Space-Time Receiver for Wideband BLAST in Rich-Scattering Wireless Channels

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Abstract— **Recent results in information theory have demonstrated the enormous capacity potential of wireless communication systems with antenna arrays at both transmitter and receiver. To exploit this potential, the Bell-laboratories LAyered Space-Time (BLAST) architecture was proposed. BLAST systems transmit parallel data streams, simultaneously and on the same frequency, in a Multiple-Input Multiple-Output (MIMO) fashion. With rich multipath propagation, these different streams can be separated and recovered at the receiver. The analysis of BLAST presented thus far had always been strictly narrowband. In this paper, we extend the formulation by presenting a receiver devised for more general frequency-selective channels. This new receiver is evaluated—via simulation—in the context of a Typical Urban (TU) channel with excellent results.**

I. Introduction

RECENT information theory results have shown the enormous link capacity potential of wireless communication systems with antenna arrays at both 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 largely uncorrelated [1][2]. To exploit this potential, the Bell-laboratories LAyered Space-Time (BLAST) architecture was proposed [3][4]. BLAST systems transmit parallel data streams, using multiple antennas, simultaneously and on the same frequency. 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 very close to the spectral efficiency supported by the channel. In this paper, we focus our attention on such rich-scattering environments where BLAST performs at its best.

The initial Diagonal BLAST (D-BLAST) architecture is theoretically capable of approaching the open-loop spectral efficiency, but at a high complexity cost [3]. A simplified version known as Vertical BLAST (V-BLAST), which still achieves a hefty portion of that efficiency, was later developed [4]. In V-BLAST, every transmit antenna radiates an equal-rate independently encoded stream of data. This independence enables the utilization, at the receiver, of interference rejection [5] and cancellation [6] techniques with the added advantage that the multiple streams are precisely synchronized. A V-BLAST receiver can be regarded, therefore, as a multi-stage synchronous multiuser detector. This type of successive cancellation methods have already proved very effective in other contexts [7]. Nonetheless, the V-BLAST formulation and analysis presented thus far had been strictly narrowband. In this paper, we extend that formulation to the more general case of frequency-selective channels, where the receiver adopts the form of a Multiple-Input Multiple-Output (MIMO) Decision-Feedback Equalizer (DFE).

With the exploding interest in space-time processing and multiuser detection in recent years, MIMO DFEs have attracted significant attention. The optimal settings—in the Minimum Mean-Square Error (MMSE) sense–for the MIMO DFE were derived in [8] and [9] within the context of Code-Division Multiple Access (CDMA). However, the settings derived therein correspond, in general, to infinite-length non-causal filters. Furthermore, 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 filters. Those optimal settings, again in the MMSE sense, have recently been reported in [10] and [11] for a single-stage receiver. In this contribution, we present the multiple-stage counterpart, which is a more natural extension of the narrowband V-BLAST.

Throughout the remainder, we utilize $(\cdot)^T$, $(\cdot)^*$ and $(\cdot)^H$ to denote matrix transposition, conjugation and hermitian transposition respectively.

II. System and Channel Models

We consider a discrete-time complex baseband model for a single-user link, assuming perfect carrier recovery and downconversion. Received signals are sampled at the symbol rate and, therefore, the receiver structure we present is symbol-spaced[1]. It is assumed that

[1]The front stage of a practical receiver implementation would almost certainly be fractionally spaced. For the purpose of exploring architectures, however, symbol spacing is a valid start. The extension to the fractionally spaced case is relatively straightforward [12].

the channel is stationary over every burst of data, although it changes from burst to burst. Perfect channel estimation at the receiver is further assumed [13].

