Differential Interference Suppression for SDMA-OFDM Based on Joint Multiple-Symbol Filtering and Detection

Li Wang, *Member, IEEE* and Lajos Hanzo, *Fellow, IEEE*

***Abstract*—This paper presents a multiple-symbol differential spatial division multiple access (MS-DSDMA) OFDM system conceived for low-complexity and high-bandwidth-efficiency applications operating in time-varying fading channels, where no channel estimation is required. A low-complexity adaptive multiple-symbol differential interference suppression (MS-DIS) scheme is proposed, which is based on the maximum signal-to-interference-plus-noise ratio (MSINR) criterion and facilitates the implementation of the powerful multiple-symbol differential sphere detection (MSDSD). Then, a practical three-stage turbo DIS receiver design framework is proposed for the MS-DSDMA system, which is constituted by concatenating the adaptive DIS filter bank, the MSDSD and the channel decoder. Both the EXtrinsic Information Transfer (EXIT) chart analysis and the Monte-Carlo-based simulation results show that the proposed three-stage turbo DIS scheme is capable of achieving a substantially enhanced performance in comparison to the conventional linear minimum mean-squared error (LMMSE) based adaptive receiver. Furthermore, two complexity reduction techniques are devised to significantly reduce the iterative detection complexity.**

I. Introduction

Orthogonal Frequency-Division Multiplexing (OFDM) [1] has become the predominant transmission technique of wideband digital communications. The prime benefit of OFDM is that the original frequency-selective wideband channel may be viewed as a set of parallel narrow-band channels created by the OFDM scheme. Thus, a high data rate may be achieved without using complex equalization techniques at the receiver. On the other hand, owing to the scarcity of spectral resources, one of the main objectives in the design of future communication systems is the efficient exploitation of the available spectrum, in order to accommodate the ever-increasing traffic demands. The most promising solution to achieve this goal is based on the exploitation of the spatial dimension, by using spatial division multiple access (SDMA) [1], where the user-specific channel impulse responses (CIRs) are estimated and invoked for differentiating the parallel uplink streams transmitted by the different users. However, it was revealed in [2] that the multiple-input multiple-output (MIMO) system's performance is highly sensitive to the channel estimation errors, which may only be mitigated at the cost of an excessive computational complexity and/or high pilot overheads in many practical time-varying fading scenarios. Fortunately, in cost- and complexity-constrained applications there are options to circumvent the channel estimation, where the multiple-access interference (MAI) may be estimated and exploited by an adaptive receiver for the desired user. For example, the adaptive minimum mean square error (MMSE) scheme [3] using the least mean square (LMS) or the recusive least squares (RLS) algorithm and the more recently proposed maximum signal-to-interference-plus-noise ratio (MSINR) based differential interference suppression (DIS) scheme [4]. For the former the interference suppression filter is adapted in order to minimize the MSE between the transmitted signal and the filter's output signal, while for the latter the filter coefficients are adjusted to maximize the SINR at its output and has been demonstrated in [4] to be able to mitigate the effects of carrier phase variations. Our novel contributions are:

1). Firstly, inspired by the block least-squares algorithm of [3] designed for standard MMSE adaptation, a new adaptive multiple-symbol DIS (MS-DIS) scheme is proposed based on our multiple-symbol differential SDMA (MS-DSDMA) OFDM system model for the sake of reducing the filter adaptation overheads and, even more importantly, for facilitating the implementation of the powerful multiple-symbol differential sphere detector (MSDSD) of [5].

2). Secondly, as a benefit of employing the MSDSD, an enhanced iteration gain may be achieved by the turbo receiver upon jointly detecting multiple consecutively received symbols. In order to further exploit the differential coding gains in the context of our adaptive MS-DIS scheme, a new channel-code-aided three-stage turbo DIS receiver is proposed, which allows a beneficial information exchange amongst the concatenated adaptive MS-DIS filter bank, the MSDSD and the channel decoder.

3). Thirdly, the new adaptive-window-duration (AWD) based MSDSD scheme is conceived, which is further aided by the proposed apriori*-LLR-threshold (ALT) technique to achieve significant complexity reductions in the turbo DIS receiver.*

The organization of the paper is as follows. Section II details our system model used in Section III dedicated to the MS-DSDMA principles. Section IV outlines our MS-DSDMA transceiver design, followed by Section V, which characterizes the performance versus complexity trade-offs, before concluding.

The authors are with the school of ECS, University of Southampton, Southampton, SO17 1BJ, UK (e-mail: {lw5,lh}@ecs.soton.ac.uk).

The finacial support of the RC UK under the auspices of the UK-India Advanced Technology Centre of Wireless Communications and of the China-UK Science Bridge in 4G wireless communications, as well as that of the EU's Concerto project is also gratefully acknowledged.

II. Multiple-Symbol DSDMA-OFDM System Model

In the context of DSDMA-OFDM, non-dispersive fading is encountered by each sub-carrier provided that the number of sub-carriers is sufficiently high. Hence, our scheme is equally applicable to single-carrier narrowband modems, if the differential encoding is carried out along the time dimension. Hence, let us consider the following per-sub-carrier-based frequency-domain (FD) system model constructed for DSDMA-OFDM systems supporting U differential-modulation-based N_t-antenna-aided mobile stations (MS) with the aid of N_r receiver antennas at the base station (BS) [1] for the nth OFDM symbol:

$$\mathbf{Y}[n] = \sum_{u=1}^{U} \mathbf{S}_u[n]\mathbf{H}_u[n] + \mathbf{W}[n], \tag{1}$$

