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Why the Incoherent Paradigm Is for the Future Wireless Networks?

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Abstract

This material is aimed to attract attention to the "incoherent approach for power NOMA-RIS-MIMO transmission in wireless channels". Such kind of approach might be successfully applied in future dense networks formed by High-Speed Vehicles (HSV networks, etc.). Those scenarios take place in doubly selective communication channels typical for such kind of radio networks. The proposal for the presented hereafter incoherent view ("paradigm") is based on several basic principles: 1) Shift from the "coherent "ideology", *i.e.* rejection of the application of any type of Channel State Information (CSI, CSIT); 2) Application of the so-called "invariant" to the communication channel's features (distortions) modulation technique together with its incoherent demodulation; 3) Orthogonal channel decomposition by means of "universal" eigen basis (in the form of Prolate Spheroidal Wave Functions, PSWF) as "artificial trajectories" of wave propagation; 4) Chaotic filtering (chaos parameter settings as UE signatures) together with sequential multiuser parallel detection algorithms for users' identification (classification). It is shown that the proposed approach might provide an effective use of the radio resource and it is relatively simple for implementation.

Keywords

Incoherent Processing, Universal Eigen Functions, Chaos Filtering, Invariant Modulation

1. Introduction

During the last decades, the Non-Orthogonal Multiple Access (NOMA) gained considerable attention as a multiple access technique, for example, for dense network applications in 5G and beyond. The following material will be dedicated to the power-domain NOMA, which still seems to be rather practical. Regarding the NOMA technology, numerous schemes have been proposed and thoroughly analyzed in the literature ([1]-[6], etc.). It was shown, for example, in [1] [2] [3] [4], that with the correct parameter settings, the symbol level power domain NOMA is able to significantly outperform the well-known Orthogonal Multiple Access (OMA) schemes with OFDMA. Moreover, in dense networks that are expected to be in service for 5G+ and beyond (6G, etc.), it is logical to predict the "huge lack" of orthogonal radio resources and though NOMA settings might be very useful.

In this regard, the application of MIMO, massive MIMO, etc. together with the channel reconfiguration techniques has to be taken into account and carefully analyzed. Note that for application in NOMA transmission, the channel reconfiguration is not a trivial problem for MIMO due to the "hard" problem: how to manage the very Aggressive Multi-user Interferences (MAI) generated by intensive service demands from multiple User Equipment (UE) both in the downlink and uplink with always limited radio-resource scenarios? In the MIMO case, this problem turns up to be immensely complex and even provokes serious doubts regarding the future application of power NOMA in MIMO channels (see for example [3] [5] [6]). The situation might be even worse, if the transmission system includes the utilization of RIS and relies on "coherent processing ideas" based on the Channel State Information (CSI, CSIT), which is always imperfect in real-life scenarios, particularly in "High-Velocity Channels" (HVCs) (see [5] [7] [8] [9] [10]), generalized hereafter through the term "Doubly Selective Channels".

Note that the prospective scenarios for wireless networks related to 5G and beyond, 6G, etc. might widely include applications of High-Speed Vehicles (HSVs), *i.e.* channels with large values for the parameters like Doppler Shifts, Frequency Offsets, time delays, etc. As it was pointed out long ago in [7] [8], such kind of effects does not "allow" to effectively maintain the so-called "coherent" or "quasi-coherent" paradigms[1] for signal processing algorithms significantly based on Channel State Information (CSI) estimation, because the time, frequency, and dynamic behavior of the channel simply do not support rather an accurate estimation of CSI. The latter is the reason to dedicate the present material to the "ideological shift" of the paradigm for MIMO-RIS-NOMA (see **Figure 1**) system design from the coherent to the incoherent one with the application of the modulation schemes named in [11] [12] as invariant to the distortions of the channel.

The material in [3] suggests that the "standard" power NOMA approach cannot be successfully applied and might be useless and "impractical" for MIMO scenarios. The latter might be explained (see below) due to the application of SIC (SICT) for MAI mitigation and UE identification. That is why [3] suggested concentrating on the Rate-Splitting Multiple Access (RSMA) approach but its application, for example, on Doubly Selective Channels was not illustrated [3]. Based on the "coherent" ideology as well, the spread spectrum concept (the so-called WSMA NOMA) for NOMA-MIMO has to be thoroughly analyzed.

[1]For terminology issues about what coherent or incoherent approach is, please refer for example to [9] (and the references therein), because this terminology will be used in the following.