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

$$\mathbf{h}_{nm} = \left[\begin{array}{cccc} h_{nm}(0) & h_{nm}(1) & \ldots & h_{nm}(L) \end{array} \right]^T . \quad (1)$$

The signal transmitted at time k is the M-dimensional vector $\mathbf{s}(\mathrm{k})$ with spatial covariance

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

where P_T is the total average transmit power, which is held constant irrespectively of the number of transmit antennas. The receiver Additive White Gaussian Noise (AWGN) can be expressed, in turn, as an N-dimensional vector $\mathbf{n}(k)$ with spatial covariance

$$E[\mathbf{n}(k)\mathbf{n}^H(k)] = \sigma^2\mathbf{I}^{N\times N}. \quad (3)$$

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

$$\mathbf{H}_m = \left[\begin{array}{c} \mathbf{h}_{1m}^T \\ \mathbf{h}_{2m}^T \\ \vdots \\ \mathbf{h}_{Nm}^T \end{array} \right] \quad ; \quad m = 1, 2, \ldots, M. \quad (4)$$

With that, we can express the N-dimensional received vector $\mathbf{x}(k)$ as

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

with the sequence of $L+1$ symbols transmitted by the m-*th* antenna denoted by

$$\mathbf{s}_m(k) = \left[\begin{array}{c} s_m(k) \\ s_m(k-1) \\ \vdots \\ s_m(k-L) \end{array} \right]. \quad (6)$$

We can now define ρ as the expected SNR—over the ensemble of all possible channel realizations—on any one of the receive antennas, which is independent of M [4] and can be expressed as

$$\rho = \frac{P_T}{M\sigma^2} \sum_{m=1}^{M} E\|\mathbf{h}_{nm}\|^2 \quad (7)$$

for any value of n. Finally, and in order to facilitate the formulation in the next section, we introduce [12] the operator $vec(\cdot)$ as

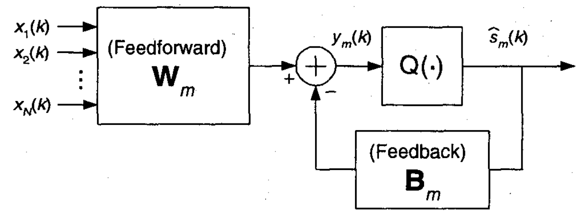

Fig. 1. Structure of an individual MISO receiver stage.

$$vec\left(\left[\begin{array}{cccc} \mathbf{v}_1 & \mathbf{v}_2 & \ldots & \mathbf{v}_K \end{array} \right]\right) = \left[\begin{array}{c} \mathbf{v}_1 \\ \mathbf{v}_2 \\ \vdots \\ \mathbf{v}_K \end{array} \right] \quad (8)$$

III. Space-Time MMSE V-BLAST Receiver

The receiver consists of M successive stages, one of which is shown in Fig. 1, each having a feedforward section $\mathbf{W}$ with K_f+1 taps per antenna and a feedback section $\mathbf{B}$ with 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 [12] 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 proved to be globally optimal at asymptotically high SNR [4]. The K_f+1 N-dimensional symbols spanned by the feedforward section can be grouped as

$$\mathbf{X}(k) = \left[\begin{array}{cccc} \mathbf{x}(k) & \mathbf{x}(k-1) & \ldots & \mathbf{x}(k-K_f) \end{array} \right]. \quad (9)$$

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

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

where the extended sequence of symbols from the m-*th* transmitter is

$$\bar{\mathbf{s}}_m(k) = \left[\begin{array}{c} s_m(k) \\ s_m(k-1) \\ \vdots \\ s_m(k-L-K_f) \end{array} \right] \quad (11)$$

and

$$\bar{\mathbf{H}}_m = \left[\begin{array}{ccc} \mathbf{H}_m & \cdots & 0 \\ & \ddots & \\ 0 & \cdots & \mathbf{H}_m \end{array} \right] \quad (12)$$

is a block Toeplitz matrix with $N(K_f+1)$ rows and $L+K_f+1$ columns.

The output of every stage at time k is an estimate $\hat{s}_m(k-d)$ for whichever stream m was selected by the ordering mechanism. The decision delay d is a parameter which—for now—will be assumed identical for all stages. The sequence of K_b most recent decisions produced by every stage is labeled as

$$\hat{\mathbf{s}}_m(k-d-1) = \begin{bmatrix} \hat{s}_m(k-d-1) \\ \vdots \\ \hat{s}_m(k-d-K_b) \end{bmatrix} \quad (13)$$

and constitutes the input to the corresponding feedback section.

