Journal of Southeast University (English Edition) Vol.20 No.2 June 2004 ISSN 1003—7985

Chip-level space-time equalization receiver scheme for MIMO HSDPA systems

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Abstract: A chip-level space-time equalization receiver scheme is proposed for multiple-input multiple-output high-speed downlink packet access (MIMO HSDPA) systems to jointly combat the co-channel interference and the inter-code interference. A fractional sample equalizer is also derived to further improve the performance of the receiver. Performance analysis and the calculation of the output signal to interference ratio (SINR) at each receiver antenna are presented to help direct the design of equalization weight in a more optimal manner. System simulations demonstrate the significant performance gain over conventional Rake receiver and high potential of MIMO HSDPA for high-data-rate packet transmission.

Key words: multiple-input multiple-output (MIMO); chip-level; interference; minimum mean square error (MMSE); weight; space-time equalization

Multiple-input multiple-output high-speed downlink packet access (MIMO HSDPA) is a Release 5 enhancement to the 3GPP W-CDMA wireless cellular standard. Multi-antenna transmission offers increased spectrum efficiency to achieve even greater downlink throughput. One of the solutions to implement MIMO is using vertical Bell labs layered space-time (V-BLAST) architecture on the downlink-shared channel (DSCH). However, in the frequency-selective fading channel, the channel delay spread corrupts the orthogonality of spreading code, resulting in severe inter-code interference, especially in HSDPA systems where a large number of orthogonal spreading codes are employed[1,2]. The interference is so severe that conventional rake receiver techniques do not account for the interference, providing unacceptably poor performance. One technique to account for the interference is the multipath interference canceller[3]. This technique is shown to be effective in a 2-path channel with equal average power paths and one chip delay offset. However, its efficacy has proven to be worse than equalization for more realistic channels consisting of several multipath components with fractional chip delays[4,5]. Meanwhile, multiple antennas transmission simultaneously brings co-channel interferences between antennas and multipath interference between the delay replicas of signals from different transmit antennas[6-8].

In this contribution, we propose an MIMO receiver architecture applying chip-level minimum mean square error (MMSE) and space-time equalization to combat jointly the co-channel, multipath and inter-code interference encountered in frequency selective fading channels. Instead of match filtering at each receiver antenna, space-time combine and VBLAST detection are in the conventional receiver.

1 System Model

Considering M-transmit N-receive antennas configuration, we assume that the transmitted signal consists of JM substreams transmitted using code reuse with J orthogonal spreading codes on the high speed downlink shared channel (HS-DSCH)[8] over the M transmit antenna, and, for simplicity, ignore the other channels such as the common pilot channel (CPICH) and dedicated physical channels (DPCHs). The high speed data stream is coded, punctured, interleaved, mapped to symbols and demultiplexed into JM equal-rate substreams where J is the number of code channels with spreading factor G. Let $b_{j,m}$ denote the symbol from the m-th antenna ($m=1, \cdots, M$) spread by the j-th code ($j=1, \cdots, J$). Then the transmitted signal from the m-th antenna during this symbol period is

$$\boldsymbol{s}_m = \sum_{j=1}^{J} \boldsymbol{c}_j b_{j,m} = \boldsymbol{C}\boldsymbol{b}_m \qquad (1)$$

where $\boldsymbol{c}_j$ is the j-th spreading code, $\boldsymbol{C} = \{\boldsymbol{c}_1, \boldsymbol{c}_2, \cdots, \boldsymbol{c}_J\}$ is the G-by-J spreading code matrix, and $\boldsymbol{b}_m = \{b_{1,m}, b_{2,m}, \cdots, b_{J,m}\}^{\mathrm{T}}$ is the vector of data from antenna m.

Received 2003-09-22.

Foundation items: The National High Technology Research and Development Program of China (863 Program) (No.2002AA123031), Grant from Nokia Company.

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2 Equalization Derivation

As shown in Fig.1, the proposed MMSE equalizer jointly processes the baseband signal from all N receive antennas in an optimum manner. On each chip interval, the output of the equalizer is an M-dimensional complex vector. In the absence of thermal noise and under the assumption of a full rank channel matrix $\boldsymbol{H}$ (defined below), the MMSE equalizer is equivalent to a zero-forcing equalizer, of which the m-th component ($m=1, \cdots, M$) over successive chip intervals is the chip sequence of the signal transmitted from the m-th antenna, which is the sum of J data substreams. Despreading this signal with respect to the J codes, the outputs are collected, and passed to the demapper, deinterleaver, and decoder.