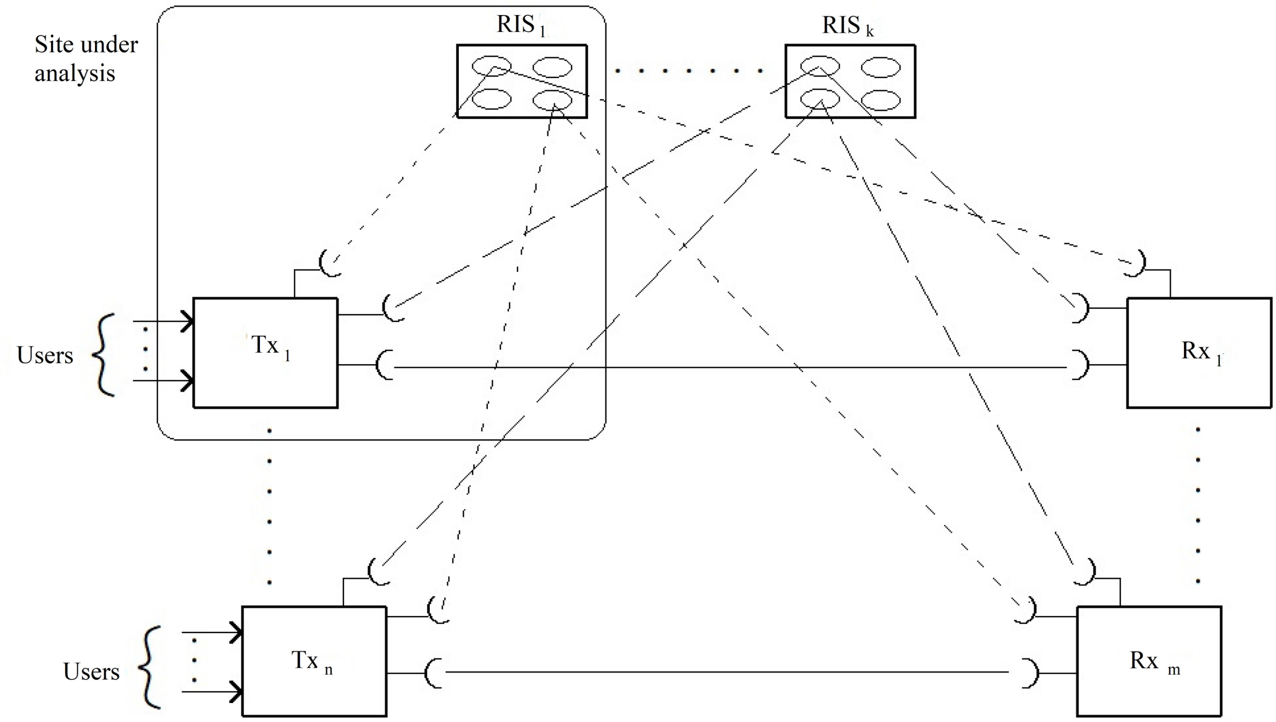

Figure 1. Block diagram of the MIMO-RIS-NOMA transmission system.

In short, how to manage multi-user interference in such kind of channels, together with Doppler Shift, Frequency Offset, etc. distortions affecting the desired information signals?

In this regard, in the following, it is suggested, first, change the paradigm for the NOMA settings, *i.e.* refuse the idea of coherent ideology for NOMA settings for HSC and "shift" it to the incoherent one, reject the application of CSI completely, as it is anyway almost impossible to apply it properly in real scenarios.

Second, apply the "invariant, robust, resistant, etc." (the term is not important here!) modulation technique to the above-mentioned distortions at the Doubly Selective Channels.

Third, in order to simplify the processing algorithms, try to "decompose the rather complex" Doubly Selective Channel System Functions (in Bello sense, see it below) into the set of so-called "artificial", "virtual trajectories" for wave propagation with its separated processing for the further addition procedure.

Fourth, based on the propagation phenomena in the Doubly Selective Channels, which actually transform the desired signals of each UE into the "almost" Gaussian Random Processes and the channel into the vector Gaussian Channel, in order to diminish the processing time, it is suggested to apply for the UE's identification, instead of the "standard" approaches, as SIC technology, the "modified" idea for the procedure, as a "parallel filtering" approach, where the UE signatures are taken from the parameters of certain chaotic attractors, which are used to model those UE's at the Gaussian Channel (see [13] [14] [15] [16] [17], etc.) in order to precisely "separate " them and then classify (identify). The latter will be presented in the following.

As it will be shown in the following, those basic ideas applied to the NOMA-RIS-MIMO transmission design in the "incoherent fashion" might significantly simplify the system implementation, minimize the SNR losses and provide realistic characteristics for the "total" noise immunity and the spectrum efficiency of the system.

The paper is organized as follows. Section 2 is totally dedicated to the so-called "orthogonal channel decomposition", based on the Universal PSWF approach applied to the MIMO-RIS design as well. Section 3 describes the filtering approach for MAI mitigation together with the sequential identification (classification) for the UE. Section 4 presents the sketch of the invariant DPSK-*k* modulation for the UE's and their incoherent autocovariance demodulation algorithm together with their noise immunity characteristics evaluation. Section 5 is dedicated to Conclusions.

2. Channel Orthogonalization, Universal Eigen Basis, and RIS-MIMO Design

2.1. Introductory Comments

General ideas for channel characterization by means of the so-called Channel System Functions were presented in the fundamental works of Bello and will be used hereafter (see below). The Channel System Functions were applied first for orthogonalization purposes by Kennedy (1969) [18] in the design framework of the "West Ford" project and were further generalized and developed (see [19] [20] [21] [22], etc. and the references therein).

Moreover, this approach was successfully applied for incoherent algorithms in SISO and MIMO systems considering Doubly Selective Channels [10] [23] [24] [25] as well as the generalized Rake principle for space-time applications [10] [22] [26]. Here, it has to be stressed the importance and originality of the works of Huges, Giannakis, etc. (see [7] [8] [10]) and their influence on this topic.

The idea of the Channel Orthogonalization is similar to the Fourier Series Analysis approach and is based on the idea of the representation of the "object" (the function of time, the field, etc.) by the finite set of the orthogonal components (basis, eigen functions).