At each stage, every undetected stream is a candidate for selection. Assuming correct decisions at all stages, i.e. $\hat{s}_m(k) = s_m(k)$ for all m and k, 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

$$y_m(k) = \mathrm{Tr}\{\mathbf{W}_m^H \mathbf{X}(k)\} - \mathbf{B}_m^H \hat{\mathbf{s}}_m(k-d-1) \quad (14)$$

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

$$\begin{aligned} y_m(k) &= \boldsymbol{w}_m^H \boldsymbol{x}(k) - \mathbf{B}_m^H \hat{\mathbf{s}}_m(k-d-1) \quad (15) \\ &= \begin{bmatrix} \boldsymbol{w}_m \\ \mathbf{B}_m \end{bmatrix}^H \begin{bmatrix} \boldsymbol{x}(k) \\ -\hat{\mathbf{s}}_m(k-d-1) \end{bmatrix} \end{aligned}$$

The MMSE Wiener-Hopf solution for $\boldsymbol{w}_m$ and $\mathbf{B}_m$ is

$$\begin{bmatrix} \boldsymbol{w}_m \\ \mathbf{B}_m \end{bmatrix} = \mathbf{R}^{-1}\mathbf{P}_m \quad (16)$$

with [12]

$$\mathbf{P}_m = E\begin{bmatrix} \boldsymbol{x}(k) \\ -\hat{\mathbf{s}}_m(k-d-1) \end{bmatrix} s_m^*(k-d) = \begin{bmatrix} (\bar{\mathbf{H}}_m)_{d+1} \\ \mathbf{0} \end{bmatrix}$$

where $(\cdot)_d$ indicates the d-th column of the corresponding matrix, and with

$$\begin{aligned} \mathbf{R} &= E\left(\begin{bmatrix} \boldsymbol{x}(k) \\ -\hat{\mathbf{s}}_m(k-d-1) \end{bmatrix} \begin{bmatrix} \boldsymbol{x}(k) \\ -\hat{\mathbf{s}}_m(k-d-1) \end{bmatrix}^H\right) \\ &= \begin{bmatrix} \sum_i \bar{\mathbf{H}}_i \bar{\mathbf{H}}_i^H + \sigma^2 \mathbf{I} & -(\bar{\mathbf{H}}_m)_{d+2\cdots d+K_b+1} \\ -(\bar{\mathbf{H}}_m)_{d+2\cdots d+K_b+1}^H & \mathbf{I} \end{bmatrix} \end{aligned}$$

where the i summation is over the undetected data streams.

Given that the MSE for stream m is

$$E|y_m(k) - s_m(k-d)|^2 = 1 - \mathbf{P}_m^H \mathbf{R}^{-1} \mathbf{P}_m \quad (17)$$

the undetected stream with the smallest MSE can be selected at every stage and extracted using the $\boldsymbol{w}_m$ and $\mathbf{B}_m$ settings defined by (16).

Finally, once a stream has been detected, its interference contribution can be removed from the input signal $\mathbf{x}(k)$ as follows

TABLE I

TU CHANNEL: PATH DELAYS AND RELATIVE POWER LEVELS

	path delay (symbols)	relative power (dB)
1	0.0	0.0
2	0.027	-4.0
3	0.054	-8.0
4	0.082	-12.0
5	0.109	-16.0
6	0.136	-20.0

$$\begin{aligned} \mathbf{x}(k) &\leftarrow \mathbf{x}(k) - (\mathbf{H}_m)_1 \hat{s}_m(k) \\ \mathbf{x}(k+1) &\leftarrow \mathbf{x}(k+1) - (\mathbf{H}_m)_2 \hat{s}_m(k) \\ &\vdots \\ \mathbf{x}(k+L) &\leftarrow \mathbf{x}(k+L) - (\mathbf{H}_m)_{L+1} \hat{s}_m(k) \end{aligned} \quad (18)$$

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.

