Layered Space–Time Receivers for Frequency-Selective Wireless Channels
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Abstract—Recent results in information theory have demonstrated the enormous potential of wireless communication systems with antenna arrays at both the transmitter and receiver. To exploit this potential, a number of layered space–time architectures have been proposed. These layered space–time systems transmit parallel data streams, simultaneously and on the same frequency, in a multiple-input multiple-output fashion. With rich multipath propagation, these different streams can be separated at the receiver because of their distinct spatial signatures. However, the analysis of these techniques presented thus far had mostly been strictly narrowband. In order to enable high-data-rate applications, it might be necessary to utilize signals whose bandwidth exceeds the coherence bandwidth of the channel, which brings in the issue of frequency selectivity. In this paper, we present a class of layered space–time receivers devised for frequency-selective channels. These new receivers, which offer various performance and complexity tradeoffs, are compared and evaluated in the context of a typical urban channel with excellent results.

Index Terms—Adaptive antennas, BLAST, equalization, frequency selectivity, interference cancellation, layered architectures, MIMO, space–time processing.

I. INTRODUCTION

RECENT information theory results have shown the enormous spectral efficiency potential of wireless communication systems with antenna arrays at both the transmitter and receiver, in particular when the channel and array structures are such that the transfer functions between different transmit and receive antenna pairs are sufficiently uncorrelated [1]–[4]. To exploit this potential, a number of layered space–time (BLAST) architectures have been proposed [5], [6]. BLAST systems transmit parallel data streams, using multiple antennas, simultaneously and in the same frequency band. With rich multipath propagation, these different streams can be separated at the receiver because of their distinct spatial signatures. Remarkably, in its original form, BLAST does not require the transmitter to possess any channel information; only the receiver is required to estimate the channel. Nonetheless, provided the scattering richness is sufficiently high, the spectral efficiency attainable—in this open-loop form—is often very close to the spectral efficiency supported with full channel information at the transmitter [7]. In this paper, we focus our attention on such rich-scattering environments where BLAST performs at its best.

A form of BLAST that is very attractive for its relative simplicity was introduced in [6], [8] and labeled as Vertical BLAST (V-BLAST). In V-BLAST, every transmit antenna radiates an equal-rate independently encoded stream of data. This transmit structure enables the utilization, at the receiver, of interference rejection and cancellation techniques [9], [10] with the added advantage that the multiple streams are precisely synchronized. A V-BLAST receiver can be regarded, therefore, as a synchronous multiuser detector with ordered successive cancellation. This type of successive cancellation method has already proved very effective in other contexts [11]–[13]. Nonetheless, the V-BLAST formulation and analysis presented thus far had been strictly narrowband. In order to extend that formulation to the more general case of frequency-selective channels, two dual approaches exist, namely, orthogonal frequency-division multiplexing (OFDM) or high-speed serial equalization. Whereas the first approach was the one chosen in [3], in this paper we concentrate on the high-speed serial case, where the receiver adopts—in its full generality—the form of a multiple-input multiple-output (MIMO) decision-feedback equalizer (DFE). Linear MIMO equalizer alternatives are also included in this framework simply by setting the length of the DFE feedback section to zero.

MIMO equalizers present a significant challenge because of the need to detect signals buried in both co-channel interference (CCI) as well as inter-symbol interference (ISI), in addition to noise. They were studied in the past in the context of cross-coupled channels and dually polarized radio systems among other problems [14]–[16]. With the exploding interest in space–time processing and multiuser detection in recent years, MIMO DFEs have again attracted significant attention. The optimal settings—in the minimum mean-square error (MMSE) sense—for the MIMO DFE were derived in [17], [18] within the framework of code-division multiple access (CDMA) and in [19], [20] for the case of space-division multiple access (SDMA). However, the settings derived therein correspond, in general, to infinite-length noncausal filters. Furthermore, in some of those analyses, the number of inputs and outputs were constrained to be identical. In our case, the number of transmit and receive antennas need not be the same, so this requirement is dropped. Also, we are interested in solutions that correspond to realizable finite-impulse-response (FIR) filter [21]. Those optimal settings, again in the MMSE sense, have recently been reported in [22]–[24] for a receiver without successive cancellation. In that receiver, only decisions on temporally preceding symbols are fed back into the detection process of each stream [Fig. 1(a)]. CCI contributions from undetected future and on-time symbols are, therefore, not canceled. In this

