transmitter. Integrating traditional PTCs into a multi-aperture MISO architecture introduces large decoding complexity at the receiver. Therefore, this paper proposes integrating the low-complexity Polarization Pairwise Coding technique²¹ to ensure a robust MISO architecture against atmospheric turbulence and PDL. By operating with the existing Alamouti diversity scheme, this approach mitigates PDL without additional complexity. For this reason, our focus is on evaluating the performance of these concatenated techniques to demonstrate the overall robustness of the system.

This paper is organized as follows. Section 2 establishes the system architecture and the mathematical channel model. Section 3 details the integration of the polarization pairwise coding within the MISO architecture and provides the theoretical noise covariance analysis. Section 4 outlines the simulation parameters and presents a comparative performance evaluation of the precoded configurations. Finally, Section 5 concludes the paper and offers perspectives for future work.

2. SYSTEM AND CHANNEL MODEL

2.1 System Architecture

As established in our previous work,⁹ the MISO 2×1 architecture utilizes two ground-based transmit apertures separated by a distance greater than the atmospheric coherence length, commonly known as the Fried parameter (r_0). For a typical ground-to-GEO scenario, this parameter ranges between 5 cm and 30 cm.⁶ Maintaining a separation distance $d > r_0$ ensures completely independent propagation paths, providing the necessary spatial diversity to mitigate turbulence-induced deep fades. Moreover, due to strict Size, Weight, and Power (SWaP) constraints on the satellite payload, the GEO receiver is equipped with only a single telescope, motivating the use of a MISO 2×1 architecture to achieve spatial diversity from the ground.

The transmission relies on Dual-Polarization Quadrature Phase Shift Keying (DP-QPSK) modulation. As demonstrated in our previous work, the Alamouti 2×1 diversity technique in a dual-polarization MISO architecture can be implemented in two distinct ways. Alamouti coding can be applied as Space-Polarization Coding (SPC), which exploits the space and polarization dimensions of a single symbol time-slot, or as Space-Time Coding (STC), which uses the spatial dimension across two consecutive symbol intervals per polarization. Although both schemes reach the same diversity and effectively combat atmospheric channel fading, we adopt the STC scheme in this work, as detailed in the next subsection. Before the Alamouti coding step, Polarization Pairwise Coding is applied. This technique combines a rotation of the constellation with a swapping of the In-phase (I) and Quadrature (Q) components between the two orthogonal polarizations (denoted X and Y for the horizontal and vertical polarizations, respectively). This precoding ensures that both polarizations experience an averaged channel degradation over the two transmitters in the MISO 2×1 system, thus mitigating the impact of PDL at the receiver. The theoretical analysis of this precoding technique is detailed in Section 3.

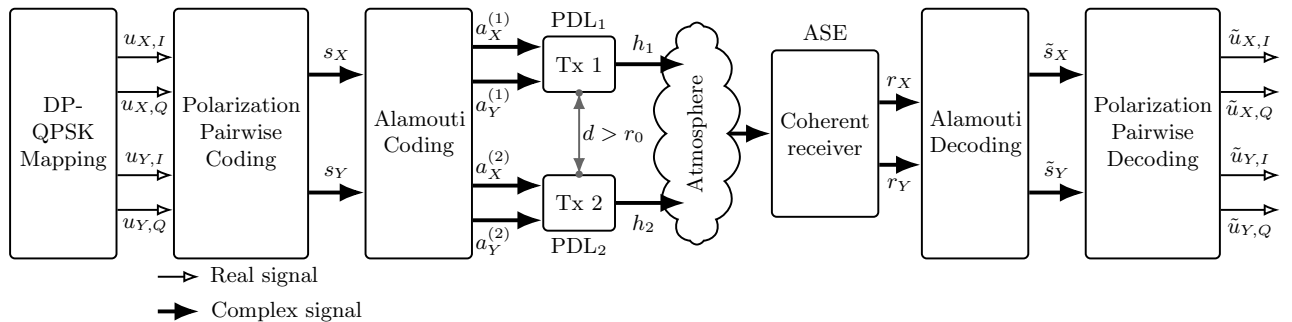


Figure 1: Block diagram of the proposed dual-polarization MISO 2×1 coherent FSO uplink architecture. The diagram illustrates the integration of Polarization Pairwise Coding with Space-Time Alamouti coding. Real-valued scalar signals are indicated by thin arrows with hollow heads, whereas complex-valued signals are represented by thick arrows with solid heads.

To evaluate the performance of this precoded architecture, the following subsections detail the complete mathematical framework of the system. First, Subsection 2.2 outlines the signal processing blocks at the transmitter. Next, Subsection 2.3 details the mathematical channel model that captures PDL on both transmitters, atmospheric turbulence, and additive Amplified Spontaneous Emission (ASE) noise at the receiver. Finally, Subsection 2.4 formulates the corresponding received signals under the assumption of ideal States of Polarization (SOP) as explained in Subsection 2.3.

2.2 Transmitter Processing Blocks

Before presenting a detailed theoretical analysis of the proposed DP-MISO system, this subsection sequentially details the different processing blocks and related transformations at the transmitter, as illustrated in Figure 1. The signal processing chain alternates between the complex domain and the real-valued domain (denoted $\mathbb{C}$ and $\mathbb{R}^2$, respectively).

2.2.1 DP-QPSK Mapping

The incoming bit stream is first mapped into independent complex symbols of a standard QPSK constellation. Let $u_X, u_Y \in \mathbb{C}$ denote the transmitted symbols on the horizontal (X) and vertical (Y) polarizations, respectively. To enable real-domain polarization processing, these complex symbols are decomposed into their In-phase (I) and Quadrature (Q) components. This defines the real-valued vector $\mathbf{u} \in \mathbb{R}^4$:

$$\mathbf{u} = \begin{pmatrix} u_{X,I} \\ u_{X,Q} \\ u_{Y,I} \\ u_{Y,Q} \end{pmatrix} = \begin{pmatrix} \Re(u_X) \\ \Im(u_X) \\ \Re(u_Y) \\ \Im(u_Y) \end{pmatrix},$$

where $\Re(\cdot)$ and $\Im(\cdot)$ are the real and imaginary parts of a complex number, respectively.

2.2.2 Polarization Pairwise Coding

The Polarization Pairwise Coding block operates in the real domain to mitigate PDL by combining constellation rotation and the swapping of the I and Q components across the two polarization states. Mathematically, this operation is defined as a generic linear transformation:

$$\mathbf{s} = \mathbf{T}\mathbf{u},$$

where $\mathbf{T} \in \mathbb{R}^{4 \times 4}$ is the overall transformation matrix detailed in Subsection 3.2. The resulting vector $\mathbf{s} \in \mathbb{R}^4$ is given by:

$$\mathbf{s} = \begin{pmatrix} s_{X,I} \\ s_{X,Q} \\ s_{Y,I} \\ s_{Y,Q} \end{pmatrix}.$$

Before entering the Alamouti encoder, this vector is mapped back into the complex domain to form the pairwise coded complex symbols for each polarization, denoted $s_X = s_{X,I} + js_{X,Q}$ and $s_Y = s_{Y,I} + js_{Y,Q}$.