Of course, the basis strongly depends on the "object" and for the MIMO channel it will be concretized hereafter.

In the following, the concept of channel orthogonalization will be described and developed. For more details, see preprint [27], etc.

2.2. Channel Orthogonalization, MIMO Case, and Universal Eigen Basis

The ongoing material considers the orthogonalization principle for MIMO channel representation and it can be applied as an expression of any of the Channel System Function (in Bello sense [28]) in terms of a reduced number of its eigen functions denoted in the following as virtual (artificial) trajectories for the propagation phenomena.

Under the Wide-Sense Stationary Uncorrelated Scattering (WSSUS) [28] assumption, in the following the focus is set on the so-called input impulse response time-delay-spread (FIR) matrix $\boldsymbol{H}(t, \tau)$ of the MIMO channel defined for "N" transmitting and "M" receiving antennas (N_{Tx}, M_{Rx}) as

$$H(t,\tau)=\int_{-\infty}^{\infty}H(t,f)\exp(j2\pi f\tau)\mathrm{d}f,$$

where t, τ and f are time, time delay and frequency; $\boldsymbol{H}(t, \tau)$ and $\boldsymbol{H}(t, f)$ are matrixes of size $N_{Tx} \times M_{Rx}$.

Then, the KLE method is an optimum procedure (see [25], etc.) (based on the so-called KLE integral equation) and is used to find a set of orthogonal eigen matrices (eigen functions), that can approximate the Gaussian $\boldsymbol{H}(t, \tau)$ with a minimum number of those eigen matrices and a predefined minimum Mean-Square Error (MSE). Mathematical foundations for this can be found in ([25], etc.) and are omitted hereafter.

In the general case, the KLE integral equation is as follows [25]:

$$\begin{aligned}&\lambda_l\boldsymbol{\Phi}_l(t,\tau,\boldsymbol{\theta}_{Tx},\boldsymbol{\theta}_{Rx})\\&=\int_{\Lambda}\boldsymbol{R}_{\boldsymbol{H}}(t,t',\tau,\tau',\boldsymbol{\theta}_{Tx},\boldsymbol{\theta}_{Rx},\boldsymbol{\theta}'_{Tx},\boldsymbol{\theta}'_{Rx})\boldsymbol{\Phi}_l(t',\tau',\boldsymbol{\theta}'_{Tx},\boldsymbol{\theta}'_{Rx})\mathrm{d}t'\mathrm{d}\tau'\mathrm{d}\boldsymbol{\theta}'_{Tx}\mathrm{d}\boldsymbol{\theta}'_{Rx},\end{aligned}\tag{2.1}$$

where $\{\lambda_l\}$ denotes the set of eigen values, $\{\phi_l(\cdot)\}$ the set of eigen matrices obtained from covariance matrix $\boldsymbol{R}_{\boldsymbol{H}}$ of the Gaussian channel and the integration domain is $[0, T_{\text{obs}}] \times [0, T_{\text{Max}}] \times [-\Delta\theta_{Tx}, \Delta\theta_{Tx}] \times [-\Delta\theta_{Rx}, \Delta\theta_{Rx}]$; $\Delta\theta_{Tx(Rx)}$ is Tx (Rx) angle beamwidth, T_{obs} is the observation time for analysis, $T_{\max}$ is the maximum delay excess time.

Equation (2.1) is rather complex for its solution and for practical scenarios it might be simplified by assuming separability for all domains of the covariance matrix $\boldsymbol{R}_H(\cdot)$, WSSUS conditions, etc. which will be pointed out later on. With these assumptions (2.1) can be simplified in the following way:

$$\left.\begin{aligned}&\lambda_\theta\boldsymbol{\Phi}_\theta(\theta)=\int_{-\frac{1}{2}\theta}^{\frac{1}{2}\theta}R_\theta(\theta,\theta')\Phi_{\theta'}(\theta')\mathrm{d}\theta'\\&\lambda_\tau\boldsymbol{\Phi}_\tau(\tau)=\int_0^{\tau_{\max}}R_\tau(\tau,\tau')\Phi_{\tau'}(\tau')\mathrm{d}\tau'\\&\lambda_t\boldsymbol{\Phi}_t(t)=\int_0^{\tau T}R(t^*,t')\Phi_t(t')\mathrm{d}t'\end{aligned}\right\}\tag{2.1.1}$$

Comparing (2.2) and (2.1.1) it is obvious that the separability hypothesis for $\boldsymbol{R}_H(\cdot)$ is a "great" proposal for the KLE simplification.

One has to notice the strong limitation of (2.1) (2.1.1): the dependence on their solutions, *i.e.* eigen matrices (eigen functions) to the concrete form of $\boldsymbol{R}_H(\cdot)$.

The latter encourages the authors of [19] [20] to propose the concept of "universal" eigen basis used under rather broad assumptions of the Channel System Functions.

2.3. Universal Eigen Basis and Generalized Kronecker Channel Model (GKCM) for MIMO Case

As it was pointed out above, KLE is an **optimum** procedure to approximate the Gaussian channel model with a minimum set of eigen matrices (eigen functions) and with a minimum value of the approximation error (MSE) (see [25]), but with one significant detail: a-priori knowledge of $\boldsymbol{R}_H(\cdot)$, which is obviously not feasible for real applications without channel sounding. So, how to avoid this problem?

