

# Non-coherent MIMO Communication for the 5th Generation Mobile: Overview and Practical Aspects

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## Abstract

Current cellular technologies are based on the concept of coherent communication, in which the channel matrix used for demodulation is estimated via reference or pilot signals. Coherent systems involve a significant increase of the signaling overhead, either when the number of Transmission Points (TP) is increased, due to the use of Coordinated Multipoint transmission/reception (CoMP) with Multiple-Input Multiple-Output (MIMO) processing, or when mobile channel changes rapidly. Another disadvantage of coherent communications is the performance degradation caused by channel estimation errors. Both drawbacks of coherent communication motivate the use of non-coherent techniques. Although there are many theoretical studies on the performance of non-coherent schemes in MIMO systems, their impact on real-world cellular systems is still unknown. This paper focuses on bringing non-coherent techniques into practical systems using CoMP and/or MIMO processing.

**Keywords:** Non-coherent, MIMO, 5G

## 1. Introduction

Cellular communication systems are continuously evolving to satisfy the highly increasing user demands. The future information society is expected to support very high data rates in dense crowds of users and in very heterogeneous scenarios. This challenging requirement has triggered the research activities towards the design of 5th Generation mobile networks (5G), where the European Union project Mobile Enablers for the Twenty-twenty Information Society (METIS) is playing a key role [1]. So far,

the METIS consortium has identified a set of scenarios and requirements to be addressed by 5G systems. As foreseen by the METIS project, wireless communication systems beyond 2020 will face a large set of requirements and use-cases that cannot be fulfilled solely by existing technologies. Of particular relevance are the needs for providing ten to one hundred times higher average user data rate per cell than today's cellular systems, while improving the energy efficiency and minimizing the cost and spectrum utilization.

It is well known that coherent communications dominate the state-of-the-art of current cellular technologies. Coherent receivers require perfect knowledge of the instantaneous channel variations (i.e., perfect Channel State Information at the Receiver (CSIR)) in order to demodulate the received information. The classical approach to obtain CSIR is to employ training-based channel estimation methods using pilots or reference signals, as done for instance in current LTE-Advanced systems. Pilot-based estimation techniques, however, involve a significant increase of both the receiver complexity and the signaling overhead, especially when the number of Transmission Points (TP) increases due to the use of Coordinated Multipoint transmission/reception (CoMP) schemes with Multiple-Input Multiple-Output (MIMO) processing. For instance, a coherent  $M \times N$  MIMO CoMP system, where the  $M$  transmitting antennas may belong to multiple spatially distributed TPs, requires at least the estimation of  $M \times N$  channels, which should be re-estimated every channel coherence time. For this reason, coherent systems are often impractical in fast-fading scenarios, due to their short coherence time. This fact discourages the use of closed-loop transmission modes, which require CSI at the Transmitter side (CSIT), for vehicular users [2]. On

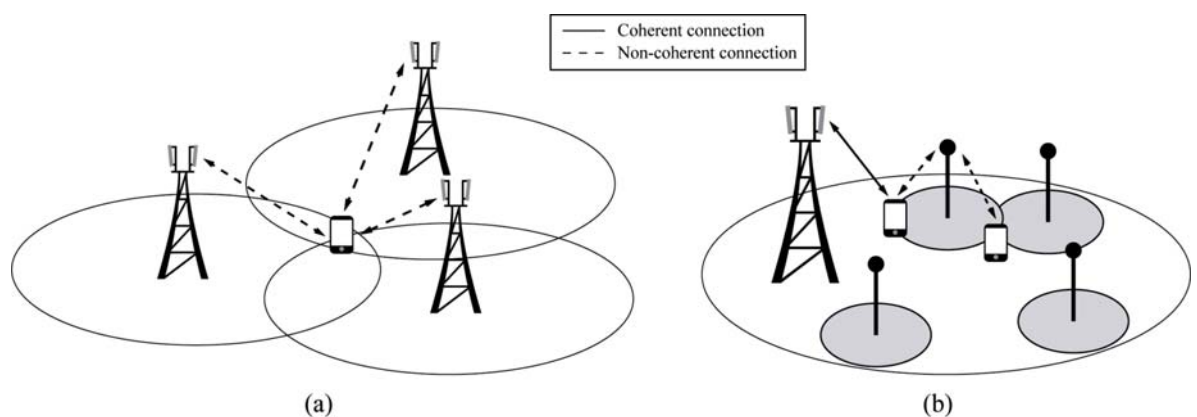
**The increasing requirements of CSIR availability in CoMP systems and the additional overhead that results from MU-MIMO transmission modes motivates the integration of non-coherent communication modes within state-of-the-art CoMP schemes.**

the other hand, in slow-fading channels, the transmitter will frequently send pilot symbols, thus leading to wasting some of the resources due to excessive pilot transmissions. Another major disadvantage of coherent communications is their degradation under channel estimation errors, which is especially dramatic under high mobility and, also, when MIMO or CoMP techniques are employed. All the aforementioned drawbacks of training-based communication motivated the increasing research on non-coherent communication techniques, which perform data detection without any knowledge of the channel coefficients at the receiver side [3][4].