where $\mathbf{Y}[n] \in \mathbb{C}^{N_t \times N_r}$, $\mathbf{S}_u[n] \in \mathbb{C}^{N_t \times N_t}$ and $\mathbf{W}[n] \in \mathbb{C}^{N_t \times N_r}$ denote the FD received and transmitted space-time signal matrices as well as the AWGN matrix having a distribution of $\mathcal{CN}(0, 2\sigma_w^2)$, respectively. The sub-carrier index is omitted here for notational simplicity. Due to pratical cost and size constraints, the employment of a single transmit antenna is assumed for each MS without loss of generality, i.e. we have $N_t = 1$. Thus, the uth single-antenna-aided MS differentially encodes its information symbols $\mathbf{V}_u[n] \in \mathcal{M}_c = \{e^{j2\pi m/M}; m = 0, 1, \cdots, M-1\}$, each of which contains $(\log_2 M)$-bit information, as $\mathbf{S}_u[n] = \mathbf{V}_u[n]\mathbf{S}_u[n-1]$. Furthermore, the FD channel transfer factor (FD-CTF) matrix $\mathbf{H}_u[n]$ is a $(N_t \times N_r)$-dimensional i.i.d. zero-mean unit-variance complex Gaussian matrix, which is also referred to as the user-specific spatial signature that has to be estimated in conventional SDMA systems. However, the FD-CTF matrix $\mathbf{H}_u[n]$ is not required at either the MS or the BS in the DSDMA-OFDM system considered for sake of circumventing the potentially excessive-complexity and yet inaccurate channel estimation.

Based on (1) we can also construct the per-sub-carrier-based *multiple-symbol* DSDMA-OFDM system model as:

$$\begin{aligned}\underline{\mathbf{Y}}[k_N] &= \sum_{u=1}^{U} \underline{\mathbf{Y}}_u[k_N] + \underline{\mathbf{W}}[k_N] \\ &= \sum_{u=1}^{U} \underline{\mathbf{S}}_u^{\mathrm{d}}[k_N]\underline{\mathbf{H}}_u[k_N] + \underline{\mathbf{W}}[k_N],\end{aligned} \tag{2}$$

where the kth received OFDM symbol block matrix $\underline{\mathbf{Y}}[k_N]$ contains N_{wind} consecutively received OFDM symbol matrices. Hence we have $\underline{\mathbf{Y}}[k_N] = [\mathbf{Y}[(N_{\mathrm{wind}}-1)(k-1)]^T \ \cdots \ \mathbf{Y}[(N_{\mathrm{wind}}-1)k]^T]^T$, and both the kth FD-CTF block matrix $\underline{\mathbf{H}}_u[k_N]$ as well as the AWGN's kth block matrix $\underline{\mathbf{W}}[k_N]$ are defined by vertically stacking the N_{wind} matrices $\mathbf{H}_u[n]$ and $\mathbf{W}[n]$ $(n = (N_{\mathrm{wind}}-1)(k-1), \cdots, (N_{\mathrm{wind}}-1)k)$, respectively. Moreover, the kth diagonal block matrix of the transmitted signal $\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N]$ of the uth MS is constructed as $\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N] = \mathrm{diag}\{\mathbf{S}_u[(N_{\mathrm{wind}}-1)(k-1)] \ \cdots \ \mathbf{S}_u[(N_{\mathrm{wind}}-1)k]\}$, which corresponds to the length-$(N_{\mathrm{wind}}-1)$ information symbol block matrix $\underline{\mathbf{V}}_u[k_N] = [\mathbf{V}_u[(N_{\mathrm{wind}}-1)(k-1)+1]^T \ \cdots \ \mathbf{V}_u[(N_{\mathrm{wind}}-1)k]^T]^T$.

III. Multiple-Symbol Differential Spatial Division Multiple Access

It is observed from both (1) and (2) that non-coherent differential detection techniques cannot be directly applied at the BS to recover the information pertaining to a specific MS without suppressing the interference imposed by all the other MSs. Therefore, we will use the designed MSINR approach for interference suppression in the DSDMA-OFDM system as advocated in [4]. However, rather than computing the uth MS's linear vector filter $\mathbf{f}_u[n]$ of a specific sub-carrier for each OFDM symbol duration n, we propose updating $\mathbf{f}_u[k_N]$ only once for N_{wind} OFDM symbol durations based on the most recently received N_{wind} signal matrices hosted by $\underline{\mathbf{Y}}[k_N]$ of (2). The resultant new multiple-symbol MSINR (MS-MSINR) criterion reduces the filter-update overhead and additionally facilitates the implementation of the powerful multiple-symbol differential sphere detector (MSDSD) in the ensuing stage, hence achieving significant performance improvements.

A. Adaptive Multiple-Symbol Differential Interference Suppression

1) Multiple-Symbol MSINR Criterion: Let us first derive the MS-MSINR criterion based on the multiple-symbol system model of (2). In order to extract N_{wind} differentially encoded OFDM symbols transmitted consecutively by the vth MS, from the kth block of N_{wind} successively received OFDM symbols hosted by $\underline{\mathbf{Y}}[k_N]$ of (2), $\underline{\mathbf{Y}}[k_N]$ is passed through a linear vector filter $\mathbf{f}_v[k_N]$ of length N_r, yielding the filter output $\mathbf{y}_v[k_N]$ of length N_{wind} as:

$$\mathbf{y}_v[k_N] = \underline{\mathbf{Y}}[k_N]\mathbf{f}_v[k_N]. \tag{3}$$

We define the multiple-symbol-based SINR as the ratio between the sum power of the N_{wind} desired filter output components of $\underline{\mathbf{Y}}_v[k_N]\mathbf{f}_v[k_N]$ and the sum power of the N_{wind} interference-plus-noise components of $(\underline{\mathbf{Y}}[k_N] - \underline{\mathbf{Y}}_v[k_N])\mathbf{f}_v[k_N]$. Our goal is to find the specific filter $\mathbf{f}_v[k_N]$ capable of maximizing the filter's output SINR, which may be mathematically expressed as:

$$\begin{aligned}\mathbf{f}_v[k_N] &= \max_{\mathbf{f}_v[k_N]} \frac{\mathbf{f}_v^{\mathrm{H}}[k_N](\mathbf{R}[k_N] - \mathbf{R}_v^i[k_N])\mathbf{f}_v[k_N]}{\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N]}, \\ &= \max_{\mathbf{f}_v[k_N]} \frac{\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}[k_N]\mathbf{f}_v[k_N]}{\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N]},\end{aligned} \tag{4}$$

where $\mathbf{R}[k_N]$ is the correlation matrix of the multiple-symbol-based received signal $\underline{\mathbf{Y}}[k_N]$ of (2), defined as

