Finite-sample performance analysis of widely-linear multiuser receivers for DS-CDMA systems

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Abstract—This paper tackles the theoretical performance analysis of widely-linear (WL) multiuser receivers for direct-sequence code-division multiple-access (DS-CDMA) systems, as well as their comparison with conventional linear (L) ones. In particular, receivers based on the minimum output-energy (MOE) criterion are considered, since they offer a good tradeoff between performance and complexity and, moreover, lend to some simplifications in the analysis. After comparing the ideal signal-to-interference-plus-noise-ratio (SINR) performances of the WL-MOE and L-MOE receivers, the paper establishes finite-sample performance results for two typical data-estimated implementations. Specifically, by adopting a first-order perturbative approach, the SINR degradation of the data-estimated WL-MOE receivers is accurately evaluated and compared with that of its linear counterpart. Simulation results are provided to validate and complement the theoretical analysis.

Index Terms—Performance analysis, proper and improper random processes, multiuser detection, direct-sequence code-division multiple-access (DS-CDMA) systems, linear and widely-linear receiving techniques.

I. INTRODUCTION

During the last two decades, starting from the seminal work of Verdi [1], a great bulk of research activities has been devoted to multiuser detection (MUD), as an effective means to combat the multiple-access interference (MAI), which is the predominant source of performance degradation in (nonorthogonal) direct-sequence (DS) code-division multiple-access (CDMA) systems. Among MUD techniques, linear MUD (L-MUD) ones, such as the decorrelating receiver [2], the minimum mean-square-error (MMSE) [3] one, and the minimum output-energy (MOE) [4] one, have been investigated in depth, since they offer convenient tradeoffs between performance, complexity, robustness, amount of a priori information, and ease of adaptive implementation.

Most L-MUD techniques assume that the complex envelope \( r(t) \) of the received signal is modeled as a proper [5] random process, exploiting hence only the information contained in its statistical autocorrelation function \( R_{rr}(t, \tau) \triangleq E[r(t)r^*(t-\tau)] \). When, however, the DS-CDMA signal and/or the disturbance are improper [5], well-established results in detection and estimation theory [6] state that linear receivers can be outperformed by widely-linear (WL) ones, which jointly elaborate the received signal \( r(t) \) and its complex conjugate \( r^*(t) \), in order to exploit also the information contained in their statistical cross-correlation function \( R_{rr}(t, \tau) \triangleq E[r(t)r(t-\tau)] \). Many digitally modulated signals of practical interest are improper, such as ASK, differential BPSK (DBPSK), offset QPSK (OQPSK), offset QAM (OQAM), MSK and its variant Gaussian MSK (GMSK). Motivated from previous observations, in recent years several papers [7], [8], [9], [10] proposed different WL-MUD techniques for DS-CDMA systems with improper signals and/or disturbances, by extending concepts from the classical L-MUD theory. In particular, WL versions of the major L-MUD receivers have been proposed and studied, such as the WL decorrelating receiver [9], [11], the WL-MMSE one [7], [8], [12], the WL-MOE one [10], [11], and the min/max WL-MOE one [13].

In all the above-mentioned papers, the performance advantage of the WL-MUD receivers over their linear counterparts has been assessed mainly by means of computer simulations. Recently, with reference to DS-CDMA systems employing BPSK modulation, a few contributions addressing the theoretical performance analysis of WL-MUD techniques appeared in the literature. In [12], the asymptotic (in the number of users) performance analysis of the WL decorrelating and WL-MMSE receivers was carried out, by extending to the WL framework classical analysis tools already developed by Tse and Hanly [15] for L-MUD techniques (a similar study was proposed in [14]). A non-asymptotic performance analysis was instead considered in [16], which provides an algebraic proof that WL-MUD receivers outperform L-MUD ones, and explicitly assesses the expected performance gain in the two-users case. The common conclusion of these studies (see also [17]) is that the performance advantage of WL-MUD receivers over L-MUD ones is twofold: the input SNR is doubled and the number of effective interferers is halved. As a consequence, for a fixed processing gain \( N \), the number of users that can be accommodated by a DS-CDMA system employing WL-MUD is doubled [12], [14], [16] compared to L-MUD. In other words, unlike L-MUD, WL-MUD can be successfully employed not only when the number of users \( J \) is smaller than or equal to \( N \) (underloaded system), but also when \( N < J \leq 2N \) (overloaded system). However, none of the aforementioned papers on WL-MUD carried out a detailed study of the conditions on the channels and codes that assure perfect MAI suppression in the absence of noise. Thus, a
first contribution of this paper is to provide conditions on the spreading codes, which guarantee complete MAI rejection for WL-MUD in both underloaded and overloaded downlink configurations, when the DS-CDMA signal dominates the background noise. In particular, we will show that even the simple Walsh-Hadamard (WH) spreading codes can be suitably modified in order to fulfill such conditions.

Another limitation of almost all the performance studies carried out so far is the idealized assumption that the receivers are perfectly implemented. However, exception made for the WL decorrelating receiver, whose synthesis is data-independent, implementation of WL-MUD receivers requires knowledge of the second-order statistics (SOS) of the received signal \( r(t) \), which can be estimated in practice from a finite number of samples. A theoretical performance analysis of the data-aided WL-MMSE and WL-MOE receivers was provided in [18], when the receivers are adaptively implemented by means of the least-mean square (LMS) algorithm, by evaluating the output signal-to-interference-plus-noise ratio (SINR). However, the SINR analysis carried out in [18] considers steady-state performances, i.e., when the sample size is infinite, and, thus, does not allow to evaluate the performance of the receivers as a function of the number of samples. This issue is important from a practical point of view because, especially when short sample-sizes are employed, the data-estimated versions of the WL-MUD receivers exhibit a severe performance degradation with respect to their ideal counterparts, reducing thus the expected performance gain over L-MUD receivers. To gain more insight about this point, this paper presents a finite-sample theoretical performance analysis of WL-MUD receivers, based on a first-order perturbative approach. In particular, taking as reference the WL-MOE receiver\(^1\), two typical data-estimated implementations are considered: the WL-SMI (sample matrix inversion) receiver, which employs a sample estimate of the data autocorrelation matrix, and the WL-SUB (subspace) receiver, which exploits the properties of the eigenvalue decomposition (EVD) of the data autocorrelation matrix to reduce the effects of estimation errors. Finally, besides deriving accurate yet simple theoretical results for the finite-sample versions of the WL-MOE receivers, we will show that even some known results for the linear receivers must be reinterpreted or modified, to allow for a fair comparison between WL-MUD and L-MUD.

The paper is organized as follows. In Section II, we present the DS-CDMA system model and introduce the WL reception strategy. L-MOE and WL-MOE receivers are introduced in Section III, whereas Section IV analyzes their ideal performances, and derives simple conditions on the spreading codes, which assure perfect MAI suppression in the downlink. Section V presents the finite-sample performance analysis of the L-MOE and WL-MOE receivers. The theoretical results reported in Sections IV and V are validated and supported by computer simulations examples, whereas their proofs are gathered in Appendix I. Finally, concluding remarks are given in Section VI.

### A. Notations

The fields of complex, real, and integer numbers are denoted by \( \mathbb{C}, \mathbb{R}, \text{and } \mathbb{Z} \), respectively; matrices [vectors] are denoted with upper case [lower case] boldface letters (e.g., \( \mathbf{A} \) or \( \mathbf{a} \)); the function of \( m \times n \) complex [real] matrices is denoted as \( \mathbb{C}^{m \times n} [\mathbb{R}^{m \times n}] \), with \( \mathbb{C}^m [\mathbb{R}^m] \) used as a shorthand for \( \mathbb{C}^{m \times 1} [\mathbb{R}^{m \times 1}] \); the superscripts \( *, T, H, \cdot -1 \) and \( \dagger \) denote the conjugate, the transpose, the Hermitian (conjugate transpose), the inverse, and the Moore-Penrose generalized inverse [19] (pseudo-inverse) of a matrix, respectively; \( \mathbf{0}_m \in \mathbb{R}^m, \mathbf{0}_{m \times n} \in \mathbb{R}^{m \times n} \) and \( \mathbf{I}_n \in \mathbb{R}^{m \times m} \) denote the null vector, the null matrix, and the identity matrix, respectively; \( \text{trace}(\cdot) \) and \( \text{rank}(\cdot) \) represent the trace and the rank; \( \mathcal{N}(\mathbf{A}), \mathcal{R}(\mathbf{A}) \), and \( \mathcal{R}^\perp(\mathbf{A}) \) denote the null space, the range (column space), and the orthogonal complement of the column space of \( \mathbf{A} \in \mathbb{C}^{m \times n} [\mathbb{R}^{m \times n}] \) in \( \mathbb{C}^m [\mathbb{R}^m] \); for any \( \mathbf{a} \in \mathbb{C}^m, ||\mathbf{a}|| \triangleq (\mathbf{a}^H \mathbf{a})^{1/2} \) denotes the Euclidean norm; \( \mathbf{A} = \text{diag}(A_{11}, A_{22}, \ldots, A_{nn}) \) is a diagonal matrix with elements \( A_{ii} \) on the main diagonal; \( \text{E}[\cdot] \) denotes ensemble averaging, \( i \triangleq \sqrt{-1} \) is the imaginary unit, and \( \text{Re}[^{\cdot}], \text{Im}[^{\cdot}] \) denote the real and imaginary parts; throughout the paper, we occasionally use the simplified notations \( \mathbf{a}_R \triangleq \text{Re}[\mathbf{a}], \mathbf{a}_I \triangleq \text{Im}[\mathbf{a}], \mathbf{A}_R \triangleq \text{Re}[\mathbf{A}], \text{and } \mathbf{A}_I \triangleq \text{Im}[\mathbf{A}] \); \( \delta_k \) denotes the Kronecker delta (i.e., \( \delta_k = 1 \) for \( k = 0 \), zero otherwise); for any stationary discrete-time random vector process \( \mathbf{x}(k) \in \mathbb{C}^m \), we denote with \( \mathbf{R}_{xx} \triangleq \text{E}[\mathbf{x}(k) \mathbf{x}^H(k)] \) in \( \mathbb{C}^{m \times m} \) and with \( \mathbf{R}_{xx}^\perp \triangleq \text{E}[\mathbf{x}(k) \mathbf{x}^T(k)] \) in \( \mathbb{C}^{m \times m} \) the autocorrelation matrix and the conjugate correlation matrix, respectively (\( \mathbf{R}_{xx} \equiv \mathbf{R}_{xx}^\perp \in \mathbb{R}^{m \times m} \) when \( \mathbf{x}(k) \in \mathbb{R}^m \)).

### II. PROBLEM FORMULATION AND WL RECEPTION TECHNIQUES

Let us consider the baseband model of a DS-CDMA system with \( J \) users, employing short spreading codes with \( N/T \) chips/symbol. After chip-matched filtering, perfect time synchronization and sampling with rate \( N/T \), the received vector \( \mathbf{r}(k) \in \mathbb{C}^N \) collecting the \( N \) samples of the incoming signal in the interval \( [kT, (k+1)T) \), with \( k \in \mathbb{Z} \), can be written [4] as follows:

\[
\mathbf{r}(k) = \sum_{j=1}^{J} \alpha_j e^{i\theta_j} \psi_j \mathbf{b}_j(k) + \mathbf{v}(k) = \mathbf{Ψ} \mathbf{A} \mathbf{b}(k) + \mathbf{v}(k)
\]

\[
= \Phi \mathbf{b}(k) + \mathbf{v}(k),
\]

where, with reference to the \( j \)th user, \( \alpha_j > 0 \) is the received amplitude (accounting for transmitted energy and channel propagation loss), \( \theta_j \in [0, 2\pi) \) is a precoding phase (which is deliberately introduced at the transmitter and whose role will be clear in the sequel), \( \psi_j \in \mathbb{C}^N \) is the unit-norm signature (encompassing spreading code and channel

\(^1\)Although we consider the WL-MOE receiver, since it tends to some simplifications in the analysis, the obtained results can be applied also to the WL-MMSE receiver, since it is well known that the MMSE and MOE approaches are equivalent [4] in terms of output SINR. Moreover, it is not difficult to extend our analysis to other categories of receivers.