The first works on non-coherent communication techniques showed that a careful design of the transmitted signals allows approaching the channel capacity of coherent systems at high signal-to-noise ratios (SNR) [5]. Among the many constellation designs for non-coherent MIMO communication, of particular interest in slow-fading scenarios are the techniques based on differential transmission [6]. To start the communication, a single reference symbol is transmitted, which is normally set to unity, and subsequent symbols are all differentially encoded, each of them based on the symbol in the previous slot. In the framework of non-coherent MIMO systems, Differential Unitary Space-Time Modulation (DUSTM) [6] can be seen as a higher dimensional extension of the standard Differential Phase-Shift Keying (DPSK) modulation to attain non-coherent communication with MIMO channels. In DUSTM, the channel is used in blocks of  $T=M$  transmissions, where  $M$  stands for the number of transmitting antennas. The transmitted signals belong to a codebook comprising a predefined set of  $M \times M$  unitary matrices. The main advantage of DUSTM is its efficient decoding, which can be carried out through Multiple-

Symbol Differential Detection (MSSD) at the receiver side [7]. Other alternatives for non-coherent MIMO communication, which are specifically designed for block-fading channels, are codebooks of unitary matrices that are isotropically distributed on the (compact) Grassmann manifold [8][9]. These codebooks exploit the MIMO channel characteristics and consider orthogonal subspaces to differentiate the transmitted symbols at the receiver side. The design of Grassmannian Constellations (GC) is gaining momentum, although so far they have been mainly studied from a theoretical point of view [9]. There exist many designs of GC, some of them systematic [8][10] and others non-systematic [9]. Regarding the decoding of GC, these constellations cannot exploit MSSD due to their non-differential construction. Nevertheless, GC can be non-coherently decoded using generalized likelihood ratio test (GLRT) receivers. Recent works show the utility of GC in Bit-Interleaved Coded Modulation (BICM) systems with Iterative Demodulation and Decoding (IDD) [11][12].

On the other hand, CoMP strategies [13] were identified as key enablers to mitigate inter-cell interference and to increase the cell edge throughput in recently developed standards such as LTE-Advanced [14][15]. If joint transmission from multiple TPs to the same user is assumed, the number of training phases followed by channel estimation stages scales linearly with the number of TPs, since the user needs information of all the channels involved in the transmission to demodulate the received data. The increasing requirements of CSIR availability in CoMP systems and the additional overhead that results from MU-MIMO transmission modes (more than 20% overhead in case of LTE-A with 2 users and 2 layers each [16]) motivates the integration of non-coherent communication modes within state-of-the-art CoMP schemes (see Figure 1(a)). Although these techniques are generally closed-loop, meaning that information about the channel coefficients is necessary at the transmitter side, some recent works claim the need for open-loop CoMP designs to be used as fallback modes in future cellular systems [2]. Figure 1(a) shows an exemplary CoMP architecture where a user has a non-coherent connection with three



■ **Figure 1.** Exemplary architectures for potential application of non-coherent communication: (a) CoMP system, (b) HetNet with macros and small cells.

different TPs. To the best of the authors' knowledge, there are no reported solutions in the literature on how to integrate non-coherent communications with multi-user and/or cooperative systems.

Still focusing on architectures involving multiple TPs, Heterogeneous Network (HetNet) architecture is considered a key element to satisfy the demands of future cellular communications [17]. HetNets may include base stations of varied coverage ranges, as well as multiple frequency bands and radio access technologies for the communication. In this framework, new cellular concepts arise, such as the macro-assisted small cell concept referred to as Phantom Cell (PC) [18]. In this kind of cells, the user is connected to the macro and phantom cells simultaneously. The key feature is that control information is delivered by the macro and data flows are delivered by the PC. Note that PCs lack legacy cell-specific signals and, hence, generally have open-loop connections with the users. Therefore, another promising application of non-coherent techniques is their exploitation in novel HetNet architectures, such as those involving PCs without legacy cell-specific signals. Figure 1(b) shows an exemplary HetNet architecture where a user has a coherent connection with the macro-cell and a non-coherent connection with the small cell.

The remainder of the paper is structured as follows. Section 2 presents an overview of non-coherent MIMO communications, where the system model and some important non-coherent techniques are introduced. Section 3 addresses the use of non-coherent techniques in practical scenarios, such as MIMO systems with antenna correlation and CoMP schemes where the power received by the UE from the different TPs is unbalanced. In Section 4, some performance results on the effect of antenna correlation and unbalanced received power are presented. Finally, some conclusions and future research directions are included in Section 5.