$$\begin{aligned}\mathbf{R}[k_N] &\triangleq E\{\underline{\mathbf{Y}}^{\mathrm{H}}[k_N]\underline{\mathbf{Y}}[k_N]\} \\ &= \sum_{u=1}^{U} \underline{\mathbf{H}}_u^{\mathrm{H}}[k_N]\underline{\mathbf{H}}_u[k_N] + 2\sigma_w^2 N_{\mathrm{wind}} N_t \mathbf{I}_{N_r}\end{aligned} \tag{5}$$

and the multiple-symbol-based interference-plus-noise correlation matrix $\mathbf{R}_v^i[k_N]$ may be expressed as

$$\begin{aligned}\mathbf{R}_v^i[k_N] &\triangleq E\{(\underline{\mathbf{Y}}[k_N] - \underline{\mathbf{Y}}_v[k_N])^{\mathrm{H}}(\underline{\mathbf{Y}}[k_N] - \underline{\mathbf{Y}}_v[k_N])\} \\ &= \sum_{u=1; u\neq v}^{U} \underline{\mathbf{H}}_u^{\mathrm{H}}[k_N]\underline{\mathbf{H}}_u[k_N] + 2\sigma_w^2 N_{\mathrm{wind}} N_t \mathbf{I}_{N_r}.\end{aligned} \tag{6}$$

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication.

Using the method of Lagrange multipliers, we may solve (4) by maximizing $\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}[k_N]\mathbf{f}_v[k_N]$ under the constraint that the interference-plus-noise component $\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N]$ is fixed. Hence, the corresponding Lagrange cost function $J_{\mathrm{SNR}}(\mathbf{f}_v[k_N])$ can be defined as $J_{\mathrm{SNR}}(\mathbf{f}_v[k_N]) \triangleq \mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}[k_N]\mathbf{f}_v[k_N]+\lambda(c_{\mathrm{SNR}}-\mathbf{f}_v^{\mathrm{H}}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N])$, where $c_{\mathrm{SNR}} > 0$ is an arbitrary positive constant, and λ represents the real-valued Lagrange multiplier. Then, differentiating $J_{\mathrm{SNR}}(\mathbf{f}_v[k_N])$ with respect to $\mathbf{f}_v^*[k_N]$ and setting the result equal to the all-zero vector of length N_r yields [6]:

$$\mathbf{R}[k_N]\mathbf{f}_v[k_N] = \lambda\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N]. \tag{7}$$

2) MS-MSINR-Based Differential Interference Suppression: It is observed from (5) and (6) that in coherent-detection-aided SDMA systems, channel estimation has to be carried out to acquire each MS's spatial signature $\underline{\mathbf{H}}_u[k_N]$ for the sake of determining the coefficients of the MS-MSINR filter $\mathbf{f}_v[k_N]$ by solving the generalized eigenvalue problem of (7) using the singular-value decomposition (SVD) [7]. Fortunately, despite dispensing with channel estimation in the DSDMA-OFDM system, the interference-plus-noise correlation matrix $\mathbf{R}_v^i[k_N]$ of (6) may be calculated by exploiting the differentially encoded transmission principles as shown in Lemma 1 below:

Lemma 1: Under the assumption of a relatively slow fading channel, the multiple-symbol-based interference-plus-noise correlation matrix $\mathbf{R}_v^i[k_N]$ may be approximately evaluated as

$$\mathbf{R}_v^i[k_N] \approx \mathcal{E}\{\underline{\mathbf{E}}_v^{\mathrm{H}}[k_N]\underline{\mathbf{E}}_v[k_N]\}, \tag{8}$$

where the multiple-symbol-based interference-plus-noise signal matrix $\underline{\mathbf{E}}_v[k_N]$ is defined as:

$$\underline{\mathbf{E}}_v[k_N] \triangleq \sqrt{\frac{1}{2}}(\underline{\mathbf{Y}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}]), \tag{9}$$

with the block index k_N^{-n} representing the kth block shifted backwards by n OFDM symbol durations and the $(N_tN_{\mathrm{wind}} \times N_tN_{\mathrm{wind}})$-element diagonal block matrix $\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N] = \mathrm{diag}\{\underline{\tilde{\mathbf{V}}}_v[k_N]\} = \mathrm{diag}\{[\mathbf{V}_v[(N_{\mathrm{wind}}-1)(k-1)]^T, \underline{\mathbf{V}}_v[k_N]^T]^T\}$ is the multiple-symbol-based transmitted information symbol matrix of the vth MS.

Proof of Lemma 1: Assuming a relatively slow fading channel, i.e. $\underline{\mathbf{H}}_u[k_N^{-1}] \approx \underline{\mathbf{H}}_u[k_N]$, $(u = 1, 2, \cdots, U)$ for each sub-carrier of the DSDMA-OFDM system considered, we have (10), based on which we may arrive at (11). Note both (10) and (11) are given at the top of the next page.

Furthermore, since all the diagonal block matrices $\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N]$, $\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]$ as well as $\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]$ are unitary, namely, we have $\left(\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N] = \left(\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N] = \left(\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N] = \mathbf{I}_{N_tN_{\mathrm{wind}}}$, finally we arrive at:

$$\mathcal{E}\left\{\left(\underline{\mathbf{Y}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}]\right)^{\mathrm{H}}\left(\underline{\mathbf{Y}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}]\right)\right\}$$
$$= 2\left(\sum_{u=1;u\neq v}^{U}\underline{\mathbf{H}}_u^{\mathrm{H}}[k_N]\underline{\mathbf{H}}_u[k_N] + 2\sigma_w^2N_{\mathrm{wind}}N_t\mathbf{I}_{N_r}\right), \tag{12}$$
$$= 2\mathbf{R}_v^i[k_N]. \tag{13}$$

Hence, the factor $\sqrt{1/2}$ in (9) is included to ensure the validity of (8). This completes the proof of Lemma 1.

