Detection algorithm and initial laboratory results using V-BLAST space-time communication architecture

G.D. Golden, C.J. Foschini, R.A. Valenzuela and P.W. Wolniansky

The signal detection algorithm of the vertical BLAST (Bell Laboratories Layered Space-Time) wireless communications architecture is briefly described. Using this joint space-time approach, spectral efficiencies ranging from 20–40bit/s/Hz have been demonstrated in the laboratory under flat fading conditions at indoor fading rates. Early results are presented.

Background: Recent information theory research has shown that the rich-scattering wireless channel is capable of enormous theoretical capacities if the multipath is properly exploited [1 – 4]. The diagonally-layered space-time architecture proposed by Foschini [1], now known as D-BLAST, uses multielement antenna arrays at both transmitter and receiver and an elegant diagonally-layered coding structure in which code blocks are dispersed across diagonals in space-time. In an independent Rayleigh scattering environment, this processing structure leads to theoretical rates which grow linearly with the number of transmit antennas, with these rates approaching 90% of Shannon capacity. However, the diagonal approach suffers from certain implementation complexities which make it inappropriate for initial implementation. In this Letter, we describe a simplified version of the BLAST detection algorithm, known as vertical BLAST, or V-BLAST, which has been implemented in realtime in the laboratory [5, 6]. Using our laboratory prototype, we have demonstrated spectral efficiencies as high as 40bit/s/Hz in an indoor slow-fading environment.

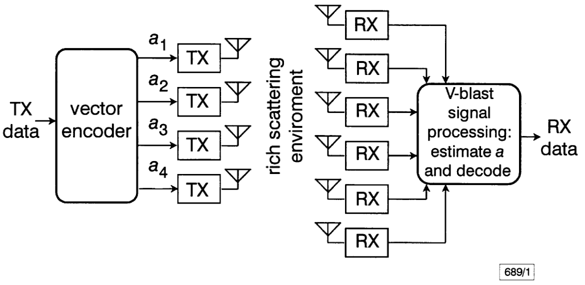

Fig. 1 *V-BLAST high-level system diagram*

Vector symbol: $\mathbf{a} \equiv (a_1, a_2, a_3, a_4)^T$
Number of transmitters = M; number of receivers = N

System description: The V-BLAST system diagram is shown in Fig. 1. QAM transmitters 1 to M operate co-channel at symbol rate $1/T$ symbol/s, with synchronised symbol timing. The collection of transmitters comprises, in effect, a vector-valued transmitter, where components of each transmitted M-vector are symbols drawn from a QAM constellation. For simplicity in the sequel, we assume that the same constellation is used for each component, and that transmissions are organised into bursts of L symbols. The power launched by each transmitter is proportional to $1/M$ so that the total radiated power is constant and independent of M.

Receivers 1 to N are, individually, conventional QAM receivers. The receivers also operate co-channel, each receiving the signals radiated from all M transmit antennas. Flat fading is assumed, and the matrix channel transfer function is $\mathbf{H}^{N\times M}$, where h_{IJ} is the (complex) transfer function from transmitter j to receiver i, and $M \le N$. We take the quasi-stationary viewpoint that the channel time variation is negligible over the L symbol periods comprising a burst, and that the channel is estimated accurately, e.g. by use of a training sequence embedded in each burst. Thus, for brevity, we will not make the distinction between $\mathbf{H}$ and its estimate. In what follows, we take a discrete-time baseband view of the detection process of a single transmitted vector symbol, assuming symbol-synchronous receiver sampling and ideal timing.

Letting $\mathbf{a} = (a_1, a_2, ..., a_M)^T$ denote the vector of transmit symbols, then the corresponding received N-vector is

$$\mathbf{r}_1 = \mathbf{H}\mathbf{a} + v \qquad (1)$$

where v is a WSS noise vector with IID components.

