Correspondence

1 OFDM-Aided Differential Space–Time Shift Keying Using Iterative Soft Multiple-Symbol Decoding

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3 Abstract—Soft-decision multiple-symbol differential sphere decoding (MSDD) is proposed for orthogonal frequency-division multiplexing (OFDM)-aided differential space–time shift keying (DSTSK)–aided transmission over frequency-selective channels. Specifically, the DSTSK signaling blocks are generated by the channel-encoded source information and the space–time (ST) blocks are appropriately mapped to a number of OFDM subcarriers. After OFDM demodulation, the DSTSK signal is noncoherently detected by our soft-decision MSDSD detector. A novel soft-decision MSDSD is designed, and the associated decision rule is derived for the DSTSK scheme. Our simulation results demonstrate that an SNR reduction of 2 dB is achieved by the proposed scheme using an MSDSD window size of 4 over the conventional soft-decision-aided differential detection benchmark, while communicating over dispersive channels and dispensing with channel estimation (CE).

4 Index Terms—Extrinsic information transfer (EXIT) chart, iterative decoding, multiple-symbol differential sphere decoding (MSDD), orthogonal frequency-division multiplexing (OFDM), space–time shift keying (STSK).

5 I. INTRODUCTION

6 Space–time shift keying (STSK) [1]–[3] has emerged as a beneficial multiple-input–multiple-output (MIMO) concept. STSK bridges the gap between the flexible diversity–multiplexing tradeoff provided by linear dispersion codes (LDCs) [4], [5] and the low-complexity design of spatial modulation (SM) [6]. Similar to the LDCs, STSK spreads the user information to both the spatial and time dimensions, but instead of simultaneously activating all the dispersion matrices (DMs), it transmits an additional $\log_2 Q$ bits by activating one out of $Q$ DMs. To overcome the performance degradation of the STSK scheme in wideband channels, orthogonal frequency-division multiplexing (OFDM)–aided STSK [7] and orthogonal frequency-division multiplexing–access/single-carrier frequency-division multiple-access–aided STSK (OFDM–aided STSK) [8] have also been proposed. The previous STSK studies [1], [2] demonstrate that coherent STSK performs well in conjunction with perfect channel state information (CSI) but exhibits a severe error floor in the presence of channel estimation (CE) errors.

7 The main contributions of this paper are as follows.

8 1) A novel soft-decision-aided MSDSD is proposed for OFDM–aided DSTSK operating in dispersive channels. The decision rule of the soft-decision MSDSD is deduced by considering the construction of DSTSK codewords based on the DMs, the Doppler frequency, the OFDM system parameters, and the generation of soft information.

9 2) A lower bound of the detection complexity is deduced, which is verified by simulations.

10 The remainder of this paper is organized as follows. In Section II, an overview of the proposed channel-coded OFDM–aided DSTSK scheme is provided. The soft-decision MSDSD is modeled in Section III. In Section IV, both the complexity imposed by the system model and the performance of the soft-decision-MSDSD–aided DSTSK scheme is investigated in Section V. Finally, we conclude in Section VI.

11 Notations: We use capital boldface letters to denote matrices, whereas $\{\cdot\}^T$, $\{\cdot\}^H$, $\text{tr}(\cdot)$, $\text{det}(\cdot)$, and $\|\cdot\|$ are used to represent the transpose, the Hermitian transpose, the trace, the determinant, and the Euclidean norm of the matrix $\{\cdot\}$, respectively. The notations $\mathbb{E}\{\cdot\}$ and $\mathbb{P}\{\cdot\}$ are used to denote the expected value and the probability of $\{\cdot\}$, respectively, whereas $\otimes$ and $\mathbb{I}_F$ represent the Kronecker product and the identity matrix, respectively. A symmetric $(N_{\text{c}} \times N_{\text{c}})$ Toeplitz matrix is denoted as $\text{toeplitz}(x_1, \ldots, x_{N_{\text{c}}})$, whereas $\text{diag}(X_1, \ldots, X_{N_{\text{c}}})$ indicates a block-diagonal matrix with the matrices $X_1, \ldots, X_{N_{\text{c}}}$ on its main diagonal. Furthermore, $C_N(\mu, \sigma^2)$ refers to the circularly symmetric Gaussian distribution with mean $\mu$ and variance $\sigma^2$.
We consider a channel-coded OFDM-aided DSTSK transceiver employing \( M \) transmit and \( N \) receive antenna elements (AEs), as shown in Fig. 1. The channel encoder/decoder blocks of Fig. 1 may incorporate a general channel coding scheme that supports soft-decision decoding at affordable complexity. A pragmatic coding architecture might be an appropriately interleaved serially concatenated recursive convolutional code (RSC) and unity rate code (URC)-aided scheme [17]–[19], as shown in Fig. 1.

The source bits are first channel encoded by the RSC encoder, before finally outputting the estimated bits [8], [19]. The RSC decoder, before finally outputting the estimated bits [8], [19].

The signal received is first OFDM demodulated and then input to the DSTSK demapper, the URC decoder, and the DSTSK soft-decision MSDSD demapper. The extrinsic soft information is then iteratively exchanged between the three soft-in–soft-out components, namely, the DSTSK demapper, the URC decoder, and the RSC decoder, before finally outputting the estimated bits [8], [19].

We observe that the STSK system is then uniquely and unambiguously described by the parameters \((M, N, T, Q, \mathcal{L})\). We may represent the codeword \( \mathbf{S}[i] \) belongs to a set \( \mathcal{S} \) of \((\mathcal{L} \cdot Q)\) codeword matrices defined by

\[
\mathcal{S} \triangleq \{ \mu \mathbf{A}_q \mid q \in \{1, \ldots, Q\}, \; l \in \{1, \ldots, \mathcal{L}\} \}.
\]

The STSK codewords are mapped to \( N_c \) parallel OFDM subcarriers, as shown in Fig. 2, before being differentially encoded. As shown in Fig. 2, \( N_c \) consecutive STSK codewords are arranged in parallel to form an OFDM-STSK frame, and OFDM modulation is carried out over each shaded symbol pipe, which constitutes an OFDM symbol. We may represent the codeword \( \mathbf{S}[i] \) by \( \mathbf{S}[n_c, k] \), so that the overall 144 codeword index is related to the OFDM frame index \( k \) and the \( \mathcal{L} \) subcarrier index \( n_c \) by \( i = kN_c + n_c, \; n_c = 0, 1, \ldots, (N_c - 1) \). Ad-ditionally, we invoke differential encoding in the TD, i.e., differential 147 encoding is performed across the consecutive OFDM symbols of the 148 same subcarrier. We have chosen TD differential encoding/decoding because we have conceived our scheme for continuous transmissions, 150 as opposed to the FD differential encoding/decoding across adjacent 151 FD subcarriers, which is more suitable for burst transmissions. To fa- 152 dilitate convenient differential encoding, we assume \( M = T \). Further- 153 more, directly generated unitary DMs are used in the proposed scheme 154 for avoiding the nonlinear Cayley transform [2], [12]. The codewords 155
The DSTSK codewords are then transmitted after the \( N_c \)-point inverse discrete Fourier transform (DFT) operation and appropriate CP insertion.