Remarks: The diagonal block matrix $\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]$ is known to the receiver during the training session or may be estimated by using the previous decisions $\hat{\mathbf{V}}_v[n], (n = (N_{\mathrm{wind}}-1)(k-1), \cdots, (N_{\mathrm{wind}}-1)k)$. Therefore, both $\mathbf{R}[k_N]$ and $\mathbf{R}_v^i[k_N]$ may be estimated for the kth length-N_{wind} block without acquiring the CSI and then tracked recursively in time-varing fading channels as follows:

$$\mathbf{R}[k_N] = \beta\mathbf{R}[k_N-1] + (1-\beta)\underline{\mathbf{Y}}^{\mathrm{H}}[k_N]\underline{\mathbf{Y}}[k_N], \tag{14}$$
$$\mathbf{R}_v^i[k_N] = \mu\mathbf{R}_v^i[k_N-1] + (1-\mu)\underline{\mathbf{E}}_v^{\mathrm{H}}[k_N]\underline{\mathbf{E}}_v[k_N], \tag{15}$$

where $0 < \beta, \mu < 1$ is the forgetting factor.

3) Adaptive Implementation of MS-DIS: In practice, rather than carrying out the high-complexity SVD to solve the generalized eigenvalue problem of (7), we apply the modified adaptive Newton algorithm of [8] to recursively update the DIS filter $\mathbf{f}_v[k_N]$. This modified adaptive Newton algorithm, which was shown in [8] to have a fast convergence and an excellent tracking capability[1], may be summarized based on (14) and (15) as follows:

$$\mathbf{P}_v[k_N+1] = \frac{\mathbf{P}_v[k_N]}{\mu} - \frac{\mathbf{P}_v[k_N]\underline{\mathbf{E}}_v^{\mathrm{H}}[k_N+1]\underline{\mathbf{E}}_v[k_N+1]\mathbf{P}_v[k_N]}{\mu\mathrm{Trace}(\mu\mathbf{I} + \underline{\mathbf{E}}_v[k_N+1]\mathbf{P}_v[k_N]\underline{\mathbf{E}}_v^{\mathrm{H}}[k_N+1])},$$
$$\mathbf{a}_v[k_N+1] = \underline{\mathbf{Y}}[k_N+1]\mathbf{f}_v[k_N],$$
$$\mathbf{r}_v[k_N+1] = \beta\mathbf{r}_v[k_N] + (1-\beta)\underline{\mathbf{Y}}^{\mathrm{H}}[k_N+1]\mathbf{a}_v[k_N+1],$$
$$b_v[k_N+1] = \beta b_v[k_N] + (1-\beta)\mathbf{a}_v^{\mathrm{H}}[k_N+1]\mathbf{a}_v[k_N+1],$$
$$\tilde{\mathbf{f}}_v[k_N+1] = \frac{\mathbf{r}_v[k_N+1]}{b_v[k_N+1]},$$
$$\mathbf{f}_v[k_N+1] = \frac{2\mathbf{P}_v[k_N+1]\tilde{\mathbf{f}}_v[k_N+1]}{1 + \tilde{\mathbf{f}}_v^{\mathrm{H}}[k_N+1]\mathbf{P}_v[k_N+1]\tilde{\mathbf{f}}_v[k_N+1]}.$$

For algorithm initialization, we simply adopt $\mathbf{P}_v[1] = 0.01\mathbf{I}_{N_r}$, $\mathbf{f}_v[1] = \mathbf{r}_v[1] = \frac{1}{N_r}[11\cdots1]^T$ and $b[1] = 1$. Accordingly, the filter $\mathbf{f}_v[k_N]$ is updated at the beginning of each N_{wind}-OFDM-symbol block and it is used unaltered within the N_{wind}-OFDM-symbol block to suppress the MAI induced by other MSs.

B. Multiple-Symbol Differential Sphere Detection

As mentioned previously, apart from its beneficial complexity reduction, the multiple-symbol-based DIS scheme also facilitates the implementation of the MSDSD detection technique as a benefit of imposing no further distortion to the phase difference between the consecutively transmitted symbols in addition to that caused by the time-varying fading channel. Let us now briefly review the soft-input soft-output (SISO) MSDSD scheme, which will be used to generate the soft-bit-information for the desired vth user following the DIS stage. Under the assumption that the interference imposed by all other MSs was significantly mitigated after

[1]Our investigations, which are omitted here owing to the lack of space, indicate that the MSINR-based DIS scheme exhibits a lower sensitivity to the quality of the feedback decision than that of its conventional RLS-LMMSE-based counterpart, resulting in a superior tracking capability.

$$\begin{aligned}\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}] &= \left(\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{S}}_v^{\mathrm{d}}[k_N^{-1}]\underline{\mathbf{H}}_v[k_N^{-1}] + \sum_{u=1;u\neq v}^{U}\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N^{-1}]\underline{\mathbf{H}}_u[k_N^{-1}]\right) + \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{W}}[k_N^{-1}],\\ &\approx \left(\underline{\mathbf{S}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{H}}_v[k_N] + \sum_{u=1;u\neq v}^{U}\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\underline{\mathbf{H}}_u[k_N]\right) + \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{W}}[k_N^{-1}],\end{aligned} \tag{10}$$

where $\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N] \triangleq \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N^{-1}]$.

$$\begin{aligned}&\mathcal{E}\left\{\left(\underline{\mathbf{Y}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}]\right)^{\mathrm{H}}\left(\underline{\mathbf{Y}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{Y}}[k_N^{-1}]\right)\right\}\\ &= \sum_{u=1;u\neq v}^{U}\underline{\mathbf{H}}_u^{\mathrm{H}}[k_N]\mathcal{E}\left\{\left(\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N] - \underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\left(\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N] - \underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\right)\right\}\underline{\mathbf{H}}_u[k_N]\\ &+ \mathcal{E}\left\{\left(\underline{\mathbf{W}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{W}}[k_N^{-1}]\right)^{\mathrm{H}}\left(\underline{\mathbf{W}}[k_N] - \underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{W}}[k_N^{-1}]\right)\right\},\\ &= \sum_{u=1;u\neq v}^{U}\underline{\mathbf{H}}_u^{\mathrm{H}}[k_N]\mathcal{E}\left\{\left(\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\mathbf{S}}_u^{\mathrm{d}}[k_N] + \left(\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\tilde{\mathbf{S}}}_{vu}^{\mathrm{d}}[k_N]\right\}\underline{\mathbf{H}}_u[k_N]\\ &+ \mathcal{E}\left\{\underline{\mathbf{W}}^{\mathrm{H}}[k_N]\underline{\mathbf{W}}[k_N] + \underline{\mathbf{W}}^{\mathrm{H}}[k_N^{-1}]\left(\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\right)^{\mathrm{H}}\underline{\tilde{\mathbf{V}}}_v^{\mathrm{d}}[k_N]\underline{\mathbf{W}}[k_N^{-1}]\right\}.\end{aligned} \tag{11}$$