\(^2\)The considered signal model is appropriate when: (i) the users are synchronous; (ii) the user channels introduce only interchip interference [4] and negligible inter symbol interference (ISI). However, all the results derived herein can be extended with straightforward modifications to account for asynchronous users and/or channels with ISI.
propagation effects), and \( b_j(k) \) is the transmitted symbol, whereas \( v(k) \in \mathbb{C}^N \) accounts for thermal noise. Moreover, in (1), we have defined \( \Psi \triangleq [\psi_1, \psi_2, \ldots, \psi_J] \in \mathbb{C}^{N \times J} \), \( A \triangleq \text{diag}(a_1, a_2, \ldots, a_J) \in \mathbb{R}^{J \times J} \), \( \Theta \triangleq \text{diag}(e^{j\theta_1}, e^{j\theta_2}, \ldots, e^{j\theta_J}) \in \mathbb{C}^{J \times J} \), and the symbol vector \( b(k) \triangleq [b_1(k), b_2(k), \ldots, b_J(k)]^T \in \mathbb{R}^J \). Finally, let \( c_j \triangleq [c_j(0), c_j(1), \ldots, c_j(N - 1)]^T \in \mathbb{C}^N \) denote the spreading vector of the \( j \)th user and let \( g_j(n) \) be the corresponding baseband chip rate discrete-time impulse response, under the assumption that \( g_j(n) \) has order \( L_{g_j} \ll N \), the signature \( \psi_j \) in (1) can be modeled [4] as

\[
\psi_j = G_j c_j,
\]

where \( G_j \in \mathbb{C}^{N \times N} \) is a Toeplitz lower triangular matrix with first column \([g_j(0), g_j(1), \ldots, g_j(L_{g_j})]^T \) and first row \([g_j(0), 0, 0, 0]^T \). Throughout the paper, we will assume that: (a1) \( b(k) \) is a binary \(^3 \) real zero-mean random vector, whose entries are independent and identically distributed (i.i.d.) random variables assuming equiprobable values in \([-1, 1]\), with \( b(k_1) \) and \( b(k_2) \) statistically independent for \( k_1 \neq k_2 \); (a2) \( v(k) \) is a complex proper [5] zero-mean Gaussian random vector, independent from \( b(k) \), where \( R_{vv} = \sigma_v^2 I_N \) and \( R_{vv^*} = O_{N \times N} \), with \( v(k_1) \) and \( v(k_2) \) statistically independent for \( k_1 \neq k_2 \). DS-CDMA systems with real \( b(k) \) have been considered in [7], [8], [9], [11]. Moreover, models similar to (1) arise also in other applications; in particular, the case of a real \( b(k) \) has been considered in array processing [20], block equalization [21], orthogonal frequency-division multiplexing (OFDM) [22], multicarrier (MC) CDMA systems [23], and multiple-input multiple-output (MIMO) systems [24]. Therefore, most results of our analysis can be extended with minor modifications also to other application areas employing WL reception techniques.

Under assumptions (a1) and (a2), the minimum-error-probability detection of \( b(k) \) is based only on \( r(k) \) (one-shot detection), and is equivalent to the maximum-likelihood (ML) rule

\[
\hat{b}(k) = \arg\min_{b(k) \in \mathbb{R}^J} \| r(k) - \Phi b(k) \|^2 = \arg\max_{b(k) \in \mathbb{R}^J} \left\{ 2 b^T(k) \text{Re}[\Phi^H r(k)] - b^T(k) \Phi^H \Phi b(k) \right\}.
\]

This equation shows that the real vector \( x_R(k) \triangleq \text{Re}[\Phi^H r(k)] \in \mathbb{R}^J \), containing the real parts of the matched filter outputs, is a sufficient statistic for recovering \( b(k) \); this was recognized in [25], and pointed out more recently in [9], [26], [27]. Unless matrix \( \Phi \) has some special structure (e.g., orthogonal columns), implementing ML detection entails a complexity that grows exponentially with the number of users \( J \). Focusing on detection of user \( j \), a common suboptimal detection strategy is to perform linear processing of \( x_R(k) \) by a weight vector \( \tilde{g}_j \in \mathbb{R}^J \):

\[
y_j(k) = \tilde{g}_j^T x_R(k),
\]

followed by a \( \text{sgn}(x) \) nonlinearity to detect \( b_j(k) \). The main drawback of (4) is that, similarly to the ML receiver, it requires knowledge of the entire matrix \( \Phi \), i.e., signatures, amplitudes and phases of all the users. However, by recalling that \( x_R(k) = \text{Re}[x(k)] = \frac{1}{2} x(k) + \frac{1}{2} x^*(k) \), the filter defined by (4) can be equivalently expressed as

\[
y_j(k) = \tilde{g}_j^T x_R(k) = \frac{1}{2} \tilde{g}_j^T x(k) + \frac{1}{2} \tilde{g}_j^T x^*(k) = \frac{1}{2} \tilde{g}_j^T \Phi^H r(k) + \frac{1}{2} \tilde{g}_j^T \Phi^T r(k),
\]

which can be simply rewritten as

\[
y_j(k) = f_{j,1}^H r(k) + f_{j,2}^H r^*(k) = f_j^H z(k),
\]

where \( f_{j} \triangleq [f_{j,1}^T, f_{j,2}^T]^T \in \mathbb{C}^{2N} \), with \( f_{j,1} \triangleq \frac{1}{2} \Phi \tilde{g}_j \in \mathbb{C}^N \) and \( f_{j,2} \triangleq \frac{1}{2} \Phi^* \tilde{g}_j \in \mathbb{C}^N \), and, moreover, \( z(k) = [r^T(k), r^H(k)]^T \in \mathbb{C}^{2N} \). Regarding \( f_{j,1} \) and \( f_{j,2} \) as free vectors to be optimized, it is apparent that filter (5) operates directly on the received vector \( r(k) \), without requiring knowledge of \( \Phi \). Eq. (5) defines a \( \text{WL transformation} \), involving both \( r(k) \) and its complex conjugate \( r^*(k) \), whereas the associated complex transformation \( z(k) \to y_j(k) \) is \( \text{linear} \) with respect to (w.r.t) \( z(k) \) and, hence, the synthesis of \( f_{j} \) is simplified by resorting to this “augmented” formulation. However, observe that the output of (5) is real only when \( f_{j,1} = f_{j,2}^* \), which is a \( \text{nonlinear} \) constraint on \( f_{j} \) which will be referred to as the \( \text{conjugate symmetry (CS) constraint} \) in the following] that must be necessarily incorporated in any optimization criterion if the equivalence between (4) and (5) has to be preserved. Nevertheless, with reference to the maximum SINR criterion adopted later on, we will show that such a constraint is automatically satisfied by the unconstrained solution, hence it does not complicate in practice the receiver synthesis. We will rely in the following on formulation (5), since it allows many advantages not only in the synthesis, but also in the performance analysis of the receivers. Note that, unlike (4), eq. (5) encompasses as a particular case the \( \text{linear receiver} \)

\[
y_j(k) = g_j^H r(k),
\]

with \( g_j \in \mathbb{C}^N \), which can be obtained indeed by setting \( f_{j,1} = g_j \) and \( f_{j,2} = 0_N \). The output of such a linear receiver is not necessarily real-valued, hence it cannot be equivalent to (4); in spite of this incongruence, it is commonly adopted in many detection problem modeled by (1), even when the vector \( b(k) \) is real-valued (see, e.g., [28]).

III. THE WL-MOE AND L-MOE RECEIVERS

The main goal of this section is to derive the WL-MOE receiver as a particular solution of the maximum SINR criterion. We start by reviewing briefly the L-MOE receiver, not only to put the necessary bases for our subsequent derivations, but also to comment on possible inconsistencies concerning the “correct” definition of the SINR to be used for linear receivers, when real symbols are employed.

In order to recover \( b_j(k) \) by a linear receiver, it is useful to rewrite (1) as follows:

\[
r(k) = \phi_j b_j(k) + \tilde{b}_j(k) + \nu(k) = \phi_j b_j(k) + \nu_j(k),
\]

where
where \( \phi_j \in \mathbb{C}^N \) is the \( j \)th column of the composite matrix \( \Phi \), whereas \( \mathbf{b}_j(k) \in \mathbb{R}^{J-1} \) denotes the vector that includes all the elements of \( \mathbf{b}(k) \) except for the \( j \)th entry \( b_j(k) \). \( \mathbf{b}_j \in \mathbb{C}^{N \times (J-1)} \) denotes the matrix that includes all the columns of \( \Phi \) except for the \( j \)th column \( \phi_j \), and, finally, \( \mathbf{p}_j \equiv \overline{\mathbf{T}}_j \mathbf{b}_j + \nu(k) \in \mathbb{C}^N \) is the interference-plus-noise (disturbance) vector. Accounting for (6), the output of a linear receiver can be expressed as

\[
y_j(k) = g_j^H r(k) = g_j^H \phi_j b_j(k) + g_j^H \mathbf{p}_j(k). \tag{7}
\]

The L-MOE receiver [4] is the solution of the following constrained optimization problem:

\[
g_{j,\text{L-MOE}} = \arg\min_{g_j \in \mathbb{C}^N} E[|y_j(k)|^2] \quad \text{subject to} \quad g_j^H \phi_j = 1, \tag{8}
\]

which can be solved by Lagrange optimization, yielding the two equivalent\(^4\) expressions:

\[
g_{j,\text{L-MOE}} = (\phi_j^H \mathbf{R}_{\mathbf{pp}} \phi_j)^{-1} \mathbf{R}_{\mathbf{pp}} \phi_j
\]

\[
= (\phi_j^H \mathbf{R}_{\mathbf{pp}} \phi_j)^{-1} \mathbf{R}_{\mathbf{pp}} \phi_j, \tag{9}
\]

where the second equality follows by applying the matrix inversion lemma\(^5\) to the autocorrelation matrix \( \mathbf{R}_{\mathbf{pp}} = \phi_j \phi_j^H + \mathbf{R}_{\mathbf{pp}} \). It can be easily shown that, among all linear receivers, the L-MOE one maximizes the SINR at its output, which, accounting for (7), can be defined as

\[
\text{SINR}(g_j) \triangleq \frac{E[|g_j^H \phi_j b_j(k)|^2]}{E[|g_j^H \phi_j|^2]} = \frac{|g_j^H \phi_j|^2}{g_j^H \mathbf{R}_{\mathbf{pp}} g_j} = \frac{|\mathbf{R}_{\mathbf{pp}} g_j|}{g_j^H \mathbf{R}_{\mathbf{pp}} g_j} \cdot \tag{10}
\]

Indeed, by using the Cauchy-Schwartz’s inequality\(^6\), any receiver maximizing (10) is given by \( g_{j,\text{max-SINR}} = \gamma_j \mathbf{R}_{\mathbf{pp}} \phi_j \), where \( \gamma_j \in \mathbb{C} - \{0\} \) is an arbitrary (nonnull) complex scalar. Hence, the L-MOE receiver is obtained by setting \( \gamma_j = (\phi_j^H \mathbf{R}_{\mathbf{pp}} \phi_j)^{-1} \), and the maximum value of (10) is

\[
\text{SINR}_{j,\text{max}} \triangleq \text{SINR}(g_{j,\text{L-MOE}}) = \frac{1}{g_j^H \mathbf{R}_{\mathbf{pp}} g_j} \mathbf{R}_{\mathbf{pp}} \phi_j = \phi_j^H \mathbf{R}_{\mathbf{pp}} \phi_j. \tag{11}
\]

Turning to the WL-MOE receiver, we preliminarily express \( z(k) \), defined in (5), as

\[
z(k) = \mathbf{H} \mathbf{b}(k) + \mathbf{d}(k), \tag{12}
\]

where \( \mathbf{H} \triangleq \begin{bmatrix} \mathbf{\Phi}_j^T & \mathbf{\Phi}_j^H \end{bmatrix} \in \mathbb{C}^{2N \times J} \) and \( \mathbf{d}(k) \triangleq \begin{bmatrix} \nu^T(k) \nu^H(k) \end{bmatrix} \in \mathbb{C}^{2N} \). Accounting for (a2), the noise \( \mathbf{d}(k) \)

\(^4\)The advantage of using \( \mathbf{R}_{\mathbf{pp}} \) instead of \( \mathbf{R}_{\mathbf{pp}} \) in (9) is that the former can be estimated from received data.

\(^5\)Given the vectors \( x, y \in \mathbb{C}^N \) and the nonsingular matrix \( X \in \mathbb{C}^{N \times N} \), the matrix inversion lemma states that \( (X + xy^H)^{-1} = X^{-1} - (1 + y^H X^{-1} x)^{-1} X^{-1} x y^H X^{-1} \).