161 \textbf{B. Channel Model}

Each link between the \( r \)-th transmit and \( r \)-th receive AE is assumed to be a frequency-selective channel, but as a benefit of OFDM-based transmission, each dispersive channel is then partitioned into \( N_c \) low-rate parallel frequency-flat subchannels. The complex-valued fading gain \( h_{m,n}[c,k] \) obeys the distribution of eq. (173) with an autocorrelation function based on Clarke’s model [23]:

\[
\varphi_{hh}[n,c,k] = \mathcal{E} \{ h[n,n+c,k] h[n,n+c,k+16] \} = J_0(2\pi f_d) \varphi_{0}\text{,}
\]

where \( f_d \) is the zeroth-order Bessel function of the first kind, and \( J_0(f_d) \) is the normalized maximum Doppler frequency, whereas \( f_d \) and \( 1/T \) represent the maximum Doppler frequency and the symbol rate, respectively. The fading is assumed quasi-static, i.e., the channel’s complex-valued envelope remains approximately constant during the transmission of an OFDM STSK frame.

Given the aforementioned assumptions, the received signal \( Y[n_c,k] \in \mathbb{C}^{T \times N} \) obtained after CP removal and DFT may be expressed as [16], [22]:

\[
Y[n_c,k] = X[n_c,k] H[n_c,k] + V[n_c,k] + W[n_c,k]\]

where \( X[n_c,k] \in \mathbb{C}^{T \times M} \) represents the codeword transmitted and \( H[n_c,k] \in \mathbb{C}^{M \times N} \) denotes the FD channel transfer matrix, with its \( (m,n) \)-th entry given by \( h_{m,n}[c,k] \). Furthermore, \( V[n_c,k] \in \mathbb{C}^{T \times N} \) is the additive white Gaussian noise (AWGN) with entries of \( \mathcal{N}(0, \sigma^2_n) \).

III. MSDSD RECEIVER

This section introduces the maximum-likelihood MSDSD (ML-MSDSD), the maximum \textit{a posteriori} MSDSD (MAP-MSDSD) algorithm, and the generation of the log-likelihood ratios (LLRs) for the soft-decision-MSDSD-aided OFDM DSTSK.

184 A. ML-MSDSD for OFDM-Aided DSTSK

The ML-MSDSD processes \( N_w \) consecutively received space–time blocks corresponding to the \( n \)-th subcarrier given by 

\[
Y[n_c,k] = X[n_c,k] H[n_c,k] + V[n_c,k] + W[n_c,k] \tag{4}
\]

where

\[
Y[n_c,k] \in \mathbb{C}^{N_w \times M \times N}, \quad V[n_c,k] \in \mathbb{C}^{N_w \times M \times N}, \quad H[n_c,k] \in \mathbb{C}^{N_w \times M \times N}, \quad X_D[n_c,k] \in \mathbb{C}^{N_w \times M \times N \times M}.\]

For the sake of notational simplicity, we omit the subcarrier index and time index \( [n_c,k] \) in the following and refer to the \( \mu \)-th submatrix of a 207 block matrix, e.g., \( B \) by the subscripted matrix \( B_\mu \). Under the assumption that \( h_{m,n} \) and \( v_{\mu,v} \) obey zero-mean Gaussian random processes, the probability density 210 function of \( Y \) conditioned on \( X_D \) is given by [25], [26]:

\[
P(Y|X_D) = \frac{1}{(\pi N_w M \det |A_\mu|)^{N/2}} \exp \left\{ - \text{tr} \left( Y^T A_\mu^{-1} Y \right) \right\} \tag{6}
\]

where \( A_\mu \) is defined by \( A_\mu = (\psi_{hh} + \sigma^2_n I_{N_w}) \), with the component autocorrelation functions \( \psi_{hh} \) being identical for all spatial channels. Applying 219 the Cholesky factorization of \( A_\mu^{-1} = LL^T \) with the upper triangular 221 matrix \( L \) and considering the identity 

\[
\text{tr}(AX^H) = \|X\|^2 \quad \text{for any vector } X, \tag{7}
\]

\[\hat{X} = \arg \max \left( \sum_{\nu=1}^{N_w} \left| Y_{\mu,\nu}^T \hat{X} + \sum_{\nu'=1}^{N_w} (Y_{\mu,\nu'}^T \hat{X})^2 \right| \right) \tag{9}
\]

where \( Y_{\mu,\nu} \) is defined by \( Y_{\mu,\nu} = Y_{\mu,\nu} u_{\mu,\nu} \), and \( u_{\mu,\nu} \) represents the \( \mu \)-th channel element of \( U \). Still referring to (9), \( \hat{X} \) denotes the estimate of \( X_D \), whereas \( \hat{X}_D \) is the \( \mu \)-th candidate submatrix of \( X_D \).

Since the ML metric of (9) is invariant to a phase shift common to all 226 elements of \( \hat{X}_D \) \( \forall \mu \) corresponding to the same \( S \) (where \( S_{\hat{X}} \in \mathbb{S}^N \)), the 227 accumulated differential matrices may be expressed as

\[
\mathbf{A}_\mu = \left\{ \prod_{\nu=1}^{N_w} S_{\hat{X}^H}^H, \quad 1 \leq \nu \leq (N_w-1) \right\} \tag{10}
\]

For the sake of reducing the complexity associated with an exhaustive search, we employ MSDSD similar to [10] and [11] to search through 231 the candidate set lying within a sphere of radius \( \rho_\mu \) as follows:

\[
\sum_{\nu=1}^{N_w} \left| Y_{\mu,\nu}^T \mathbf{A}_\mu + \sum_{\nu'=1}^{N_w} (Y_{\mu,\nu'}^T \mathbf{A}_\nu) \right|^\nu \leq \rho_\mu^{N_w} \tag{11}
\]

B. MAP-MSDSD

Assuming the STSK codewords to be mutually independent, (6) and (9) yield [15]:

\[
- \ln \left( P(S) \right) - \ln \left( P(Y|S) \right) = - \ln \left( P(S) \right)
\]

\[
\sum_{\mu=1}^{N_w} \left| Y_{\mu,\nu}^T \mathbf{A}_\mu + \sum_{\nu'=1}^{N_w} (Y_{\mu,\nu'}^T \mathbf{A}_\nu) \right|^\nu \leq \ln \left( P(S) \right) \tag{12}
\]
The MAP-MSDSD may be thus expressed as
\[
\sum_{\mu=1}^{(N_w-1)} \left( \left\| \sum_{\nu=1}^{N_w} (Y^H_{\mu,\nu} A_\nu) \right\|^2 - \ln \left( P(S_\mu) \right) \right) \leq \rho^2 - \|u_{N_w,n} Y_{N_w}\|^2 \overset{\mathrm{def}}{=} \rho^2. \tag{13}
\]

Clearly, the codeword \( S_\mu \) obeys the specific distance criterion \([10], [15]\), that the current partial Euclidean distance (PED) \( d^2_{\mu} \) is the 239 sum of the previous PED \( d^2_{\mu+1} \) and the distance increment \( \Delta^2 \), i.e.,
\[
d^2_\mu \overset{\approx}{=} \Delta^2 + d^2_{\mu+1} = \left\| u_{\mu,b} Y_{\mu,b} A_{\mu+1} \right\|^2 + \sum_{\nu=1}^{N_w} u_{\mu,\nu} Y_{\nu,b} A_{\mu+1} - \ln \left( P(S_\mu) \right)
\]
\[
+ \sum_{\nu=1}^{N_w} u_{\mu,\nu} Y_{\nu,b} A_\nu - \ln \left( P(S_\nu) \right) \leq \rho^2. \tag{14}
\]

Similar to the MAP-MSDSD principle described in \([10], [15]\), the 240 MAP-MSDSD is initialized with \( \mu = (N_w - 1) \) and then proceeds by 241 applying the search criterion of (14) until \( \mu = 1 \), where the search ra-
dius is updated to \( \rho^2 = d^2_{\mu} \), and the search is repeated by commencing 244 from \( \mu = 2 \) until \( \mu = (N_w - 1) \) is reached. If the new search does not 245 provide a better estimate, the previous estimate is retained.