the DIS processing, the probability density function (PDF) of the DIS filter's output signal $\mathbf{y}_v[k_N]$ of (3) was conditioned on the transmitted signal $\mathbf{S}_v^{\mathrm{d}}[k_N]$, which may be approximated for Rayleigh fading channels as [9] (the N_{wind}-symbol block index k_N is omitted here for the sake of notational simplicity): $p(\mathbf{y}_v|\underline{\mathbf{S}}_v^{\mathrm{d}}) \approx \frac{\exp(-\mathbf{y}_v^{\mathrm{H}}\Psi^{-1}\mathbf{y}_v)}{\det(\pi\Psi)}$, where we have $\Psi = \mathcal{E}\{\mathbf{y}_v\mathbf{y}_v^{\mathrm{H}}|\underline{\mathbf{S}}_v^{d}\} = \underline{\mathbf{S}}_v^{\mathrm{d}}\Sigma_v\underline{\mathbf{S}}_v^{\mathrm{d}H} + 2\sigma_w^2\mathbf{I}_{T_b}$, in which $\Sigma_v = \mathcal{E}\{\underline{\mathbf{H}}_v\underline{\mathbf{H}}_v^{\mathrm{H}}\}$ denotes the vth MS's channel covariance matrix. According to its definition, a reduced-complexity computation of the *a posteriori* Log-Likelihood-Ratio (LLR) associated with the ith transmitted bit x_i at the output of the maximum-*a-posterori* (MAP) based MSDSD [5] may be finally expressed with the aid of Bayes' theorem and the "max-sum" approximation as:

$$L_D(x_i) = \ln\frac{\Pr(x_i = 1|\mathbf{y}_v)}{\Pr(x_i = -1|\mathbf{y}_v)}, \tag{16}$$

$$\begin{aligned}&\approx -\left\|\mathbf{U}\hat{\mathbf{s}}_{\mathrm{MAP}}^{x_i=+1}\right\|^2 + \ln\left[\Pr(\hat{\mathbf{x}}_{\mathrm{MAP}}^{x_i=+1})\right]\\ &+ \left\|\mathbf{U}\hat{\mathbf{s}}_{\mathrm{MAP}}^{x_i=-1}\right\|^2 - \ln\left[\Pr(\hat{\mathbf{x}}_{\mathrm{MAP}}^{x_i=-1})\right],\end{aligned} \tag{17}$$

where $\mathbf{U}$ is an upper-triangular matrix, which may be obtained as $\mathbf{U} \triangleq (\mathbf{F}\ \mathrm{diag}\{\mathbf{y}_v\})^*$, and $\mathbf{F}$ is also an upper-triangular matrix generated using the Cholesky factorization of the matrix $(\Sigma_v + 2\sigma_w^2\mathbf{I}_{N_{\mathrm{wind}}})^{-1}$. Thus, thanks to the upper-triangular structure of the matrix $\mathbf{U}$, when evaluating Eq. (17), we may find the multiple-symbol-based vectors $\hat{\mathbf{s}}_{\mathrm{MAP}}^{x_i=b}$ and $\hat{\mathbf{x}}_{\mathrm{MAP}}^{x_i=b}$, $(b = -1 \text{ or } +1)$, which host the MAP symbol and the corresponding bit estimates, respectively, by the sphere detection (SD) algorithms. Furthermore, $\Pr(\mathbf{x})$ of Eq. (17) is the *a priori* probability, which may be computed based on the *a priori* LLRs delivered by the outer channel decoder in an iterative receiver. In the sequel, the extrinsic LLR, $L_E(x_i)$, delivered by the MSDSD, may be calculated by excluding the corresponding *a priori* LLR, $L_A(x_i) = \frac{\Pr(x_i=1)}{\Pr(x_i=-1)}$, from the *a posteriori* LLR of (17), according to $L_E(x_i) = L_D(x_i) - L_A(x_i)$, which is then exploited by the outer decoder after passing it through the interleaver.

IV. MS-DSDMA Transceiver Design

A. Turbo DIS Filter Optimization

In order to improve the quality of the feedback decision and hence circumventing the need for a potentially excessive training sequence overhead, we propose a channel-code aided turbo DIS receiver for the DSDMA-OFDM system supporting U MSs, as depicted in Fig. 1. Specifically, the BS receiver of Fig. 1 is constituted by three modules, namely the DIS filter bank, the MSDSD and the channel decoder, where the extrinsic information may be exchanged amongst the three concatenated components in a number of consecutive iterations. As shown in Fig. 1, $A(\cdot)$ represents the *a priori* information expressed in terms of the LLRs, while $E(\cdot)$ denotes the corresponding *extrinsic* information, whereas the labels u and c represent the uncoded and coded bits, respectively, corresponding to the specific module indicated by the subscript.