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Fig. 1. Detection strategies in the space–time plane. Example depicting the
detection process at the time of extracting the symbol marked with ‘O’, which is
symbol /107 of one of four data streams. Symbols marked with an ‘X’ have already
been detected and their interference contribution canceled. (a) MIMO-DFE. (b)
Partially connected OSIC-DFE.

The paper presents two different receiver structures that operate
on the basis of ordered successive interference cancellation
(OSIC) and are hence a more natural extension of narrowband
V-BLAST. In these OSIC receivers, the data streams are
successively detected in an ordered manner. Thus, as shown in
Fig. 1(b), each stream is detected with the entire CCI contri-
bution from every previously detected stream already canceled
out. Furthermore, the detection ordering is organized so that
strong streams—more resilient to interference—are detected
earlier in the process [6]. With that, the weak streams—which
mostly determine the aggregate error rate—can be detected
more reliably. As a result, the overall performance improves
markedly. In order to evaluate these OSIC receivers, we first
review the MIMO-DFE and formulate our OSIC-DFE receiver in both
its partially connected and fully connected forms. The perform-
ance and complexity of these various receivers are evaluated
in Sections IV and V, respectively, and the results summarized
in Section VI.

Throughout the remainder, we utilize (·)T, (·)*, and (·)H to
denote matrix transposition, conjugation, and Hermitian trans-
position, respectively.

II. SYSTEM AND CHANNEL MODELS

We consider a discrete-time complex baseband model for a
single-user link, assuming perfect carrier recovery and down-
conversion. Received signals are sampled at the symbol rate and,
therefore, the receiver structures we present are symbol-spaced.2
It is assumed that the channel is stationary over every burst of
data, although it may change from burst to burst. Perfect channel
estimation at the receiver is further assumed [27].

We use M × N to signify a configuration with M transmit
and N receive antennas and L + 1 to indicate the number of
symbol-spaced taps in the channel response. Thus, the sampled
channel response from transmitter m to receiver n, including
transmit and receive filters, is denoted by

\[ \mathbf{h}_{mn} = [h_{mn}(0) \ h_{mn}(1) \ \ldots \ h_{mn}(L)]^T \]  (1)

with no time dependence within every individual burst. The
signal transmitted at time k is the M × 1 vector \( \mathbf{s}(k) \) with spatial
covariance

\[ E[\mathbf{s}(k)\mathbf{s}^H(k)] = \frac{P_T}{M} \mathbf{I}_M \]  (2)

where \( P_T \) is the total average transmit power, which is held con-
stant irrespective of the number of transmit antennas. The re-
ceiver additive white Gaussian noise (AWGN) can be expressed,
in turn, as an N × 1 vector \( \mathbf{n}(k) \) with spatial covariancem

\[ E[\mathbf{n}(k)\mathbf{n}^H(k)] = \sigma^2 \mathbf{I}_N. \]  (3)

We can assemble the \( \mathbf{h}_{mn} \) vectors into M separate matrices of
size \( N \times (L+1) \) as follows:

\[ \mathbf{H}_m = \begin{bmatrix} \mathbf{h}_{m0}^T \\ \mathbf{h}_{m1}^T \\ \vdots \\ \mathbf{h}_{mM}^T \end{bmatrix}, \quad m = 1, 2, \ldots, M. \]  (4)