Detection algorithm: Let the ordered set

$$S \equiv \{k_1, k_2, ..., k_M\} \qquad (2)$$

be a permutation of the integers 1, 2, ..., M specifying the order in which components of the transmitted symbol vector $\mathbf{a}$ are extracted. The detection algorithm operates on $\mathbf{r}_1$, progressively computing decision statistics $y_{K_1}, y_{K_2}, ..., y_{K_M}$, which are then sliced to form estimates of the underlying data symbols $\hat{a}_{K_1}, \hat{a}_{K_2}, ..., \hat{a}_{K_M}$. Thus, decision statistic y_{K_1} is computed first, then y_{K_2}, and so on. Later we show how to determine a particular ordering S_{OPT} which is optimal in a certain sense; for now, we discuss the general detection procedure with respect to an arbitrary ordering S.

The detection process uses linear combinatorial nulling and symbol cancellation to successively compute the y_{K_I}, proceeding generally as follows:

Step 1: Using nulling vector $\mathbf{w}_{K_1}$, form a linear combination of the components of $\mathbf{r}_1$ to yield y_{K_1}:

$$y_{k_1} = \mathbf{w}_{k_1}^T \mathbf{r}_1 \qquad (3)$$

Step 2: Slice y_{K_1} to obtain $\hat{a}_{K_1}$:

$$\hat{a}_{k_1} = Q(y_{k_1}) \qquad (4)$$

where $Q(\cdot)$ denotes the quantisation (slicing) operation appropriate to the constellation in use.
Step 3: Assuming that $\hat{a}_{K_1} = a_{K_1}$, cancel a_{K_1} from the received vector $\mathbf{r}_1$, resulting in modified received vector $\mathbf{r}_2$:

$$\mathbf{r}_2 = \mathbf{r}_1 - \hat{a}_{k_1}(\mathbf{H})_{k_1} \qquad (5)$$

where $(\mathbf{H})_{K_1}$ denotes the k_1th column of $\mathbf{H}$. Steps 1–3 are then performed for components $k_2, ..., k_M$ by operating in turn on the progression of modified received vectors $\mathbf{r}_2, \mathbf{r}_3, ..., \mathbf{r}_M$.

The specifics of the detection process depend on the criterion chosen to compute the nulling vectors $\mathbf{w}_{K_I}$, the most common choices being minimum mean-squared error (MMSE) and zero-forcing (ZF). The detection process is described here with respect to the latter since it is simpler. The k_Ith ZF-nulling vector is defined as the unique minimum norm vector satisfying

$$\mathbf{w}_{k_i}^T(\mathbf{H})_{k_j} = \begin{cases} 0 & j > i \\ 1 & j = i \end{cases} \qquad (6)$$

Thus, the k_Ith ZF-nulling vector is orthogonal to the subspace spanned by the contributions to $\mathbf{r}_I$ due to those symbols not yet estimated and cancelled. It is not difficult to show that the unique vector satisfying eqn. 6 is just the k_Ith row of $\mathbf{H}^{+}_{\overline{K_{I-1}}}$ where the notation $\mathbf{H}_{\overline{K_I}}$ denotes the matrix obtained by zeroing columns $k_1, k_2, ..., k_I$ of $\mathbf{H}$ and $^+$ denotes the Moore-Penrose pseudoinverse [7].

MMSE nulling is discussed in more detail in the adaptive array literature, e.g. [8]. In either case, however, the noise power of the k_Ith decision statistic y_{K_I} is just proportional to $\|\mathbf{w}_{K_I}\|^2$, and thus the post-detection SNRs are proportional to $1/\|\mathbf{w}_{K_I}\|^2$.

The full ZF detection algorithm can be described compactly as a recursive procedure, including determination of the optimal ordering, as follows:

$$\text{initialisation:} \quad \mathbf{G}_1 = \mathbf{H}^+ \qquad (7a)$$

$$i = 1 \qquad (7b)$$

$$\text{recursion:} \quad k_i = \underset{j \notin \{k_1 \dots k_{i-1}\}}{\operatorname{argmin}} \|(\mathbf{G}_i)_j\|^2 \qquad (7c)$$