1) Channel-Code-Aided Turbo DIS: At the early stage of the iterative detection process, namely when less confident *a priori* information is gleaned from the channel decoder in comparison to that provided by the MSDSD (namely when the mutual information (MI) $I_{E(c_1)}$ between the *extrinsic* value $E(c_1)$ and the bit c_1 is smaller than that between the *extrinsic* value $E(u_2)$ and the bit u_2, i.e. $I_{E(u_2)}$), $\underline{\tilde{\mathbf{V}}}_v[k_N]$ of (9) should be obtained based on the output of the MSDSD by toggling the decision-directed mode switch to the 'a' location of Fig. 1, if the system is working in the decision-directed mode. However, as soon as the *a priori* information becomes more confident during the iterative detection process, namely when we have

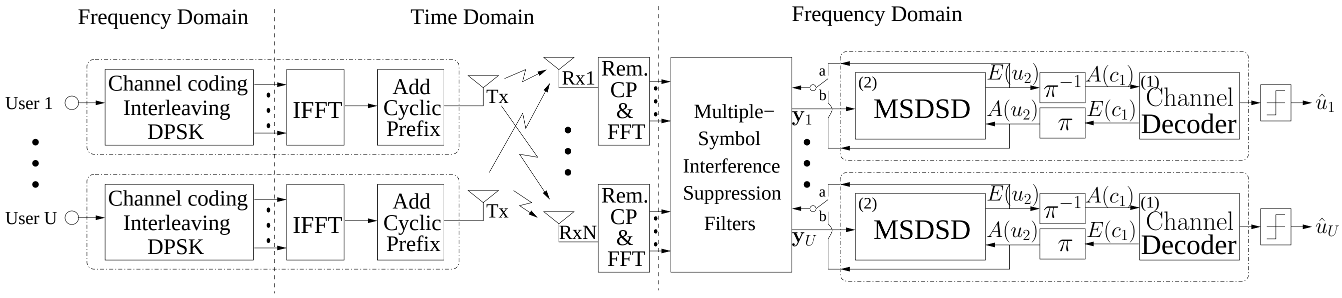


Fig. 1. Multiple-symbol DSDMA-OFDM transceiver architecture.

$I_E(c_1) > I_E(u_2)$, it is preferred to switch to the "channel-code-aided" decision-directed mode by toggling the switch to the 'b' location in Fig. 1, so that $\underline{\tilde{\mathbf{V}}}_v[k_N]$ of (9) is calculated from the *a priori* information provided by the channel decoder, for the sake of enhancing the optimization of the DIS filter bank.

2) Soft-Symbol-Decision-Direct DIS: Based on the idea of retaining the valuable soft-information contained in the *a posteriori* LLRs, which would be simply discarded by the action of subjecting the LLRs to hard decisions, soft-symbol-decision-direct (SSDD) DIS is advocated, where the soft- rather than hard-decision symbol is caculated based on the *a priori* LLRs delivered either by the MSDSD or by the channel decoder, which in turn is used as our estimate of the transmitted symbol in (9) to adjust the DIS filter's coefficients. In order to assess the MAI suppression capability of the multiple-symbol-based adaptive DIS scheme, we examine the SINR loss (SINRL) at the output of the DIS filter $\mathbf{f}_v[k_N]$ in comparison to that achieved by the theoretically optimal filter $\mathbf{f}_v^o[k_N]$, which is obtained under the assumption of having perfect channel knowledge at the receiver. More specifically, the SINRL for the k_Nth N_{wind}-symbol block can be expressed in dB as: $10\log\left(\frac{\mathbf{f}_v^{oH}[k_N]\mathbf{R}[k_N]\mathbf{f}_v^o[k_N]\mathbf{f}_v^{\text{H}}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v[k_N]}{\mathbf{f}_v^{oH}[k_N]\mathbf{R}_v^i[k_N]\mathbf{f}_v^o[k_N]\mathbf{f}_v^{\text{H}}[k_N]\mathbf{R}[k_N]\mathbf{f}_v[k_N]}\right)$. In Fig. 2 we visualize the benefits of the SSDD scheme by plotting the filter's SINRL versus both the filter optimization iteration index and the *a priori* MI measured at its decision feedback branch seen in Fig. 1. For simplicity, we consider a (2×2)-element DSDMA-OFDM system with its system parameters being summarized in Table I. As observed in Fig. 2, significant SINRL performance gains may be achieved by the SSDD scheme, over its hard-symbol-decision-directed (HSDD) counterpart, which indicates the enhanced robustness of SSDDs against error-prone feedback decisions. More specifically, the effects of the sharply-degraded SINRL experienced by the HSDD-DIS filter when the *a priori* MI is low may be substantially mitigated by employing the SSDD-based regime. On the other hand, it is also observed from Fig. 2 that the adaptive DIS filter's tracking capability is enhanced for both the DSDD- and SSDD-based techniques, when more confident *a priori* information becomes available which is expected to become available owing to the beneficial information exchange amongst the concatenated blocks of the three-stage turbo DIS receiver of Fig. 1.

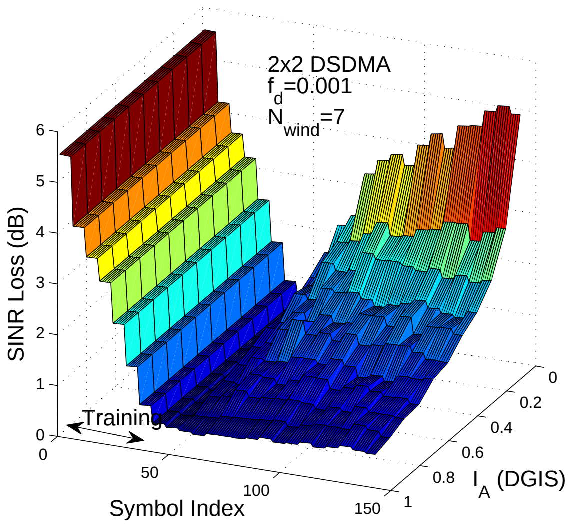
TABLE I
SUMMARY OF SYSTEM PARAMETERS

Modulation	DQPSK in Time Domain
Users Supported	2
Normalized Doppler Freq.	0.001
System	DSDMA-OFDM Uplink
Sub-Carriers	1024
Rx at BS	2
Channel Code	Half-Rate RSC(2,1,3) (5/7)
TDL Channel Model	Typical Urban 6-Tap Channel Model
Channel Delay Profile	[0 2 6 16 24 50]

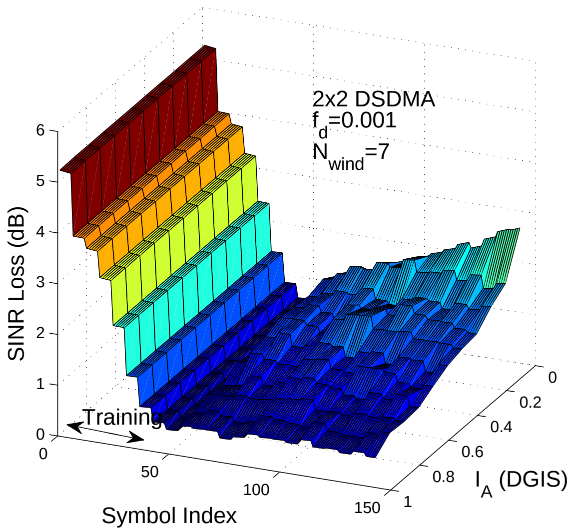
B. Reduced-Complexity MSDSD Design

In order to reduce the complexity imposed by the MSDSD decoder for the turbo DIS receiver of Fig. 1, two novel complexity reduction techniques, namely the adaptive-window-duration (AWD) and the *apriori*-LLR-threshold (ALT), are devised in this section by taking advantage of the iterative detection mechanism.