With that, we can express the \( N \times 1 \) received vector \( \mathbf{x}(k) \) as

\[ \mathbf{x}(k) = \sum_{m=1}^{M} \mathbf{H}_m \mathbf{s}_m(k) + \mathbf{n}(k) \]  (5)

2The front stage of a practical receiver implementation would almost cer-
tainly be fractionally spaced. For the purpose of exploring architectures, how-
ever, symbol spacing is a valid start. The extension to the fractionally spaced
case is relatively straightforward [26].
with the sequence of $L + 1$ symbols transmitted by the $n$th antenna denoted by

$$s_m(k) = \begin{bmatrix} s_m(k) \\ s_m(k-1) \\ \vdots \\ s_m(k-L) \end{bmatrix}. \quad (6)$$

We can now define $\rho_n$ as the expected SNR—over the ensemble of all possible channel realizations—on the $n$th receive antenna as

$$\rho_n = \frac{P_T}{M\sigma^2} \sum_{m=1}^{M} E|H_{nm}|^2 \quad (7)$$

which is—because of local stationarity—identical for all $n$ and thus we can write $\rho = \rho_n$ for every value of $n$. Finally, and in order to facilitate the formulation, we introduce $[28]$ the operator $\text{vec}(\cdot)$ as

$$\text{vec}(\begin{bmatrix} v_1 \\ v_2 \\ \vdots \\ v_K \end{bmatrix}) = \begin{bmatrix} v_1 \\ v_2 \\ \vdots \\ v_K \end{bmatrix} \quad (8)$$

where $v_1, \ldots, v_K$ are equal size column vectors.

### III. SPACE–TIME RECEIVERS: FORMULATION

#### A. MIMO-DFE

The MIMO-DFE, shown in Fig. 2, consists of a feedforward filter with a temporal span of $K_f + 1$ taps and a feedback filter with a temporal span of $K_b$ taps [22], [23]. It is a fully connected receiver in that past decisions on all data streams are fed back into the detection process for each stream. We will use $(K_f+1, K_b)$ to denote a receiver with $(K_f+1) \times N \times M$ feedforward taps and $M \times K_b \times M$ feedback taps.

The $K_f + 1$ $N$-dimensional symbols spanned by each of the $M$ components of the feedforward filter can be grouped as

$$X(k) = [x(k) \ x(k-1) \ \ldots \ x(k-K_f)]. \quad (9)$$

We can define $\mathbf{x}(k) = \text{vec}(X(k))$, which can be conveniently expressed as

$$\mathbf{x}(k) = \sum_{m=1}^{M} \mathbf{H}_m \mathbf{s}_m(k) + \mathbf{n}(k) \quad (10)$$

where the extended sequence of symbols from the $m$th transmitter is

$$\mathbf{s}_m(k) = \begin{bmatrix} s_m(k) \\ s_m(k-1) \\ \vdots \\ s_m(k-L-K_f) \end{bmatrix} \quad (11)$$

and

$$\mathbf{H}_m = \begin{bmatrix} H_m & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & H_m \end{bmatrix} \quad (12)$$

is a block Toeplitz matrix with $N(K_f+1)$ rows and $L+K_f+1$ columns. In addition, we have

$$\mathbf{n}(k) = \text{vec}([\mathbf{n}(k) \ \mathbf{n}(k-1) \ \ldots \ \mathbf{n}(k-K_f)]). \quad (13)$$

The sequence of $K_b$ most recent decisions on every stream is labeled as

$$\hat{s}_m(k-d-1) = \begin{bmatrix} \hat{s}_m(k-d-1) \\ \vdots \\ \hat{s}_m(k-d-K_b) \end{bmatrix} \quad (14)$$

with the decision delay $d$ being a parameter so that

$$\hat{S}(k-d-1) = [\hat{s}_1(k-d-1) \ \ldots \ \hat{s}_M(k-d-1)] \quad (15)$$

or, equivalently, $\hat{S}(k-d-1) = \text{vec}(\hat{S}(k-d-1))$ constitute the input to the feedback filter.