2.2.3 Alamouti Space-Time Coding

To provide spatial diversity against atmospheric turbulence, the complex symbols s_X and s_Y are processed by an Alamouti Space-Time Coder. Because the atmospheric channel and the PDL are assumed to be quasi-static over at least two symbol durations, the encoding is performed over two consecutive time slots k and $k+1$.

To achieve spatial diversity, the symbols are mapped across the two ground transmitters according to the standard Alamouti scheme. For a given polarization (e.g., the X polarization), the mapping over the two transmitters and time-slots is defined as:

$$\begin{pmatrix} a_X^{(1)}[k] & a_X^{(1)}[k+1] \\ a_X^{(2)}[k] & a_X^{(2)}[k+1] \end{pmatrix} = \begin{pmatrix} s_X[k] & -s_X^*[k+1] \\ s_X[k+1] & s_X^*[k] \end{pmatrix},$$

where $(\cdot)^*$ denotes the complex conjugate and the superscript $(\cdot)^{(i)}$ designates the i -th transmitter. The Y polarization is encoded identically.

As previously introduced, spacing the apertures by a distance d that exceeds the Fried parameter ($d > r_0$) guarantees that the transmitted optical beams experience independent and uncorrelated atmospheric fading. This statistical independence is the fundamental prerequisite for realizing the theoretical spatial diversity gains of the Alamouti scheme.

2.3 Channel Model

The optical signals from the two transmitters propagate through the atmospheric channel, where they are subject to atmospheric turbulence, deterministic propagation losses, and hardware-induced losses. In this paper, the deterministic losses are taken into account in the static Received Optical Power (ROP) of the system.

The first and second channel coefficients for the optical paths from the first and second transmitting apertures are denoted by the complex coefficients h_1 and h_2 . Each coefficient h_i equally affects the X and Y polarizations of the i -th transmitter. Given the spatial separation of the ground telescopes, h_1 and h_2 are modeled as independent and identically distributed (i.i.d) variables with a distribution and a time correlation dependent on the scenario (aperture size, adaptive optics radial order, etc.) as detailed in Section 4.

In both transmitters, propagation through optical components introduces PDL. The PDL acts as an asymmetric attenuation between the orthogonal polarization states and can be modeled by a diagonal matrix given by

$$\mathbf{H}_{\text{PDL}} = \begin{pmatrix} \sqrt{1+\gamma} & 0 \\ 0 & \sqrt{1-\gamma} \end{pmatrix},$$

where the parameter γ quantifies the amplitude of the PDL ($-1 < \gamma < 1$). In the logarithmic scale, the PDL value in dB is expressed as

$$\Gamma_{dB} = 10 \log_{10} \left(\frac{1+\gamma}{1-\gamma} \right).$$

In MISO 2×1 , the two transmitters 1 and 2 are affected by their respective PDL matrices $\mathbf{H}_{\text{PDL}_1}$ and $\mathbf{H}_{\text{PDL}_2}$.

Finally, at the receiver side, the optical pre-amplification introduces Amplified Spontaneous Emission (ASE) noise which is modeled as an Additive White Gaussian Noise (AWGN) added to the received signal.

Given the system's vulnerability to State of Polarization (SOP) rotations,⁹ this model assumes that the SOP of the optical signals is controlled and perfectly aligned with the principal axes at both transmitting telescopes. Furthermore, we assume that the unknown SOP rotation at the receiver is perfectly estimated and compensated in the coherent receiver prior to the Alamouti Decoding stage depicted in Figure 1. This results in simplified received signals that can be easily combined and decoded.

2.4 Received Signals

Although this paper focuses on MISO 2×1 architecture, we first recall the standard Single-Input Single-Output (SISO) model to establish a baseline for PDL impairment.

For a SISO system, the received vector in the complex domain $\mathbb{C}$ over a single symbol period (time index omitted) is expressed as:

$$\begin{pmatrix} r_X \\ r_Y \end{pmatrix} = h \mathbf{H}_{\text{PDL}} \begin{pmatrix} s_X \\ s_Y \end{pmatrix} + \begin{pmatrix} n_X \\ n_Y \end{pmatrix},$$

where s_X and s_Y denote the QPSK symbols transmitted on the X and Y polarizations, respectively. The scalar h is the complex atmospheric fading coefficient, $\mathbf{H}_{\text{PDL}}$ models the PDL, and $(n_X, n_Y)^T \sim \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I}_2)$ is the AWGN vector, with N_0 representing the noise variance per polarization, and $\mathbf{I}_2$ denoting the 2×2 identity matrix.

For the Alamouti 2×1 DP system, the received signal vectors in the complex domain $\mathbb{C}$ over two consecutive discrete time-slots k and $k + 1$ are given by the coupled system:

$$\left\{ \begin{array}{l} \begin{pmatrix} r_X[k] \\ r_Y[k] \end{pmatrix} = h_1 \mathbf{H}_{\text{PDL}_1} \begin{pmatrix} a_X^{(1)}[k] \\ a_Y^{(1)}[k] \end{pmatrix} + h_2 \mathbf{H}_{\text{PDL}_2} \begin{pmatrix} a_X^{(2)}[k] \\ a_Y^{(2)}[k] \end{pmatrix} + \begin{pmatrix} n_X[k] \\ n_Y[k] \end{pmatrix} \\ \quad = h_1 \mathbf{H}_{\text{PDL}_1} \begin{pmatrix} s_X[k] \\ s_Y[k] \end{pmatrix} + h_2 \mathbf{H}_{\text{PDL}_2} \begin{pmatrix} s_X[k+1] \\ s_Y[k+1] \end{pmatrix} + \begin{pmatrix} n_X[k] \\ n_Y[k] \end{pmatrix} \\ \begin{pmatrix} r_X[k+1] \\ r_Y[k+1] \end{pmatrix} = h_1 \mathbf{H}_{\text{PDL}_1} \begin{pmatrix} a_X^{(1)}[k+1] \\ a_Y^{(1)}[k+1] \end{pmatrix} + h_2 \mathbf{H}_{\text{PDL}_2} \begin{pmatrix} a_X^{(2)}[k+1] \\ a_Y^{(2)}[k+1] \end{pmatrix} + \begin{pmatrix} n_X[k+1] \\ n_Y[k+1] \end{pmatrix} \\ \quad = -h_1 \mathbf{H}_{\text{PDL}_1} \begin{pmatrix} s_X^*[k+1] \\ s_Y^*[k+1] \end{pmatrix} + h_2 \mathbf{H}_{\text{PDL}_2} \begin{pmatrix} s_X^*[k] \\ s_Y^*[k] \end{pmatrix} + \begin{pmatrix} n_X[k+1] \\ n_Y[k+1] \end{pmatrix} \end{array} \right., \quad (1)$$

where the coefficients h_i and the matrices $\mathbf{H}_{\text{PDL}_i}$ for $i \in \{1, 2\}$ correspond to the atmospheric fading and PDL of the i -th transmitter, respectively. The noise vectors at each time-slot are independent and identically distributed as $(n_X[k], n_Y[k])^T \sim \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I}_2)$.