\(^6\)Given the vectors \( x, y \in \mathbb{C}^n \), the Cauchy-Schwartz’s inequality states that \( |x^H y|^2 \leq \|x\|^2 \|y\|^2 \), where the upper bound is achieved by \( y = \gamma x \), with \( \gamma \in \mathbb{C} \).

\(^7\)When \( N \) and \( J \) are large enough, this assumption is well-satisfied for maximum-SINR equalizers (see, e.g. [30]).

is an improper Gaussian random vector, with \( \mathbf{R}_{dd} = \sigma_d^2 \mathbf{I}_{2N} \) and \( \mathbf{R}_{dd} = \sigma_d^2 \mathbf{J}_{2N} \), where

\[
\mathbf{J}_{2N} \triangleq \begin{bmatrix} \mathbf{O}_{N \times N} & \mathbf{I}_N \\ \mathbf{I}_N & \mathbf{O}_{N \times N} \end{bmatrix} \in \mathbb{R}^{2N \times 2N}, \tag{13}
\]

is a block permutation matrix [29]. Accounting for (12), eq. (5) can be written as

\[
y_j(k) = f_j^H h_j b_j(k) + f_j^H \mathbf{H} \mathbf{b}_j(k) + d(k)
\]

\[
= f_j^H h_j b_j(k) + f_j^H \mathbf{q}_j(k), \tag{14}
\]

where \( h_j = [\phi_j^T, \phi_j^H]^T \in \mathbb{C}^{2N} \), \( \mathbf{H}_j = [\mathbf{F}_j^T, \mathbf{J}_j^H]^T \in \mathbb{C}^{2N \times (J-1)} \), and \( \mathbf{q}_j(k) \triangleq [p_j^T(k), p_j^H(k)]^T = \mathbf{H}_j \mathbf{b}_j(k) + d(k) \in \mathbb{C}^{2N} \) is the augmented disturbance vector. To establish a general framework encompassing both linear and WL receivers, we refer to the scheme in Fig. 1, wherein linear receivers can be obtained by setting \( f_{j,2} = 0_N \), and the \( \text{Re}[] \) operation is needed only when \( y_j(k) \) is complex, as it happens for linear receivers, or even for WL ones possibly not satisfying the CS constraint. It should be observed that the L-MOE receiver maximizes the SINR given by (10), which is evaluated before the \( \text{Re}[] \) block. Since \( b_j(k) \) is real, a more appropriate performance measure is the SINR after the \( \text{Re}[] \) block, which can be written, accounting for (14), as

\[
\text{SINR}(f_j) \triangleq \frac{E[\mathbf{R}^2 f_j^H h_j b_j(k)]}{E[\mathbf{R}^2 f_j^H \mathbf{q}_j(k)]} = \frac{\text{Re}^2 f_j^H h_j}{\text{Re}^2 f_j^H \mathbf{q}_j(k)}, \tag{15}
\]

Indeed, if the disturbance contribution \( f_j^H \mathbf{q}_j(k) \) at the receiver output can be approximated as a Gaussian random variable\(^7\), maximizing (15) w.r.t. \( f_j \) amounts to minimizing the error probability

\[
P_{e,j} \triangleq \Pr \{ b_j(k) \neq \mathbf{b}_j(k) \} \approx Q(\sqrt{\text{SINR}(f_j)}),
\]

where \( Q(x) \triangleq (1/\sqrt{2\pi}) \int_x^{\infty} e^{-u^2/2} du \) denotes the \( Q \) function.

Since maximization of (15), due to the presence of the \( \text{Re}[] \) operator, is not as standard as maximizing (10), we discuss it briefly in the following Lemma.

**Lemma 1:** Any WL receiver (5) maximizing (15) can be expressed as

\[
f_{j,\text{max-SINR}} = \xi_j \mathbf{R}_{\mathbf{qq}}^{-1} h_j + f_{j,a}, \tag{16}
\]

where \( \xi_j \in \mathbb{R} - \{0\} \) is an arbitrary (nonnull) real scalar and \( f_{j,a} \) is an arbitrary antisymmetric vector, i.e., \( f_{j,a} \in \mathcal{A} \triangleq \{ f = \}

\[ \]
 SINR is given by
\[ \text{SINR}_{j,\text{max}} \triangleq \text{SINR}(f_{j,\text{max-SINR}}) = h_j^H R_{\mathbf{q}_j,\mathbf{q}_j}^{-1} h_j. \]  

Proof: See Appendix A.

Note that the maximum SINR solution (16) differs from that of the linear case for the fact that the scalar \( \xi \) for both linear and WL receivers, following [18], we evaluate the SINR whose synthesis is based on perfect knowledge of the SOS inversion lemma (see footnote 5) to the autocorrelation matrix \( R_{\mathbf{xx}} = \mathbf{h}_j^H h_j + R_{\mathbf{q}_j,\mathbf{q}_j} \). By reasoning as in the proof of Lemma 1, it can be shown that (18) is obtained equivalently as the unique solution of the following WL-MOE criterion:
\[ f_{j,\text{WL-MOE}} = \arg\min_{f_j \in \mathbb{C}^{2N}} \mathbb{E}[\Re \{ y_j(k) \}] \text{ subject to } f_j^H h_j = 1. \]  

IV. IDEAL PERFORMANCES OF THE L-MOE AND WL-MOE RECEIVERS

In this section, we compare the SINR performances of the ideal WL-MOE and the L-MOE receivers, i.e., those receivers whose synthesis is based on perfect knowledge of the SOS of the received signal. The analysis for the data-estimated versions of the receivers will be carried out in Section V.

In order to carry out a meaningful performance comparison between linear and WL receivers, following [18], we evaluate for both receivers the SINR after the Re[\cdot] block, given by (15). Since the WL-MOE receiver maximizes such a SINR (see Lemma 1), one simply has:
\[ \text{SINR}_{j,\text{WL-MOE}} \triangleq \text{SINR}(f_{j,\text{WL-MOE}}) = h_j^H R_{\mathbf{q}_j,\mathbf{q}_j}^{-1} h_j. \]  

Instead, observe that evaluating the SINR given by (15) for the L-MOE receiver leads to a result generally different from (11). By observing that the L-MOE receiver can be viewed as a WL receiver with augmented weight vector \( f_{j,\text{L-MOE}} \triangleq [g_{j,\text{L-MOE}}^T, \mathbf{0}_N^T]^T \), recalling that \( g_{j,\text{L-MOE}}^H \mathbf{\phi}_j = 1, \) and applying the straightforward identity \( \Re \{ z \} = \frac{1}{2} \{ z + \Re \{ z \} \}, \forall z \in \mathbb{C} \), the SINR (15) for the L-MOE receiver can be written as
\[ \text{SINR}_{j,\text{L-MOE}} \triangleq \text{SINR}(f_{j,\text{L-MOE}}) = \frac{1}{2} \mathbb{E}[\Re \{ g_{j,\text{L-MOE}}^H P_j(k) \}] = \frac{g_{j,\text{L-MOE}}^H R_{\mathbf{p}_j,\mathbf{p}_j} g_{j,\text{L-MOE}}^* + \Re \{ g_{j,\text{L-MOE}}^H R_{\mathbf{p}_j,\mathbf{p}_j} g_{j,\text{L-MOE}}^* \}}{g_{j,\text{L-MOE}}^H R_{\mathbf{p}_j,\mathbf{p}_j} g_{j,\text{L-MOE}}^* + \Re \{ g_{j,\text{L-MOE}}^H R_{\mathbf{p}_j,\mathbf{p}_j} g_{j,\text{L-MOE}}^* \}}. \]  

On one hand, since the WL-MOE is a maximum-SINR receiver, it results that \( \text{SINR}_{j,\text{L-MOE}} \leq \text{SINR}_{j,\text{WL-MOE}} \). On the other hand, since \( \Re^2 \{ z \} \leq |z|^2, \forall z \in \mathbb{C} \), accounting for (11), one has \( \text{SINR}_{j,\text{L-MOE}} \geq \text{SINR}_{j,\text{max}} \). Overall, we maintain that
\[ \text{SINR}_{j,\text{WL-MOE}} \geq \text{SINR}_{j,\text{L-MOE}} \geq \text{SINR}_{j,\text{max}}. \]  

Although the first inequality in (22) concisely states that the performance of the WL-MOE receiver is not worse than that of its linear counterpart, it does not allow us to quantify the relative performance gain. Indeed, no clear insight on the performance comparison between the WL-MOE and L-MOE receivers can be drawn out from the SINR formulas (20) and (21). To overcome this conceptual difficulty, we carry out in the next subsection the performance comparison in the high-SNR regime, by deriving the analytical expressions of \( \text{SINR}_{j,\text{WL-MOE}} \) and \( \text{SINR}_{j,\text{L-MOE}} \) as the noise variance \( \sigma_n^2 \) approaches zero. It should be observed that, more generally, the results reported in Subsection IV-A turn out to be useful in all those situations wherein the DS-CDMA signal dominates the background noise, which is a common occurrence in many practical environments.

A. Analysis in the high-SNR regime

The discussion carried out in this subsection is mainly based on some mathematical results whose proofs are reported in Appendix B. Such results show that, in the limiting case of vanishingly small noise, i.e., as \( \sigma_n^2 \to 0 \), the performance comparison between the L-MOE and WL-MOE receivers heavily depends on the rank properties of \( \Phi \) and \( \mathbf{H} \), respectively.

As regards linear processing, it is shown that, in the high-SNR regime, the L-MOE receiver is able to achieve perfect MAI suppression for each active user, that is, \( \lim_{\sigma_n^2 \to 0} \text{SINR}_{j,\text{L-MOE}} = \lim_{\sigma_n^2 \to 0} \text{SINR}_{j,\text{max}} = +\infty, \forall j \in \{1,2,\ldots,J\} \), and if only if (iff) the matrix \( \Phi \) is full-column rank, i.e., \( \text{rank}(\Phi) = J \). Moreover, in such a case, it results that
\[ \lim_{\sigma_n^2 \to 0} \frac{\text{SINR}_{j,\text{L-MOE}}}{\text{SINR}_{j,\text{max}}} = 2, \forall j \in \{1,2,\ldots,J\}, \]  

which shows that, as intuitively expected, since the Re[\cdot] block in Fig. 1 discards one-half of the noise-plus-MAI power in \( y_j(k) \), \( \text{SINR}_{j,\text{L-MOE}} \) is asymptotically greater than \( \text{SINR}_{j,\text{max}} \) of exactly 3 dB. Note that this simple result holds only when \( \text{rank}(\Phi) = J \). If the matrix \( \Phi \) is not full-column rank, the L-MOE receiver is unable to perfectly suppress the MAI, even in the absence of noise; in this case, both \( \text{SINR}_{j,\text{max}} \) and \( \text{SINR}_{j,\text{L-MOE}} \) take on finite values, which depend on \( \mathbf{\phi}_j \) and the eigenstructure of the MAI autocorrelation matrix \( \mathbf{\Phi}^H \mathbf{\Phi} \). Therefore, the assumption \( \text{rank}(\Phi) = J \) is crucial and deserves a brief comment. By virtue of nonsingularity of the diagonal matrices \( \mathbf{A} \) and \( \mathbf{\Theta} \), it follows that \( \text{rank}(\mathbf{A} \mathbf{\Theta}) = \text{rank}(\mathbf{\Theta}) \). Henceforth, the matrix \( \mathbf{F} \) is full-column rank iff the signatures \( \psi_1, \psi_2, \ldots, \psi_J \) are linearly independent, a condition which can be fulfilled only if the number of users \( J \) is smaller than or equal to the processing gain \( N \) (underloaded systems). It is noteworthy that the linear independence of the signatures \( \psi_1, \psi_2, \ldots, \psi_J \) depends on both the spreading codes and the channel impulse responses of all the active users. Thus, in general, it is difficult to give
easily interpretable conditions assuring that $\Psi$ is full-column rank. A substantial simplification occurs in the downlink, wherein all the user signals propagate through a common multipath channel, i.e., $g_j(n) = g(n)$, with order $L_{g_j} = L_g$, for each user. In this case, the signature $\psi_j$ given by (2) becomes $\psi_j = G e_j$, where the common Toeplitz channel matrix $G = G_\Psi$ turns out to be nonsingular under the mild assumption that $g(0) \neq 0$, which is assumed to hold hereinafter. Accounting for this model, the matrix $\Psi$ becomes