We can arrange the structure of the feedforward filter into $M$ matrices $\mathbf{W}_m$, $m = 1, 2, \ldots, M$, with dimensionality $(K_f+1) \times N$. Each of these matrices has the role of projecting—in the MMSE sense—one of the incoming data streams away from its own uncanceled (pre-cursor) ISI as well as the uncanceled CCI from all other streams. Similarly, we can arrange the feedback filter into $M$ matrices $\mathbf{B}_m$, $m = 1, 2, \ldots, M$, with dimensionality $M \times K_b$. With that, we can express the estimate for stream $m$ at time $k$ as

$$y_m(k) = \text{Tr} \left\{ \mathbf{W}_m^H \mathbf{x}(k) \right\} - \text{Tr} \left\{ \mathbf{B}_m^H \hat{S}(k-d-1) \right\} \quad (16)$$

where $\text{Tr}$ denotes the trace of a matrix. Using $\mathbf{w}_m = \text{vec}(\mathbf{W}_m)$ and $\mathbf{b}_m = \text{vec}(\mathbf{B}_m)$, (16) can be rewritten as

$$y_m(k) = \mathbf{w}_m^H \mathbf{x}(k) - \mathbf{b}_m^H \hat{S}(k-d-1) = \left[ \begin{bmatrix} \mathbf{w}_m \\ \mathbf{b}_m \end{bmatrix}^H \right] \left[ \begin{bmatrix} \mathbf{x}(k) \\ \hat{S}(k-d-1) \end{bmatrix} \right]. \quad (17)$$

Let us assume, for the time being, that previous decisions are correct, i.e., $\hat{s}_m(k-d) = s_m(k-d)$ for all $m$ and $k$. The MMSE Wiener–Hopf solution for $\mathbf{w}_m$ and $\mathbf{b}_m$ that minimizes

$$E[|y_m(k) - s_m(k-d)|^2] \quad (18)$$

is then

$$\left[ \begin{bmatrix} \mathbf{w}_m \\ \mathbf{b}_m \end{bmatrix} \right] = \mathbf{R}^{-1} \mathbf{p}_m \quad (19)$$

3In general, the delays can be different for every data stream. In our formulation, as in those of [22] and [23], we consider them equal ($d = d_m$ for $m = 1 \ldots M$).
with [26]

\[
P_m = E \left[ -s_m(k-d-1) s_m^*(k-d) \right] = \frac{P_T}{M} \left[ \begin{array}{c} \mathbf{0} \\ \mathbf{f}_m^{(d+1)} \end{array} \right] \tag{20}
\]

where \((\cdot)^{d+1}\) indicates the \((d+1)\)th column of the corresponding matrix, and with (21), shown at the bottom of the page, where \((\cdot)^{d+2, \ldots, K_{b}+1}\) indicates columns \(d+2\) to \(d+K_{b}+1\). Notice that the highly structured form of \(\mathbf{R}\) may be exploited to simplify its inverse operation in (19).

**B. Partially Connected OSIC-DFE**

After reviewing the MIMO-DFE, we now proceed to introduce the first of our OSIC structures. The partially connected receiver consists of \(M\) successive stages, the first two of which are shown in Fig. 3, each having a feedforward filter \(\mathbf{W}\) with a temporal span of \(K_{f} + 1\) taps per receive antenna and a feedback filter \(\mathbf{b}\) with a temporal span of \(K_{b}\) taps. Hence, every \((K_{f} + 1, K_{b})\) stage resembles a multiple-input single-output (MISO) DFE, for which a large body of literature exists (see, for instance, [29]–[32] and the references therein). At each stage, the “best” data stream—in the MMSE sense—is extracted, detected, and canceled out. This simple ordering strategy was shown to be globally optimal, in the sense of minimum error rate, at asymptotically high SNR [6].