## 2. Non-coherent MIMO communications

### 2.1 System model

We consider a MIMO downlink single-user transmission from a set of  $M$  antennas, either belonging to the same Base Station (BS) or to a set of coordinated TPs, to a User Equipment (UE) with  $N$  antennas. The channel block length (number of time slots where the channel coefficients are considered unchanged) is assumed to be  $T$  consecutive uses of the channel. During channel block  $t$ , we assume a transmission through a channel  $\mathbf{H}^{(t)} \in \mathbb{C}^{M \times N}$ , containing Gaussian i.i.d. elements with zero-mean and unit-variance, which are unknown at both the transmitter and receiver sides. The signal transmitted through the  $M$  antennas during the  $t$ -th channel block is represented by the matrix  $\mathbf{X}^{(t)} \in \mathbb{C}^{T \times M}$ . The possible transmitted matrices are uniformly drawn from a finite-length constellation  $\mathcal{C} = \{\mathbf{S}_j, j=1, \dots, |\mathcal{C}|\} \subset \mathbb{C}^{T \times M}$ , the elements of which have

covariance  $\Gamma_{\mathbf{x}}$  and zero mean. The signal received by the UE,  $\mathbf{Y}^{(t)} \in \mathbb{C}^{T \times N}$ , is

$$\mathbf{Y}^{(t)} = \sqrt{\frac{\rho T}{\text{Tr}(\Gamma_{\mathbf{x}})}} \mathbf{X}^{(t)} \mathbf{H}^{(t)} + \mathbf{Z}^{(t)} \quad (1)$$

where  $\rho$  is the SNR of the received signal and  $\mathbf{Z}^{(t)} \in \mathbb{C}^{T \times N}$  is the additive white Gaussian noise (AWGN) matrix with i.i.d. zero-mean, unit-variance elements.

In the system model (1), the key element is the transmitted matrix  $\mathbf{X}^{(t)}$ , which must be carefully designed to allow a non-coherent detection at the receiver side. The next two subsections describe several alternatives for the construction of constellations that enable a non-coherent communication.

### 2.2 Schemes for slowly-varying channels

As stated in the introduction, DUSTM is an extension of the differential phase shift-keying (DPSK) constellation to support systems with multiple transmit and receive antennas. As in every differentially-encoded constellation, each transmitted signal in this scheme works as a reference for the following one. This scheme is intended for slowly-varying channels, where the channel can be assumed constant during any two consecutive channel uses.

In the case of MIMO systems, the codebook of symbols is composed of a set of matrices. A differential modulation scheme based on unitary matrices ( $\mathbf{S}_i^H \mathbf{S}_i = \mathbf{I}$ ) was proposed in [5], where the signal matrix to be transmitted at time slot  $t$  is differentially encoded as:

$$\mathbf{X}^{(t)} = \mathbf{V}^{(t)} \mathbf{X}^{(t-1)} \quad (2)$$

Here  $\mathbf{V}^{(t)}$  belongs to a codebook of  $L=2^{RM}$  unitary diagonal matrices of size  $M \times M$ , being  $R$  the number of bits carried by each diagonal entry. Note that  $T=M$  is assumed in this scheme and then  $\mathbf{X} \in \mathbb{C}^{M \times M}$ . To initialize the communication,  $\mathbf{X}^{(0)}$  is considered as an  $M \times M$  identity matrix.

The group of  $R \cdot M$  input bits to be transmitted during  $M$  time slots is assigned an index  $l$ , which is obtained following a Gray labelling. This index  $l$  has a direct correspondence with the element of the codebook of  $\mathbf{V}^{(t)}$  matrices that must be selected. The set of  $\mathbf{V}$  matrices can be constructed in a cyclic way starting from  $\mathbf{V}_l$ , which gives the first element of the codebook as:

$$\mathbf{V}_1 = \begin{pmatrix} e^{\frac{j2\pi u_1}{L}} & \dots & 0 \\ 0 & \ddots & 0 \\ 0 & \dots & e^{\frac{j2\pi u_M}{L}} \end{pmatrix}, \quad (3)$$

$$u_m \in \{0, 1, \dots, L-1\}, m=1, 2, \dots, M.$$

The rest of elements are subsequently obtained as  $\mathbf{V}_l = (\mathbf{V}_1)^l$ ,  $l=0, \dots, L-1$ . Recall that the transmitted information is contained in the symbol index  $l$ . Details on

Another promising application of noncoherent techniques is their exploitation in novel HetNet architectures, such as those involving Phantom Cells without legacy cell-specific signals.

how to design the  $u_m$  coefficients can be found in [6].

Demodulation requires collecting two successive received matrices to form an equivalent received matrix with  $P=2M$  rows. With the approximation that the channel is constant during  $P$  time slots corresponding to two channel blocks, i.e.  $\mathbf{H}^{(t)} \approx \mathbf{H}^{(t-1)} = \mathbf{H}$ , the received signals in two consecutive blocks are:

$$\mathbf{Y}^{(t-1)} = \sqrt{\frac{\rho T}{\text{Tr}(\Gamma_X)}} \mathbf{X}^{(t-1)} \mathbf{H} + \mathbf{Z}^{(t-1)} \quad (4)$$

$$\mathbf{Y}^{(t)} = \sqrt{\frac{\rho T}{\text{Tr}(\Gamma_X)}} \mathbf{X}^{(t)} \mathbf{H} + \mathbf{Z}^{(t)} \quad (5)$$

Substituting (2) into (5) and combining with (4) we obtain

$$\mathbf{Y}^{(t)} = \mathbf{V}^{(t)} \mathbf{Y}^{(t-1)} + \mathbf{Z}^{(t)} - \mathbf{V}^{(t)} \mathbf{Z}^{(t-1)} = \mathbf{V}^{(t)} \mathbf{Y}^{(t-1)} + \sqrt{2} \mathbf{Z}^{(t)} \quad (6)$$

which is the fundamental differential receiver equation. Although here the desired signal is corrupted by noise with twice the variance, the channel matrix is no longer necessary for the detection stage. This results in the well-known 3 dB performance loss in effective SNR when the channel is unknown in comparison to when it is known.