$$\Psi = G \left[ e_1, e_2, \ldots, e_J \right] = G C,$$

which, by virtue of nonsingularity of $G$, implies that

$$\text{rank}(\Phi) = \text{rank}(\Psi) = \text{rank}(C).$$

Consequently, in the downlink scenario, the linear independence of the spreading vectors $e_1, e_2, \ldots, e_J$ is a necessary and sufficient condition for assuring the full-column rank property of $\Phi$ and, hence, allowing the L-MOE receiver to completely reject the MAI in the high-SNR region. Let us focus attention on the performance comparison between the L-MOE and WL-MOE receivers. As a first result, it is shown in Appendix B that, if $\Phi$ (or, equivalently, $\Psi$) is full-column rank, then

$$\lim_{\sigma^2_{\Psi} \to 0} \frac{\text{SINR}_{\text{WL-MOE}}}{\text{SINR}_{\text{L-MOE}}} = \frac{\| \phi \|^2 - \text{Re}[\phi^H \Phi] \{ \text{Re}[\Phi^H \Phi] \}^{-1} \text{Re}[\Phi^H \phi]}{\| \phi \|^2 - \phi^H \Phi \{ \Phi^H \Phi \}^{-1} \Phi^H \phi},$$

which, in addition to (22), evidences that, since

$$\lim_{\sigma^2_{\Psi} \to 0} \text{SINR}_{\text{L-MOE}} = +\infty \quad \text{when} \quad \text{rank}(\Phi) = J,$$

the WL-MOE receiver also suppresses the MAI exactly in the high-SNR regime, i.e., $\lim_{\sigma^2_{\Psi} \to 0} \text{SINR}_{\text{WL-MOE}} = +\infty$, if $H$ is full-column rank. If $\Phi$ is full-column rank (a condition that can hold only when the system is underloaded), then $H$ is full-column rank, too. However, the matrix $H$ can be full-column rank even when $N < J \leq 2N$, wherein $\Phi$ is structurally rank-deficient; in this overloaded environment, it results that

$$\lim_{\sigma^2_{\Psi} \to 0} \frac{\text{SINR}_{\text{WL-MOE}}}{\text{SINR}_{\text{L-MOE}}} = +\infty, \quad \forall j \in \{1, 2, \ldots, J\}.$$  

In other words, provided that $\text{rank}(H) = J$, the performance gap between the WL-MOE and L-MOE receivers becomes arbitrarily large for vanishingly small noise, when $N < J \leq 2N$. This fact strongly motivates us to provide conditions assuring that $H$ be full-column rank in overloaded scenarios. To this aim, we provide the following Theorem, by focusing attention directly on the downlink scenario in an effort to give simple and insightful conditions.

**Theorem 1**: When $N < J \leq 2N$, the code matrix can be decomposed as $C = C_{\text{left}} [I_N \quad \Pi]$, where

$$C_{\text{left}} \triangleq \left[ e_1, e_2, \ldots, e_N \right] \in \mathbb{C}^{N \times N} \text{is nonsingular and} \quad \Pi \in \mathbb{C}^{N \times (J-N)} \text{is a tall matrix. In this overloaded scenario, under the assumption that} \Psi \text{exhibits the form given by (24), the matrix} \quad \Pi = \Theta_1^H \Theta_2^H \text{is full-column rank iff} \quad \Pi - (\Theta_1^H \Theta_2^H) \text{is full-column rank, where} \quad \Theta_1 \triangleq \text{diag}(e^{j\theta_1}, e^{j\theta_2}, \ldots, e^{j\theta_N}) \in \mathbb{C}^{N \times N} \text{and} \quad \Theta_2 \triangleq \text{diag}(e^{j\theta_{N+1}}, e^{j\theta_{N+2}}, \ldots, e^{j\theta_J}) \in \mathbb{C}^{(J-N) \times (J-N)}.$$

**Proof**: See Appendix C.

Theorem 1 deserves some interesting comments, aimed at clarifying in particular the role of the preceding phases in (1), which are at the designer’s disposal. First of all, it is apparent that the full-column rank property of $H$ does not depend
on the channel impulse response\textsuperscript{10}, but depends on both the spreading codes of all the active users and their precoding phases $\theta_1, \theta_2, \ldots, \theta_J$. To this respect, it is interesting to investigate how such phases influence the full-column rank property of $H$ in overloaded systems, focusing attention to the case wherein Walsh-Hadamard (WH) spreading codes are employed. To do this, without loss of generality, assume that $c_{N+j} = c_j$, for $j \in \{1, 2, \ldots, J - N\}$, and let $C_{\text{left}}$ denote the common Hadamard matrix of order $N$. In this case, it is easily verified that $\Pi = [e_1, e_2, \ldots, e_{J-N}]$, with $e_j$ denoting the $j$th column of $I_N$. Thus, if WH spreading vectors are used, the matrix $\Pi$ is real-valued (i.e., $\Pi = \Pi^*$) and, moreover, one has $(\Theta_1^2)^* \Pi = \Pi (\Theta_1^2)^*$, where $\Theta_{1,\text{red}} \triangleq \text{diag}(e^{i\theta_1}, e^{i\theta_2}, \ldots, e^{i\theta_{J-N}}) \in \mathbb{C}^{(J-N) \times (J-N)}$. In light of these observations, by additionally remembering that $\Pi$ is full-column rank, it follows that $\text{rank}(\Pi^* (\Theta_1^2)^* \Pi \Theta_2^2) = \text{rank}(\Pi^* (I_{J-N} - (\Theta_1^2)^* \Theta_2^2)) = \text{rank}(I_{J-N} - (\Theta_1^2)^* \Theta_2^2)$. Since the matrix $I_{J-N} - (\Theta_1^2)^* \Theta_2^2$ is diagonal with diagonal entries $1 - e^{i2(\theta_{N+j} - \theta_j)}$, $\forall j \in \{1, 2, \ldots, J - N\}$, by virtue of Theorem 1, it can be stated that, when $N < J \leq 2N$, the augmented matrix $H$ is full-column rank iff

$$\theta_{N+j} - \theta_j \neq h\pi, \ \forall j \in \{1, 2, \ldots, J - N\} \text{ and } h \in \mathbb{Z}. \ (30)$$

As an immediate implication of (30), it is worth pointing out that, if no precoding is performed at the transmitter, i.e., $\theta_1 = \theta_2 = \cdots = \theta_J$, and common WH spreading codes are employed, the WL-MOE receiver is unable to achieve perfect MAI suppression in overloaded systems, even in the absence of noise. Henceforth, in order to allow WL-MUD to successfully work in an overloaded downlink, while employing WH spreading sequences, incorporation of precoding phases is crucial. This is the reason that motivated us to introduce the phases $\theta_1, \theta_2, \ldots, \theta_J$ in (1). It is worthwhile to observe that condition (30) does not uniquely specify the precoding phases and, thus, different choices can be pursued. To corroborate the previous considerations, we provide a numerical example.

\textbf{Example 1:} Consider a DS-CDMA downlink with $\alpha_1 = \alpha_2 = \cdots = \alpha_J = 1$ and processing gain $N = 16$, and without loss of generality, assume that the desired user is the first one (i.e., $j = 1$). The SNR, which is defined as $1/\sigma^2_h$, is set to 15 dB, and the signatures are generated according to (24). The system uses unit-norm WH vector codes and operates over a channel of order $L_h = 5$, whose taps $g(0), g(1), \ldots, g(5)$ are modeled as i.i.d. complex proper zero-mean Gaussian random variables, normalized so that $\|\psi_j\|^2 = 1, \forall j \in \{1, 2, \ldots, J\}$. Fig. 2 reports the ideal SINR performance of the WL-MOE receiver as a function of the number of users $J$, ranging from an underloaded ($1 < J \leq N$) system to an overloaded ($N < J \leq 2N$) one. Specifically, we report $\text{SINR}_{-\text{WL-MOE}}$ [see (20)] in two different situations: in the former one, there is no precoding at the transmitter, i.e., $\theta_1 = \theta_2 = \cdots = \theta_J = 0$ (referred to as “without precoding”); in the latter one, we use a precoding strategy (30), by setting $\theta_1 = \theta_2 = \cdots = \theta_N = 0$ and $\theta_{N+1} = \theta_{N+2} = \cdots = \theta_{2N} = \pi/4$ (referred to as “with precoding”). The results of Fig. 2 are obtained by carrying out $10^4$ independent Monte Carlo trials, with each run using only a different channel realization. It can be observed that, if WH spreading sequences are employed and condition (30) is not accounted for, the WL-MOE receiver does not work at all, when the system becomes overloaded. In contrast, the proposed precoding strategy allows the WL-MOE receiver to achieve a satisfactory performance even when $N < J \leq 2N$.

\section{Finite-Sample Performances of the L-MOE and WL-MOE Receivers}

In this section, two different data-estimated versions of the L-MOE and WL-MOE receivers are introduced, i.e., the WL receiver employing sample-matrix inversion (WL-SMI) and the WL receiver employing subspace decomposition (WL-SUB), and their finite-sample performance are evaluated by adopting a first-order perturbative approach [31], [32].

Starting from $K$ samples of the received vector $r(k)$, the WL-SMI receiver is obtained by replacing $R_{zz}$ in (18) with its sample estimate

$$\hat{R}_{zz} = \frac{1}{K} \sum_{k=1}^{K} z(k) z^H(k), \quad (31)$$

thus obtaining

$$f_{j,\text{WL-SMI}} \triangleq (h_j^H \hat{R}_{zz}^{-1} h_j)^{-1} \hat{R}_{zz}^{-1} h_j. \quad (32)$$

Note that, in the sequel, we assume that the desired channel impulse response (and thus $h_j$) is exactly known at the receiver. The WL-SUB receiver resorts to the EVD of $R_{zz}$ to mitigate the performance degradation due to finite-sample-size effects. To this end, it is required\textsuperscript{11} that the

\textsuperscript{10}In the uplink scenario, the full-column rank property of $H$ and, thus, the performance of the WL-MOE receiver, depends not only on the precoding phases, but also on the channel impulse responses of all the active users.