$$\mathbf{w}_{k_i} = (\mathbf{G}_i)_{k_i} \qquad (7d)$$

$$\mathbf{y}_{k_i} = \mathbf{w}_{k_i}^T \mathbf{r}_i \qquad (7e)$$

$$\hat{a}_{k_i} = Q(y_{k_i}) \qquad (7f)$$

$$\mathbf{r}_{i+1} = \mathbf{r}_i - \hat{a}_{k_i}(\mathbf{H})_{k_i} \qquad (7g)$$

$$\mathbf{G}_{i+1} = \mathbf{H}^{+}_{\overline{k_i}} \qquad (7h)$$

$$i = i + 1 \qquad (7i)$$

where $(\mathbf{G}_I)_J$ is the jth row of $\mathbf{G}_I$. Thus, eqn. 7*c* determines the elements of S_{OPT}, the optimal ordering, discussed below. Eqn. 7*d*–*f* compute, respectively, the ZF-nulling vector, the decision statistic, and the estimated component of $\mathbf{a}$. Eqn. 7*g* performs cancellation of the detected component from the received vector, and eqn. 7*h* computes the new pseudoinverse for the next iteration. Note that this new pseudoinverse is based on a 'deflated' version of **H**, in which columns k_1, k_2, ..., k_I have been zeroed. This is because these columns correspond to components of **a** which have already been estimated and cancelled, and thus the system becomes equivalent to a 'deflated' version of Fig. 1 in which transmitters k_1, k_2, .., k_I have been removed, or equivalently, a system in which $a_{K_1} = ... = a_{K_I} = 0$.

Determination of S_{OPT}*:* Recall that all components of **a** are assumed to utilise the same constellation. Under this assumption, the y_{K_I} with the lowest post-detection SNR will dominate the error performance of the detection process. An important aspect of the nonlinear processing in this scheme is that, due to symbol cancellation, these post-detection SNRs depend on the order in which the decision statistics are computed. Thus, an obvious figure of merit for this system, though not the only one possible, is the maximisation of the worst, i.e. the minimum, of these post-detection SNRs. It can be shown that the local optimisation (eqn. 7*c*) of choosing the component with the best SNR at each stage, leads, somewhat surprisingly, to the global optimum S_{OPT} in this maximin sense. The proof is given in [6].

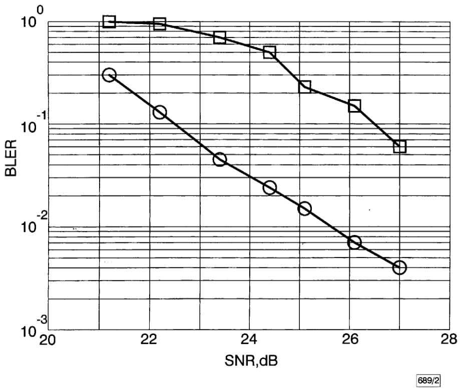

Fig. 2 *Block error rate (BLER) against SNR for M = 8, N = 12 V-BLAST laboratory prototype operating at 25.9bit/s/Hz*

□ nulling alone (no symbol cancellation)
○ nulling with optimally-ordered symbol cancellation

Laboratory results: Fig. 2 shows results obtained using the system of Fig. 1 with $M = 8$ transmitters and $N = 12$ receivers. The horizontal axis is spatially averaged received SNR, i.e. $1/N\Sigma_{i=1}^{N} SNR_i$, where SNR_i is the the ratio of received signal power (from all M transmitters) to noise power at the ith receiver. The system was operated at a carrier frequency of 1.9GHz and symbol rate 24.3ksymbols/s in a bandwidth of 30kHz, utilising uncoded 16-QAM on each transmitter, yielding a raw spectral efficiency of

$$\frac{(8\,\text{xmtrs}) \times (4\,\text{bit/sym/xmtr}) \times (24.3\,\text{ksym/s})}{30\,\text{kHz}}$$
$$= 25.9\,\text{bit/s/Hz}$$

The burst length L is 100 symbols, 20 of which are used to estimate the channel on each burst, so that the payload efficiency is 80% of the raw spectral efficiency, or 20.7bit/s/Hz. At 34dB SNR, spectral efficiencies as high as 40bit/s/Hz have been demonstrated at similar error rates. All results were obtained in a short-range indoor environment with negligible delay spread.

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