### 2.3 Schemes for block-fading channels

#### 2.3.1 Geometric interpretation

As several studies show [7], at high SNR the coherent capacity in block-fading channels can be achieved if the input signals are isotropically-distributed unitary matrices of size  $T \times M$ , provided that  $T \geq \min\{M, N\} + M$ . Following the results in [7], several constellation design methods targeting the high SNR capacity-achieving isotropic distribution were proposed in [8][9]. These unitary matrices can be seen as  $M$ -dimensional linear subspaces lying inside a  $T$ -dimensional complex Euclidean space [7]. The

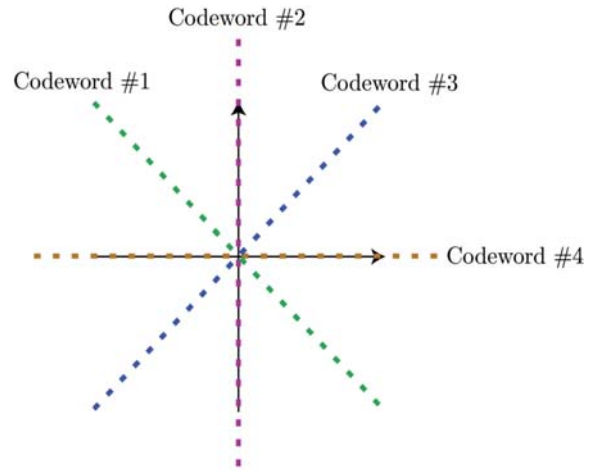


Figure 2. Exemplary Grassmannian codebook for  $M=1$  antenna,  $T=2$  time slots: 4 different directions in a plane.

columns of the proposed signal matrices form a basis of an  $M$ -dimensional subspace. Furthermore, each matrix  $\mathbf{X}$  is a point in the Grassmann manifold, which is the set of all  $M$ -dimensional linear subspaces of  $\mathbb{C}^T$ . Figure 2 shows an exemplary GC composed of four different directions in a plane, which can be represented by four  $2 \times 1$  matrices, i.e. four one-dimensional subspaces in a two-dimensional space.

When the input signal matrix  $\mathbf{X}$  is passed through the channel, the column vectors that span the original  $M$ -dimensional subspace are rotated and scaled, but they still lie within the same subspace. On the other hand, the noise does affect the subspace, but its effect can be neglected at high SNR. The particular subspace basis rotation is not detectable by a receiver without channel knowledge; however, the  $M$ -dimensional linear subspace spanned by this basis can be detected by using a Generalized-Likelihood Receiver Test (GLRT) [8]. The GLRT criterion projects the received signal on the different subspaces that compose the GC. Then, it calculates the energies of all the projections and selects the projection that maximizes the energy as follows

$$\hat{\mathbf{X}} = \arg \max_i \text{Tr}(\mathbf{Y}^H \mathbf{S}_i \mathbf{S}_i^H \mathbf{Y})$$

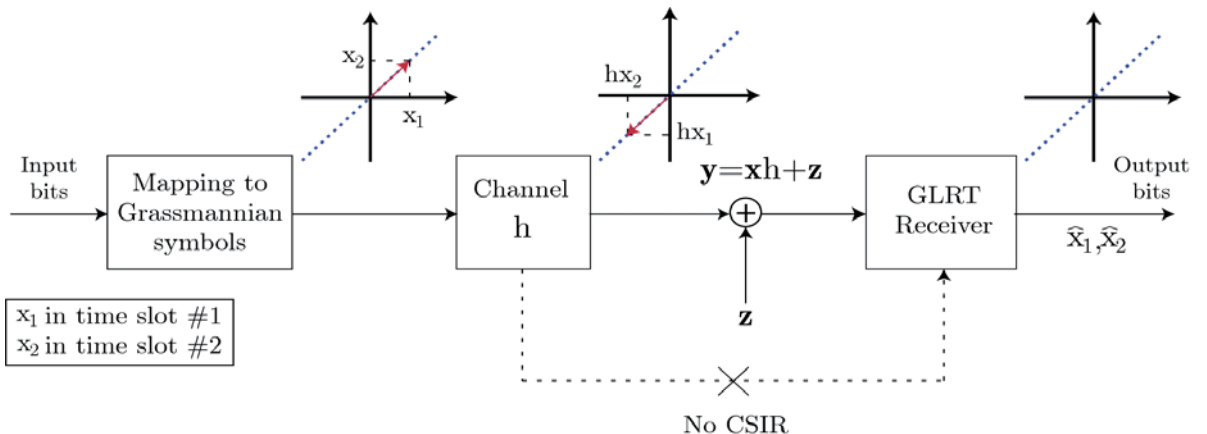


Figure 3. Block diagram of a non-coherent transceiver with  $M=1$  antenna and  $T=2$  time slots.