\textsuperscript{11}As a matter of fact, this assumption is not required for the WL-SMI receiver and it is necessary only for the WL-SUB one. However, since the WL-MOE receiver is not able to ensure perfect MAI suppression, for each user, when $H$ is rank-deficient, we maintain the assumption $\text{rank}(H) = J$ for both the two data-estimated WL receivers.
augmented matrix $H$ is full-column rank (an issue that has been discussed in Section IV), which necessarily requires that $J \leq 2N$. Under this assumption, accounting for (12) and recalling that $R_{dd} = \sigma_d^2 I_{2N}$, the EVD of $R_{zz}$ is given by $R_{zz} = U_q \Lambda_q U_q^H + \sigma_n^2 U_n U_n^H$, where $U_q \in \mathbb{C}^{2N \times J}$ collects the eigenvectors associated with the $J$ largest (signal- and noise-dependent) eigenvalues $\lambda_1, \lambda_2, \ldots, \lambda_J$ of $R_{zz}$ (arranged in decreasing order), whose columns span the signal subspace, i.e., the column space $\mathcal{R}(H)$ of $H$, while $U_n \in \mathbb{C}^{2N \times (2N-J)}$ collects the eigenvectors associated with the eigenvalue $\sigma_n^2$, whose columns span the noise subspace, i.e., the orthogonal complement $\mathcal{R}^\perp(H)$ in $\mathbb{C}^{2N}$ of the signal subspace and, finally, $\Lambda_q = \text{diag}(\lambda_1, \lambda_2, \ldots, \lambda_J)$. By substituting the EVD of $R_{zz}$ in (18) and exploiting the orthogonality between signal and noise subspaces, one obtains
\[
\hat{f}_{j,\text{WL-MOE}} = (h_j^H U_q \Lambda_q^{-1} U_q^H h_j)^{-1} U_q \Lambda_q^{-1} U_q^H h_j.
\] Since in practice the EVD is performed on $\hat{R}_{zz}$ given by (31), by denoting the sample matrices corresponding to $U_q$ and $\Lambda_q$, with $\hat{U}_q$ and $\hat{\Lambda}_q$, respectively, we have:
\[
\hat{f}_{j,\text{WL-SUB}} = (\hat{h}_j^H \hat{U}_q \hat{\Lambda}_q^{-1} \hat{U}_q^H h_j)^{-1} \hat{U}_q \hat{\Lambda}_q^{-1} \hat{U}_q^H h_j.
\] It is worth noting that the weight vector $f_{j,\text{WL-SUB}}$ is not equal to $f_{j,\text{WL-MOE}}$, since $U_q^H h_j \neq 0$ for $j \neq \text{index of receivers}$ due to the finite-sample-size effects. This implies that the two receivers WL-SMI and WL-SUB might exhibit different SINR performances. To carry out the performance analysis for WL-SMI and WL-SUB in an unified framework, let us denote with $\hat{f}_j$ any data-estimated WL-MOE receiver, i.e., $\hat{f}_j = f_{j,\text{WL-SMI}}$ or $\hat{f}_j = f_{j,\text{WL-SUB}}$, and set $\hat{f}_j = f_{j,\text{WL-MOE}}$ for simplicity, where $f_{j,\text{WL-MOE}}$ is given by (18) or (33). Adopting a perturbation perspective, the vector $\hat{f}_j$ can be expressed as
\[
\hat{f}_j = f_j + \delta f_j,
\] where $\delta f_j$ is a small (i.e., $||\delta f_j|| \ll 1$) zero-mean perturbation term. Since any data-estimated version of the WL-MOE receiver must satisfy the constraint $h_j^H h_j = 1$, it results that $\delta f_j^H h_j = 0$, thus the SINR (15) for the data-estimated receivers can be written as
\[
\text{SINR}(\hat{f}_j) = \frac{1}{E_{f_j,q_j}\{\text{Re}\{\delta f_j^H q_j(k)\}\}},
\] where the symbol $E_{f_j,q_j}\{\cdot\}$ denotes joint average w.r.t to $\hat{f}_j$ and $q_j(k)$ of the quantity in brackets. A simplifying and reasonable assumption [33] is that $\hat{f}_j$ is independent from $q_j(k)$. In this case, by accounting for the CS property of $f_j$, substituting (35) into (36), performing the average w.r.t to $q_j(k)$, and recalling that, due to assumptions (a1) and (a2), the vector $q_j(k)$ is zero-mean, one has:
\[
\text{SINR}(\hat{f}_j) = \frac{1}{f_j^H R_{q_j,q_j} f_j + E_{\delta f_j}\{\delta f_j^H R_{q_j,q_j} \delta f_j\}},
\] where only the average w.r.t to $\delta f_j$ must be evaluated. To perform this calculation, we need explicit expressions for the perturbation $\delta f_j$ of the WL-SMI and WL-SUB receivers, which are provided by the following Lemma.

**Lemma 2**: Assume that $H$ is full-column rank and let $\hat{R}_{zz}$ be estimated by (31). The first-order perturbation term of the WL-SMI and WL-SUB receivers can be expressed as
\[
\delta f_j = -\Gamma_{j,\text{WL}} \hat{r}_{q_j,b_j},
\] where $\hat{r}_{q_j,b_j} \triangleq \frac{1}{K} \sum_{k=0}^{K-1} q_j(k) b_j(k)$ is the sample estimate of the cross-correlation between the disturbance vector $q_j(k)$ and the desired symbol $b_j(k)$, and
\[
\Gamma_{j,\text{WL}} = \begin{cases}
P_{j,\text{WL}} R_{q_j,q_j}^{-1} & \text{(WL-SMI)} \\
(P_{j,\text{WL}} R_{q_j,q_j}^{-1} - \gamma_{j,\text{WL}} U_n U_n^H) & \text{(WL-SUB)}
\end{cases}
\] with $P_{j,\text{WL}} \triangleq I_{2N} - (h_j^H R_{q_j,q_j} h_j)^{-1} R_{q_j,q_j} h_j h_j^H \in \mathbb{C}^{N \times N}$ denoting an oblique projection matrix [33], and $\gamma_{j,\text{WL}} \triangleq \sigma_n^{-2} + (h_j^H R_{\text{zz}} h_j)^{-1} h_j^H U_n \Omega_{\text{WL}} U_n^H R_{\text{zz}} h_j$, where $\Omega_{\text{WL}} \triangleq \Lambda_q - \sigma_n^2 I_J \in \mathbb{R}^{J \times J}$.

**Proof**: See Appendix D.

It should be observed that Lemma 2 provides a compact characterization of the perturbation terms, obtained under the simplifying assumption [33] that the predominant error in estimating $R_{zz}$ is due to $r_{q_j,b_j}$ (see Appendix D for details). This approximation will allow us to obtain simple yet accurate results, which will be validated in Section VI. Accounting for Lemma 2, the average in (37) can be expressed as (we drop the subscript $\delta f_j$ in $E_{\delta f_j}\{\cdot\}$ for notational simplicity)
\[
E\{\delta f_j^H R_{q_j,q_j} \delta f_j\} = E\{\delta f_j^H \Gamma_{j,\text{WL}} \delta f_j\} = \text{trace}\{\Gamma_{j,\text{WL}} R_{q_j,q_j} \Gamma_{j,\text{WL}} \hat{r}_{q_j,b_j} \hat{r}_{q_j,b_j}^H\},
\] where, by accounting for assumptions (a1) and (a2), one has:
\[
E\{\hat{r}_{q_j,b_j} \hat{r}_{q_j,b_j}^H\} = \frac{1}{K^2} \sum_{k,h=1}^{K} E\{q_j(k) b_j(k) b_j(h) q_j^H(h)\} = \frac{1}{K^2} \sum_{k,h=1}^{K} E\{q_j(k) q_j^H(h)\} E\{b_j(k) b_j(h)\} = \frac{1}{K^2} \sum_{k,h=1}^{K} E\{q_j(k) q_j^H(h)\} \delta_{k-h}.
\]
By substituting (41) in (40), the result back in (37), and recalling that $\Gamma_{j,\text{WL}} R_{q_j,q_j} f_j = \text{SINR}_{j,\text{WL-MOE}}$, where $\text{SINR}_{j,\text{WL-MOE}}$ is given by (20), we get
\[
\text{SINR}(\hat{f}_j) = \frac{\text{SINR}_{j,\text{WL-MOE}}}{1 + \text{trace}(\Gamma_{j,\text{WL}} R_{q_j,q_j} \Gamma_{j,\text{WL}} R_{q_j,q_j} \delta f_j) / \text{SINR}_{j,\text{WL-MOE}}}.
\] The final result is obtained by evaluating the trace(·) term in (42), on the basis of the $\Gamma_{j,\text{WL}}$ expressions given in Lemma 2. To do this, it is convenient to consider the SMI and SUB cases separately. With reference to the WL-SMI receiver, since...
\( \Gamma_{j,\text{WL}} = P_{j,\text{WL}} R_{q,q}^{-1} \), by using the properties of the trace operator, after some algebraic manipulations, one obtains:

\[
\text{trace}(\Gamma_{j,\text{WL}}^2) = \frac{\text{trace}(P_{j,\text{WL}} P_{j,\text{WL}}^H)}{K} = \frac{2N-1}{K},
\]

which can be substituted in (42), thus leading to

\[
\text{SINR}_{j,\text{WL-SMI}} \triangleq \text{SINR}(f_{j,\text{WL-SMI}}) = \frac{\text{SINR}_{j,\text{WL-MOE}}}{1 + \frac{2(N-1)}{K} \text{SINR}_{j,\text{WL-MOE}}}. \tag{44}
\]

As regards the WL-SUB receiver, since \( \Gamma_{j,\text{WL}} = P_{j,\text{WL}} R_{q,q}^{-1} - \gamma_j U_n U_n^H \), by using again the properties of the trace operator and observing that \( U_n^H h_j = 0_{2N-J} \), after some algebra, one has:

\[
\text{trace}(\Gamma_{j,\text{WL}}^2) = 2N-1 - (\gamma_{j,\text{WL}} \sigma_e^2 + \gamma_j \sigma_{\text{WL}}^2 - \gamma_{j,\text{WL}} \sigma_e^2) (2N-J)
\]

\[
= (J-1) + (2(N-J) - 1) \gamma_{j,\text{WL}} \sigma_e^2. \tag{45}
\]

After substituting (45) into (42), one gets:

\[
\text{SINR}_{j,\text{WL-SUB}} \triangleq \text{SINR}(f_{j,\text{WL-SUB}}) = \frac{\text{SINR}_{j,\text{WL-MOE}}}{1 + \frac{(J-1)+(2N-J) - 1}{K} \gamma_{j,\text{WL}} \sigma_e^2}. \tag{46}
\]

The expression (46) for the WL-SUB receiver can be further simplified by observing that, for \( \sigma_e^2 \to 0 \), one has \( \gamma_{j,\text{WL}} \sigma_e^2 \to 1 \), hence the trace in (45) reduces to \( J - 1 \). By accounting for this observation, for moderate-to-high values of the SNR, eq. (46) can be approximatively written as

\[
\text{SINR}_{j,\text{WL-SUB}} = \frac{\text{SINR}_{j,\text{WL-MOE}}}{1 + \frac{J-1}{K} \text{SINR}_{j,\text{WL-MOE}}}. \tag{47}
\]

It is worth noting that, despite of the apparent similarity between (44)-(47) and the SINR formulas reported in [18, eqs. (14) and (25)], our results are not directly comparable with those of [18]. Indeed, the results of [18] report the SINR performances of the LMS-based adaptive implementation of the WL-MMSE and WL-MOE receivers only for \( K \to +\infty \) (steady-state performances); in this latter case the performance penalty paid by the WL-MUD receivers with respect to their ideal counterparts is exclusively due to gradient-noise effects.

The finite-sample performance analysis of the L-MOE receivers is now in order. Specifically, the L-MOE receiver is given by

\[
g_{j,\text{L-MOE}} \triangleq (\phi_j H \bar{R}_{\text{tr}}^{-1} \phi_j)^{-1} \bar{R}_{\text{tr}}^{-1} \phi_j \tag{48}
\]

and is obtained by replacing \( R_{\text{tr}} \) in (9) with its sample estimate

\[
\hat{R}_{\text{tr}} = \frac{1}{K} \sum_{k=1}^K r(k) r^H(k). \tag{49}
\]

Instead, the L-SUB receiver is defined as

\[
g_{j,\text{L-SUB}} \triangleq (\phi_j H \tilde{V}_s \Sigma_s^{-1} \tilde{V}_s^H \phi_j)^{-1} \tilde{V}_s \Sigma_s^{-1} \tilde{V}_s^H \phi_j. \tag{50}
\]

where the columns of \( \tilde{V}_s \in \mathbb{C}^{N \times J} \) coincide with the eigenvectors corresponding to the \( J \) largest eigenvalues \( \hat{\mu}_1, \hat{\mu}_2, \ldots, \hat{\mu}_J \) (arranged in decreasing order) of the sample autocorrelation matrix \( \bar{R}_{\text{tr}} \), and \( \Sigma_s \triangleq \text{diag}(\hat{\mu}_1, \hat{\mu}_2, \ldots, \hat{\mu}_J) \in \mathbb{R}^{J \times J} \). In order to carry out the performance analysis of the L-SMI and L-SUB receivers, it should be stressed that, since the relevant SINR is the one after the Re[\( \gamma \)] part, one cannot simply apply results available in the literature (e.g., [32]), since they refer to the SINR evaluated before the Re[\( \gamma \)] part. However, to avoid to be overwhelmed by mathematical derivations, we will report only the final results and defer to Appendix E for their proofs. Under the assumption that the matrix \( \Phi \) is full-column rank (see Section IV for a brief discussion regarding this issue), which necessarily requires that \( J \leq N \), it turns out that, in the high-SNR regime, the output SINR (15) of the L-SMI and L-SUB receivers can be approximatively written as

\[
\text{SINR}_{j,\text{L-SMI}} \triangleq \text{SINR}(g_{j,\text{L-SMI}}) = \frac{\text{SINR}_{j,\text{L-MOE}}}{1 + \frac{J-1}{K} \text{SINR}_{j,\text{L-MOE}}}. \tag{54}
\]

Equations (44), (47), (51) and (52) allow one to easily compare the finite-sample performances of WL-MOE and L-MOE receivers. By comparing (47) and (52) for the subspace receivers, since \( \text{SINR}_{j,\text{WL-MOE}} \geq \text{SINR}_{j,\text{L-MOE}} \) by (22), it turns out that \( \text{SINR}_{j,\text{WL-SUB}} \geq \text{SINR}_{j,\text{L-SUB}} \) for any value of \( K \) and for \( J \leq N \). A similar conclusion does not hold for the SMI receivers. Indeed, it can be easily proven that, for \( J < N \) it results that \( \text{SINR}_{j,\text{WL-SMI}} \geq \text{SINR}_{j,\text{L-SMI}} \) only for \( K \geq K_{\text{min}} \), where

\[
K_{\text{min}} \triangleq \frac{3N-J}{2 (\text{SINR}_{j,\text{L-MOE}} - \text{SINR}_{j,\text{L-MOE}}^1)} > 0 \tag{53}
\]

is a threshold sample-size. In other words, it can be inferred that, in underloaded scenarios, the WL-SMI receiver assures the expected performance advantage over the L-SMI one only if a sufficient number of samples are processed. This loss of performance is due to the increase of the dimension of the autocorrelation matrix to be estimated from \( N \to 2N \), which entails a diminished estimation accuracy, requiring hence a larger number of data samples for achieving a satisfactory performance, without resorting to subspace concepts.