IV. Performance Evaluation

In this section, we will evaluate—via simulation—the performance of the proposed receiver in a stationary (no Doppler) Typical Urban (TU) channel [14] with spatial rich scattering. The TU channel profile, with a normalized rms delay spread τ=0.2886 symbols, is detailed in Table I. We use QPSK modulation and square-root raised cosine transmit and receive filters with 35% excess bandwidth. There is no coding. Since detailed analysis of spectral efficiency versus number of antennas were presented—for the narrowband case—in [4], here we concentrate on studying the effect of different delay spreads for a given configuration, which is chosen to be 4×6.

The integer decision delay d and the sampling phase are free parameters. In a practical implementation, they would be adjusted by some synchronization mechanism. In our simulations, they are always chosen to minimize the Bit Error Rate (BER). As an example, we present in Fig. 2 the BER as a function of the combined delay and sampling phase—normalized by the symbol time—for a (4,2) receiver. The different local minima, representing different alignments of the channel and the feedforward taps, correspond to various values of d with the optimal sampling phase .

In Fig 3, the BER is shown as a function of SNR for a multiplicity of receivers, namely (1,0),(2,1),(4,2) and (6,3). The (1,0) case, corresponding to the narrowband receiver, displays a rather poor behavior whereas a simple (2,1) configuration handles the frequency selectivity rather well. In fact, the returns diminish very rapidly at around (4,2) and little gain is obtained with additional taps. Notice also that, with enough taps, the flat-fading equivalent—shown in dashed—is outperformed. This is an indication that the space-time

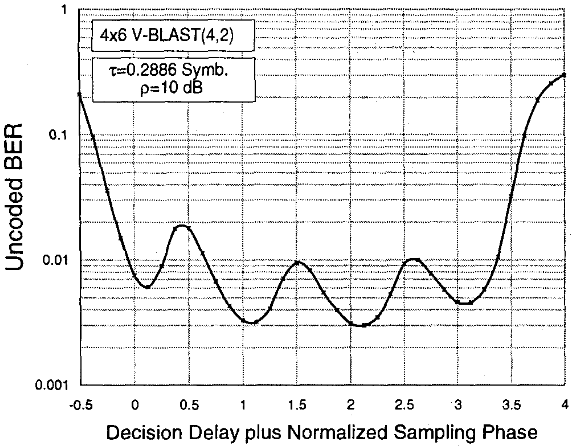

Fig. 2. BER vs. combined decision delay and normalized sampling phase with respect to the first ray for a 4×6 BLAST(4,2) with ρ=10 dB.

receiver is exploiting the frequency diversity provided by the TU channel.

A fundamental issue is that of the 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 larger the level of 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 overly pessimistic results. Span analysis for MISO systems can be found in the literature. (see [12], [15], and [16], for instance). Nonetheless, it is not clear how well these techniques extend to multi-stage MIMO systems, where errors in the interference cancellation process may propagate through multiple stages. In our approach, based on simulation, the required span is that which yields a desired BER level. Accordingly, we present in Fig. 4 the BER as a function of the normalized delay spread in the asymptotically high SNR regime ($\rho \to \infty$), where the MMSE receiver behaves in zero-forcing mode. Based on these results, we outline some rules—which mostly agree with those reported in [10] for single-stage MIMO—relating K_f, K_b, and d:

1. The decision delay, d, should not exceed K_f, but should be larger than L if possible. Making $d>L$ leads to negligible improvement.
2. K_b should be large enough to cancel all postcursor taps, that is, $K_b \simeq K_f + L - d$. If the receiver has to operate at relatively large BER levels, it might be advantageous to reduce K_b so as to minimize the impact of error propagation.
3. K_f should only be as large as necessary (Fig. 4).

V. Summary

We have formulated a space-time V-BLAST receiver, optimized in the MMSE sense, which can operate in frequency-selective environments. The receiver, which is a generalization of the original V-BLAST receiver presented in [4], is architectured as multi-stage with interference cancellation at every stage. Each stage adopts the form of a MISO DFE, with the stage ordering optimized according to the MMSE criterion. The performance of this receiver has been evaluated, using a TU channel, with excellent results. As future work, it is essential to extend the BLAST channel estimation analysis of [13] to the wideband case and to address its impact on the required receiver span. In addition, it would be interesting to explore the possibility of using different filter spans and delay decisions at different stages in the detection process.