The ongoing idea was inspired by the fundamental characteristics of PSWF (see [19] [20] [22] [25] and references therein), which shows that PSWF might be suitable to propose **a generic** principle for channel modeling as it only requires an a-priori knowledge of a minimum set of parameters of the channel: channel bandwidth, $F_{\max}$, maximum delay spread, $T_{\max}$, and the angular bandwidth, $\Delta\theta$.

Once again, assuming the separability for $\boldsymbol{R}_H(\cdot)$ in space-frequency-time domains it follows that the required number of PSWF for the approximation for each domain can be easily predicted in advance, as following:

- For time delay domain,

$$M = 2F_{\max}T_{\max} + 1.$$

- For spatial domain (assuming ULA arrays for MIMO),

$$M_{Tx(Rx)} = 4\pi\Delta\theta_{Tx(Rx)}\left(L-1\right)\frac{d}{\lambda}+1,$$

where L is the number of elements of the ULA (see the following); d is the separation between the ULA elements; λ is the working wavelength; ULA stands for Uniform Linear Antennas applied at MIMO.

Note, that in [20], it was shown that **any** Planar Antenna (PA) apertures can be successfully approximated (in covariance sense) by a non-uniform linear aperture (n-ULA). Furthermore, the latter (also in covariance sense) can be approximated with a predefined error (MSE) by ULA[2]. That is why for the above presented expressions of the number of PSWF and as an approximation ULA was finally chosen.

Note, that the approximation with PSWF obviously always requires more functions than the KLE solution, because the latter is exact and optimum and the universal PSWF basis is introduced heuristically (see practical examples of the standard at [27], which shows a negligible increment of the number of functions for approximation to KLE solutions, but it is **invariant to the statistics of the channel**.

Next, consider hereafter, (as it was proposed in ([20] [27], etc.)) the NOMA-RIS-MIMO channel with the dispersion and fading (including Reconfigurable Intelligent Surfaces (RISs)) to be approximated by generic Generalized Kronecker Channel Model (GKCM).

Then, as it was shown in [20] [22] generally the $\boldsymbol{H}(t, \tau)$ of the GKCM might be

[2]It must be mentioned that in [32], the same issue was proved from geometrical considerations.

represented in the form:

$$\boldsymbol{H}(t,\tau)=\boldsymbol{V}_{Tx}\left[\tilde{\boldsymbol{\Omega}}\cdot\boldsymbol{G}(t,\tau)\right]\boldsymbol{V}_{Rx}^{H}, \tag{2.2}$$

where H indicates here the Hermitian transpose of the matrix; $\tilde{\boldsymbol{\Omega}}$ is the element wise square root of the Coupling Matrix (CM) Ω, which physically represents, how the eigen modes (eigen matrices) of the transmit and receive correlation matrixes (in our case PSWF) are connected through the scattering environment of the RIS; $\boldsymbol{V}$ with certain indexes is "orthogonalization matrixe", $\boldsymbol{G}(t, \tau)$ is an!! $N_{Tx} \times M_{Rx}$ i. i. d. and zero means Gaussian CIR Matrix. So, the essence of the GKSM, including the MIMO-RIS-NOMA channel is rather simple: the model contains a set of "independent" SISO channels, artificially created from the appropriate eigen basis (in our case-PSWF) accumulated in a model by the CM (Coupling Matrix). And RIS is nothing else, but a special case of CM.

2.4. GKCM and Reconfigurable Intelligent Surface (RIS) Design

The RIS, as an element of the NOMA-MIMO transmission system was proposed rather recently, to improve the characteristics of NOMA transmission against the traditional OMA setting with OFDMA (see [29] [30] [31], etc), see also **Figure 1**.

To the best of our knowledge, the first attempt for the artificial modification of the propagation media to improve the characteristics of the information transmission was first proposed long ago by R. Kennedy in the framework of the already mentioned West Ford project [18].

The possible scenario for MIMO-RIS-NOMA is illustrated in **Figure 1** and it can be seen, that the {UE's} are located usually in small clusters surrounding each RIS in the system and the RIS is actually providing a beamforming mechanism for certain UE's. But taking into account the above material this "beamforming" might be implemented (certainly, in the artificial way!) through the "Coupling Matrix" (CM) of the GKCM, which can be used to provide the "propagation tracks" for each UE's.

So, the RIS algorithm is nothing else but an algorithm for "predefined connections" between the eigen matrices (eigen functions), as artificial beams, at Tx and Rx in order to provide the required conditions (SNR values) for identification of {UE's} (user's decoding) at the Rx's.

Summarizing all this, one can consider that the RIS design is the synthesis of the CM by the RIS controller for required conditions of {UE's}, whose identification algorithms will be presented at the next section.

Returning back to the RIS design issue, one has to assume that (see **Figure 1**) there are $i=\overline{1,n}$ RIS located in different sites of the MIMO system; each site is assigned to several {UE's} ($j=\overline{1,m}$). Then as it was shown in ([1] [13], etc.) on each site (in Power NOMA), for successful identification at the Rx, the conditions for each UE's have to be significantly different and are assumed to form a variation series:

$$h_1^2 < h_2^2 < \cdots < h_m^2, \tag{2.3}$$

where $\{h_j^2\}$ are SNR's values required for each {UE's} to be successfully identified with a predefined precision

Assuming the diagonal Ω, its "*n-th*" element is

$$\Omega_{n,n} = \left\langle \left| \boldsymbol{V}_{Tx_n}^H \theta \boldsymbol{V}_{Rx_n}^* \right|^2 \right\rangle, \tag{2.4}$$

where "*n*" is a PSWF index associated with $\{UE_n\}$, θ might be associated with PAS (Power Azimuth Spectrum) and denotes directions on RIS or mutual "losses" of the connected eigen modes for example in the metamaterial, or in the controller processing, etc.; < > denotes the statistical average operator.