Another interesting conclusion that can be drawn from (44) through (52) is that all finite-sample receivers exhibit a SINR saturation effect, i.e., a bit-error-rate (BER) floor, for vanishingly small noise. Indeed, when \( \sigma_e^2 \to 0 \) and \( H \) is full-column rank (\( J \leq 2N \)), it has been shown in Subsection IV-A that \( \text{SINR}_{j,\text{WL-MOE}} \) grows without bound. Thus, accounting for (44) and (47), we get:

\[
\lim_{\sigma_e^2 \to 0} \text{SINR}_{j,\text{WL-SMI}} = \frac{K}{2(N-1)}; \tag{54}
\]

which show that, in the high-SNR regime, the performance of the WL-SMI receiver does not depend on the number of
consider both SMI- and SUB-based data-estimated versions to assumptions (a1) and the additive noise vector \( v \). In addition, the symbol vector is worthwhile to note that in this latter case WL processing inherits an increased computational complexity compared with more sophisticated subspace-based methods based on EVD. It is evident that the advantages of WL receivers could be lost by employing simple estimation methods such as the SMI, whereas it is mandatory to resort to more sophisticated subspace-based methods based on EVD. It is worthwhile to note that in this latter case WL processing incurs an increased computational complexity compared with linear one, due to the increased dimension of the augmented correlation matrix, with respect to the conventional data autocorrelation matrix.

**A. Numerical examples**

Herein, we present the results of Monte Carlo computer simulations and compare them with the analytical results derived in Section V [see (44), (47), (51) and (52)]. Specifically, in all the examples, the same simulation setting considered in Example 1 is adopted (downlink scenario and \( N = 16 \)), with \( \theta_1 = \theta_2 = \cdots = \theta_N = 0 \) and \( \theta_{N+1} = \theta_{N+2} = \cdots = \theta_{2N} = \pi/4 \) (we recall that this precoding strategy assures the full-column rank property of the augmented matrix \( H \) in overloaded scenarios). In addition, the symbol vector \( b(k) \) and the additive noise vector \( v(k) \) are generated according to assumptions (a1) and (a2). For the sake of comparison, we consider both SMI- and SUB-based data-estimated versions of the L-MOE and WL-MOE receivers (wherein the channel impulse response is assumed to be exactly known), as well as their exact counterparts (wherein, besides the channel impulse response, perfect knowledge of the autocorrelation matrices \( R_{rr} \) and \( R_{zz} \) is assumed). Finally, as performance measure, in addition to the SINR given by (15) and averaged over \( 10^4 \) Monte Carlo runs, we resort to the average BER at the output of the considered receivers. More specifically, after estimating the receiver weight vectors on the basis of the given data record \( K \), for each run (wherein, besides the channel impulse response, independent sets of noise and data sequences are randomly generated), an independent record of \( K_{\text{ber}} = 10^3 \) symbols is considered to evaluate the BER.

**Example 2:** In this example, we evaluate both the (average) SINR and BER performances of the considered receivers as a function of the SNR. The number of users is set equal to \( J = 10 \) (underloaded system) and the sample size is kept fixed to \( K = 500 \) symbols. Let us first consider the SINR performances, which are reported in Fig. 3. It can be seen that the analytical expressions (44), (47), (51) and (52) for the data-estimated linear and WL receivers agree very well with their corresponding simulation results, for all values of the SNR. In particular, in this underloaded scenario, while the L-SUB and WL-SUB receivers perform comparably, the WL-SMI receiver pays a significant performance loss with respect to the L-SMI one. Indeed, in the high-SNR region, the difference between the saturation values of SINR\(_{1,L-SMI}\) and \( \text{SINR}_{1,WL-SUB} \) is about 4 dB, which is in good agreement with (54) and (55). The unsatisfactory performance of the WL-SMI receiver is also apparent from Fig. 4, which depicts the BER curves of the data-estimated receivers under comparison. It is evident that the curves of the WL-SUB, L-SUB and L-SMI receivers go down very quickly as the SNR increases, thus assuring a huge performance gain with respect to the WL-SMI receiver, which instead exhibits a marked BER floor.

**Example 3:** Fig. 5 reports the SINR as a function of the number of users \( J \). The SNR is set equal to 15 dB and \( K = 500 \) symbols are considered. Besides confirming the very good agreement between analytical and experimental
results for all the data-estimated receivers, results of Fig. 5 show that the performances of all the linear receivers worsen very quickly when the system tends to be overloaded, i.e., \( J \) approaches \( N = 16 \). Beyond this value, the WL receivers assure a significant performance gain with respect to their corresponding linear counterparts, by approaching the curve of the L-SMI receiver. It is worth observing that the WL-SUB receiver outperforms the L-SUB one, for all the considered values of \( K \).

Example 4: In this last experiment, we report the SINR performances of the considered data-estimated receivers as a function of the sample size \( K \). The SNR is set equal to 15 dB and \( J = 14 \) users (underloaded system) are considered. It can be observed from Fig. 6 that the accuracy of the formulas (44), (47), (51) and (52) improves as \( K \) increases. Additionally, it is worth observing that the WL-SUB receiver outperforms the L-SUB one, for all the considered values of \( K \). In contrast, the WL-SMI receiver performs worse than its corresponding linear counterpart, by approaching the curve of the L-SMI receiver only when the sample size \( K \) is as large as 1500 symbols, which agrees very well with the value \( K_{\text{min}} = 1686 \) predicted by (53).

VI. CONCLUSIONS

We developed performance comparisons between ideal and data-estimated WL-MOE and L-MOE receivers. With reference to the ideal implementation, we investigated the relative performances of the WL-MOE and L-MOE receivers in the high-SNR regime. In this case, we provided a necessary and sufficient condition on the spreading codes, which allows the WL-MOE receiver to achieve perfect MAI suppression even in overloaded downlink configurations. As regards the data-estimated versions of the WL-MOE and L-MOE receivers, we derived easily interpretable formulas, which allow one to obtain clear insights about the effects of different parameters on performances. In a nutshell, compared with the L-MOE one, the performance of the WL-MOE receiver turns out to be more sensitive to finite-sample-size effects, and the performance gains predicted by the theory can be achieved in practice only by resorting to the more sophisticated subspace-based implementation. Finally, in this paper the channel impulse response was assumed to be exactly known at the receiving side and the precoding phases are not optimized. The assessment of the effects of channel-estimation errors and the optimization of the precoding phases are the topic of our current research and will be addressed in a forthcoming paper.

APPENDIX

Proofs

A. Proof of Lemma 1

Any vector \( f_j \in \mathbb{C}^{2N} \) can be uniquely decomposed as \( f_j = f_{j,s} + f_{j,a} \), where we defined the symmetric part \( f_{j,s} \in S \triangleq \{ f = [f_1^T, f_2^T]^T \in \mathbb{C}^{2N} | f_1 = f_2^T \in \mathbb{C}^N \} \) and the antisymmetric part \( f_{j,a} \in A \triangleq \{ f = [f_1^T, f_2^T]^T \in \mathbb{C}^{2N} | f_1 = -f_2^T \in \mathbb{C}^N \} \). Since both \( h \) and \( q(k) \) (in 14) are symmetric, i.e., they belong to \( S \), one has \( \text{Re}[f_j^H h_j] = f_j^H h_j \) and \( \text{Re}[f_j^H h_j] = f_j^H h_j \) in (15), that is, the SINR (15) is not affected by the antisymmetric part \( f_{j,a} \). Hence, the weight vector \( f_{j,\text{max-SINR}} \) maximizing \( \text{SINR}(f_j) \) given by (15) can equivalently be obtained by maximizing the following constrained cost function:

\[
\text{SINR}'(f_j) \triangleq \frac{|f_j^H h_j|^2}{E[|f_j^H \tilde{q}(k)|^2]} = \frac{|f_j^H h_j|^2}{f_j^H R_{q,k} f_j},
\]

subject to \( f_j \in S \). (56)

Note that in general \( \text{SINR}(f_j) \neq \text{SINR}'(f_j) \), but they coincide for \( f_j \in S \). The unconstrained maximization of \( \text{SINR}'(f_j) \) leads [29] to the solution \( f_j,\text{max-SINR} = \gamma_j R_{q,k}^{-1} h_j \), with \( \gamma_j \in \mathbb{C} - \{0\} \). At this point, we have to impose that \( f_j,\text{max-SINR} \) satisfies the constraint \( f_j,\text{max-SINR} \in S \). To this respect, it can be verified that \( R_{q,k}^{-1} h_j \in S \), hence, fulfillment of the constraint is ensured by imposing that \( \gamma_j \) be real, i.e., \( \gamma_j = \bar{\gamma}_j \). In conclusion, we can state that the general expression of the weight vector \( f_j,\text{max-SINR} \) maximizing \( \text{SINR}(f_j) \) is given by

![Fig. 5. Average SINR versus number of users (K = 500 symbols and SNR = 15 dB).](image1)

![Fig. 6. Average SINR versus sample size K (J = 14 users and SNR = 15 dB).](image2)
\[ f_{j,\max} \text{SINR} = \xi_j R_{\text{q}_j}^{-1} \text{h}_j, \quad \text{with } \xi_j \triangleq \text{Re}[\gamma_j] \in \mathbb{R} - \{0\}. \]

The corresponding maximum value of SINR\(f_j\) turns out to be

\[ \text{SINR}(f_{j,\max}\text{SINR}) = h_j^H R_{\text{q}_j}^{-1} \text{h}_j. \]

B. Relationships between SINR\(_{j,\text{max}}\), SINC_{j,\text{L-MOE}} and SINC_{j,\text{WL-MOE}} in the high-SNR regime

First of all, we derive the expression of SINC\(_{j,\text{max}}\) [see (11)] in terms of \(\sigma_v^2\). Under assumptions (a1)-(a2), one has \(R_{p_j p_j} = \Phi_j \Phi_j^H + \sigma_v^2 I_N\). Hence, by resorting to the EVD of \(\Phi_j, \Phi_j^H\), one obtains \(R_{p_j p_j} = \Phi_j \Lambda_j \Phi_j^H + \sigma_v^2 I_N\), where \(\Phi_j, \Lambda_j \in \mathbb{C}^{N \times J}\) collects the eigenvectors associated with the \(r_j\) nonnull eigenvalues \(\mu_{j,1}, \mu_{j,2}, \ldots, \mu_{j,r_j}\) of \(\Phi_j \Phi_j^H\) (arranged in decreasing order), with \(r_j := \text{rank}(\Phi_j) \leq \min\{N, J - 1\}\) and \(\Lambda_j = \text{diag}(\mu_{j,1}, \mu_{j,2}, \ldots, \mu_{j,r_j}) \in \mathbb{R}^{J \times J}\). Relying on this decomposition and reasoning as in [34], the following series expansion of SINC\(_{j,\text{max}}\) holds:

\[
\text{SINC}_{j,\text{max}} = \frac{\sigma_v^2}{\sigma_v^2} \Phi_j^H U_{j,s} V_{j,s}^H \Phi_j + \Phi_j^H U_{j,s} \Lambda_{j,s}^{-1} V_{j,s}^H \Phi_j + o(\sigma_v^2), \tag{57}
\]

where \(V_{j,s}, U_{j,s} \in \mathbb{C}^{N \times (N-r_j)}\) collects the eigenvectors of \(\Phi_j \Phi_j^H\) associated with its \(N - r_j\) null eigenvalues. Eq. (57) shows that, as \(\sigma_v^2 \to 0\), \(\text{SINC}_{j,\text{max}} \to +\infty\) if and only if \(\Phi_j \neq 0\), which implies that \(\Phi_j \notin \mathcal{N}(V_{j,n}^H) \equiv \mathcal{R}(\Phi_j^H)\). It is noteworthy that this condition holds, \(\forall j \in \{1, 2, \ldots, J\}\), iff the matrix \(\Phi \in \mathbb{C}^{N \times J}\) is full-column rank, i.e., rank(\(\Phi\)) = \(J\), which imposes that the number of users \(J\) must be smaller than or equal to the processing gain \(N\) (underloaded system). On the other hand, when \(\Phi_j \neq 0\), it results that \(\text{lim}_{\sigma_v^2 \to 0} \text{SINC}_{j,\text{max}} = \Phi_j^H U_{j,s} \Lambda_{j,s}^{-1} V_{j,s}^H \Phi_j\), which evidences that, as \(\sigma_v^2 \to 0\), \(\text{SINC}_{j,\text{max}}\) takes a finite value.