3. POLARIZATION PAIRWISE CODING FOR PDL MITIGATION

To ensure mathematical clarity throughout this section, we adopt a notational convention to distinguish between the operation domains. Variables denoted with the subscript eq (e.g., $\mathbf{r}_{eq} \in \mathbb{C}^4$, $\mathbf{H}_{eq} \in \mathbb{C}^{4 \times 4}$) represent the equivalent space-time MISO system operating over two consecutive time-slots in the complex domain. In contrast, variables without this subscript (e.g., $\mathbf{s} \in \mathbb{R}^4$, $\mathbf{R} \in \mathbb{R}^{4 \times 4}$) strictly refer to real-valued vectors and matrices acting on the In-phase and Quadrature components within a single time-slot.

3.1 Motivation for Polarization Pairwise Coding

As stated in Section 1, the primary challenge of implementing advanced polarization-time codes (PTCs) in a satellite payload is their inherent computational complexity. Classical 2-timeslot polarization codes (such as the Golden and Silver codes¹³) require a Maximum Likelihood (ML) decoder, e.g., a Sphere Decoder or similar, to perform at their best. This high decoding complexity makes them unsuited for our objective of minimizing the satellite's computational load. In contrast, Polarization Pairwise Coding²¹ offers a highly effective and significantly simpler alternative. It achieves efficient PDL mitigation through low-complexity linear decoding. Recent studies on other single-time-slot coding methods, such as the IQ-code, confirm that single-interval processing significantly reduces decoding complexity compared to multi-slot spatial coding techniques, and maintains complete compatibility with existing DSP blocks, such as low-complexity Zero-Forcing (ZF) decoders.²⁰

Beyond DSP complexity, the pairwise approach is expected to offer complementary operational advantages. By avoiding the constellation expansion seen in classical PTCs, it maintains a low Peak-to-Average Power Ratio (PAPR), comparable to low-PAPR Number Theory (NT) code¹⁴ studied in fiber propagation. Moreover, symbol-by-symbol processing does not introduce any additional decoding delay into the system.

3.2 Integration of Polarization Pairwise Coding within the MISO Architecture

Originally proposed for DP-SISO systems, Polarization Pairwise Coding mitigates PDL by physically swapping the In-phase (I) of the Y -polarization and Quadrature (Q) of the X -polarization components. In our extended MISO architecture, the decision to apply this technique before Alamouti encoding is motivated by two primary advantages at the receiver. First, as will be detailed later, performing Alamouti decoding first allows the system to recover the spatial diversity, thereby increasing the average Signal-to-Noise Ratio (SNR) before compensating for the PDL. Second, this arrangement enables a natural decoupling of the receiver operations (i.e., Alamouti

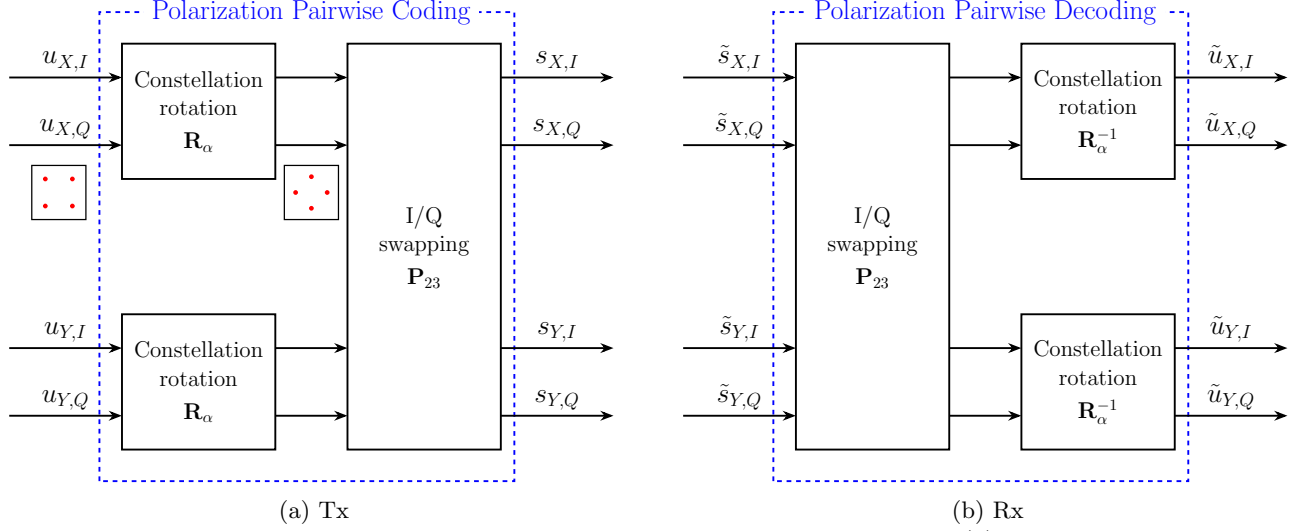


Figure 2: Block diagram of the Pairwise Polarization Coding at the transmitter Tx (a) and Pairwise Polarization Decoding at the receiver Rx (b).

decoding followed by pairwise decoding). Reversing this order at the transmitter would substantially increase receiver complexity, as joint processing would be required to avoid significant performance degradation.

As depicted in Figure 1, the I/Q swapping in the MISO system is performed prior to Alamouti coding. It consists of two steps: a constellation rotation over the transmitted symbols before Alamouti coding. Then, the I/Q swapping between the In-phase (I) and Quadrature (Q) components over the X and Y polarizations is performed.

Hereafter we summarize the main operations applied to the real-valued (4D) transmitted signals over the I and Q components in $\mathbb{R}^4$ before Alamouti coding (see Figure 2a):

- **Constellation rotation:** this can be represented using a real rotation matrix of angle α , $\mathbf{R}_\alpha = \begin{pmatrix} \cos \alpha & -\sin \alpha \\ \sin \alpha & \cos \alpha \end{pmatrix}$.