Then, if h_j^2 for UE is a priori predefined, then:

$$\Omega_{n,n} \cong \frac{h_j^2}{\overline{\varphi_{n_{Tx}}^2}\,\overline{\varphi_{n_{Rx}}^2}\,\theta_{n,n}\lambda_n^2}. \tag{2.5}$$

Formula (2.5) is nothing else but a RIS controller algorithm.

Considering, that for the known characteristics and parameters of the MIMO system, eigen functions and $\{\lambda_n\}$ can be calculated a-priory, though, $\Omega_{n,n}$ can be easily calculated mainly a-priori and the algorithm for RIS is as easy as possible (see also some comments at [27]).

Example of the CM calculus is presented in **Figure 2** (see also [27]).

One can see that the presented algorithm for RIS design can be easily generalized in a "straightforward" way for the multiple RIS design case and for Massive Access in dense networks for multiple {UE's} [33] [34].

The latter gives an optimistic "hope" that such kind of RIS design, due to its simplicity, might be found as a prospective one for dense networks with large number of both {UE's} and sites.

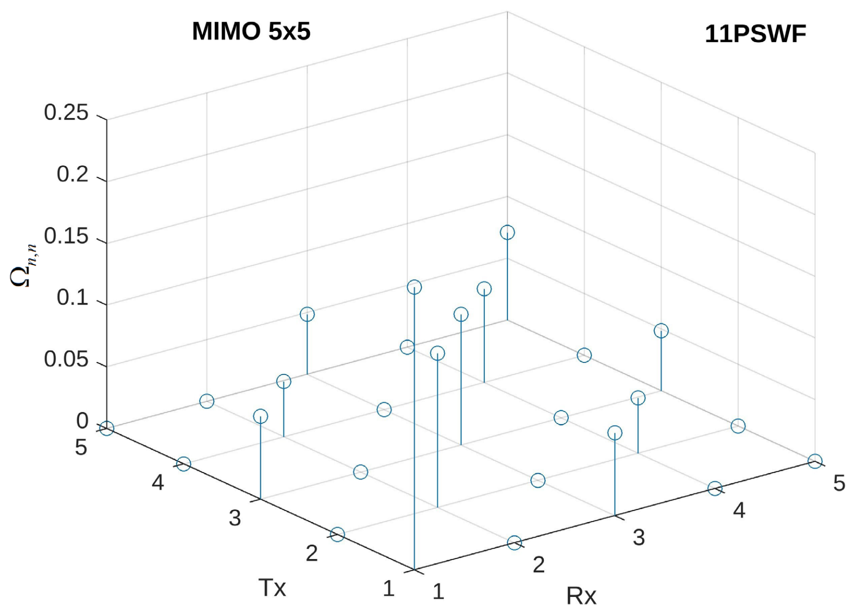

Figure 2. Calculus of CM.

3. Parallel Filtering or Filter Bank Approach, for NOMA User's Identification (MAI Cancelation)

3.1. Chaos Filtering and Its Properties

Keeping in mind that the current material is related to the scenarios corresponding to Double Selective Channels, the methods proposed for power NOMA Transmission will be first analyzed the so-called SIC (Successive Interference Cancelation) design. One must note that the idea of SIC was proposed and implemented long ago for MAI mitigation (see, for example [35] and references therein).

The SIC design has been thoroughly investigated and developed (see [1] [13] etc.); as it was already mentioned above (see Section 1) and the low feasibility of an accurate user identification provoked doubts regarding its application in power NOMA for MIMO transmission systems, particularly in Double Selective channels.

In this regard, the Filter Bank approach was proposed for the UE's identification at [13] for HSC. Its application is based on specific chaos filtering methods for "parallel filtering" of all {UE's} in a simultaneous fashion and to pass the filtered results to the multiple hypothesis sequential testing for UE's decoding (identification). This approach is characterized by reduced processing times [13] [36].

It is reasonable to remind once more, but briefly, that due to the physical phenomena at Double Selective channels both in time and frequency domains, the transmitted signals are "destroyed" into almost random processes with "double selective" shapes, characterized by such common distortions, as Doppler Shifts, Frequency Offsets, Frequency Shifts, etc. The latter gives an opportunity to apply the chaotic models (see [14] [15] [16] [17] [37] [38] [39], etc. and references therein) as a UE's signatures instead of random models in order to take advantage of the "singular" features (*i.e.* practical "invariance" to SNR value of the corresponding processing algorithms).

3.2. Chaos Filtering and {UE's} Identification (Classification)

The users' identification by means of their simultaneous filtering or estimation seems to be a rather opportunistic idea for the Doubly Selective Channels (see **Figure 3**).

At the same time, one has to notice that the Filter Bank Solution has to be based on the principles of precise algorithms as the aim is to classify, or "separate" (identify) the many different users by applying only their "signatures" in power NOMA transmission systems.