The input signal \(\mathbf{x}(k) = \mathbf{w}(\mathbf{X}(k))\) is, like before, defined by (10).

The output of every stage at time \(k\) is an estimate \(\hat{s}_m(k-d)\) for symbol \(k-d\) of whichever stream \(m\) was selected by the ordering mechanism. The decision delay \(d\) is—as previously stated—assumed identical for all stages. The sequence of \(K_{b}\) most recent decisions produced by that stage is labeled as

\[
\hat{s}_m(k-d-1) = \begin{bmatrix} \hat{s}_m(k-d-1) \\ \vdots \\ \hat{s}_m(k-d-K_{b}) \end{bmatrix}
\]

and constitutes the input to the corresponding feedback filter.

At each stage, every undetected stream is a candidate for selection. Let us denote by \(\mathbf{W}_m\) and \(\mathbf{b}_m\) the feedforward and feedback settings that, at a certain stage, extract stream \(m\) as

\[
y_m(k) = \text{Tr} \{ \mathbf{W}_m^H \mathbf{X}(k) \} - \mathbf{b}_m^H \hat{s}_m(k-d-1) \tag{23}
\]

which, using \(\mathbf{u}_m = \mathbf{w}(\mathbf{W}_m)\), is equivalent to

\[
y_m(k) = \mathbf{u}_m^H \mathbf{x}(k) - \mathbf{b}_m^H \hat{s}_m(k-d-1)
\]

Assuming again correct decisions at all stages, the MMSE Wiener–Hopf solution for \(\mathbf{u}_m\) and \(\mathbf{b}_m\) is

\[
\begin{bmatrix} \mathbf{u}_m \\ \mathbf{b}_m \end{bmatrix} = \mathbf{R}_m^{-1} \mathbf{p}_m
\]

with [26]

\[
P_m = E \left[ -s_m(k-d-1) s_m^*(k-d) \right] = \frac{P_T}{M} \left[ \begin{array}{c} \mathbf{0} \\ \mathbf{f}_m^{(d+1)} \end{array} \right] \tag{26}
\]

where \((\cdot)^{d+1}\) indicates, as before, the \((d+1)\)th column of the corresponding matrix, and with

\[
\mathbf{R}_m = E \left[ \begin{bmatrix} \mathbf{x}(k) \\ -s_m(k-d-1) \end{bmatrix} \begin{bmatrix} \mathbf{x}(k) \\ -s_m(k-d-1) \end{bmatrix}^H \right] = P_T \left[ \sum_{j=1}^{M} \mathbf{f}_j \mathbf{f}_j^H + \sigma^2 \mathbf{I} - (\mathbf{f}_1)^{d+2, \ldots, K_{b}+1} \right] \tag{27}
\]

where the summation is over the yet-to-be-detected data streams. Again, the structured form of \(\mathbf{R}_m\) may lead to simpli-
The undetected stream with the smallest MSE\(^4\) can be selected at every stage and extracted using the \(\mathbf{w}_m\) and \(\mathbf{b}_m\) settings defined by (25). Finally, once a stream has been detected, its interference contribution can be removed from the input signal \(\mathbf{x}(k)\) as follows:

\[
\begin{align*}
\mathbf{x}(k) &\leftarrow \mathbf{x}(k) - (\mathbf{H}_m)_1 \hat{s}_m(k) \\
\mathbf{x}(k+1) &\leftarrow \mathbf{x}(k+1) - (\mathbf{H}_m)_2 \hat{s}_m(k) \\
\vdots &\vdots \\
\mathbf{x}(k+L) &\leftarrow \mathbf{x}(k+L) - (\mathbf{H}_m)_{L+1} \hat{s}_m(k).
\end{align*}
\] (29)

Notice that, since the channel response spans \(L+1\) symbols, the interference arising from \(s_m(k)\) has to be canceled from \(L+1\) consecutive entries.