We apply the rotation matrix to the In-phase (I) and Quadrature (Q) components of both polarizations as follows

$$\mathbf{R} = \text{diag}(\mathbf{R}_\alpha, \mathbf{R}_\alpha) = \begin{pmatrix} \mathbf{R}_\alpha & \mathbf{0}_{2 \times 2} \\ \mathbf{0}_{2 \times 2} & \mathbf{R}_\alpha \end{pmatrix}.$$

- **Permutation matrix:** it corresponds to the swapping of I and Q components. In the case of Zhu et al.'s Polarization Pairwise Coding,²¹ the permutation matrix used is

$$\mathbf{P}_{23} = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix}.$$

This permutation swaps coordinate 2 (X, Q) with coordinate 3 (Y, I) in the 4D-flattened transmitted signal. While $\mathbf{P}_{23}$ is used in the original Pairwise Polarization Coding,²¹ alternative permutation matrices such as $\mathbf{P}_{13}$, $\mathbf{P}_{14}$, or $\mathbf{P}_{24}$ can be used as discussed in the next subsection, where $\mathbf{P}_{xy}$ denotes the permutation matrix associated with the permutation of coordinates x and y of a given 4D vector.

The transmitted signals are then :

$$\mathbf{s} = \mathbf{P}_{23}\mathbf{R}\mathbf{u}, \quad (2)$$

where $\mathbf{u} \in \mathbb{R}^4$ is the real-valued vector of the DP-QPSK symbols and $\mathbf{T} = \mathbf{P}_{23}\mathbf{R} \in \mathbb{R}^{4 \times 4}$ is the overall transformation matrix of this precoding, as defined in Subsection 2.2.2.

3.3 Theoretical Performance Analysis

3.3.1 Alamouti Decoding and Noise Stability

To analyze the performance of the PDL mitigation technique used in our MISO architecture, it is convenient to rewrite the received signals over two consecutive time-slots in Equation 1 into an equivalent 4×1 complex vector representation $\mathbf{r}_{eq}$. Taking the complex conjugate of the received signal vector during the second Alamouti time slot, we can express the global system in a vector form to expose explicitly the transmitted symbols as follows

$$\begin{aligned} \mathbf{r}_{eq} &= \begin{pmatrix} r_X[k] \\ r_Y[k] \\ r_X^*[k+1] \\ r_Y^*[k+1] \end{pmatrix} = \begin{pmatrix} h_1 \mathbf{H}_{PDL_1} & h_2 \mathbf{H}_{PDL_2} \\ h_2^* \mathbf{H}_{PDL_2} & -h_1^* \mathbf{H}_{PDL_1} \end{pmatrix} \begin{pmatrix} s_X[k] \\ s_Y[k] \\ s_X[k+1] \\ s_Y[k+1] \end{pmatrix} + \begin{pmatrix} n_X[k] \\ n_Y[k] \\ n_X^*[k+1] \\ n_Y^*[k+1] \end{pmatrix} \\ &= \mathbf{H}_{eq} \mathbf{s}_{eq} + \mathbf{n}_{eq}, \end{aligned} \quad (3)$$

where $\mathbf{H}_{eq} \in \mathbb{C}^{4 \times 4}$ is the equivalent space-time channel matrix capturing the combined effects of fading and PDL, defined explicitly as:

$$\mathbf{H}_{eq} = \begin{pmatrix} h_1 \mathbf{H}_{PDL_1} & h_2 \mathbf{H}_{PDL_2} \\ h_2^* \mathbf{H}_{PDL_2} & -h_1^* \mathbf{H}_{PDL_1} \end{pmatrix},$$

and $\mathbf{n}_{eq}$ is the corresponding equivalent complex noise. Since the complex conjugation of circularly symmetric Gaussian noise does not change its statistical properties, $\mathbf{n}_{eq} \sim \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I}_4)$, where $\mathbf{I}_4$ is the 4×4 identity matrix.

To apply this coding to the MISO 2×1 architecture, the I/Q permutation is performed before the Alamouti coding step. At the receiver, the first step is to decouple the spatial streams using standard Alamouti linear decoding. Using the equivalent STC system matrix $\mathbf{H}_{eq}$ defined above, the receiver multiplies the combined signal vector $\mathbf{r}_{eq}$ by the Hermitian transpose of the channel matrix, $\mathbf{H}_{eq}^H$. The decoded signal vector $\tilde{\mathbf{s}}_{eq}$ is given by:

$$\begin{aligned} \tilde{\mathbf{s}}_{eq} &= \mathbf{H}_{eq}^H \mathbf{r}_{eq} \\ &= (\mathbf{H}_{eq}^H \mathbf{H}_{eq}) \mathbf{s} + \mathbf{H}_{eq}^H \mathbf{n}_{eq} \\ &= \mathbf{D}_{eq} \mathbf{s}_{eq} + \tilde{\mathbf{n}}_{eq} \end{aligned} \quad (4)$$

Because the real diagonal PDL matrices commute ($\mathbf{H}_{PDL_1} \mathbf{H}_{PDL_2} = \mathbf{H}_{PDL_2} \mathbf{H}_{PDL_1}$), the off-diagonal 2×2 blocks of $\mathbf{H}_{eq}^H \mathbf{H}_{eq}$ perfectly cancel out. This results in a block-diagonal effective channel matrix $\mathbf{D}_{eq}$:

$$\mathbf{D}_{eq} = \mathbf{H}_{eq}^H \mathbf{H}_{eq} = \begin{pmatrix} |h_1|^2 \mathbf{H}_{PDL_1}^2 + |h_2|^2 \mathbf{H}_{PDL_2}^2 & \mathbf{0}_{2 \times 2} \\ \mathbf{0}_{2 \times 2} & |h_1|^2 \mathbf{H}_{PDL_1}^2 + |h_2|^2 \mathbf{H}_{PDL_2}^2 \end{pmatrix} = \begin{pmatrix} g_X & 0 & 0 & 0 \\ 0 & g_Y & 0 & 0 \\ 0 & 0 & g_X & 0 \\ 0 & 0 & 0 & g_Y \end{pmatrix},$$

where $g_X = |h_1|^2(1 + \gamma_1) + |h_2|^2(1 + \gamma_2)$ and $g_Y = |h_1|^2(1 - \gamma_1) + |h_2|^2(1 - \gamma_2)$ are the effective gains.

The covariance matrix of the resulting noise vector $\tilde{\mathbf{n}}$ after Alamouti decoding is:

$$\mathbf{K}_{\tilde{\mathbf{n}}_{eq}} = \mathbb{E} [\tilde{\mathbf{n}}_{eq} \tilde{\mathbf{n}}_{eq}^H] = \mathbf{H}_{eq}^H \mathbb{E} [\mathbf{n}_{eq} (\mathbf{n}_{eq})^H] \mathbf{H}_{eq} = N_0 \mathbf{H}_{eq}^H \mathbf{H}_{eq} = N_0 \mathbf{D}_{eq} \quad (5)$$

As $\mathbf{D}_{eq}$ is diagonal and the effective noise powers (given by $N_0 g_X$ and $N_0 g_Y$, respectively) repeat identically across the two time-slots, the noise terms remain completely independent and have the same variance per-polarization at time-slots k and $k+1$. This property justifies analyzing the Polarization Pairwise Coding over a single time-slot.