C. Log-Likelihood Ratio and Soft-Decision-MSDSD-Aided 248 OFDM DSTSK

The extrinsic LLR \( L_e(\cdot) \) for \( b_j \) is now evaluated by combining the 259 a posteriori and a priori LLR: \( L_e(b_j) = L_{ap}(b_j) - L_a(b_j) \). The ex-
trinsic information extracted from the soft-decision MSDSD demapper 261 is iteratively exchanged with the URC decoder of Fig. 1, which forms the 262 inner iteration, whereas the exchange of extrinsic information 263 between the URC decoder and the RSC decoder of Fig. 1 may be 264 termed as the outer iteration. Note that, for each outer iteration be-
deen the RSC decoder and the URC decoder, several inner iterations 266 may be invoked between the URC and the soft-decision-MSDSD- 267 aided DSTSK demapper \([2], [19]\). Finally, the RSC decoder generates 268 a posteriori LLRs, from which the source bits are estimated.

IV. COMPLEXITY

Here, the complexity of the proposed scheme is detailed, and the 271 complexity imposed by the MAP-MSDSD is quantified.

Equation (2) shows that there exists \( (L \cdot Q) \) legitimate code-
word matrices for each \( \log_2(Q) \) bits of source information. The 274 exhaustive-search-based solution to (9) involves a search in a \( (L \cdot 275 Q) (N_w - 1) \) element space of candidate matrices \( X \) corresponding to all 276 possible choices of \( S \). The ML-MSDSD associated with chosen sphere 277 radius \( \rho \) imposes average complexity, which is lower bounded by \([11]\)

\[
C \geq \frac{(L \cdot Q)^{N_w - 1} - (L \cdot Q)}{(L \cdot Q) - 1} \tag{16}
\]

where

\[
\zeta \overset{\Delta}{=} \frac{\sigma^2 (1 + \epsilon)}{2 (1 + \sigma^2)} = (1 + \epsilon) MN w, \quad \epsilon > 0. \tag{17}
\]

To quantify the complexity of the MAP-MSDSD scheme, we con-
side the number of real-valued multiplication operations (RMOs) 281 required for obtaining a single soft-output value, which is used as 282 our complexity metric. The lower bound of the complexity may be 283 obtained if the number of RMOs required for computing the soft 284 outputs corresponding to the first codeword estimate \( \hat{S} \) is counted 285 and if a single constrained estimate \( \hat{S} \) is taken into account \([15]\). 286 Considering the upper diagonal nature of the matrix \( U \), we observe 287 that \( Y^H_{\mu,\nu} \) is defined only for \( \nu \geq \mu \) in the context of (9), although each 288 \( Y^H_{\mu,\nu} \) terms in (15) thus involves a total of \( 2 M N + (2 + 290 299) N_w \) RMOs, assuming real-valued autocorrela-
tion functions of \( \varphi_{\nu h} [n_c, \kappa] \). To compute the \( a \) posteriori LLRs given 292 by (15), the number of RMOs associated with the computation of each 293 \( \sum_{\nu=1}^{N_w} (Y^H_{\mu,\nu} A_\nu)^2 \) is \( 4 M^2 N (N_w - \mu + 1) + 2 \). The number of 294 RMOs required for generating \( \log_2(Q) \) soft outputs corresponding 295 to a single codeword estimate \( \hat{S} \) is thus given by

\[
\text{RMO}[\hat{S}] = MN w (N_w + 1) + \sum_{\nu=1}^{N_w} [4 M^2 N (N_w - \mu + 1) + 2] \tag{18}
\]

\[
= MN (2 M + 1) N_w (N_w + 1) + 2 N w.
\]

On the other hand, the number of RMOs related to each bit \( \hat{b} \) of 297 the constrained estimate \( \hat{S} \) is found to be \( 2 N w M^2 F (N_w + 1) + 1 \). 298 The lower bound for the number of RMOs associated with the gen-
299 eration of a single soft output is thus \( MN (2 M + 1) / \log_2(Q) \cdot 300 + 300 2 M^2 N w^2 \) for large \( N_w \). The complexity of the scheme, however, 301 depends on a number of parameters, such as on the channel SNR, on 302 the autocorrelation function of the channel’s fading plus noise, and 303 most importantly, on the \( a \) priori mutual information \( I_A \) of the inner 304 decoder \([15]\). In Section V, the complexity of the MAP-MSDSD will 305 be investigated as a function of the observation window width \( N_w \) 306 parameterized by the available \( a \) priori information \( I_A \).
TABLE 1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
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<tbody>
<tr>
<td>Dispersive channel model</td>
<td>COST207-TU12</td>
</tr>
<tr>
<td>Fast fading envelope</td>
<td>Correlated Rayleigh fading</td>
</tr>
<tr>
<td>Normalized Doppler spread, ( f_d )</td>
<td>0.01</td>
</tr>
<tr>
<td>Number of subcarriers, ( N_c )</td>
<td>128</td>
</tr>
<tr>
<td>Length of cyclic prefix</td>
<td>32</td>
</tr>
<tr>
<td>Overall symbol duration</td>
<td>300 ns</td>
</tr>
<tr>
<td>STSK (( M, N, T, Q, \xi ))</td>
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</tr>
<tr>
<td>RSC encoder and decoder</td>
<td>Half rate</td>
</tr>
<tr>
<td>Constraint length=2</td>
<td></td>
</tr>
<tr>
<td>Generator polynomial</td>
<td>(011, 010)_2</td>
</tr>
<tr>
<td>Length of interleavers</td>
<td>200,000 bits</td>
</tr>
<tr>
<td>Outer iterations</td>
<td>11</td>
</tr>
<tr>
<td>Inner iterations</td>
<td>2</td>
</tr>
</tbody>
</table>

![Fig. 3](image-url) Simulated BER performance of the soft-decision-MSDSD-aided OFDM DSTSK (2, 2, 2, 4, 4) scheme for transmission over dispersive COST207-TU12 channel with normalized Doppler frequency \( f_d = 0.01 \) and different observation window sizes \( N_w = 2, 4, 6, 10 \). The BER falls sharply after \( I_{\text{outer}} = 9 \) outer iterations as a benefit of employing the URC and the performance approaches that of the coherent scheme with perfect CSI with an increasing value of \( N_w \).