(7)

From the perspective of average symbol error probability minimization, in general, the GLRT provides a suboptimal result compared to the Maximum-Likelihood (ML) criterion. However, for the case of unitary codebooks assumed in this work, it can be shown that the GLRT provides ML detection performance [10].

An exemplary procedure for transmission and detection of GC is described next. Figure 3 shows the block diagram of the associated non-coherent transceiver which uses  $M=1$  antenna,  $T=2$  time slots and the GC in Figure 2. First of all, the information bits to be transmitted,  $x_1$  and  $x_2$ , are mapped e.g. to Codeword #3 (see Figure 2). After the codeword is transmitted, its underlying basis (in red) is transformed by the channel, but it remains in the same subspace. Although the non-coherent receiver cannot detect the particular transformation caused by the channel, it can indeed detect the subspace spanned by this basis if  $\mathbf{x}\mathbf{h} \gg \mathbf{z}$ . Therefore, the transmitted information can be recovered without any CSIR.

### 2.3.2 Codebook construction

#### Systematic designs

Most systematic designs of unitary matrices are based on the mapping of a linear Space-Time Block Code (STBC) into a unitary code by means of a certain transformation. One example of these techniques is the code designed in [10], where the exponential transform is applied to a non-unitary codeword matrix  $\tau\mathbf{B}$ , with  $\tau=0.30$ , as

$$\mathbf{S}_i = \begin{pmatrix} 0 & \tau\mathbf{B} \\ -\tau\mathbf{B}^H & 0 \end{pmatrix} \begin{pmatrix} \mathbf{1} \\ 0 \end{pmatrix} \quad (8)$$

where

$$\mathbf{B} = \begin{pmatrix} s_1 + s_2 \zeta_8 & \zeta_{16}(s_3 + s_4 \zeta_8) \\ \zeta_{16}(s_3 - s_4 \zeta_8) & s_1 - s_2 \zeta_8 \end{pmatrix} \quad (9)$$

For the construction of  $\mathbf{B}$ , the input bits need to be previously grouped and mapped to symbols of a Quadrature Amplitude Modulation (QAM), denoted by  $s_i$ .

The notation  $\zeta_k = e^{j\frac{2\pi}{k}}$  was used in (9).

Another systematic design was proposed in [19], where the first step is to construct a non-unitary codeword  $\mathbf{S}_i^*$  as

$$\mathbf{S}_i^* = \begin{pmatrix} \mathbf{A} \\ \mathbf{B} \end{pmatrix} \quad (10)$$

where

$$\mathbf{A} = \begin{pmatrix} a_1 + ja_2 & \zeta_8(a_3 - ja_4) \\ \zeta_8(a_3 + a_4) & a_1 - ja_2 \end{pmatrix}, \quad (11)$$

$$\mathbf{B} = \begin{pmatrix} b_1 + jb_2 & \zeta_8(b_3 - jb_4) \\ \zeta_8(b_3 + b_4) & b_1 - jb_2 \end{pmatrix}$$

Here the input bits must be mapped to two different constellations with different sizes, one for  $\mathbf{A}$  ( $a_i$  symbols) and another for  $\mathbf{B}$  ( $b_i$  symbols). Further details can be found in [19]. Once the non-unitary codeword is constructed, its corresponding unitary matrix is obtained using the QR decomposition, such as  $\mathbf{S}_i^* = \mathbf{QR}$ . The unitary codeword

used for transmission is directly  $\mathbf{S}_i = \mathbf{QR}$ .

The main advantage of systematic codebooks is that they do not need to be stored at the transmitter and at receiver side and only their design rule is necessary. However, their performance is still poorer than the one of the next introduced non-systematic designs.

#### Non-systematic designs

Non-systematic codebooks are constructed using numerical tools to optimize a certain distance metric on the Grassmann manifold. An overview of distance metrics can be found in [9], where the chordal Frobenius norm between two subspaces was chosen for the design of GC. The chordal Frobenius norm between the two subspaces that are spanned by the columns of two codewords  $\mathbf{S}_i$  and  $\mathbf{S}_j$  is defined as

$$d^2(\mathbf{S}_i, \mathbf{S}_j) = 2M - 2\text{Tr} \left( \sum \mathbf{s}_i^H \mathbf{s}_j \right) \quad (12)$$

where  $\sum \mathbf{s}_i^H \mathbf{s}_j$  is obtained from the SVD decomposition  $\mathbf{S}_i^H \mathbf{S}_j = \mathbf{U} \mathbf{\Sigma} \mathbf{V}^H$ .