3.3.2 Post-Decoding Noise Covariance Matrix

Since the effective noise variance and channel gains repeat identically across k and $k + 1$, the two time-slots are statistically equivalent after Alamouti decoding. This property allows us to omit the time index and formulate an equivalent system model over a single time-slot without loss of generality. The real-valued effective channel gain matrix extracted from $\mathbf{D}_{eq}$ is given by $\mathbf{D} = \text{diag}(g_X, g_X, g_Y, g_Y)$. The post-Alamouti Decoding received signal in the 4D real domain is:

$$\tilde{\mathbf{s}} = \mathbf{D}\mathbf{s} + \tilde{\mathbf{n}} = \mathbf{D}\mathbf{P}_{23}\mathbf{R}\mathbf{u} + \tilde{\mathbf{n}} \quad (6)$$

To recover the original symbols $\mathbf{u}$, the receiver applies the inverse of the transmitter's pairwise coding operations (Figure 2b). It first de-interleaves the signal by applying the inverse permutation $\mathbf{P}_{23}^T$ as we have $\mathbf{P}_{23}^T = \mathbf{P}_{23}^{-1} = \mathbf{P}_{23}$, and then de-rotates the constellation by applying $\mathbf{R}^T$ (with $\mathbf{R}^T = \mathbf{R}^{-1}$):

$$\tilde{\mathbf{u}} = \mathbf{R}^T \mathbf{P}_{23}^T \tilde{\mathbf{s}} = (\mathbf{R}^T \mathbf{P}_{23}^T \mathbf{D} \mathbf{P}_{23} \mathbf{R}) \mathbf{u} + \tilde{\mathbf{n}}_{decod}, \quad (7)$$

where $\tilde{\mathbf{n}}_{decod} = \mathbf{R}^T \mathbf{P}_{23}^T \tilde{\mathbf{n}}$ is the final noise after I/Q de-interleaving.

To prove that the PDL effect is symmetrically distributed between the components I and Q , we analyze the covariance matrix of this decoded noise vector $\mathbf{K}_{\tilde{\mathbf{n}}_{decod}}$. As the global complex noise variance is N_0 , the variance is halved to $N_0/2$ per branch when separating the signal into its I and Q components. Thus, the covariance of the real-valued post-Alamouti decoding noise $\tilde{\mathbf{n}}$ is $\frac{N_0}{2} \mathbf{D}$. We obtain:

$$\mathbf{K}_{\tilde{\mathbf{n}}_{decod}} = \mathbb{E} [\tilde{\mathbf{n}}_{decod} \tilde{\mathbf{n}}_{decod}^T] = \mathbf{R}^T \mathbf{P}_{23}^T \left(\frac{N_0}{2} \mathbf{D} \right) \mathbf{P}_{23} \mathbf{R}. \quad (8)$$

By performing matrix multiplication and substituting the explicit definitions of g_X and g_Y , we can isolate the total spatial channel gain $\Sigma = |h_1|^2 + |h_2|^2$ and the effective PDL imbalance $\Delta = \gamma_1 |h_1|^2 + \gamma_2 |h_2|^2$. The covariance matrix simplifies to:

$$\mathbf{K}_{\tilde{\mathbf{n}}_{decod}} = \frac{N_0}{2} \begin{pmatrix} \Sigma + \Delta \cos(2\alpha) & -\Delta \sin(2\alpha) & 0 & 0 \\ -\Delta \sin(2\alpha) & \Sigma - \Delta \cos(2\alpha) & 0 & 0 \\ 0 & 0 & \Sigma + \Delta \cos(2\alpha) & -\Delta \sin(2\alpha) \\ 0 & 0 & -\Delta \sin(2\alpha) & \Sigma - \Delta \cos(2\alpha) \end{pmatrix}. \quad (9)$$

This transformation obtained proves the elegance of the technique. The permutation $\mathbf{P}_{23}$ forces both the X and Y polarizations to experience a contrasted noise variance according to the rotation angle α . A mathematically equivalent behavior is observed when employing any of the alternative permutation matrices ($\mathbf{P}_{13}$, $\mathbf{P}_{14}$, or $\mathbf{P}_{24}$) introduced in the previous subsection. *For the optimal rotation angle $\alpha = \frac{\pi}{4}$, the asymmetric terms on the diagonal are eliminated ($\cos(2\alpha) = 0$), which perfectly symmetrizes the noise distribution across the I and Q components of the polarizations X and Y at the receiver.*

3.4 Preliminary AWGN Performance Analysis

Before introducing atmospheric fading, it is essential to validate the effectiveness of the Polarization Pairwise Coding technique in the MISO context. To isolate the effects of this precoding technique and the asymmetric PDL, we first establish a preliminary baseline using an Additive White Gaussian Noise (AWGN) channel. In this scenario, the atmospheric fading coefficients are completely static, and the total transmitted power is equally split between the two MISO apertures.

Although the instantaneous PDL introduced by optical components is a random variable, practical system designs, such as those adopted by Thales Alenia Space, are constrained by maximum PDL values imposed by hardware specifications. Because the MISO system duplicates the exact same components for both transmitters, it is relevant to evaluate the system in a symmetric worst-case scenario. Therefore, we assume that both transmitters operate at their maximum specified PDL limit (i.e., $\gamma_1 = \gamma_2 = \gamma$). Under this assumption and

without precoding, the theoretical loss of SNR is determined by the degraded polarization of the system (the Y-polarization when $\gamma > 0$), resulting in an SNR loss of $10 \log_{10}(1 - \gamma)$. However, by averaging the channel gains across both polarizations, proper PDL mitigation precoders using joint coding with parallel and independent decoding¹⁵ reduce this penalty to at most $10 \log_{10}(1 - \gamma^2)$.

Table 1 quantifies this theoretical gain for the symmetric impairment scenario. At a severe PDL of 6 dB, averaging the channel provides an SNR gain of almost 2 dB, demonstrating the massive potential to avoid the worst-case alignment.

Table 1: SNR loss due to PDL in an AWGN channel.

PDL value	SNR loss without precoding: $10 \log_{10}(1 - \gamma)$	SNR loss with precoding: $10 \log_{10}(1 - \gamma^2)$	SNR gain (Δ)
PDL = 1.5 dB	0.81 dB	0.13 dB	0.68 dB
PDL = 3 dB	1.75 dB	0.5 dB	1.25 dB
PDL = 6 dB	3.90 dB	1.92 dB	1.98 dB

Figure 3 illustrates the achievable rate (in bits/symbol) as a function of the constellation rotation angle α for the two symmetric PDL regimes evaluated in Table 1. Without precoding, the severely degraded Y polarization (red squares) defines the worst-case scenario, suffering a capacity loss that creates a severe system bottleneck. The overall non-precoded performance is represented by the mean achievable rate (dashed line).