### C. Fully Connected OSIC-DFE

Finally, we introduce the fully connected OSIC-DFE, which consists of \(M\) stages each exactly equivalent to an individual MIMO-DFE as described in Section III-A. The fully connected OSIC-DFE receiver is displayed in Fig. 4. In a \((K_f + 1, K_b)\) configuration, every stage consists of \((K_f + 1) \times N \times M\) feedforward taps and \(M \times K_b \times M\) feedback taps. Like in the partially connected case, the \(M\) stages are ordered so that the data streams are extracted and decoded in sequence of decreasing MSE. Unlike the partially connected receiver, though, all \(M\) streams are implicitly decoded within every stage and their symbol decisions fed back internally. However, only the stream selected by the ordering algorithm at every stage is retained while the others are discarded and decoded again in a later stage. Thus, the potential advantage of the fully connected receiver arises from having a more powerful feedback section within every stage. At the same time, this enhancement is also a potential drawback because it makes the receiver more vulnerable to error propagation, in particular given that potentially unreliable tentative decisions from weaker streams are utilized to estimate the symbols of stronger streams.

Since the MIMO-DFE was formulated in Section III-A, it will not be repeated here. The extension to the fully connected OSIC-DFE is a straightforward cascade of \(M\) such identical structures.

### IV. PERFORMANCE EVALUATION

A rigorous comparison of the OSIC-DFE and MIMO-DFE receivers, including the effects of error propagation, requires detailed simulations. In this section, we present the results of a series of simulations conducted in a stationary (no Doppler) TU channel [25] with rich spatial scattering. The (average) temporal channel profile, with a normalized rms delay spread \(\tau = 0.2886\) symbols, is detailed in Table I. The channel response consists of six independently faded Rayleigh paths with specific delays and average powers.

The propagation scenario we consider, typical of a mobile system, is based on the existence of an area of local scattering around the terminal. Little or no local scattering is presumed around the base station. Accordingly, the angular spread is expected to be very large—possibly as large as 360°—at the terminal while small at the base [33], [34]. Hence, negligible correlation among the terminal antennas is basically guaranteed. The antennas at the base station, on the other hand, can be decorrelated by spacing them sufficiently apart [35]–[37]. For each path, consequently, the transfer coefficients from the various transmit antennas to the various receive antennas are assumed uncorrelated.

We employ QPSK modulation and square-root raised cosine transmit and receive filters with 35% excess bandwidth as in IS-136 TDMA and its evolutionary convergence toward enhanced data rates for GSM evolution (EDGE) [38], [39]. There is no coding. Since detailed analysis of spectral efficiency versus number of antennas were presented—for the narrowband case—in [6], here we concentrate on studying the effect of different delay spreads for a given antenna configuration, which is chosen to be \(4 \times 6\).
The integer decision delay $d$ and the sampling phase (referenced to the first channel path and modulo the symbol period) are free parameters, which we lump together into a combined fractional delay $\delta$, expressed in symbol periods. In a practical implementation, $\delta$ would be adjusted by some synchronization mechanism. In our simulations, it is always chosen to minimize the bit error rate (BER). As an example, we present in Fig. 5 the BER as a function of $\delta$ with (4,2) taps. The four different local minima, representing different alignments of the channel and the four feedforward taps, correspond to various values of $\delta$ with the optimal sampling phase. Notice how both OSIC-DFE architectures are markedly superior to the MIMO-DFE, with the uncoded BER being almost an order of magnitude better with proper synchronization. The MIMO-DFE, in contrast, is much less sensitive to synchronization offsets. Nonetheless, sensitivity to sampling phase offsets would be drastically reduced—in all cases—with fractional spacing [40]. Both OSIC-DFE receivers perform very similarly in this example.