**V. PERFORMANCE RESULTS**

Here, the performance of the proposed scheme is investigated using the parameters listed in Table I. We have employed the COST207-TU12 channel model for the links between each transmit–receive antenna pair. The power delay profile characterizing the 12 taps of the COST207-TU12 channel is detailed in [30] and [31]. As mentioned in Table I, we employ an RSC (2, 1, 2) outer code having octally represented generator polynomials of \( (g_r, g) = (3, 2)_8 \) as well as two random interleavers with a length of 200,000 bits.

Fig. 3 characterizes the achievable bit error rate (BER) of the proposed soft-decision-MSDSD-aided OFDM DSTSK scheme associated with observation window sizes of \( N_w = 2, 4, 6, 10 \) and compares to that achievable by the corresponding coherent scheme relying on perfect CSI. We observe that the proposed scheme has the benefit of dispensing with CE due to differential encoding, while mitigating the performance erosion of classic STSK by employing OFDM. Further, the multiple-symbol detection partially mitigates the inherent performance penalty imposed by noncoherent detection. We observe in Fig. 3 that, as \( N_w \) increases, the BER performance gradually approaches that of the perfect CSI-oriented coherent scheme. Note that all the performance characteristics exhibit a vanishingly low BER after \( I_{\text{outer}} = 9 \) outer iterations, which is confirmed by the decoding trajectory for \( N_w = 6 \). The EXIT charts with \( N_w = 2, 4 \) are, however, “pinched off”; thus, the BER at this SNR do not converge.

Fig. 4. EXIT charts of the inner decoders of the soft-decision-MSDSD-aided OFDM DSTSK (2, 2, 2, 4, 4) scheme at \( SNR = 1 \) dB with normalized Doppler frequency \( f_d = 0.01 \) and different observation window sizes \( N_w = 2, 4, 6, 10 \) and of the corresponding coherent system having perfect CSI at the receiver. At this SNR, we observe the inner EXIT charts with \( N_w = 6, 10 \) have an open EXIT tunnel and converge to the (1.0, 1.0) point of perfect convergence, indicating a sharp fall in the BER curve after \( I_{\text{outer}} = 9 \) outer iterations, which is confirmed by the decoding trajectory for \( N_w = 6 \). The EXIT charts with \( N_w = 2, 4 \) are, however, “pinched off”; thus, the BER at this SNR do not converge.

More, the multiple-symbol detection partially mitigates the inherent performance penalty imposed by noncoherent detection. We observe in Fig. 3 that, as \( N_w \) increases, the BER performance gradually approaches that of the perfect CSI-oriented coherent scheme. Note that all the performance characteristics exhibit a vanishingly low BER after \( I_{\text{outer}} = 9 \) outer iterations, which is the explicit benefit of using the URC in the system. The URC is a low-complexity code, which has an infinite impulse response and hence assists the inner decoder in efficiently spreading the soft information [2], [19]. As a result, the extrinsic information transfer (EXIT) charts of Figs. 4 and 5 converge to the (1.0, 1.0) point of perfect decoding convergence, leading to a vanishingly low BER, thus eliminating the system’s error floor. The maximum achievable rates for the corresponding scheme, where the 336 scheme still exhibits an infinitesimally low BER were computed by exploiting the area property of EXIT charts [18], [32], [33] and are shown in Fig. 3 as the ultimate benchmark of the scheme.

To elaborate further, Figs. 4 and 5 portray the EXIT charts of our 340 proposed scheme at \( SNR = 1 \) and 4 dB, respectively. We observe in Fig. 4 that the inner decoder’s EXIT charts recorded at \( SNR = 1 \) dB for \( N_w = 2, 4 \) are “pinched off,” i.e., there remains no “open” EXIT 341 tunnel, indicating a high residual BER. By contrast, the BER associated with \( N_w = 6, 10 \) may be expected to decrease sharply at this SNR 345 after \( I_{\text{outer}} = 9 \) outer iterations, which is confirmed by the staircase- 346 shaped Monte Carlo-simulation-based decoding trajectory [8], [13]. Fig. 5, on the other hand, shows the EXIT charts at \( SNR = 4 \) dB, 348 where all the curves associated with \( N_w = 2, 4, 6, 10 \) exhibit an open 349 EXIT tunnel, implying an infinitesimally low BER after \( I_{\text{outer}} = 9 \) 350 iterations. The EXIT charts of the soft-decision-MSDSD-aided OFDM 351 DSTSK recorded both for \( SNR = 1 \) dB and \( SNR = 4 \) dB are further 352 compared in Figs. 4 and 5 to the ultimate benchmark of the coherent 353 detector assuming perfect CSI at the receiver.

Fig. 6 characterizes the complexity associated with the MAP- 355 MSDSD of the OFDM-aided DSTSK (2, 2, 2, 4, 4) scheme at \( SNR = 356 \\ 4 \) dB as a function of the window size \( N_w \), parameterized by the 357...
We have proposed a soft-decision-MSDSD-aided multicarrier DSTSK scheme for communications over wideband channels. The 374 OFDM-aided DSTSK provides a flexible diversity versus multiplexing gain tradeoff by spreading the source information across both the spatial and time dimensions, while mitigating the potential performance degradation imposed by the frequency selectivity of the channel. The 378 turbo-principle-based soft-decision MSDSD facilitates joint decisions 379 over a number of DSTSK codewords, while exploiting the fading- 380 plus-noise statistics of the channel. We have demonstrated that the 381 proposed soft-decision-MSDSD-aided DSTSK scheme provides sub- 382 stantial flexibility at moderate complexity owing to dispensing with 383 CE. Furthermore, the MSDSD mitigates the performance degradation 384 inflicted by the CDD scheme without an undue increase in computa- 385 tional complexity.

VI. Conclusion

We have proposed a soft-decision-MSDSD-aided multicarrier 373 DSTSK scheme for communications over wideband channels. The 374 OFDM-aided DSTSK provides a flexible diversity versus multiplexing gain tradeoff by spreading the source information across both the spatial and time dimensions, while mitigating the potential performance degradation imposed by the frequency selectivity of the channel. The 378 turbo-principle-based soft-decision MSDSD facilitates joint decisions 379 over a number of DSTSK codewords, while exploiting the fading- 380 plus-noise statistics of the channel. We have demonstrated that the 381 proposed soft-decision-MSDSD-aided DSTSK scheme provides sub- 382 stantial flexibility at moderate complexity owing to dispensing with 383 CE. Furthermore, the MSDSD mitigates the performance degradation 384 inflicted by the CDD scheme without an undue increase in computa- 385 tional complexity.

The BER curve after (1.0, 1.0) point as a benefit of employing the URC, indicating a sharp fall in
EXIT charts have a quite open EXIT tunnel at this SNR and converge to the
a priori decoding trajectory. As expected, the complexity rapidly escalates upon increasing
of RMOs in Section IV is also shown as a benchmarker in Fig. 6.368

Fig. 5. EXIT charts of the inner decoders of the soft-decision-MSDSD-aided
DSTSK (2, 2, 4, 4) scheme at SNR = 4 dB with normalized Doppler
frequency $f_d = 0.01$ and different observation window sizes $N_w = 2, 4, 6, 10$ and that of the corresponding coherent inner decoder as a benchmark. All the
EXIT charts have a quite open EXIT tunnel at this SNR and converge to the
(1,0,1,0) point as a benefit of employing the URC, indicating a sharp fall in the BER curve after $I_{outer} = 9$ outer iterations, which is confirmed by the decoding trajectory.