At this point, we are able to establish the relationship existing between SINC\(_{j,\text{max}}\) and SINC\(_{j,\text{L-MOE}}\) [see (11) and (21)], in the limiting case of vanishingly small noise. Preliminary, we observe that, under assumptions (a1)-(a2), one has \(R_{p_j p_j} = \Phi_j \Phi_j^H\). By substituting (9) in (21) and accounting for (11), after some algebraic manipulations, one obtains

\[
\text{lim}_{\sigma_v^2 \to 0} \frac{\text{SINC}_{j,\text{L-MOE}}}{\text{SINC}_{j,\text{max}}} = \frac{2}{1 + \text{lim}_{\sigma_v^2 \to 0} \text{Re}[\phi_j^H R_{p_j p_j} \Phi_j^H R_{p_j p_j}^{-1}] \phi_j^H]. \tag{58}
\]

By resorting to the limit formula for the Moore-Penrose inverse \cite{19}, it can be seen that \(\text{lim}_{\sigma_v^2 \to 0} R_{p_j p_j}^{-1} = \Phi_j^H \Phi_j\) and \(\text{lim}_{\sigma_v^2 \to 0} \Phi_j^H R_{p_j p_j} = \Phi_j^H\). Consequently, we get

\[
\text{lim}_{\sigma_v^2 \to 0} \text{Re}[\phi_j^H R_{p_j p_j} \Phi_j^H R_{p_j p_j}^{-1}] \phi_j^H] = \text{Re}[\phi_j^H \Phi_j^H \Phi_j] \phi_j^H, \tag{59}
\]

which can only assume finite values. Therefore, based on the previous discussion regarding the asymptotic expression of SINC\(_{j,\text{max}}\), by virtue of (57) and (58), we can conclude that, if \(\Phi\) is full-column rank, then

\[
\text{lim}_{\sigma_v^2 \to 0} \frac{\text{SINC}_{j,\text{L-MOE}}}{\text{SINC}_{j,\text{max}}} = 2, \quad \forall j \in \{1, 2, \ldots, J\}, \tag{59}
\]

which additionally implies that, as \(\sigma_v^2 \to 0\), SINC\(_{j,\text{L-MOE}} \to +\infty\), \(\forall j \in \{1, 2, \ldots, J\}\).

Let us now derive the expression of SINC\(_{j,\text{WL-MOE}}\) [see (20)] in terms of \(\sigma_v^2\). Under assumptions (a1)-(a2), one has \(R_{q_j q_j} = \mathbf{H}_j \mathbf{H}_j^H + \sigma_v^2 I_{2N}\). Reasoning as previously done for SINC\(_{j,\text{max}}\), we express SINC\(_{j,\text{WL-MOE}}\) explicitly in terms of \(\sigma_v^2\) as follows:

\[
\text{SINC}_{j,\text{WL-MOE}} = h_j^H \text{R}_{\text{q}_j,\text{q}_j}^{-1} \text{h}_j = h_j^H U_{j,s} U_{j,s}^H \text{h}_j + h_j^H A_{j,s}^{-1} \text{U}_{j,s}^H \text{h}_j + o(\sigma_v^2), \tag{60}
\]

where \(U_{j,s}, A_{j,s} \in \mathbb{C}^{2N \times J}\) collects the eigenvectors associated with the \(J\) nonnull eigenvalues \(\lambda_{j,1}, \lambda_{j,2}, \ldots, \lambda_{j,J}\) of \(\mathbf{H}_j \mathbf{H}_j^H\) (arranged in decreasing order), with \(\nu_j := \text{rank}(\mathbf{H}_j) \leq \min\{2 N, J - 1\}\) and \(A_{j,s} = \text{diag}(\lambda_{j,1}, \lambda_{j,2}, \ldots, \lambda_{j,J}) \in \mathbb{R}^{J \times J}\). This means that, in the absence of noise, the WL-MOE receiver is able to achieve perfect MAI suppression for each active user iff rank(\(\mathbf{H}_j\)) = \(J\). The matrix \(\mathbf{H}_j\) turns out to be full-column rank iff the null spaces of the matrices \(\Phi\) and \(\Phi^\perp\) intersect only trivially (see, e.g., [23]), that is, \(\mathcal{N}(\Phi) \cap \mathcal{N}(\Phi^\perp) = \{0_J\}\). If \(\Phi\) is full-column rank, which necessarily requires that \(J \leq N\) (underloaded system), this condition is trivially satisfied and, hence, the augmented matrix \(\mathbf{H}\) is full-column rank as well. However, the converse statement is not true, that is, \(\mathbf{H}\) may be full-column rank even when \(N < J \leq 2 N\) (overloaded system). To point out a first consequence of this result, let us focus attention on the case when \(N < J \leq 2 N\). In this overloaded scenario, the matrix \(\Phi\) cannot be full-column rank and, thus, it results that, as \(\sigma_v^2 \to 0\), SINC\(_{j,\text{WL-MOE}}\) takes on a finite value. In contrast, since \(\mathbf{H}\) can still be full-column rank in an overloaded system, relying on the results provided before, we can infer that, if \(\mathbf{H}\) is full-column rank, then

\[
\text{lim}_{\sigma_v^2 \to 0} \frac{\text{SINC}_{j,\text{WL-MOE}}}{\text{SINC}_{j,\text{max}}} = +\infty, \tag{61}
\]

\(\forall j \in \{1, 2, \ldots, J\}\), with \(N < J \leq 2 N\). Let us now consider an overloaded scenario \((J \leq N)\) and assume that \(\Phi\) is full-column rank. Since in this case the matrix \(\mathbf{H}\) is full-column rank, too, it follows that both SINC\(_{j,\text{L-MOE}}\) and SINC\(_{j,\text{WL-MOE}}\) diverge, in the limiting case of vanishingly small noise, and thus \(\text{lim}_{\sigma_v^2 \to 0} \frac{\text{SINC}_{j,\text{L-MOE}}}{\text{SINC}_{j,\text{max}}}\) assumes an indeterminate form. To overcome this mathematical difficulty, we preliminarily develop the relationship existing between SINC\(_{j,\text{WL-MOE}}\) and SINC\(_{j,\text{max}}\).
in the high-SNR regime, by resorting to the series expansions (57) and (60). So doing, we get:

$$\lim_{\sigma^2 \to 0} \frac{\text{SINR}_{J,\text{L-MOE}}}{\text{SINR}_{J,\text{max}}} = \frac{h_j^H U_{j,n} H_{j,n} h_j}{\phi_j^H V_{j,n} H_{j,n} \phi_j} = \frac{\|U_{j,n} h_j\|^2}{\|V_{j,n} \phi_j\|^2},$$

(62)

where, since both $\Phi$ and $H$ are full-column rank, it follows that $\|V_{j,n} \phi_j\| \neq 0$ and $\|U_{j,n} h_j\| \neq 0, \forall j \in \{1, 2, \ldots, J\}$. It is worth observing that $V_{j,n} V_{j,n}^H$ and $U_{j,n} U_{j,n}^H$ represent the orthogonal projections [19] onto the subspaces $\mathbb{R}^J(\Phi_j)$ and $\mathbb{R}^J(\Pi_j)$, respectively, which can be equivalently expressed [19] as $V_{j,n} V_{j,n}^H = I_N - \Phi_j (\Phi_j^H \Phi_j)^{-1} \Phi_j$ and $U_{j,n} U_{j,n}^H = I_N - \Pi_j (\Pi_j^H \Pi_j)^{-1} \Pi_j$. By substituting this two relations in (62), and remembering that $\Pi_j = [\Phi_j^T, \Phi_j^H]^T$ and $h_j = [\phi_j^T, \phi_j^H]^T$, after some algebraic manipulations, one has:

$$\lim_{\sigma^2 \to 0} \frac{\text{SINR}_{J,\text{L-MOE}}}{\text{SINR}_{J,\text{max}}} = 2 \frac{\|\phi_j\|^2 - \text{Re}[\phi_j^H [\text{Re}[\Phi_j^H \Phi_j]]^{-1} \text{Re}[\Phi_j^H \phi_j]]}{\|\phi_j\|^2 - \text{Re}[\phi_j^H [\Phi_j^H \Phi_j]^{-1} \Phi_j^H \phi_j]}.$$  

(63)

Therefore, if $\Phi$ is full-column rank, accounting for (59) and (63), we can state that:

$$\lim_{\sigma^2 \to 0} \frac{\text{SINR}_{J,\text{L-MOE}}}{\text{SINR}_{J,\text{max}}} = \lim_{\sigma^2 \to 0} \frac{\text{SINR}_{J,\text{L-MOE}}}{\text{SINR}_{J,\text{L-MOE}}} = \frac{\|\phi_j\|^2 - \text{Re}[\phi_j^H [\Phi_j^H \Phi_j]^{-1} \Phi_j^H \phi_j]}{\|\phi_j\|^2 - \text{Re}[\phi_j^H [\Phi_j^H \Phi_j]^{-1} \Phi_j^H \phi_j]}.$$  

(64)

D. Proof of Lemma 2

First, let us consider the SMI implementation of the WL-MOE receiver. By substituting (12) in (31), the sample autocorrelation matrix $R_{zz}$ of the augmented vector $z(k)$ can be expressed as

$$\hat{R}_{zz} = h_j h_j^H + h_j q_j b_j + \hat{r}_{q_j} b_j h_j^H + \hat{R}_{q_j} q_j,$$

(67)

where $\hat{r}_{q_j} b_j \triangleq \frac{1}{N} \sum_{k=0}^{N-1} q_j(k) b_j(k)$ and $\hat{R}_{q_j} q_j \triangleq \frac{1}{N} \sum_{k=0}^{N-1} q_j(k) q_j^H(k)$ represent sample estimates of the cross-correlation between the disturbance vector $q_j(k)$, and the desired symbol $b_j(k)$, and the autocorrelation matrix of $q_j(k)$, respectively. It is shown in [33] that, for moderate-to-high values of the sample size, i.e., $K \geq 6N$, the predominant cause of SINR degradation is represented by $\hat{r}_{q_j} b_j$ and, thus, replacing $\hat{R}_{q_j} q_j$ with $R_{q_j} q_j$ in (67) has a very marginal effect on the SINR. Therefore, remembering that $R_{zz} = h_j h_j^H + R_{q_j} q_j$, eq. (67) can be rewritten as $\hat{R}_{zz} \approx R_{zz} + h_j \hat{r}_{q_j} b_j + \hat{r}_{q_j} b_j h_j^H$. Its inverse admits [19] the following first-order approximation $R_{zz}^{-1} \approx R_{zz}^{-1} - R_{zz}^{-1} h_j \hat{r}_{q_j} b_j + \hat{r}_{q_j} b_j h_j^H R_{zz}^{-1}$, which can be substituted in (32), thus obtaining

$$f_{j,\text{L-SMI}} \approx f_{j,\text{L-MOE}} - P_{j,\text{L-MOE}} R_{zz}^{-1} \hat{r}_{j,q_j} b_j$$

(68)

with $P_{j,\text{L-MOE}} \triangleq I_{2N} - (h_j h_j^H + R_{q_j} q_j)^{-1} h_j h_j^H = I_{2N} - (h_j h_j^H + R_{q_j} q_j)^{-1} R_{q_j} q_j h_j^H = I_{2N} - (h_j h_j^H + R_{q_j} q_j)^{-1} R_{q_j} q_j h_j^H \in \mathbb{C}^{N \times N}$, where here and in the sequel the symbol $\approx$ denotes first-order equality, i.e., we neglect all the summands that tend to zero, as the sample size $K$ approaches infinity, faster than the norm of the corresponding perturbation term. It is easily verified that $P_{j,\text{L-MOE}} R_{zz}^{-1} = P_{j,\text{L-MOE}} R_{q_j} q_j$.