When the Polarization Pairwise Coding is applied (blue crosses and green diamonds), the achievable rates of the X and Y polarizations perfectly overlap, demonstrating that the asymmetric channel impairment is successfully balanced. Both polarizations exhibit a sinusoidal dependence on α . This behavior directly validates the theoretical decoded noise covariance matrix $\mathbf{K}_{\tilde{n}_{decod}}$ derived in Section 3.3. For instance, at $\alpha = 0$, the $\cos(2\alpha)$ term maximizes the noise variance asymmetry across the I and Q components. Because achievable rate measures mutual information, the orthogonal I and Q capacities combine to reach the geometric mean capacity.

Furthermore, as predicted by the theoretical model, setting the rotation angle to the optimal value of $\alpha = \pi/4$ forces the $\cos(2\alpha)$ terms to zero. At this optimal operating point, the achievable rate slightly exceeds the mean non-precoded baseline due to the noise correlation introduced by the off-diagonal terms of the covariance matrix. This proves that the Polarization Pairwise Coding successfully and robustly mitigates the asymmetric PDL even in our MISO system.

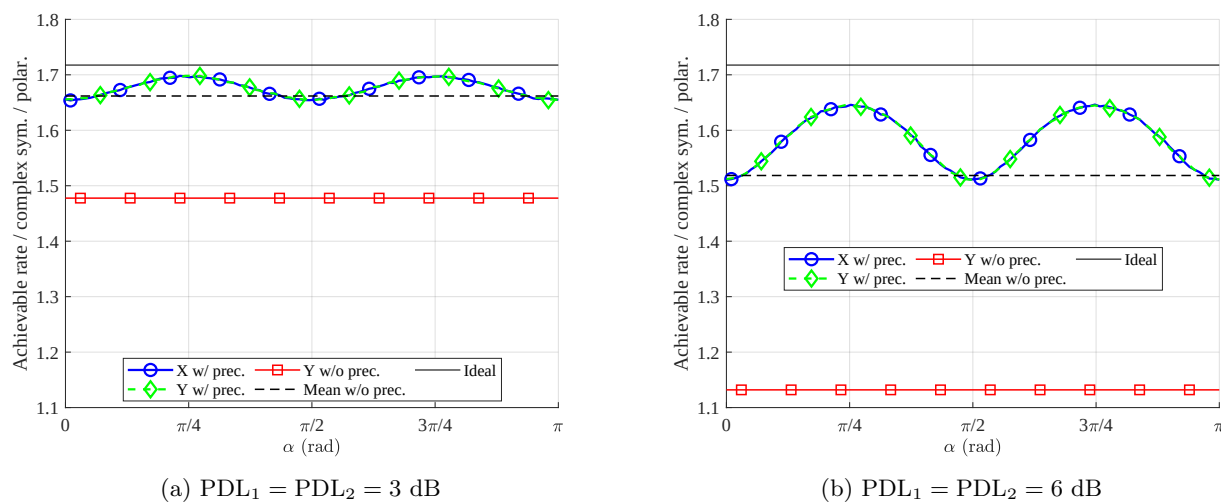


Figure 3: Achievable rate per polarization and per symbol (bits/symbol) of the MISO 2×1 system over an AWGN channel as a function of the rotation angle α (rad), under symmetric PDL regimes at SNR = 5 dB.

Having validated this mitigation technique under static AWGN conditions, it is now necessary to evaluate the system under the dynamic fading conditions characteristic of real-world optical ground-to-GEO links.

4. SIMULATION RESULTS AND PERFORMANCE EVALUATION

This section presents a performance comparison between SISO and MISO architectures over a realistic atmospheric fading channel, evaluating the impact of the Polarization Pairwise Coding technique. We report numerical results for various system configurations and discuss the architectural trade-offs revealed by the simulations.

4.1 Simulation Parameters and Key Performance Metrics

To evaluate the system under realistic dynamic conditions, the Free Space Optical (FSO) link is simulated using advanced time-series used in industrial system design at Thales Alenia Space. These time-series accurately emulate a realistic ground-to-GEO link, capturing both the dynamically varying atmospheric turbulence and the residual signal fading after Adaptive Optics (AO) correction.

Two distinct uplink scenarios are studied: a 20 cm Low Order Adaptive Optics (LO-AO) scenario and a 50 cm High Order Adaptive Optics (HO-AO) scenario. It is important to note that these dimensions refer strictly to the transmit aperture diameter (D_{Tx}), while the receiver aperture (D_{Rx}) remains fixed at 30 cm in both configurations.

The primary parameters for these scenarios are summarized in Table 2. The 20 cm LO-AO configuration results in a static Received Optical Power (ROP) of -41 dBm, while the larger 50 cm HO-AO configuration achieves a static ROP of -33 dBm. Because all other environmental and geographical parameters (such as the elevation angle and the Fried parameter r_0) remain identical, this methodology perfectly isolates the specific effects of the transmit aperture size and its corresponding AO correction capabilities on the overall link robustness.

Table 2: Time-Series Uplink Scenarios

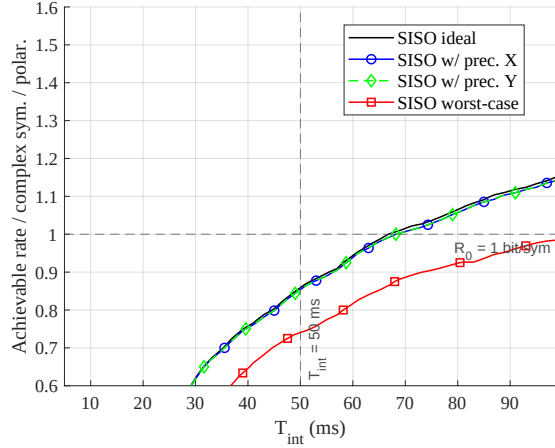
Parameter	Scenario 20 cm LO-AO	Scenario 50 cm HO-AO
D_{Tx} (cm)	20	50
D_{Rx} (cm)	30	30
Elevation ($^\circ$)	30	30
r_0 (cm)	5.11	5.11
AO radial orders	4	12
ROP (dBm)	-41	-33

Because the turbulent channel causes deep signal fades, temporal interleaving is required to ensure reliable communication. This is emulated by averaging the instantaneous Mutual Information (MI) over a moving time window of duration T_{int} expressed in milliseconds (ms). The outage probability is defined as $P_{out} = \Pr(MI < R_0)$, where R_0 is the per-polarization target rate with $R_0 = 2R_c$ for QPSK modulation and R_c is the Forward Error Correction (FEC) code rate ($0 < R_c < 1$). Consistent with FSO Very High Throughput Systems (VHTS) requirements²² adopted in Thales Alenia Space designs, our primary performance metric is the achievable target rate per polarization and per QPSK symbol in bits per symbol plotted as a function of the interleaving depth T_{int} , evaluated at a fixed target availability of 99.9% ($P_{out} = 10^{-3}$).