Basically, the constellation design problem consists in selecting unitary matrices with maximum pairwise chordal Frobenius norm between the subspaces they span. The main advantage of these approaches is that it allows exploiting all the system degrees of freedom without restricting the constellation to have a specific structure. However, designing large constellations using a direct optimization may be of prohibitive complexity. Hence, the use of other design methods such as the greedy or rotation-based ones is encouraged [9].

#### Coherent codes used as a baseline

To prove the interest of non-coherent communication based on GC, these schemes need to be compared to codes designed for the same total number of antennas that need to be coherently decoded (with CSIR). To be consistent with previous works, we will select two coherent (training-based) baselines. In particular, for 2 transmitter antennas, the well-known Alamouti code [20], with the following  $T \times M$  code matrix with  $T=2$ ,  $M=2$

$$\mathbf{S}_i = \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix} \quad (13)$$

and also the Golden code [21], constructed as

$$\mathbf{S}_i = \frac{1}{\sqrt{5}} \begin{pmatrix} \alpha(s_1 + s_2\theta) & \alpha(s_3 + s_4\theta) \\ i\gamma(s_3 + s_4\bar{\theta}) & \gamma(s_1 + s_2\bar{\theta}) \end{pmatrix} \quad (14)$$

where  $\theta = \frac{1 + \sqrt{5}}{2}$  is the Golden number,  $\gamma = 1 + i\theta$ ,  $\bar{\theta} = 1 - \theta$ ,  $\alpha = 1 + i\bar{\theta}$ .

### 3. Non-coherent schemes in practical scenarios

As previously mentioned, there are many theoretical studies on the design and performance of non-coherent schemes in MIMO systems. However, their impact over real-world cellular systems is still unknown. Before these techniques are mature enough to be considered by wireless communications standards, extensive research on their performance in realistic scenarios must be undertaken. In this direction, we present preliminary results of the performance of non-systematic GC in two realistic scenarios: a MIMO system with spatial correlation and a CoMP system with unbalanced TPs. Here the term unbalanced refers to TPs that are simultaneously serving the same UE but they are located at different distances from it.

#### 3.1 Scenario 1: MIMO system with antenna correlation

In practice, the channels between the different antennas of a MIMO system are often correlated and thus the potential multi antenna gains are not always attainable. This problem is more dominant in mobile terminals due to their compact size.

In this paper, a spatially-correlated Rayleigh block-fading channel is modeled by the Kronecker model [22]. This channel is constant during  $T$  consecutive channel uses but there exists spatial correlation among the antennas. Hence, the system model in (1) is applicable. This model assumes separable statistics at the transmitter and the receiver and its realizations can be generated from the channel without correlation,  $\mathbf{H}$ , as:

$$\mathbf{H}_{\text{CORR}} = \sqrt{\mathbf{R}_{\text{TX}}} \mathbf{H} \sqrt{\mathbf{R}_{\text{RX}}} \quad (15)$$

where the channel correlation matrices are modeled in the same way as in [20]

$$\mathbf{R}_{\text{TX}} = \begin{pmatrix} 1 & \alpha_{12} & \cdots & \alpha_{1M} \\ \alpha_{21} & 1 & \cdots & \alpha_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ \alpha_{M1} & \alpha_{M2} & \cdots & 1 \end{pmatrix}, \quad 0 \leq \alpha_{ij} \leq 1 \quad (18)$$

$$\mathbf{R}_{\text{RX}} = \begin{pmatrix} 1 & \beta_{12} & \cdots & \beta_{1M} \\ \beta_{21} & 1 & \cdots & \beta_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ \beta_{M1} & \beta_{M2} & \cdots & 1 \end{pmatrix}, \quad 0 \leq \beta_{ij} \leq 1 \quad (19)$$

Here  $\alpha_{ij}$  and  $\beta_{ij}$  are the transmit and receive correlation parameters between the  $i$ -th and  $j$ -th antennas, respectively. We note that the considered correlation between transmit and receive antennas is assumed to be dual, that is,  $\alpha_{ij} = \alpha_{ji}$  and  $\beta_{ij} = \beta_{ji}$ . Following the usual assumptions, in this paper we consider two antennas are highly correlated when their correlation parameter is equal to

0.7 and low when it is equal to 0.3.

#### 3.2 Scenario 2: CoMP system with unbalanced transmission points

In practical CoMP systems, the UE is rarely located at the same distance from all the TPs involved in the transmission. Therefore, the signal is usually received with different power levels. Focusing on the transmission of GC matrices jointly from multiple TPs, we consider that there is a power imbalance among the TP of  $0 \leq \gamma \leq 1$  that weighs the columns (basis) of the transmitted constellation symbol in a different way. For an exemplary GC of  $4 \times 2$  symbol matrices, which could be transmitted from two TPs with a single antenna each, the original constellation point is composed of two columns,  $\mathbf{c}_1$  and  $\mathbf{c}_2$ , for  $m=1, 2$

$$\mathbf{S}_j = [\mathbf{c}_1 \mathbf{c}_2] \quad (13)$$

and the unbalanced constellation point is

$$\mathbf{S}_j = [\mathbf{c}_1 \sqrt{1+\gamma} \quad \mathbf{c}_2 \sqrt{1-\gamma}] \quad (14)$$

where it can be noted that for  $\gamma \neq 0$  the subspace basis is no longer orthonormal, it is just orthogonal.