At this point, we focus attention on the subspace implementation of the WL-MOE receiver. Preliminarily, we observe that the EVD of $R_{zz}$ is given by

$$\tilde{R}_{zz} = \tilde{U}_s \tilde{A}_s \tilde{U}_s^H + \tilde{U}_n \tilde{A}_n \tilde{U}_n^H,$$

(69)

where $\tilde{U}_s, \tilde{A}_s, \tilde{U}_n$ and $\tilde{A}_n$ are sample estimates of $U_s, A_s, U_n$ and $A_n$, respectively. When $R_{zz}$ is estimated from the received data as in (31), for a sufficiently large sample size $K$, the estimate can be decomposed as $R_{zz} = R_{zz} + \delta R_{zz}$, where $\delta R_{zz}$ is a small additive perturbation (in the Frobenius norm...
sense). Consequently, the matrices $\hat{U}_s$ and $\hat{A}_s$ can be written as $\hat{U}_s = U_s + \delta U_s$ and $\hat{A}_s = A_s + \delta A_s$, where $\delta U_s$ and $\delta A_s$ represent the resulting perturbation in the estimated signal subspace, whose norm is of the order of $\|\delta R_{zz}\|$. It results [31], [32] that $\delta U_s \approx U_s \Omega_{WL}^{-1} \delta R_{xx} \Omega_{WL}^{-1}$, with $\Omega_{WL} \triangleq \Sigma_s - \sigma_s^2 I_J$, and $\delta A_s \approx U_s^H \delta R_{xx} U_s$. By substituting the above expressions of $\hat{U}_s$ and $\hat{A}_s$ in (34), and remembering that $P_{j, WL} R_{zz} = P_{j, WL} R_{q_j q_j}$, we get:

$$
\mathbf{f}_{j, WL-MOE} \approx \mathbf{f}_{j, WL-MOE} \triangleq \left( P_{j, WL} R_{q_j q_j} - \gamma_{j, WL} U_n U_n^H \right) \hat{r}_{q_j b_j},
$$

$$
\delta f_{j, WL-MOE} = \mathbf{f}_{j, WL-MOE} + \delta f_{j, WL-MOE},
$$

(70)

where $\gamma_{j, WL} \triangleq \sigma_s^{-2} + (h_j^H R_{xx}^{-1} h_j)^{-1} h_j^H U_s \Omega_{WL}^{-1} U_s^H R_{xx}^{-1} h_j$.

### E. Performance analysis of the L-SMI and L-SUB receivers

From a unified perspective, let us denote with $\hat{g}_j$ any data estimated L-MOE receiver, i.e. $\hat{g}_j = g_{j, L-SMI}$ or $\hat{g}_j = g_{j, L-SUB}$, and set $\hat{g}_j = g_{j, L-MOE}$ for simplicity. Adopting a perturbation approach, the vector $\hat{g}_j$ can be expressed as

$$
\hat{g}_j = g_j + \delta g_j,
$$

(71)

where $\delta g_j$ is a small (in the Frobenius norm sense) zero-mean additive perturbation. Since any data-estimated version of the L-MOE receiver must satisfy the constraint $g_j^H \phi_j = 1$, it results that $\delta g_j^H \phi_j = 0$. Thus, using the identity $\mathbf{E}[z^2] = \frac{1}{2} \{ \mathbf{E}[z^2] \}$, $\forall z \in \mathbb{C}$, the SINR (15) for data-estimated linear receivers becomes

$$
\text{SINR}(\hat{g}_j) = \frac{2}{E_{\hat{g}_j, p_j} \{ |g_j^H p_j(k)|^2 \} + E_{\hat{g}_j, p_j} \{ \text{Re}(g_j^H p_j(k))^2 \}}.
$$

(72)

Similarly to the WL case, we assume that $\hat{g}_j$ is independent from $p_j(k)$. In this case, by substituting (71) into (72), performing the average w.r.t. to $p_j(k)$, and recalling that, due to assumptions (a1) and (a2), the vector $p_j(k)$ is zero-mean, one has

$$
\text{SINR}(\hat{g}_j) = \frac{1}{2} \{ g_j^H R_{p_j p_j}^H g_j + E_{\delta g_j} [ g_j^H R_{p_j p_j} \delta g_j ] \}
\text{Re}(g_j^H R_{p_j p_j} g_j) + E_{\delta g_j} [ g_j^H R_{p_j p_j} \delta g_j ]
$$

(73)

The characterization of the perturbation term $\delta g_j$ is given by the following Lemma.

**Lemma 3:** Assume that $\Phi$ is full-column rank and let $R_{rr}$ be estimated by (49). Moreover, let $R_{rr} = V_s \Sigma_s V_s^H + \sigma_v^2 V_n V_n^H$, where $V_s \in \mathbb{C}^{N \times J}$ collects the eigenvectors associated with the $J$ largest eigenvalues $\mu_1, \mu_2, \ldots, \mu_J$ of $R_{rr}$ (arranged in decreasing order), while $V_n \in \mathbb{C}^{N \times (N-J)}$ collects the eigenvectors associated with the eigenvalue $\sigma_v^2$, and, finally, $\Sigma_s \triangleq \text{diag}(\mu_1, \mu_2, \ldots, \mu_J)$. The first-order perturbation term of the L-SMI and L-SUB receivers can be expressed as

$$
\delta g_j = -G_{j, L} \hat{r}_{p_j b_j},
$$

(74)

where $G_{j, L} \triangleq \frac{1}{K} \sum_{k=1}^{K} P_{j, L}^T \delta p_j(k) b_j(k)$ is the sample cross-correlation between the interference and the desired signal, and

$$
\Gamma_{j, L} = \begin{cases}
P_{j, L} R_{p_j p_j}^T, & (L-SMI) \\
\gamma_{j, L} V_n V_n^T, & (L-SUB)
\end{cases}
$$

(75)

with $P_{j, L} \triangleq I_N - (\phi_j^H R_{p_j p_j} \phi_j)^{-1} R_{p_j p_j} \phi_j \phi_j^H = I_N - (\phi_j^H R_{rr} \phi_j)^{-1} R_{rr} \phi_j \phi_j^H$ and $\gamma_{j, L} \triangleq \sigma_v^2 + (\phi_j^H R_{rr} \phi_j)^{-1} \phi_j^H V_s \Omega_{L}^{-1} V_s^H R_{rr} \phi_j$, where $\Omega_{L} \triangleq \Sigma_s - \sigma_v^2 I_J \in \mathbb{R}^{J \times J}$.

**Proof:** The proof is omitted since it is similar to that of Lemma 2.

By virtue of Lemma 3, we are now able to evaluate the averages in (73). Specifically, dropping the subscript $\delta g_j$ in $E_{\delta g_j}$ [for notational simplicity, we have:

$$
E[\delta g_j^H R_{p_j p_j} \delta g_j] = \text{trace}[\Gamma_{j, L}^T \delta g_j^H R_{p_j p_j} \delta g_j] = \frac{1}{K} \text{trace}(\Gamma_{j, L}^T R_{p_j p_j} \Gamma_{j, L} R_{p_j p_j}),
$$

(76)

$$
E[\delta g_j^H R_{p_j p_j}^* \delta g_j^*] = \text{trace}[\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*] = \frac{1}{K} \text{trace}(\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*).
$$

(77)

By substituting (76) and (77) into (73), and recalling that $\text{SINR}_{j, L-MOE}^{-1} = (g_j^H R_{p_j p_j} g_j + \text{Re}[(g_j^H R_{p_j p_j}^* g_j)])^2/2$, we get:

$$
\text{SINR}(\hat{g}_j) = \frac{1}{2} \{ \text{trace}(\Gamma_{j, L}^T R_{p_j p_j} \Gamma_{j, L} R_{p_j p_j}) \text{Re}[(\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*)] \text{SINR}_{j, L-MOE}^{-1} \}
$$

(78)

Along the same lines of the WL case, it can be shown that

$$
\text{trace}(\Gamma_{j, L}^T R_{p_j p_j} \Gamma_{j, L} R_{p_j p_j}) = \left \{ \begin{array}{ll}
N - 1, & (L-SMI) \\
J - 1 + (N - J)[1 - \gamma_{j, L} \sigma_v^2]^2, & (L-SUB)
\end{array} \right.
$$

(79)

On the other hand, evaluation of the term $\text{trace}(\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*)$ is more complicated and, for its calculation, it is convenient to consider the SMI and SUB cases separately. With reference to the L-SMI receiver, since $\Gamma_{j, L} = P_{j, L} R_{p_j p_j}^T$, after simple algebra, one obtains

$$
\text{trace}(\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*) = \text{trace}(R_{p_j p_j}^T \Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*) = \text{trace}(P_{j, L} R_{p_j p_j}^T \Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*) = \text{trace}(P_{j, L} R_{p_j p_j}^T R_{p_j p_j}^* R_{p_j p_j}^* R_{p_j p_j}^*),
$$

(80)

where we used the identities $P_{j, L} R_{p_j p_j}^* = R_{p_j p_j}^* P_{j, L}$ and $P_{j, L} R_{p_j p_j}^* R_{p_j p_j}^* = P_{j, L} R_{p_j p_j}^* R_{p_j p_j}^*$. To obtain a more manageable expression of $\text{trace}(\Gamma_{j, L}^T R_{p_j p_j}^* \Gamma_{j, L}^* R_{p_j p_j}^*)$, we consider its asymptotic value as $\sigma_v^2 \to 0$, i.e., in the high-SNR
regime. By accounting for the expression of $P_{j\perp}$ given by Lemma 3, observing that, under assumptions (a1) and (a2),

$$\lim_{\sigma_j^2 \to 0} P_{j\perp} R_{j\perp}^{-1} R_{j\perp} = \lim_{\sigma_j^2 \to 0} \left\{ I_N - \left( \Phi \Phi^H + \sigma_j^2 I_N \right)^{-1} \Phi j \phi_j^H \right\} \left( \Phi \Phi^H + \sigma_j^2 I_N \right)^{-1} \Phi^T$$

$$= \left[ I_N - \left( \Phi \Phi^H + \sigma_j^2 I_N \right)^{-1} \Phi^T \right] \Phi^T = \left( \Phi^T \right) S_j \Phi^T \right)$$

(81)

where $I_j \triangleq \left[ 0, \ldots, 0, 1, 0, \ldots, 0 \right]^T \in \mathbb{R}^{j \times j}$ and $S_j \triangleq I_j - I_j J_j \in \mathbb{R}^{J_j \times J_j}$. Accounting for (81), the asymptotic value of (80) is given by

$$\lim_{\sigma_j^2 \to 0} \text{trace} (\Gamma_{j\perp} \Phi^T \Phi^T) = \text{trace} \left( S_j \Phi_j \Phi_j^T \right) = J_j - 1.$$

As regards the L-SUB receiver, since $\Gamma_{j\perp} = P_{j\perp} R_{p_j p_j}^{-1} - \gamma_{j\perp} V_{n_j} V_{n_j}^H$, recalling that $R_{p_j p_j}^{-1} = \Phi_j \Phi_j^T$, and observing that $V_{n_j} \Phi_{j} = O(n \to j \times (j-1))$, it follows that

$$\text{trace} (P_{j\perp} R_{p_j p_j}^{-1} \Gamma_{j\perp} R_{p_j p_j}^{-1}) = \text{trace} \left( \Phi_j \Phi_j^T \Phi_j^T \right)$$

where $P_{j\perp} \Phi_j \Phi_j^T \Phi_j^T = \Phi_j \Phi_j^T$. This turns out to be exactly equal to (80). Thus, in conclusion, by substituting (79) and (82) into (78), eqs. (51) and (52) are obtained by additionally observing that, with reference to the L-SUB receiver, $\gamma_{j\perp} \sigma_j^2 \to 1$ as $\sigma_j^2 \to 0$.

REFERENCES


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