Fig. 6. Complexity in terms of the numbers of RMOs for the proposed
DSTSK (2, 2, 4, 4) scheme at SNR = 4 dB using the parameters of Table I as
a function of observation window size $N_w$ parameterized against the a priori
information $I_A$. The complexity shoots up with $N_w > 6$, although the rate of increase in complexity slows down with increased
a priori information.

As expected, the complexity rapidly escalates upon increasing $N_w$.

The complexity is parameterized against the
observation window size $N_w$.

<table>
<thead>
<tr>
<th>$I_A$</th>
<th>Lower bound</th>
<th>Complexity (no. of RMOs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2</td>
<td>10^3</td>
<td>10^4</td>
</tr>
<tr>
<td>0.4</td>
<td>10^5</td>
<td>10^6</td>
</tr>
<tr>
<td>0.6</td>
<td>10^7</td>
<td>10^8</td>
</tr>
<tr>
<td>0.8</td>
<td>10^9</td>
<td>10^10</td>
</tr>
<tr>
<td>1.0</td>
<td>10^11</td>
<td>10^12</td>
</tr>
</tbody>
</table>

$N_w$ may be adaptively 364 selected depending on the quality of the soft input. To be specific, 365 the value of $I_A$ increases during the consecutive decoding iterations, 366 and we can flexibly increase $N_w$ when the value of $I_A$ is higher. The 367 theoretical lower bound of the complexity quantified by the number 368 of RMOs in Section IV is also shown as a benchmark in Fig. 6. 369

A priori information $I_A$ provided by the outer decoder for the demap-
per of Fig. 1. The a priori information $I_A$ is measured by the average 360 mutual information [13] between the a priori LLL $L_{apriori}(b_j)$ and the 361 a posteriori LLL $L_{aposteriori}(b_j)$ of Fig. 1. The influence of the a priori 362 information $I_A$ on the complexity may be beneficially exploited in the 363 context of adaptive system design [15], where $N_m$ may be adaptively 364 selected depending on the quality of the soft input. To be specific, 365 the value of $I_A$ increases during the consecutive decoding iterations, 366 and we can flexibly increase $N_w$ when the value of $I_A$ is higher. The 367 theoretical lower bound of the complexity quantified by the number 368 of RMOs in Section IV is also shown as a benchmark in Fig. 6.

As expected, the complexity rapidly escalates upon increasing $N_w$,


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OFDM-Aided Differential Space–Time Shift Keying
Using Iterative Soft-Multiple-Symbol Decoding

Mohammad Ismat Kadir, Sheng Chen, K. V. S. Hari, K. Giridhar, and Lajos Hanzo

Abstract—Soft-decision multiple-symbol differential sphere decoding (MSDD) is proposed for orthogonal frequency-division multiplexing (OFDM)-aided differential space–time shift keying (DSTSK)-aided transmission over frequency-selective channels. Specifically, the DSTSK signaling blocks are generated by the channel-encoded source information and the space–time (ST) blocks are appropriately mapped to a number of OFDM subcarriers. After OFDM demodulation, the DSTSK signal is noncoherently detected by our soft-decision MSDD detector. A novel soft-decision MSDD detector is designed, and the associated decision rule is derived for the DSTSK scheme. Our simulation results demonstrate that an SNR reduction of 2 dB is achieved by the proposed scheme using an over the conventional soft-decision-aided differential detection benchmark, while communicating over dispersive channels and dispensing with channel estimation (CE).

Index Terms—Extrinsic information transfer (EXIT) chart, iterative decoding, multiple-symbol differential sphere decoding (MSDD), orthogonal frequency-division multiplexing (OFDM), space–time shift keying (STSK).

I. INTRODUCTION

Space–time shift keying (STSK) [1]–[3] has emerged as a beneficial multiple-input–multiple-output (MIMO) concept. STSK bridges the gap between the flexible diversity–multiplexing tradeoff provided by linear dispersion codes (LDCs) [4], [5] and the low-complexity design of spatial modulation (SM) [6]. Similar to the LDCs, STSK spreads the user information to both the spatial and time dimensions, but instead of simultaneously activating all the dispersion matrices (DMs), it transmits an additional log2 Q bits by activating one out of Q DMs. To overcome the performance degradation of the STSK scheme in wideband channels, orthogonal frequency-division multiplexing (OFDM)-aided STSK [7] and orthogonal frequency-division multiplexing–single-carrier frequency-division multiple-access-aided STSK (STSK) [8] have also been proposed. The previous STSK studies [1], [2] demonstrate that coherent STSK performs well in conjunction with perfect channel state information (CSI) but exhibits a severe error floor in the presence of channel estimation (CE) errors.

Differential STSK (DSTSK) employing conventional differential 41 detection (CDD) has also been proposed for the sake of dispensing 42 with the CE [2] and thus to eliminate the potentially high-Doppler- 43 dependent pilot overhead. However, CDD suffers from a typical 3-dB 44 performance penalty in low-Doppler scenarios. Furthermore, an ir- 45 reducible error floor may be observed in a high-mobility scenario 46 characterized by a high Doppler frequency. To circumvent the per- 47 formance degradation of CDD, multiple-symbol differential detection 48 (MSDD) was proposed for differential phase-shift keying (DPSK) in 49 [9]. MSDD uses the fading-plus-noise statistics of the channel for 50 jointly detecting information symbols from Nw number number 51 of consecutively received symbols, where Nw is usually referred to 52 as the observation window size. The performance improvement of 53 MSDD is, however, achieved at the cost of increased complexity, 54 which increases exponentially with Nw. For mitigating this potent- 55 tially excessive complexity, sphere decoding (SD) was invoked 56 for MSDD in the context of multiple-symbol differential sphere decoding 57 (MSDD) in [10] and [11]. Hard-decision MSDD was conceived 58 in [12] for a DSTSK scheme operating in nondispersive channels. 59 As a further advance, inspired by the near-capacity performance of 60 turbo detection [13], [14], a soft-decision MSDD scheme was also 61 designed for DPSK in [15]. Furthermore, the concept of differential 62 space–frequency modulation employing MSDD in conjunction with 63 a specific subcarrier allocation was proposed in [16] for exploiting both 64 the achievable spatial- and frequency-domain diversity. However, the 65 conception of the soft-decision-MSDD-aided DSTSK designed for 66 realistic dispersive scenarios constitutes an unexplored open problem. 67 Against this background, we conceive a novel soft-decision MSDD 68 for OFDM-based DSTSK operating in frequency-selective channels. 69 The main contributions of this paper are as follows.

1) A novel soft-decision-aided MSDD is proposed for OFDM- 70 aided DSTSK operating in dispersive channels. The decision rule of the soft-decision MSDD is deduced by considering the construction of DSTSK codewords based on the DMs, the Doppler frequency, the OFDM system parameters, and the generation of soft information.

2) A lower bound of the detection complexity is deduced, which is verified by simulations.