1) Adaptive-Window-Duration Based MSDSD: Unlike the conventional MSDSD scheme using a fixed observation window size of N_{msdsd} during the entire iterative detection process, the observation window size employed by the MSDSD was initially set to $N_{\text{msdsd}} = 2$ for the sake of low complexity, which will be slightly increased, as soon as the iterative decoding process exchanging extrinsic information between the combined "DIS-MSDSD" decoder and the channel decoder converges. The proposed AWD-aided MSDSD scheme is characterized with the aid of the EXIT chart seen in Fig. 3(a) in the context of a (2×2)-element DQPSK modulated DSDMA-OFDM system, where we may observe the transition of the decision-direct mode from the MSDSD-based mode to the channel-code-based mode at the second iteration, as we discussed in Section IV-A1. Indeed, the complexity imposed by the MSDSD is significantly reduced by the AWD scheme, as observed in Fig. 3(b), where the complexity imposed by the MSDSD in terms of the number of the PED evaluations per bit is plotted versus the SNR for the systems operating both with and without the AWD scheme. Remarkably, the complexity imposed by the MSDSD is substantially reduced in Fig. 3(b) with the aid of the AWD scheme, namely by as much as 66% at the SNR of 4.5dB, when the open EXIT tunnel created by having $N_{\text{msdsd}} = 7$ is rather narrow. This is not unexpected, since although an increased number of iterations may be needed between the "DIS-MSDSD" decoder and the

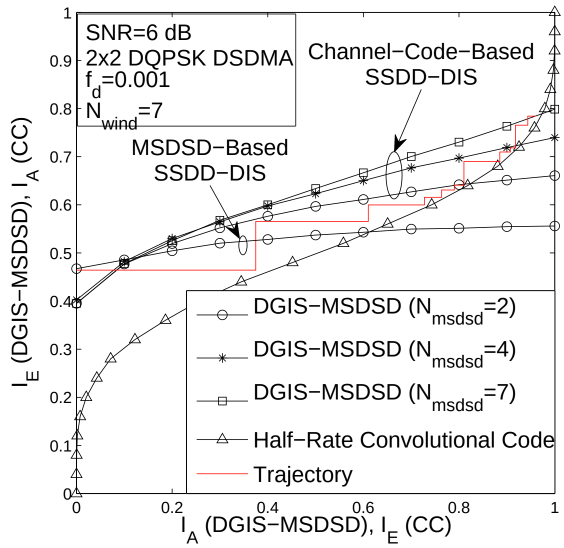


(a) Hard-symbol-decision-direct.

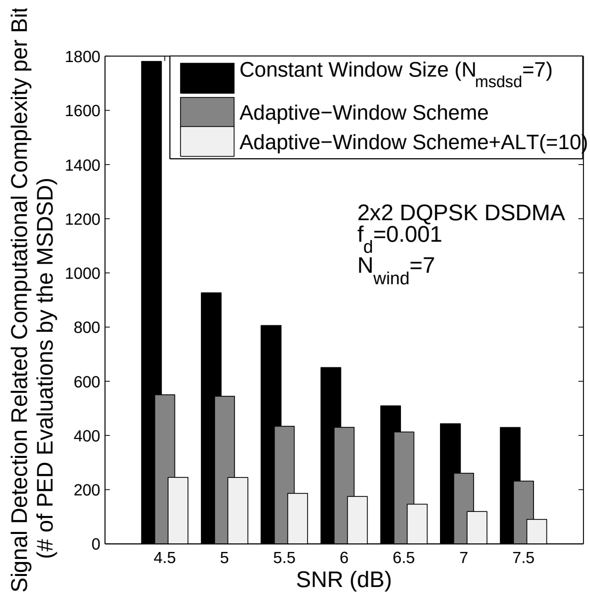


(b) Soft-symbol-decision-direct.

Fig. 2. SINRL performance of the hard-symbol-decision- and soft-symbol-decision-direct multiple-symbol DIS filters.



(a) EXIT trajectory.



(b) Complexity reduction achieved.

Fig. 3. Characterization of the adaptive-window aided scheme for the MSDSD.

CC decoder to achieve the same amount of iteration gain, when the AWD scheme is employed, the per-iteration complexity imposed by the MSDSD using a reduced N_{msdsd} is expected to be exponentially reduced, yielding a potentially reduced overall complexity.

2) Apriori-LLR-Threshold Aided MSDSD: In order to further reduce the complexity imposed by the MSDSD decoder during the iterative decoding process, an *apriori*-LLR-threshold-based scheme is proposed for the MSDSD. First of all, let us review the definition of the *a priori* LLRs, which is the logarithm of the ratio of the bit probabilities associated with $+1$ and -1, that may be expressed as $L_A(x_j) = \ln \frac{P[x_j=+1]}{P[x_j=-1]}$. Therefore, the sign of the resultant LLRs indicates whether the current bit is more likely to be $+1$ or -1, whereas the magnitude reflects how reliable the decision concerning the current bit is. In the light of this, the search space of the MSDSD may be significantly reduced by invoking an ALT controlled technique. To be specific, when calculating the *a posteriori* LLR $L_D(x_i)$ of (16) for the ith bit component x_i of the bit vector $\mathbf{x}$, the MSDSD search space may be reduced by a factor of 2^J, if the *a priori* LLRs of J number of bit elements x_j $(j \neq i, j \in \mathcal{J})$ delivered by the channel decoder exhibit high magnitudes, which are higher than the preset threshold T_{ALT}, as the iterative detection proceeds. This is because the vector candidates $\mathbf{x}$ associated with x_j $(j \in \mathcal{J})$ having values opposite to those indicated by the sign of $L_A(x_j)$ $(j \in \mathcal{J})$ are unlikely to be the genuine transmitted bit vector, which may be excluded from the search