In Fig. 6, the BER is shown as a function of the SNR for a multiplicity of receiver settings, namely (1,0), (2,1), (4,2) and (6,3). Only the MIMO-DFE and the partially connected OSIC-DFE are shown. The fully connected OSIC-DFE performs only slightly better (a small fraction of a decibel) than its partially connected counterpart. The (1,0) case, corresponding to a narrowband V-BLAST receiver, displays a rather poor behavior in spite of the relatively small delay spread of the TU channel. A simple (2,1) configuration handles the frequency selectivity much better in the OSIC-DFE case, which is not so in the MIMO-DFE. The difference between the two is significant (over 3 dB at BER = 10^{-3}). However, the difference shrinks as the number of taps increases and it falls below 0.5 dB with (6,3). For each receiver, the relative returns diminish rapidly after (4,2) and little gain is obtained by adding additional taps, particularly in the OSIC case.

A fundamental issue is that of the temporal span required to handle a given level of delay spread. The longer the span, the larger the number of coefficients that have to be estimated and the higher the complexity. Given a certain delay spread, the span can be analytically estimated from the standpoint of the receiver being able to “invert” the channel at asymptotically high SNR, although that may lead to inaccurate results. Span analyses for MISO systems can be found in the literature (see [26], [32], and [41], for instance). Nonetheless, it is not clear how well these techniques extend to MIMO systems, in particular in the OSIC case, where errors in the interference cancellation process may propagate through multiple receiver stages. In our approach, based on simulation, the required span is that which yields a desired BER level. Accordingly, we present in Fig. 7 the BER as a function of the normalized delay spread in the asymptotically high SNR regime ($\rho \to \infty$), where MMSE receivers behave in zero-forcing mode. The curves shown correspond to receivers with (1,0), (2,1) and (4,2) settings. The (1,0) receiver—corresponding to narrowband V-BLAST—can only handle delay
spreads up to about 0.2 symbols, after which the BER explodes.\(^6\) The (2,1) receivers can handle spreads up to a full symbol at BER \(= 10^{-5}\). Finally, the (4,2) receivers can handle over 1.5 symbols at that BER. In all cases, the OSIC-DFE structures outperform the MIMO-DFE. The relative performance of the partially connected and fully connected OSIC-DFE versions, however, varies depending on the tap settings. Clearly, error propagation across stages has a definite role at determining whether the additional feedback available in the fully connected case is beneficial or, instead, detrimental. In any case, the difference between the two appears to be small and does not seem to justify a significantly higher level of complexity if such was needed to implement a fully connected structure.

V. COMPLEXITY EVALUATION

Having established the comparative performance of the various receivers, we now proceed to evaluate their relative complexity. Such comparison will be performed under the following conditions.

1) An initial estimate of the channel is periodically obtained by means of embedded training sequences or some other form of pilot and the receiver settings are initialized on base of that estimate. If an OSIC receiver is used, the detection order is also determined and held constant until the next such initialization.

2) Between those initialization procedures, the receiver settings are adaptively adjusted on a symbol-by-symbol basis using a recursive least-squares (RLS) algorithm \([43]\). Notice that the receiver settings can only be adaptively adjusted up to the point when the detection order needs to be modified. Hence, the periodicity of the initializations should be chosen accordingly depending on the normalized Doppler rate of the channel and other considerations, but that is beyond the scope of this paper. Our goal is to assess the complexity of the initialization and adaptation procedures for the various receivers.

Let us define for convenience

\[
\begin{align*}
N_{fc} &= N(K_f + 1) + MK_b \\
N_{pc} &= N(K_f + 1) + K_b
\end{align*}
\]

with \(N_{fc}\) the total number of coefficients in the MIMO-DFE as well as the number of coefficients in each of the \(M\) stages of the fully connected OSIC-DFE, and with \(N_{pc}\) the number of coefficients in each stage of the partially connected OSIC-DFE. A comparison of receiver complexities in number of multiply–add operations is presented in Table II.