For the performance evaluation, the following power difference among the TPs (in dB) will be assumed:

$$\text{Power difference (dB)} = 10 \log \left( \frac{1+\gamma}{1-\gamma} \right) \quad (15)$$

## 4. Performance evaluation of non-coherent techniques in practical scenarios

We consider a block-fading channel that remains constant during  $T=4$  time slots to evaluate the joint non-coherent communication from 2 TPs with a single antenna each, i.e.  $M=2$  transmitter antennas, to a single user with  $N=2$  antennas. The selected codebook is a GC of 4096 points, which leads to a transmission rate of 3 bpcu (12 bits are transmitted in 4 time slots). The GC scheme is compared to two open-loop coherent schemes designed for the same total number of antennas. The coherent baselines are the Alamouti code and the Golden code, the code matrices of which are (13) and (14), respectively.

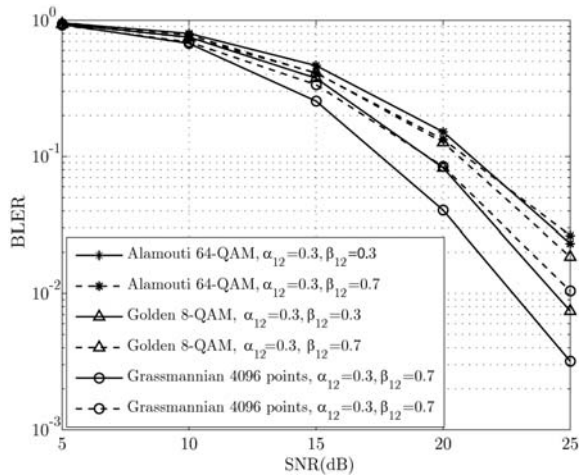
Provided that  $M$  time slots are necessary for training the coherent schemes, 64-QAM and 8-QAM symbols are chosen to set a transmission rate of 3 bpcu in the Alamouti and Golden schemes, respectively. The Alamouti and Golden schemes will be evaluated here under MMSE channel estimation.

#### 4.1 Scenario 1: MIMO system with antenna correlation

We here consider different correlation levels both at the transmitter and at the receiver sides. Figure 4 shows the Block Error Rate (BLER) performance versus SNR of the investigated schemes. It can be observed that, at low

SNR, the three evaluated schemes have better performance when  $\beta_{12}=0.7$ , hence receiver antenna correlation is beneficial in this scenario. As the SNR gets higher, the setup with  $\beta_{12}=0.3$  achieves better BLER performance for all the schemes. These two results are in concordance with [23] where the authors argue that a fully correlated channel matrix maximizes the mutual information at sufficiently low SNR. In contrast, at high SNR, a fully uncorrelated channel matrix is showed to be optimal. In particular, the crossing point is SNR=22.5 dB for the Alamouti code, SNR=10.2 dB for the Golden code and SNR=7 dB for the GC. Regarding the performance degradation of each individual scheme, for a target  $BLER = 10^{-1}$ , the Golden code and GC suffer a performance degradation of approximately 2 dB when the receiver antenna correlation is increased. On the other hand, the Alamouti code is quite robust to the receiver correlation, showing a degradation of less than 0.1 dB. The reason for this result may be that, while the Alamouti code can be decoded with acceptable performance using a single receiver antenna, the Golden code and GC require at least two receiver antennas to be correctly decoded. Therefore, keeping the receiver antenna correlation as low as possible is more crucial for the Golden code and GC than for the Alamouti code.

Regarding the comparison among the three investigated schemes, it can be observed a performance gain of 2 dB for GC over Golden codes when  $\alpha_{12}=0.3$  and  $\beta_{12}=0.3$ . This gain is also maintained when  $\beta_{12}$  is increased to 0.7. Therefore, in all cases, GC transmission outperforms the coherent baselines.



■ **Figure 4.** Effects of the spatial correlation over the evaluated coherent and non-coherent schemes for  $\alpha_{12}=0.3$ .

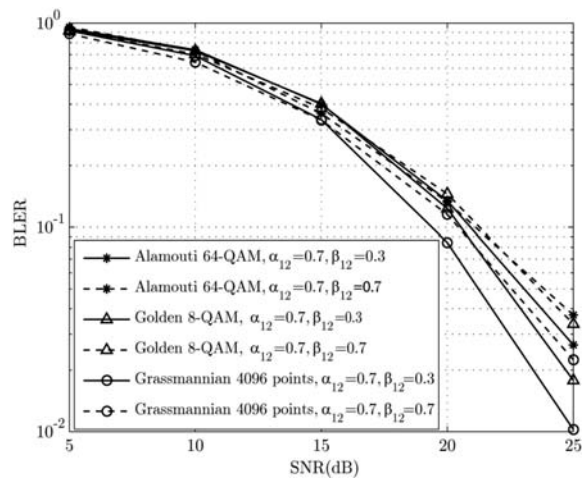
In Figure 5, we consider a setup similar to the previous one but with a higher correlation at the transmitter side, given by the value  $\alpha_{12}=0.7$ . From this figure, it can be seen that the BLER of all the schemes at low SNR is lower when  $\beta_{12}=0.7$ , hence receiver antenna correlation is also beneficial in this scenario. At high SNR, the  $\beta_{12}=0.3$  case is more advantageous, as happened also in the  $\alpha_{12}=0.3$  case. The