The remainder of this paper is organized as follows. In Section II, an overview of the proposed channel-coded OFDM-aided DSTSK scheme is provided. The soft-decision MSDD is modeled in Section III. In Section IV, both the complexity imposed by the system is quantified. The performance of the soft-decision-MSDD-aided DSTSK scheme is investigated in Section V. Finally, we conclude in Section VI.

Notations: We use capital boldface letters to denote matrices, whereas \( \{ \}^T \), \( \{ \}^H \), \( \text{tr} \{ \} \), \( \text{det} \{ \} \), and \( \| \cdot \| \) are used to represent the 87 transpose, the Hermitian transpose, the trace, the determinant, and the Euclidean norm of the matrix \( \{ \} \). Respectively, the notations \( \mathcal{E} \{ \} \), \( \mathcal{P} \{ \} \) are used to denote the expected value, the complex conjugate, and the probability of \( \{ \} \) respectively, whereas \( \otimes \) and \( \mathcal{I}_r \) represent the Kronecker product and the (\( \times \) \) \( \times \)) identity matrix, respectively. A symmetric \( (N_w \times N_w) \) Toeplitz matrix is denoted as \( \text{toeplitz} \{ x_1, \ldots, x_{N_w} \} \), whereas \( \text{diag} \{ X_1, \ldots, X_{N_w} \} \) indicates a block-diagonal matrix with the matrices \( X_1, \ldots, X_{N_w} \) on its main 95 diagonal. Furthermore, \( \mathcal{C}(\mu, \sigma^2) \) refers to the circularly symmetric 96 complex Gaussian distribution with mean \( \mu \) and variance \( \sigma^2 \).
Fig. 1. Transceiver architecture of the proposed concatenated channel-coding-aided DSTSK scheme relying on the soft-decision MSDSD as the inner decoder.

98 II. System Model

99 We consider a channel-coded OFDM-aided DSTSK transceiver employing $M$ transmit and $N$ receive antenna elements (AEs), as shown in Fig. 1. The channel encoder/decoder blocks of Fig. 1 may incorporate a general channel coding scheme that supports soft-decision decoding at affordable complexity. A pragmatic coding architecture might be an appropriately interleaved serially concatenated recursive convolutional code (RSC) and unitary rate code (URC)-aided scheme [17]–[19], as shown in Fig. 1.

100 The source bits are first channel encoded by the RSC code, and the encoded bits are then interleaved by a random bit interleaver $\Pi_1$. Following URC precoding, the interleaved bits are further interleaved by a second interleaver $\Pi_2$. The resultant bits are then mapped to DSTSK codewords, which are further mapped to $N_c$ parallel subcarriers and then differentially encoded in the time domain (TD), i.e., across the consecutive OFDM symbols of the same subcarrier. The DSTSK codewords are then OFDM modulated, while incorporating appropriately cyclic prefixes (CPs).

101 The signal received is first OFDM demodulated and then input to the DSTSK soft-decision MSDSD demapper. The extrinsic soft information is then iteratively exchanged between the three soft-in–soft-out components, namely, the DSTSK demapper, the URC decoder, and the RSC decoder, before finally outputting the estimated bits [8], [19].

102 A. DSTSK Architecture and OFDM Layout

103 The STSK encoder generates space–time (ST) codewords from the source information by activating a single DM in any symbol duration. The source bits are first channel encoded by the RSC code, and the encoded bits are then interleaved by the pairwise error probability of the codewords [5], [12], [20], [21] under the power constraint in [2] expressed by $\text{tr}(A^H_q A_q) = T \forall q$. The resultant STSK system is then uniquely and unambiguously described by the parameters $(M, N, T, Q, \mathcal{L})$.

104 We observe that the STSK codeword $S[i]$ belongs to a set $\mathcal{S}$ of $157$ $(L \cdot Q)$ codeword matrices defined by

$$\mathcal{S} \doteq \{ n_q A_q | (q \in \{1, \ldots, Q\}, \ l \in \{1, \ldots, L \} \} \,. \quad (2)$$

105 The STSK codewords are mapped to $N_c$ parallel subcarriers, as shown in Fig. 2, before being differentially encoded. As shown in Fig. 2, $N_c$ consecutive STSK codewords are arranged in parallel to form an OFDM-STSK frame, and OFDM modulation is carried out over each shaded symbol pipe, which constitutes an OFDM symbol. We may represent the codeword $S[i]$ by $S[n_q, k]$, so that the overall 144 codeword index $i$ is related to the OFDM frame index $k$ and the 145 subcarrier index $n_q$ by $i = kN_c + n_q$, $n_q = 0, 1, \ldots, (N_c - 1)$. Ad-146ditionally, we invoke differential encoding in the TD, i.e., differential 147 encoding is performed across the consecutive OFDM symbols of the 148 same subcarrier. We have chosen TD differential encoding/decoding because we have conceived our scheme for continuous transmissions, 149 and compared to the FD differential encoding/decoding across adjacent 150 FD subcarriers, which is more suitable for burst transmissions. To fa-152cilitate convenient differential encoding, we assume $M = T$. Further-153more, directly generated unitary DMs are used in the proposed scheme 154 for avoiding the nonlinear Cayley transform [2], [12]. The codewords 155
\[ S[n_c, k] \text{ are thus differentially encoded to form the transmit blocks } \]
\[ X[n_c, k] = \begin{cases} X[n_c, k - 1 | S[n_c, k], & k = 1, 2, \ldots, N_w \\ \emptyset, & k = 0. \end{cases} \]  
(3)

The DSTSK codewords are then transmitted after the \( N_c \)-point inverse discrete Fourier transform (DFT) operation and appropriate CP insertion.

\[ P(Y | X_D) = \frac{1}{(\pi N_w M) \det(\Lambda_Y)} \exp\left\{ -\text{tr} \left( Y^H \Lambda_Y^{-1} Y \right) \right\} \]  
(6)

where \( \Lambda_Y \) is defined by \( \Lambda_Y \triangleq E(Y Y^H) | X_D \). The ML estimate \( \hat{X} \) under the assumption of quasi-static fading and unitary \( X_D \) reduces to [26], [27]
\[ \hat{X} = \arg\max_x P(Y | X_D) = \arg\min_x \left\{ \text{tr} \left( Y^H \Lambda_Y^{-1} Y \right) \right\}. \]  
(7)

Here, the conditional covariance matrix \( \Lambda_Y \) is related to the channel parameters [25], [26] by
\[ \Lambda_Y^{-1} = \frac{1}{N} N \bar{D} (\Lambda - 1 \otimes I_M) \bar{X}_D. \]  
(8)