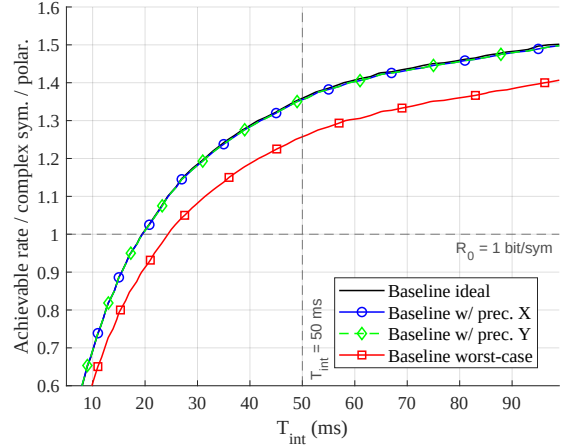
4.2 Compared Configurations and PDL Assumptions

Taking into account the hardware architectures and power allocation strategies established in our previous work,⁹ we evaluated three MISO configurations against a Reference SISO 50 cm HO-AO system. To ensure a fair comparison, the assumed PDL impairment for each configuration is derived directly from its specific optical amplification requirements:

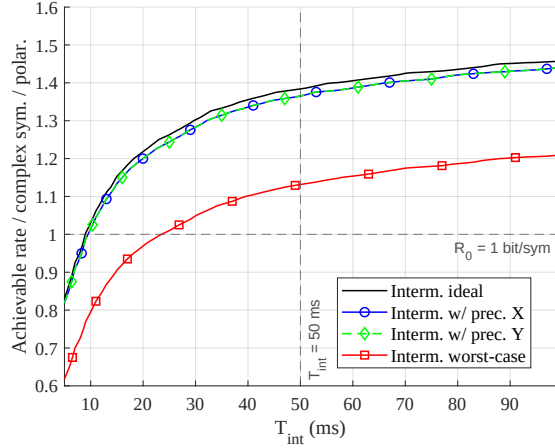
- **Reference SISO (50 cm HO-AO) configuration:** A standard single-aperture transmission chain utilizing a high-gain optical amplifier. We assume a baseline component PDL of 3 dB.



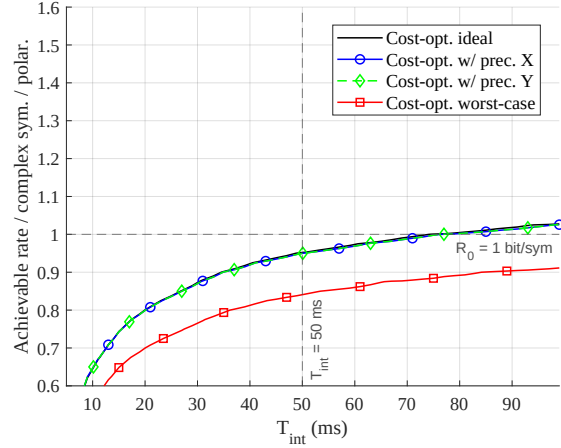
(a) Reference SISO (PDL = 3 dB)



(b) Baseline MISO (PDL = 1.5 dB)



(c) Intermediate MISO (PDL = 3 dB)



(d) Cost-optimized MISO (PDL = 1.5 dB)

Figure 4: Achievable rate per polarization and per symbol vs interleaving depth T_{int} (ms) at $P_{\text{out}} = 10^{-3}$.

- **Power-Constrained MISO configurations:** This category includes the **Baseline MISO** (with two 50 cm HO-AO apertures) and **Cost-optimized MISO** (with two 20 cm LO-AO apertures). Both configurations conserve the total transmitted optical power, which means that each MISO aperture operates at half the power (-3 dB) of the SISO baseline. This reduced power requirement permits the use of lower-gain DP amplifiers on each transmitter, which introduce less PDL. Therefore, we assume that $\text{PDL}_1 = \text{PDL}_2 = 1.5$ dB.
- **Power-Augmented MISO configuration (Intermediate MISO with two 20 cm LO-AO apertures):** This configuration doubles the total transmit power by assigning each 20 cm aperture high-gain amplifier, identical to the one used in the Reference SISO. As a result, each branch experiences the same severe PDL ($\text{PDL}_1 = \text{PDL}_2 = 3$ dB).

4.3 Performance Evaluation of the Precoded MISO Configurations

Before comparing the distinct MISO architectures against the SISO baseline, it is crucial to quantify the internal performance gains provided by the Polarization Pairwise Coding within each specific configuration.

Figure 4 demonstrates the critical role of pairwise coding in preventing the system from being limited by the worst-case polarization state. By effectively balancing the polarizations, the precoding ensures that all

configurations maintain an achievable rate nearly identical to their respective ideal PDL-free scenarios.

To illustrate the practical impact, Table 3 compares the required interleaving duration (T_{int}) needed to reach a target spectral efficiency of 1 bit/symbol (corresponding to a standard Forward Error Correction (FEC) code rate of 1/2).

Table 3: Comparison of required interleaving depth (T_{int}) to achieve a target spectral efficiency of 1 bit/symbol.

Configuration	Total Tx Power	Assumed PDL	Required T_{int} (ms)*	
			Without Precoding	With Precoding
Reference SISO (Fig. 4a)	Ref.	3.0 dB	> 100	≈ 68
Baseline MISO (Fig. 4b)	Ref.	1.5 dB	≈ 26	≈ 19.5
Intermediate MISO (Fig. 4c)	$2 \times \text{Ref.}$	3.0 dB	≈ 23.5	≈ 9.5
Cost-Optimized MISO (Fig. 4d)	Ref.	1.5 dB	> 100	≈ 77

* lower is better

As summarized in Table 3, the absence of precoding leaves the system limited by the severely degraded polarization, demanding a significantly longer interleaving depth to reach the target rate. In fact, for the SISO and Cost-Optimized configurations, the target rate cannot even be reached within a 100 ms window without precoding. This mitigating behavior applies consistently across all architectures. For the Baseline configuration, the precoding successfully closes the gap caused by its 1.5 dB PDL, reducing the necessary memory from 26 ms to 19.5 ms. The impact is most pronounced in the Intermediate scenario, where precoding reduces the required memory and latency by a factor of more than two (from approximately 24 ms to 9.5 ms).

In addition, the advantages of pairwise coding can be observed from a capacity perspective. Table 4 highlights the capacity improvements at a fixed interleaving depth of 50 ms.

Table 4: Comparison of achievable rate per polarization and per symbol at a fixed interleaving depth of $T_{\text{int}} = 50$ ms.

