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


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 multi-element 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 real-time in our laboratory prototype, we have demonstrated spectral efficiencies as high as 40bit/s/Hz in an indoor slow-fading environment.

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 vector symbol, assuming symbol-synchronous communication architecture, e.g. [8]. In either case, however, the noise power of the decision statistic \( y_k = w_k^T r_k \) is just proportional to \( ||r_k||^2 \), and thus the post-detection SNR is proportional to \( 1/||w_k||^2 \).

The specifics of the detection process depend on the criterion chosen to compute the nulling vectors \( w_k \), the most common choices being minimum mean-square error (M M SE) and zero-forcing (ZF). The detection process is described here with respect to an arbitrary ordering of \( r_k \) to simplify. The kth ZF-nulling vector is defined as the unique minimum norm vector satisfying

\[
\begin{align*}
\mathbf{w}_k^H (\mathbf{H})_{kk} &= \begin{cases} 0 & j > i \\ 1 & j = i \end{cases} \\
\end{align*}
\]

Thus, the kth ZF-nulling vector is orthogonal to the subspace spanned by the contributions to \( r_k \) 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 kth row of \( \mathbf{H} \), where the notation \( \mathbf{H} \) denotes the matrix obtained by zeroing columns \( k_1, k_2, \ldots, k_0 \) of \( \mathbf{H} \) and \( * \) denotes the M ore-Penrose pseudoinverse [7].

M M SE nulling is discussed in more detail in the adaptive array literature, e.g. [8]. In either case, however, the noise power of the kth decision statistic \( y_k \) is just proportional to \( ||w_k||^2 \), and thus the post-detection SNR is proportional to \( 1/||w_k||^2 \).

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

\begin{align*}
\text{initialisation:} & \quad G_1 = \mathbf{H}^+ \\
& \quad i = 1 \\
\text{recursion:} & \quad \mathbf{l}_k = \text{argmin}_{j \in \{1,2,\ldots,i-1\}} \| (\mathbf{G}_j)_{ik} \|^2 \\
& \quad \mathbf{w}_{ik} = (\mathbf{G}_j)_{ik} \\
& \quad y_{ik} = \mathbf{w}_{ik}^H r_k \\
& \quad \tilde{a}_{ik} = Q(y_{ik}) \\
& \quad r_{i+1} = r_k - \tilde{a}_{ik} (\mathbf{H})_{kk} \\
& \quad \mathbf{G}_{i+1} = \mathbf{H}^+_{\mathbf{l}_i} \\
& \quad i = i + 1
\end{align*}

where \( (\mathbf{G}_j)_{ik} \) is the jth row of \( \mathbf{G}_j \). Thus, eqn. 7c determines the elements of \( \mathbf{S}_{opt} \), the optimal ordering, discussed below. Eqn. 7d-f compute, respectively, the ZF-nulling vector, the decision statistic, and the estimated component of \( \mathbf{a} \). Eqn. 7g performs cancellation of the detected component from the received vector, and eqn. 7h computes
the new pseudoinverse for the next iteration. Note that this new pseudoinverse is based on a ‘deflated’ version of \( H \), in which columns \( k_1, k_2, \ldots, k_L \) 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 \( k_1, k_2, \ldots, k_L \) have been removed, or equivalently, a system in which \( a_{k_1} = \ldots = a_{k_L} = 0 \).

Determination of \( S_{\text{OPT}} \): Recall that all components of \( a \) are assumed to utilise the same constellation. Under this assumption, the \( y_{k_1} \) 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. 7c) of choosing the component with the best SNR at each stage, leads, somewhat surprisingly, to the global optimum \( S_{\text{OPT}} \) in this maximin sense. The proof is given in [6].

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. \( 1/N \sum_{i=1}^{N} \text{SNR}_i \), where SNR is the ratio of received signal power (from all \( M \) transmitters) to noise power at the \( i \)-th receiver. The system was operated at a carrier frequency of 1.9 GHz and symbol rate 24.3ksym/s in a bandwidth of 30kHz, utilising uncoded 16-QAM on each transmitter, yielding a raw spectral efficiency of

\[
\frac{(8 \times \text{bits}) \times (41.8 \text{sym} \times \text{snr}) \times (24.3 \text{ksym/s})}{30 \text{kHz}} = 25.9 \text{bits/s/Hz}
\]

The burst length \( L \) is 100 symbols, 20 of which are used to estimate the channel on each burst, but 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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References