Notice the strong increase in initialization complexity for the OSIC receivers (square in the number of transmit antennas). This is a direct consequence of the need to consider every possible detection order in order to select the one that yields the smallest MSE. In the MIMO DFE, there is no need to order and thus the initialization complexity is much more modest.

As far as the adaptation is concerned, the relative increase in complexity associated with adjusting the settings of an OSIC-DFE structure over the MIMO-DFE is smaller (linear with the number of transmit antennas). The complexity cost of running the filters and performing the interference cancellation is comparatively very small and thus it is not accounted for. Furthermore, the feedback sections can often be implemented without the need for multipliers.

The relative complexity of the fully connected OSIC-DFE with respect to its partially connected counterpart is defined by powers of the ratio

\[
r = \frac{N_{fc}}{N_{pc}} = \frac{N(K_f + 1) + MK_b}{N(K_f + 1) + K_b}.
\]

Since, in general, \(N \geq M\) and \(K_f + 1 > K_b\), the ratio is bounded by

\[
1 \leq r < \frac{2N}{1 + N} \leq 2.
\]

Some particular cases are as follows.

- If \(N(K_f + 1) \gg MK_b\), the ratio becomes \(r \approx 1\) and thus the complexities of the partially connected and fully connected OSIC-DFE receivers become similar.

- If both \(M\) and \(N\) are large, the ratio becomes \(r \approx 1 + MK_b/N(K_f + 1)\). Furthermore, if \(M = N\), it particularizes to \(r \approx 1 + K_b/K_f + 1\).

VI. SUMMARY

We have formulated a class of vertical layered space-time (V-BLAST) receivers, based on ordered successive interference cancellation (OSIC) and optimized in the MMSE sense, which can operate in frequency-selective environments. These receiver structures are generalizations of the original narrowband V-BLAST receiver presented in [6]. Their performance has been evaluated, using a TU channel profile and a variety of normalized delay spreads, with excellent results. Furthermore, these new receivers have been compared against a MIMO-DFE [22], [23] and proved superior, in particular when the temporal span of the filters is relatively short. With the standard normalized delay spread of the TU channel, the OSIC receivers show

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<td>(O(N_f^2))</td>
</tr>
<tr>
<td>Partially Connected OSIC-DFE</td>
<td>(O(M^2 \cdot N_f^2))</td>
<td>(O(MN_f^2))</td>
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<td>Fully Connected OSIC-DFE</td>
<td>(O(M^2 \cdot N_f^2))</td>
<td>(O(MN_f^2))</td>
</tr>
</tbody>
</table>

\(^6\)Interestingly, this value coincides with what has been historically regarded as the threshold for equalization requirements in single-antenna receivers [42].
a 4-dB advantage at uncoded BER = 10^{-3} with (2,1) filter settings. With (4,2), the difference shrinks to about 1.5 dB. With even longer filters, the difference saturates at about 1 dB. This improved performance, however, comes at a significant cost in terms of complexity. Such increase is, specifically, a direct function of the number of transmit antennas.

Both the partially connected and the fully connected forms of the OSIC receiver perform similarly in most cases. The fully connected is slightly superior when the effects of error propagation across stages are small (low BER regime and/or short feedback sections) whereas the partially connected is slightly superior when error propagation becomes significant. Given the additional complexity required by the fully connected implementation—up to 100%—and the likely increase in the impact of error propagation associated with imperfect channel estimation, we conclude that the partially connected receiver constitutes the more attractive alternative.

As future work, it is essential to extend existing channel estimation analyses to the wideband case and to assess the impact of channel estimation on the required receiver span. In addition, it would be interesting to explore the possibility of using different filter spans and decision delays at different stages in the detection process. Synchronization, which was assumed ideal—in the sense of minimum BER—throughout this work, is another topic for further research.

REFERENCES


[38] JS136A TDMA Cellular/PCS, Dual-Mode Mobile Station—Base Station Compatibility Standard, EIA/TIA 136, 1996.


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