Configuration	Total Tx Power	Assumed PDL	Achievable Rate** (bits/sym)	
			Without Precoding	With Precoding
Reference SISO (Fig. 4a)	Ref.	3.0 dB	≈ 0.74	≈ 0.85
Baseline MISO (Fig. 4b)	Ref.	1.5 dB	≈ 1.26	≈ 1.36
Intermediate MISO (Fig. 4c)	$2 \times \text{Ref.}$	3.0 dB	≈ 1.13	≈ 1.36
Cost-Optimized MISO (Fig. 4d)	Ref.	1.5 dB	≈ 0.84	≈ 0.95

** higher is better

As demonstrated in Table 4, operating at a fixed interleaving depth of 50 ms without precoding results in a noticeable penalty due to the worst-case polarization. By applying precoding, the system effectively increases the achievable rates across all configurations to nearly match their respective ideal PDL-free configurations.

Figure 5 compares the overall performance of the precoded SISO and MISO architectures. As summarized in Table 5, MISO architectures generally outperform the Reference SISO in the low latency regime (i.e., small T_{int} values). For the MISO Intermediate and MISO Baseline configurations, the required interleaver depth at the target rate of 1 bit/symbol is significantly reduced to approximately 9.5 ms and 19.5 ms, respectively, compared to the Reference SISO which requires nearly 68 ms. Although the MISO Cost-Optimized configuration requires slightly more interleaver memory than the Reference SISO to reach 1 bit/symbol (~ 77 ms), it still outperforms the reference at interleaving depths up to $T_{\text{int}} \approx 65$ ms. This shows that spatial diversity can effectively compensate for reduced transmit power and smaller ground apertures at shorter interleaving durations.

At a fixed interleaving depth of $T_{\text{int}} = 50$ ms, both the MISO Baseline and MISO Intermediate configurations achieve nearly identical high performance, reaching an achievable rate per polarization and per symbol of

Table 5: Performance comparison of Reference SISO and MISO configurations at key thresholds.

Configuration	Total Tx Power	Assumed PDL	Required T_{int} at 1 bit/sym (ms) *	Achievable Rate at 50 ms (bits/sym) **
Reference SISO	Ref.	3.0 dB	≈ 68.0	≈ 0.85
MISO Baseline	Ref.	1.5 dB	≈ 19.5 (−71%)	≈ 1.36 (+60%)
MISO Intermediate	$2 \times \text{Ref.}$	3.0 dB	≈ 9.5 (−86%)	≈ 1.36 (+60%)
MISO Cost-Optimized	Ref.	1.5 dB	≈ 77.0 (+13%)	≈ 0.95 (+12%)

* lower is better

** higher is better

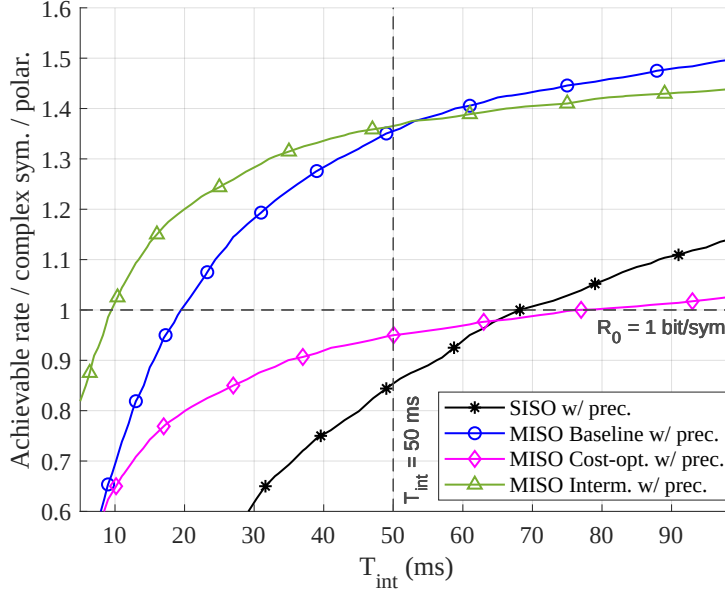


Figure 5: Achievable rate per polarization and per symbol (bits/symbol) vs interleaving depth T_{int} (ms) at $P_{\text{out}} = 10^{-3}$, comparing configurations with PDL = 3 dB (SISO and MISO Intermediate) against those with PDL = 1.5 dB (MISO Baseline and MISO Cost-optimized).

approximately 1.36 bits/symbol. The MISO Intermediate configuration operates with smaller 20 cm transmit apertures, reduced adaptive optics radial orders, and suffers from a severe 3 dB PDL penalty due to its high-gain amplifiers. However, its doubled transmit power, combined with the spatial diversity and the robust Polarization Pairwise Coding, allows it to perfectly match the performance of the MISO Baseline system (which relies on massive 50 cm transmit apertures and experiences only 1.5 dB of PDL).

Ultimately, extending our previous hardware trade-off studies, these results show that the proposed MISO architecture provides flexible design choices regarding power amplification, aperture size, adaptive optics complexity, and required interleaver memory. Additionally, it successfully mitigates the Polarization-Dependent Loss (PDL), the hardware impairment addressed in this study, which otherwise limits the overall system link budget.

5. CONCLUSION AND PERSPECTIVES

This study evaluated the performance of MISO ground-to-GEO optical feeder links under Polarization-Dependent Loss (PDL). Because both the Reference SISO and MISO architectures are vulnerable to this degradation, which limits the overall system link budget, finding an efficient mitigation strategy without increasing receiver complexity is essential. Extending our previous system-level trade-off analysis, we integrated the Polarization Pairwise Coding technique into the MISO system to mitigate the capacity loss caused by worst-case PDL.

A major advantage of this integrated technique is its remarkably low complexity and simple implementation. By strategically switching the In-phase (I) and Quadrature (Q) components between the orthogonal X and Y polarizations, the coding perfectly distributes the noise power over the I and Q components of each polarization. Furthermore, the decoupling of the data streams requires only linear processing at the receiver, entirely avoiding the massive computational complexity typically associated with classical Polarization Time Coding (PTC) techniques.

Our realistic time-series simulations confirm that this simple precoding minimizes PDL penalties, enabling a system link budget that closely matches the ideal PDL-free scenario. By effectively averaging the channel gains across both polarizations, the coding ensures that all compared MISO configurations maintain an achievable rate nearly identical to their ideal PDL-free scenarios. Consequently, precoded MISO architectures can significantly mitigate PDL while preserving the flexibility to relax hardware constraints regarding power amplification, aperture size, adaptive optics, and required memory (interleaving depth).

Although this work successfully demonstrates the resilience of the MISO architecture against both atmospheric turbulence and PDL, future studies must evaluate its robustness against other practical hardware imperfections such as synchronization skews, I/Q imbalances, and phase noise, moving toward a realistic channel model for next-generation ground-to-GEO optical links.

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