In [13], the SNR and SINR conditions, for effective users decoding (identification), were experimentally proposed (see also [13] for details).

It was shown there, that to neglect the influence of {UE's} classification errors for the final Noise Immunity of the transmission system, the so-called "significant difference" between the users (see [1]) has to be almost similar to their SNR values.

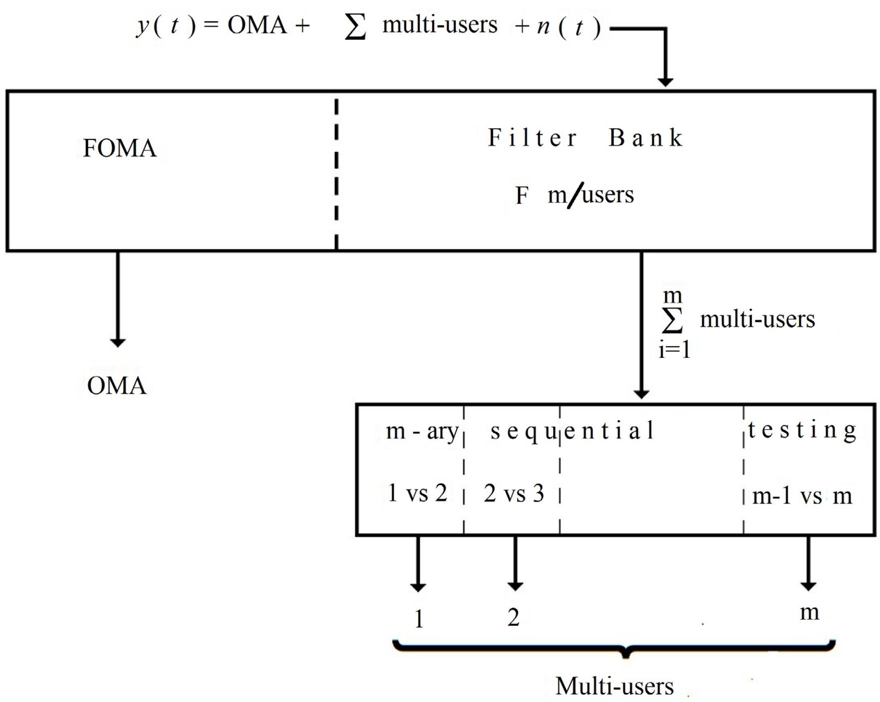

Figure 3. Block diagram for the UE's filtering classification (identification).

One has to notice, that the corresponding background for chaos filtering was exhaustively presented in the already mentioned references, though there is no sense to repeat it!

The most important extractions from those references are presented in the following:

- As an "optimum" option hereafter the Extended Kalman Filter (EKF) is chosen, implemented in the so-called "one moment" (1MMEKF) or "two moments" fashion (2MMEKF) [14] [16]. This option offers a good balance between filtering accuracy and computational complexity.
- The 1MMEKF and 2MMEKF filtering algorithms show some differences in terms of filtering accuracy, but in terms of computational complexity and processing time the difference is practically negligible; both of them are "simple" and "fast".
- Note, that the accuracy of the 1MMEKF and 2MMEKF does not demonstrate completely their "singular" properties in the same way as the optimum filtering [25] [37] but anyway those algorithms are rather accurate, efficient and universal.

The extractions (and some relevant comments) of the simulations results are presented in the following.

Under the influence of additive white noise, the simulations illustrate the efficiency of the Filter Bank approach for identification of OMA, {UE´s} and their further(final) classification by means of multiple hypothesis testing with sequential analysis algorithms [36] [40].

To illustrate the efficiency of the 1MMEKF and 2MMEKF algorithms for filtering OFDMA signals as OMA in presence of NOMA (MAI) interferences and additive white noise (as it is pointed here in **Figure 3**).

For simulations, the OMA and NOMA signals were modeled considering 50 carriers (see [13] for details related to the current Standard). From the correspondent figures it follows, that for SNR = 13 - 15 dB (usually applied in wireless communications) and SIR of the same order, the Normalized (regarding the signal power) Mean Square Error (NMSE) of the filtering is less than 5%.

This NMSE value shows, that the filtered OFDMA (OMA signal) preserves 95% of its average power after its filtering, while the SNR losses are less than 1 dB and the noise immunity of OMA is almost preserved.

Next, it follows from **Figure 4**, that the concept of sufficiently different channel conditions for the identification of OMA and NOMA users (see [1]) might be considered as around 9 - 12 dB, *i.e.* it is almost similar to the SNR. So the OMA signal can be successfully eliminated from the aggregate input signal in order to classify the NOMA users. The final efficiency of filtering for the individual users {UE´s} is illustrated in Table I of [27], which shows that it is also less than 5%.

The interested Reader might find in [13] various examples of the accuracy of OMA, NOMA filtering, which shows actually the same features, as in **Figure 4**.

After the separation of all UEs (OMA, NOMA) by means of the filter bank (see **Figure 4**), the "filtering results" have to be identified with each {UEs} model, *i.e.* it is required a classification procedure.