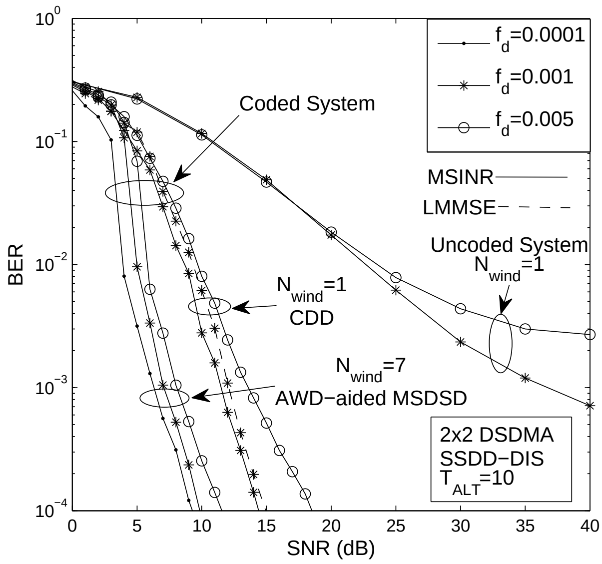


Fig. 4. BER performance of the MS-DSMA OFDM system using the ALT- and AWD-aided MSDSD.

space. For example, our investigation suggests that it is desirable to have $T_{\mathrm{ALT}} = 10$ for the (2×2) DSDMA-OFDM system considered above, which is capable of achieving a significant complexity reduction for the MSDSD without suffering any substantial system performance losse. Noticeably, as seen in Fig. 3(b), the combination of the proposed AWD and ALT schemes allows the MSDSD to achieve an identical iterative gain at SNR$=4.5$ dB, despite imposing a substantially reduced computational complexity, which is only about $\frac{1}{9}$ of that required by the conventional MSDSD scheme.

V. Simulation Results and Discussions

In Fig. 4 the BER performance of the proposed turbo MS-DIS-aided DSDMA system of Fig. 1 is plotted in comparison to those of its LMMSE-based and MSINR-based single-symbol-DIS-aided counterparts, in the specific context where two single-antenna-aided users are assumed to transmit simutaneously to the two-antenna-aided BS. The simulation parameters are summarized in Table I. It is observed in Fig. 4 that for $N_{\mathrm{wind}} = 1$ the coded RLS-based-LMMSE DSDMA-OFDM system is slightly inferior to its MSINR-based counterpart in terms of the BER performance within the SNR range of interest. Furthermore, when the MS-DIS scheme operates in conjunction with $N_{\mathrm{wind}} = 7$, the MSINR-based system using the ALT- and AWD-aided MSDSD is capable of achieving an SNR gain of 5 dB over its LMMSE-based counterpart at the BER target of 10^{-4} in the channel-coded scenario associated with $f_d = 0.001$. Finally, observe in Fig. 4 that the error-floor induced by a more severely time-selective channel may be significantly mitigated by the proposed MSINR-based MS-DIS scheme in conjunction with the ALT- and AWD-aided MSDSD. More specifically, an SNR gain of about 7 dB can be achieved by the proposed turbo MS-DIS-aided three-stage receiver employing $N_{\mathrm{wind}} = 7$ in comparison to the conventional MSINR-based DIS-assisted system using $N_{\mathrm{wind}} = 1$ in the time-varying fading channel associated with $f_d = 0.005$.

VI. Conclusion

A turbo MS-DIS-aided three-stage receiver employing the MSDSD was proposed for the DSDMA-OFDM system, which is suitable for low-complexity and high-bandwidth-efficiency applications in time-varying fading channels, where no channel estimation is required. With the aid of the ALT- and AWD-aided MSDSD scheme devised, the MS-DSDMA-OFDM system is capable of achieving a significant performance improvements over both its conventional LMMSE- and MSINR-based DIS assisted counterparts, while imposing an affordable computational complexity.

References

[1] L. Hanzo, M. Munster, B. J. Choi, and T. Keller, *OFDM and MC-CDMA for Broadband Multi-User Communications, WLANs and Broadcasting*. IEEE Press, 2003.

[2] T. Weber, A. Sklavos, and M. Meurer, "Imperfect channel-state information in mimo transmission," *IEEE Transactions on Communications*, vol. 54, pp. 543–552, Mar. 2006.

[3] H. V. Poor and G. W. Wornell, *Wireless Communications: Signal Processing Perspectives*. Prentice Hall PTR, 1998.

[4] S. K. Cheung and R. Schober, "Differential spatial multiplexing," *IEEE Transactions on Wireless Communications*, vol. 5, pp. 2127–2135, Aug. 2006.

[5] L. Lampe, R. Schober, V. Pauli, and C. Windpassinger, "Multiple-symbol differential sphere decoding," *IEEE Transactions on Communications*, vol. 12, pp. 1981–1985, Dec. 2005.

[6] H.-J. Su and E. Geraniotis, "Maximum signal-to-noise ratio array processing for space-time coded systems," *IEEE Transactions on Communications*, vol. 50, pp. 1419–1422, Sept. 2002.

[7] R. A. Horn and C. R. Johnson, *Matrix Analysis*. The Press Syndicate of the University of Cambridge, 1999.

[8] J. Yang, F. Yang, H.-S. Xi, W. Guo, and Y. Sheng, "Robust adaptive modified newton algorithm for generalized eigendecomposition and its application," *EURASIP Journal on Advances in Signal Processing*, vol. 2007, pp. 1–10, June 2007.

[9] R.-R. Chen, R. Koetter, U. Madhow, and D. Agrawal, "Joint noncoherent demodulation and decoding for the block fading channel: a practical framework for approaching shannon capacity," *IEEE Transactions on Communications*, vol. 51, pp. 1676–1689, Oct. 2003.