Despite the fact that spatial correlation is detrimental for coherent and non-coherent systems, the latter outperformed the coherent ones.

crossing points are now at SNR=8 dB for the Alamouti

code, SNR=16.2 dB for the Golden code and SNR=15 dB for the GC. Regarding the performance degradation of each individual scheme, for a target  $BLER = 10^{-1}$ , the Golden code and GC suffer a performance degradation of approximately 1 dB, which is half the degradation observed for  $\alpha_{12}=0.3$ . The Alamouti code is proven to be more robust to the receiver correlation, showing a degradation of about 0.2 dB. In general, it can be observed that GC-based transmissions outperform the coherent baselines.

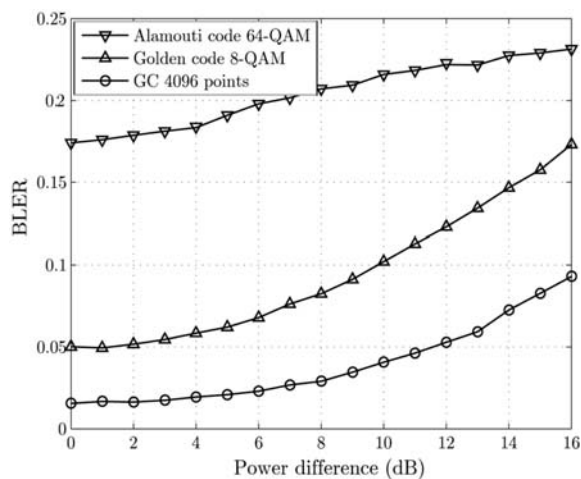
Overall, when comparing Figures 4 and 5, it can be seen that the gap between the performance curves decreases as  $\alpha_{12}$  increases. As said above, both schemes show cross over points between the curves for high and low receive correlation. Also, by comparing the curves for  $\alpha_{12}=0.3, \beta_{12}=0.7$  with those for  $\alpha_{12}=0.7, \beta_{12}=0.3$  it can be noted that they show the same performance results. Therefore, correlation at the transmitter has the same effect as correlation at the receiver for the three evaluated schemes.



■ **Figure 5.** Effects of the spatial correlation over the evaluated coherent and non-coherent schemes for  $\alpha_{12}=0.7$ .

#### 4.2 Scenario 2: CoMP system with unbalanced transmission points

Figure 3 shows the BLER of the GC, Alamouti and Golden schemes for an SNR of 20 dB, within a range of values of the power difference among TPs. It can be seen that the three schemes are degraded by the power imbalance, but the GC still outperforms the Alamouti and Golden schemes. It is worth noting that more than 17 dB of power imbalance are needed for the GC BLER performance be degraded one order of magnitude. Therefore, subspace transmission is a quite robust scheme within



■ **Figure 6.** BLER performance of GC, Alamouti and Golden codes at 20 dB of SNR with increasing power difference between two TPs. the normal operation range.

## 5. Conclusion

This paper presented concept of non-coherent MIMO communication together with some of the main techniques that enable this type of communication. In addition, we conducted a performance comparison between coherent and non-coherent signalling schemes under practical channel conditions. First, we evaluated the performance of GC and two coherent benchmark schemes, the Alamouti and Golden codes, in a spatially correlated block-fading MIMO system. Despite the fact that spatial correlation is detrimental for coherent and non-coherent systems, the latter outperformed the coherent ones. However, as the correlation between the receive antennas increases, the performance gap between both systems is reduced. Furthermore, we observed that when  $\alpha_{12}$  is fixed, the receive correlation coefficient  $\beta_{12}$  tends to improve the system performance at low and moderate values of SNR. On the contrary, correlation between the receive antennas is harmful at high SNR for both systems.

We also evaluated the impact of the power imbalance over the received signal when the user is not at the same distance from all the TPs involved in the cooperative transmission. The three evaluated schemes (Alamouti, Golden and GC) were shown to be slightly degraded by the power imbalance. Nevertheless, GC outperformed the other schemes and its BLER degradation was less than one order of magnitude until 17 dB of power imbalance. Therefore, subspace transmission is robust within a wide range of power imbalance values.

Further work includes investigating constellation designs robust to spatial correlation and power imbalance among TPs. In addition, the design of MU-MIMO techniques for non-coherent coordinated communication will be also included in our way forward. Another relevant aspect to be investigated in future work is how GC could be fitted

within an OFDM system.

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