One way to achieve it, was proposed at [13] through the application of sequential m-hypothesis testing algorithms, which are characterized by reduced time consuming (see [13] [36] and references therein). The latter is detailed in

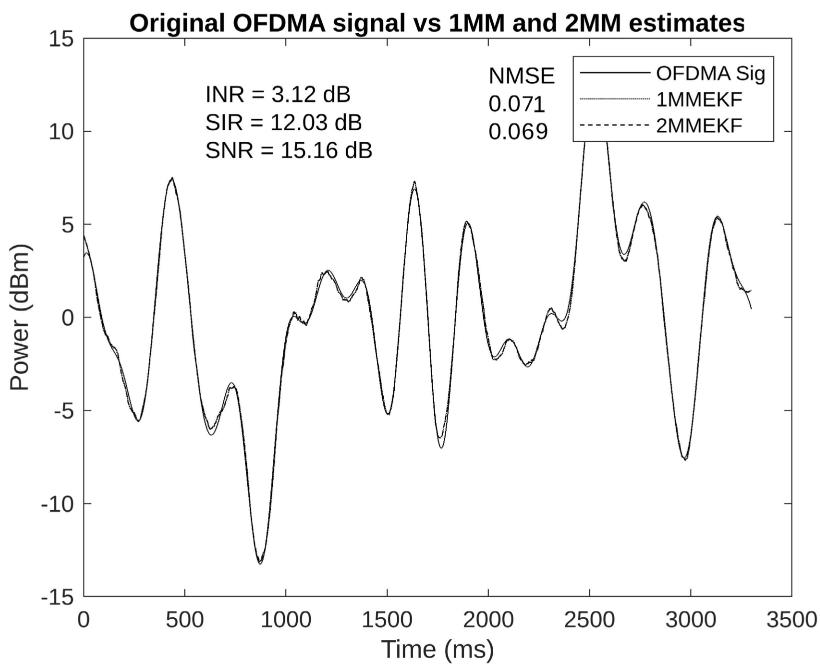

Figure 4. Illustration of the efficiency for OMA filtering in presence of MAI and noise.

[13] and is omitted in the following.

In [13], it was also mentioned that applying for classification errors characterization the Chernoff Upper bound with the help of so called minimum Kullback-Leibler distance between Gaussian hypothesis for the scenarios of significant difference between them, the classification error is much less than P_{err} and might be neglected [13]. This is the essence of the application of the "significant difference requirement" for the {UE's) (see above).

4. Incoherent ("Invariant") MIMO Transmission

4.1. Preface

The ongoing material is dedicated to the incoherent concept and moreover, associated with invariant features in relation to the channel properties.

One has to notice that the "incoherent concept" for application in Doubly Selective Channels has been already discussed in previously cited references [7] [8] [23] [24] [25], where the utilization of DPSK (in generalized fashion) was considered. Nevertheless the "invariant" (robust) aspects were not properly pointed out and so, here, after a brief sketch of this topic is provided, mainly based on the material of [11], stressing questions related to DPSK or more exactly to the differential phase shift keying of high order, *i.e.* DPSK-k.

The modulation DPSK-k is a generalization of the DPSK, that introduces k higher orders for the phase differences, was thoroughly investigated in [12] and it might have opportunistic practical applications for scenarios, where the goal is to achieve invariant properties to different channel distortions such as Doppler Shift, Frequency Offset, channel dispersions, etc[3].

Unfortunately, the ideas of [12] were not considered for NOMA transmission, so the following material is aimed to "fill this gap" (see also [11] for details). The interested Reader can find in [11] the simulation results for several scenarios for the channels illustrating the "invariant" features of DPSK-k.

The ideas proposed hereafter were inspired by [12], as an attempt to generalize DPSK-k for MIMO transmission over Doubly Selective Channels applying the channel orthogonalization together with incoherent paradigm for MIMO reception.

The good question is how all those issues might be useful for HSV channels, dense networks, etc. which doubtless will be part of future generations of communication systems such as 5G+, 6G and beyond. Some answers follow from the results presented below.

4.2. Artificial Trajectories and MIMO Reception: Generalized DPSK-k Modulation

The artificial trajectories for the MIMO channels with fading and dispersion presented above (see Section 2) were already broadly applied for multi-carrier

[3]Remind that the original DPSK or DPSK-1 is invariant only to the slow variations of the initial phase of channel (signal).

system design in Doubly Selective Channels (see [23] [25] and references therein) both for SISO and MIMO cases together with incoherent DPSK modulation. The following is completely dedicated to the analysis of invariant MIMO reception in Doubly Selective Channels based on the DPSK-k generalization [11].

The DPSK-k generalizations from the SISO (see [11]) to the MIMO scenarios require some mathematical work whose result yields rather cumbersome expressions [11]. Here, only a simplified explanation is offered.

For the $N_T \times N_R$ MIMO system with the application of Space Time Block Codes (STBCs) or Orthogonal SBTC (OSTBC), the same logic as at (18) is valid with the difference, that "scalar" description at [12] have to be "substituted "by matrixes (see those details at [11]).

Though, the constellation symbols form the matrix Δ_b (for the time instant "b") and the transmitted difference modulation matrix is:

$$\Gamma_b^k = f_w\left(\Gamma^{k-1}\right) = f_w\left(\Gamma^{k-2}\right) = \cdots = f_2\left(\Gamma^1\right) = f_1\left(\Delta\right). \tag{4.1}$$

Then, for demodulation, the estimated date Δ_b is calculated in the way:

$$\hat{\Delta}_b = f_z\left(\Psi^k\right) = \cdots = f_2\left(\Psi^2\right) = f_1\left(\Psi^1\right), \tag{4.2}$$

where Ψ^k is the received space modulated matrix.