For the sake of notational simplicity, we omit the subcarrier index and time index \([n_c, k] \) in the following and refer to the \( \mu \)th submatrix of a 207 block matrix, e.g., \( B \) by the subscripts matrix \( B_\mu \). Under the assumption that \( h_{m,n} \) and \( \nu_{m,n} \) obey zero-mean Gaussian random processes, the probability density function of \( Y \) conditioned on \( X_D \) is given by [25], [26]

\[ Y[n_c, k] \in \mathbb{C}^{N_w M \times N_w} \quad V[n_c, k] \in \mathbb{C}^{N_w M \times N_w} \quad H[n_c, k] \in \mathbb{C}^{N_w M \times N_w} \quad X_D[n_c, k] \in \mathbb{C}^{N_w M \times N_w M}. \]

\[ Y[n_c, k] = X_D[n_c, k] H[n_c, k] + V[n_c, k] \]  
(5)

where

\[ Y, X_D, H, V \] are the accumulated differential matrices may be expressed as \[ P(Y | X_D) = P(Y | X_D) = P(Y | X_D) \]  
(6)

\[ \hat{X} = \arg\max_x \left\{ \sum_{\nu=1}^{N_w} \left( Y_{\mu, \nu} \bar{X}_\nu + \sum_{\nu'=1}^{N_w} (Y_{\mu, \nu'} \bar{X}_{\nu'}) \right)^2 \right\} \]  
(9)

\[ \hat{X} = \arg\max_x \left\{ \sum_{\nu=1}^{N_w} \left( Y_{\mu, \nu} \bar{X}_\nu + \sum_{\nu'=1}^{N_w} (Y_{\mu, \nu'} \bar{X}_{\nu'}) \right)^2 \right\} \]  
(9)

\[ \mathcal{A}_w = \left\{ \prod_{\nu=\nu_1}^{\nu_2} S_{\nu}^{H \prime}, \quad 1 \leq \nu \leq (N_w - 1) \right\} \]  
(10)

For the sake of reducing the complexity associated with an exhaustive 230 search, we employ MSDSD similar to [10] and [11] to search through the candidate set lying within a sphere of radius \( \rho_2 \) as follows:

\[ \sum_{\mu=1}^{N_w} \left( Y_{\mu, \nu} \bar{A}_\nu + \sum_{\nu'=1}^{N_w} (Y_{\mu, \nu'} \bar{A}_{\nu'}) \right)^2 \leq \rho_2^2. \]  
(11)

\[ \bullet \quad \text{MAP-MSDSD} \]

Assuming the STSK codewords to be mutually independent, (6) and (9) yield [15]:

\[ - \ln \left( P(S | Y) \right) \quad \alpha \ln \left( P(Y | S) \right) - \ln \left( P(S) \right) \]

\[ \alpha \ln \left( P(Y | S) \right) - \ln \left( P(S) \right) \]  
(12)
The MAP-MSDSD may be thus expressed as

$$\sum_{\mu=1}^{(N_w-1)} \left( \left\| \sum_{v=\mu}^{N_w} (Y^H_{\mu,v}A_v) \right\|^2 - \ln \left( P(S_\mu) \right) \right) \leq \rho^2 - \left\| u_{N_w,N_w} Y_{N_w} \right\|^2 \triangleq \rho^2 \cdot (13)$$

Clearly, the codeword \( S_\mu \) obeys the specific distance criterion [10], [15], the MAP-MSDSD is initialized with \( \mu = (N_w - 1) \) and proceeds by applying the search criterion of (14) until \( \mu = 1 \), where the search radius is updated to \( \rho^2 = d^2_{\mu+1} \), and the search is repeated by commencing from \( \mu = 2 \) until \( \mu = (N_w - 1) \) is reached. If the new search does not provide a better estimate, the previous estimate is retained.

The extrinsic LLR \( L_{\nu}(b_j) \) for \( b_j \) is now evaluated by combining the 259 \textit{a posteriori} and \textit{a priori} LLR: \( L_{\nu}(b_j) = L_p(b_j) - L_{\nu}(b_j) \). The ex-260 trinsic information extracted from the soft-decision MSDSD demapper 261 is iteratively exchanged with the URC decoder of Fig. 1, which forms 262 the \textit{iter} inner iteration, whereas the exchange of extrinsic information 263 between the URC decoder and the RSC decoder of Fig. 1 may be 264 termed as the \textit{outer} iteration. Note that, for each outer iteration be- 265 tween the RSC decoder and the URC decoder, several inner iterations 266 may be invoked between the URC and the soft-decision-MSDSD-aided DSTSK demapper [2], [19]. Finally, the RSC decoder generates 268 \textit{a posteriori} LLRs, from which the source bits are estimated.

IV. COMPLEXITY

Here, the complexity of the proposed scheme is detailed, and the 271 complexity imposed by the MAP-MSDSD is quantified.

Equation (2) shows that there exists \((L \cdot Q)\) legitimate code- 272 word matrices for each \( \log_2(L \cdot Q) \) bits of source information. The 274 exhaustive-search-based solution to (9) involves a search in a \( (L \cdot 275 Q)(N_w - 1) \) element space of candidate matrices \( X \) corresponding to all 276 possible choices of \( S \). The ML-MSDSD associated with chosen sphere 277 radius \( \rho \) imposes average complexity, which is lower bounded by [11] 278

$$C \geq \frac{(L \cdot Q)^{N_w-1} - (L \cdot Q)}{(L \cdot Q) - 1}$$

where

$$\zeta \triangleq \frac{(1 + \epsilon)^2}{2(1 + \epsilon^2)} \rho^2 = (1 + \epsilon)MN_{N_w}, \quad \epsilon > 0 \cdot (17)$$

To quantify the complexity of the MAP-MSDSD scheme, we con- 280 sider the number of real-valued multiplication operations (RMOs) 281 required for obtaining a single soft-output value, which is used as 282 our complexity metric. The lower bound of the complexity may be 283 obtained if the number of RMOs required for computing the soft 284 outputs corresponding to the first codeword estimate \( \hat{S} \) is counted 285 and if a single constrained estimate \( \hat{S}^b \) is taken into account [15], [16]. 286 Considering the upper diagonal nature of the matrix \( U \), we observe 287 that \( Y^H_{\mu,v} \) is defined only for \( v \geq \mu \) in the context of (9), although each \( \rho^2 \) \( Y^H_{\mu,v} \) is an \((N \times T)\)-element matrix, where \( T = M \). The computation 289 of the \( Y^H_{\mu,v} \) terms in (15) thus involves a total of \( 2MN(1 + 2 + \cdots + 290 N_w) = MN_{N_w}(N_w + 1) \) RMOs, assuming real-valued autocorrela- 291 tion functions of \( \varphi_{kk}[n, \kappa] \). To compute the \textit{a posteriori} LLRs given 292 by (15), the number of RMOs associated with the computation of each 293 \( \sum_{v=\mu}^{N_w} (Y^H_{\mu,v}A_v) \) is 4\( M^2N(N_w - \mu + 1) + 2 \). The number of 294 RMOs required for generating \( \log_2(L \cdot Q) \) soft outputs corresponding 295 to a single codeword estimate \( \hat{S}^b \) is thus given by

$$\text{RMO}[\hat{S}^b] = MN_{N_w}(N_w + 1) + \sum_{\mu=1}^{N_w} [4M^2N(N_w - \mu + 1) + 2]$$

where \( \hat{S}^b \) and \( \hat{S}^b \) represent the MAP-MSDSD estimate and the 298 constrained estimate associated with \( b_j = b \), respectively.