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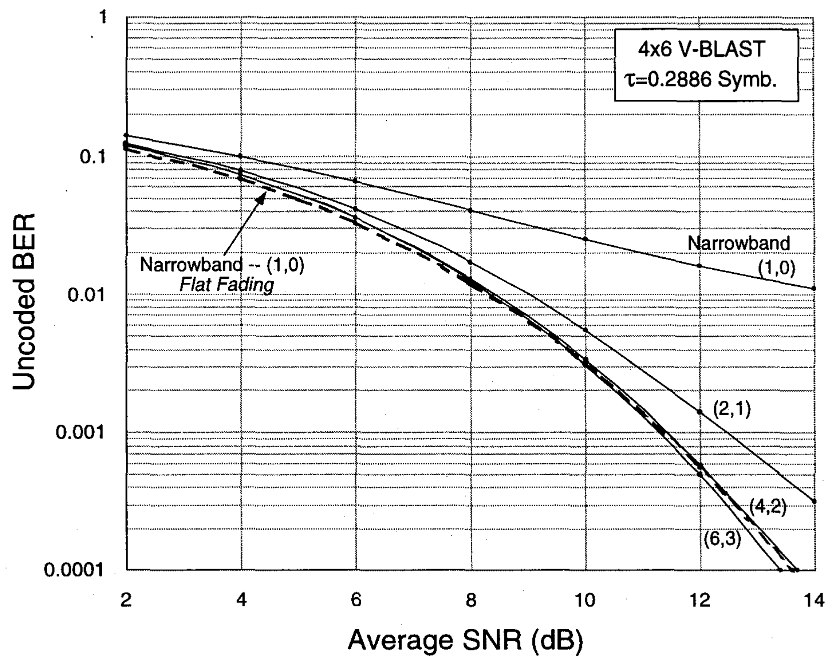

Fig. 3. BER vs. SNR for different space-time configurations (including narrowband) are shown in solid. The performance of the narrowband receiver in a flat-fading channel with identical SNR is shown in dashed.

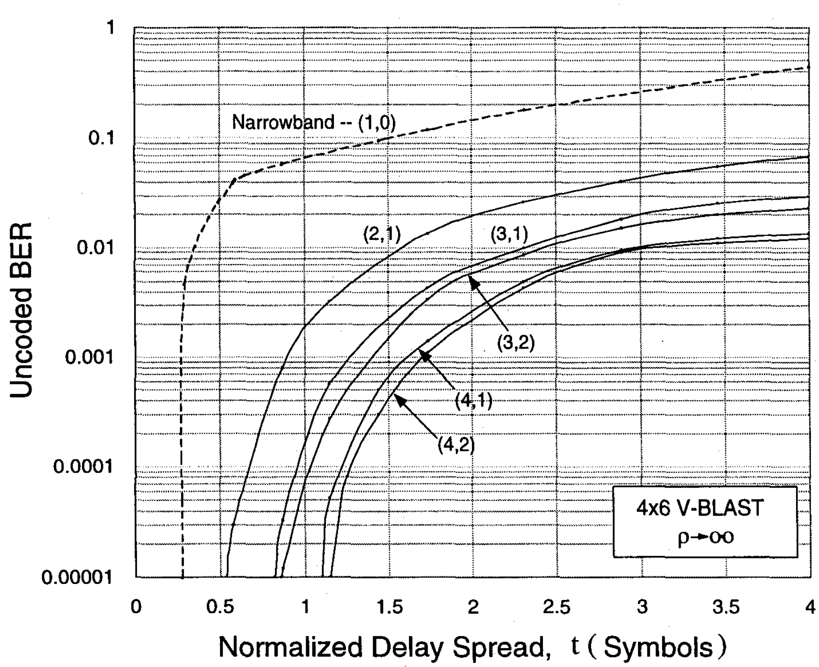

Fig. 4. BER vs. normalized delay spread for different configurations with $\rho \to \infty$. The dashed line corresponds to the narrowband receiver.