The structures of these matrixes are based on the binomial coefficients (see analogy with scalar case at [12]) and show the necessary multiplications of Γ^1. The same logic is for $f_z\left(\Psi^k\right)$ and the necessary multiplication for Ψ^k (see details in [11]) if the autocovariance demodulator is considered.

This "straightforward" analogy to the "scalar" case might be useful in the following, as the further material, based on application of the concept of the processing of the "statistically independent" virtual trajectories for MIMO.

Though, one can consider from (4.1) and (4.2) that the autocovariance demodulator in the MIMO case is a representation of the autocovariance demodulation principle for the estimated transmitted matrix over a "b" time interval (instant) and the q-th artificial trajectory (see mathematical details in [11]).

It should be considered, that the autocovariance demodulator computationally the simplest option (compared for example with the ML version proposed in [8]), but for sure at the same time is the option with the lowest noise immunity properties.

So, how to encounter an adequate "balance" between noise immunity requirements and an attractive implementation? Somehow it follows from the next text.

4.3. Noise Immunity Evaluations and Simulation Results

First, it has to notice that the exact analytical calculation of the noise immunity for the DPSK-k autocovariance demodulation for the MIMO case is almost impossible to achieve because even for the rather simple SISO case it is also hardly possible to achieve (see [12]).

Here, after is applied the well-known Chernoff bounds (see [41]) as an upper bound for the BER in the binary case as it was applied in [11] with its further simulation validation, as it was presented there, in [11].

By applying the last option, it is possible to find the final expression for the Chernoff bound of the BER for the autocovariance demodulation in the form:

$$P_{\text{chernoff\_DPSK-k}} = \exp\left(-\alpha_k h^2\right), \tag{4.3}$$

where h^2 is the SNR at the demodulator input and the α can be found for any DPSK-*k*. The detailed explanation of this was already presented at Appendix B in [11].

Consequently: for the DPSK-1 $\alpha_1 = \frac{1}{4}$ (see [11]) and for DPSK-2 $\alpha_2 = \frac{1}{8}$ and for the DPSK-3 $\alpha_3 = \frac{1}{16}$ (see [11]) for Equation 4.3, etc.

Now, let us see the MIMO case with the assumption of quadratic addition algorithm for homogeneous conditions at "diversity" branches together with the same assumption for the artificial trajectory's addition.

One has to notice that the application of the artificial trajectories diminishes the necessary order of the number of diversity branches to achieve the channel hardening (see calculus in [22]),which finally yields to reduction on the fading strength (see for examples at [22] [42]) and "converts" the Doubly Selective Channel into the "constant" channel without fading and with aggregate SNR from all diversity branches (together with the virtual ones) in order to improve the noise immunity of the MIMO system and apply less complex channel error correcting codes (see the simulation results at [11]). Though [41], the final BER is:

$$P_e = \frac{Q}{\prod_{k=1}^{Q}\left(i + \alpha_k h^2\right)}. \tag{4.4}$$

Equation (4.4) is valid for statistically independent fading at all the diversity "branches" (physical and virtual).

In the following, the simulation results presented in [11] serve to verify the "robustness" of the upper bounds of the expressions (4.1), (4.4). They show, that such upper bounds offer no more than a 3 dB difference between the simulation data and the upper bound for the SNR around $P_{err} \sim 10^{-4}$. Such a kind of accuracy for the approximations of the upper bounds might be considered acceptable.

As it was mentioned before, due to the noise immunity losses (produced by the application of autocovariance demodulators), the error correcting codes (BCH, for simple example) might be a plausible solution for "cancelling" those losses for MIMO systems over Doubly Selective Channels, which together with the channel hardening can reduce the negative impact of the imminent noise increment for the autocovariance demodulators for DPSK-*k*.

Though the use of the error correcting code allows to "neglect" the consequences of the application of simple invariant demodulators. This issue is also illustrated in [11] (see also some comments in [27]).

5. Conclusions

In order to avoid any "misleading", it might be reasonable to stress from the very beginning of the "Conclusions", that the above-presented material doesn't have any "attempt" to deny the coherent paradigm. The material above is ONLY (!) to remind, that in Double Selective Channels (SISO, MIMO, etc) with or without NOMA, a coherent approach is almost impossible (or better, may be extremely hard) to implement and it might be reasonable to look for incoherent methods for implementation.

For power NOMA-MIMO transmission system design at Doubly Selective Channels, the presented material illustrates different aspects of the incoherent approach without any relation to the CSI, CSIT of the channel.

- Artificial (virtual) trajectories approach for the decomposition of the Channel System Functions in order to achieve a channel hardening at diversity combining at RX terminal, as fast as possible.
- Application of the artificial trajectories for RIS design by means of the Coupling Matrix of the GKCM as a model of the MIMO-RIS-NOMA channel.
- Filtering method together with an m-hypothesis sequential testing for effective classification of the {UE's}.
- Invariant DPSK-k modulation together with autocovariance demodulation to achieve an incoherent {UE's} demodulation processing, as simple as possible.

It is possible to characterize each of these approaches (see above) as somehow fast and simple, which makes them as a good option for operation at the Doubly Selective Channels.

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Conflicts of Interest

The author declares no conflicts of interest regarding the publication of this paper.

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