On the other hand, the number of RMOs related to each bit \( b \) of 297 the constrained estimate \( \hat{S}^b \) is found to be \( 2MN[4M^2N(N_w + 1) + 1] \). 298 The lower bound for the number of RMOs associated with the gen- 299 eration of a single soft output is thus \( \frac{MN(2M + 1)}{1 + (M^2N)[N_w^2]} \) for large \( N_w \). The complexity of the scheme, however, 301 depends on a number of parameters, such as on the channel SNR, on 302 the autocorrelation function of the channel’s fading plus noise, and 303 most importantly, on the \textit{a priori} mutual information \( I_A \) of the inner 304 decoder [15]. In Section V, the complexity of the MAP-MSDSD will 305 be investigated as a function of the observation window width \( N_w \) 306 parameterized by the available \textit{a priori} information \( I_A \).
TABLE I

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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</thead>
<tbody>
<tr>
<td>Dispersive channel model</td>
<td>COST207-TU12</td>
</tr>
<tr>
<td>Fast fading envelope</td>
<td>Correlated Rayleigh fading</td>
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<tr>
<td>Normalized Doppler spread, ( f_d )</td>
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</tr>
<tr>
<td>Number of subcarriers, ( N_c )</td>
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</tr>
<tr>
<td>Length of cyclic prefix</td>
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<tr>
<td>Overall symbol duration</td>
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<td>STSK ( (M, N, T, Q, \Lambda) )</td>
<td>( (2, 2, 2, 4, 4) )</td>
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<tr>
<td>RSC encoder and decoder</td>
<td>Half rate</td>
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</tr>
<tr>
<td>Generator polynomial</td>
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<td>Length of interleavers</td>
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<td>Outer iterations</td>
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</tr>
<tr>
<td>Inner iterations</td>
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</table>

V. PERFORMANCE RESULTS

Fig. 3 characterizes the achievable bit error rate (BER) of the proposed scheme at SNR = 1 and 4 dB, respectively. We observe in Fig. 4 that the inner decoder’s EXIT charts recorded at SNR = 1 dB for \( N_w = 6 \) exhibit an open EXIT tunnel, indicating a high residual BER. By contrast, the BER associated with \( N_w = 6 \), 10 may be expected to decrease sharply at this SNR and compares to the (1.0, 1.0) point of perfect decoding convergence, leading to a vanishingly low BER, thus eliminating the system’s error floor. The maximum achievable rates for the corresponding scheme, where the 336 scheme still exhibits an infinitesimally low BER were computed by 337 exploiting the area property of EXIT charts [18], [32], [33], and 338 shown in Fig. 3 as the ultimate benchmark of the scheme.

To elaborate further, Figs. 4 and 5 portray the EXIT charts of our 340 proposed scheme at SNR = 1 and 4 dB, respectively. We observe in 341 Fig. 4 that the inner decoder’s EXIT charts recorded at SNR = 1 dB for \( N_w = 2, 4, 6, 10 \) and of the corresponding coherent scheme having perfect CSI at the receiver. At this SNR, we observe the inner EXIT charts with \( N_w = 6, 10 \) have an open EXIT tunnel and converge to the (1.0, 1.0) point of perfect convergence, indicating a sharp fall in the BER curve after \( I_{outer} = 9 \) outer iterations, which is confirmed by the decoding trajectory for \( N_w = 6 \). The EXIT charts with \( N_w = 2, 4 \) are, however, “pinched off”; thus, the BER at this SNR do not converge.

more, the multiple-symbol detection partially mitigates the inherent 324 performance penalty imposed by noncoherent detection. We observe 325 Fig. 3 that, as \( N_w \) increases, the BER performance gradually 326 approaches that of the perfect CSI-oriented coherent scheme. Note 327 that all the performance characteristics exhibit a vanishingly low BER 328 after \( I_{outer} = 9 \) outer iterations, which is the explicit benefit of using 329 the URC in the system. The URC is a low-complexity code, which 330 has an infinite impulse response and hence assists the inner decoder 331 in efficiently spreading the soft information [2], [19]. As a result, the 332 extrinsic information transfer (EXIT) charts of Figs. 4 and 5 converge 333 to the (1.0, 1.0) point of perfect decoding convergence, leading to a 334 vanishingly low BER, thus eliminating the system’s error floor. The 335 maximum achievable rates for the corresponding scheme, where the 336 scheme still exhibits an infinitesimally low BER were computed by 337 exploiting the area property of EXIT charts [18], [32], [33] and are 338 shown in Fig. 3 as the ultimate benchmark of the scheme.

Fig. 4. EXIT charts of the inner decoders of the soft-decision-MSDSD-aided OFDM DSTSK \( (2, 2, 2, 4, 4) \) scheme at SNR = 1 dB with normalized Doppler frequency \( f_d = 0.01 \) and different observation window sizes \( N_w = 2, 4, 6, 10 \) and of the corresponding coherent system having perfect CSI at the receiver. 354

Fig. 6 characterizes the complexity associated with the MAP- 355 MSDSD of the OFDM-aided DSTSK \( (2, 2, 2, 4, 4) \) scheme at SNR = 356 4 dB as a function of the window size \( N_w \), parameterized by the 357
the BER curve after EXIT charts have a quite open EXIT tunnel at this SNR and converge to the and that of the corresponding coherent inner decoder as a benchmark. All the a priori decoding trajectory. I as a function of observation window size f. As expected, the complexity rapidly escalates upon increasing of RMOs in Section IV is also shown as a benchmarker in Fig. 6. DSTSK (2, 2, 2, 4, 4) scheme at SNR 4 dB using the parameters of Table I as a function of observation window size Nw parameterized against the a priori mutual information of the inner decoder I_A. The complexity shoots up with Nw > 6, although the rate of increase in complexity slows down with increased a priori information.

Fig. 5. EXIT charts of the inner decoders of the soft-decision-MSDSD-aided OFDM DSTSK (2, 2, 2, 4, 4) scheme at SNR = 4 dB with normalized Doppler frequency f_d = 0.01 and different observation window sizes N_w = 2, 4, 6, 10 and that of the corresponding coherent inner decoder as a benchmark. All the EXIT charts have a quite open EXIT tunnel at this SNR and converge to the (1, 1) point as a benefit of employing the URC, indicating a sharp fall in the BER curve after I_outer = 9 outer iterations, which is confirmed by the decoding trajectory.

Fig. 6. Complexity in terms of the numbers of RMOs for the proposed DSTSK (2, 2, 2, 4, 4) scheme at SNR = 4 dB using the parameters of Table I as a function of observation window size N_w.

VI. CONCLUSION

We have proposed a soft-decision-MSDSD-aided multicarrier 373 DSTSK scheme for communications over wideband channels. The 374 OFDM-aided DSTSK provides a flexible diversity versus multiplexing 375 gain tradeoff by spreading the source information across both the spatial and time dimensions, while mitigating the potential performance 377 degradation imposed by the frequency selectivity of the channel. The 378 turbo-principle-based soft-decision MSDSD facilitates joint decisions 379 over a number of DSTSK codewords, while exploiting the fading-380 plus-noise statistics of the channel. We have demonstrated that the 381 proposed soft-decision-MSDSD-aided DSTSK scheme provides sub-382 stantial flexibility at moderate complexity owing to dispensing with 383 CE. Furthermore, the MSDSD mitigates the performance degradation 384 inflicted by the CDD scheme without an undue increase in computa-

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albeit it does not become excessively high, provided that the a priori information gleaned from the outer decoder is in the range of I_A ≥ 0.8.


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