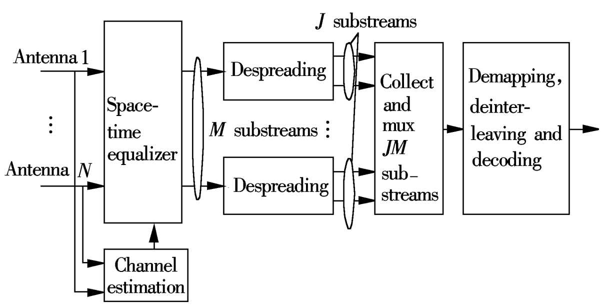

Fig.1 Block diagram of equalization receiver

2.1 MIMO received signal model

$\boldsymbol{x}_m=\{x_m(1), x_m(2), x_m(3), \cdots\}^{\mathrm{T}}$ denotes the vector of data transmitted over a given frame over the m-th antennas ($m=1, \cdots, M$) corresponding $\boldsymbol{s}_1$ in Eq.(1). Let E be the span of the equalizer measured in units of the chip period, $\boldsymbol{x}_m(k)$ be the $(E+L-1)$-dimensional subvector of $\boldsymbol{x}_m$ starting with the k-th term $x_m(k)$. Let L be the delay spread of the channel measured in units of the chip period, and P be the oversampling factor. $\boldsymbol{H}_{n,m}$ denotes the channel matrix composed of channel coefficients $h_{n,m,p}(l)$ as the channel coefficient between the m-th transmitter ($m=1, \cdots, M$) and the n-th receiver ($n=1, \cdots, N$) corresponding to the l-th path and the p-th oversample ($l=0, \cdots, L-1$, and $p=1, \cdots, P$). $n_{n,p}(k)$ is the additive zero-mean, complex Gaussian noise at the n-th antenna on the p-th sample of the k-th chip with variance $\sigma_n^2/2$ per dimension. Defining $\boldsymbol{n}_n(k)=\{n_{n,1}(k), \cdots, n_{n,P}(k), \cdots, n_{n,1}(k+E-1), \cdots, n_{n,P}(k+E-1)\}^{\mathrm{T}}$, then the received signal vector at the n-th antenna can be written as

$$\boldsymbol{y}_n(k)=\sum_{m=1}^{M}\boldsymbol{H}_{n,m}\boldsymbol{x}_m(k)+\boldsymbol{n}_n(k) \tag{2}$$

where

$$\boldsymbol{H}_{n,m}=\begin{bmatrix} h_{n,m,1}(L-1) & \cdots & h_{n,m,1}(0) & & & \\ \vdots & & \vdots & & & \\ h_{n,m,P}(L-1) & \cdots & h_{n,m,P}(0) & & & \\ & & & \ddots & & \\ & & & h_{n,m,1}(L-1) & \cdots & h_{n,m,1}(0) \\ & & & \vdots & & \vdots \\ & & & h_{n,m,P}(L-1) & \cdots & h_{n,m,P}(0) \end{bmatrix} \tag{3}$$

By stacking the received vectors and generalizing the definition of $\boldsymbol{y}(k)=\{\boldsymbol{y}_1^{\mathrm{T}}(k), \cdots, \boldsymbol{y}_N^{\mathrm{T}}(k)\}^{\mathrm{T}}$, we can obtain

$$\begin{Bmatrix} \boldsymbol{y}_1(k) \\ \vdots \\ \boldsymbol{y}_N(k) \end{Bmatrix}=\begin{bmatrix} \boldsymbol{H}_{1,1} & \cdots & \boldsymbol{H}_{1,M} \\ \vdots & & \vdots \\ \boldsymbol{H}_{N,1} & \cdots & \boldsymbol{H}_{N,M} \end{bmatrix}\begin{Bmatrix} \boldsymbol{x}_1(k) \\ \vdots \\ \boldsymbol{x}_M(k) \end{Bmatrix}+\begin{Bmatrix} \boldsymbol{n}_1(k) \\ \vdots \\ \boldsymbol{n}_N(k) \end{Bmatrix} \tag{4}$$

$$\boldsymbol{y}(k)=\boldsymbol{H}\boldsymbol{x}(k)+\boldsymbol{n}(k) \tag{5}$$

The sizes of $\boldsymbol{y}(k)$, $\boldsymbol{H}$, $\boldsymbol{x}(k)$, and $\boldsymbol{n}(k)$ are PEN-by-1, PEN-by-$M(E+L-1)$, $M(E+L-1)$-by-1, and PEN-by-1, respectively.

2.2 Fractional-sample space-time equalization

Let the components of $\boldsymbol{x}_{\mathrm{d}}(k)$ be an M-dimensional vector of which the m-th component is the transmitted signal from the m-th antenna with delay with respect to sample k, $\boldsymbol{x}_{\mathrm{d}}(k)=\{x_1(k+d), \cdots, x_M(k+d)\}^{\mathrm{T}}$. Given $\boldsymbol{y}(k)$, the MMSE equalizer[9] $\boldsymbol{W}_{\mathrm{d}}$ (an M-by-PEN complex matrix) minimizes the mean-square error between the equalizer output $\boldsymbol{W}_{\mathrm{d}}\boldsymbol{y}(k)$ and the desired M-dimensional output vector $\boldsymbol{x}_{\mathrm{d}}(k)$: $E[\|\boldsymbol{W}_{\mathrm{d}}\boldsymbol{y}(k)-\boldsymbol{x}_{\mathrm{d}}(k)\|^2]$. Assuming that the noise $\boldsymbol{n}(k)$ and the desired vector $\boldsymbol{x}_{\mathrm{d}}(k)$ are independent, the Weiner solution is given by

$$\begin{aligned}\boldsymbol{W}_{\mathrm{d}}&=E[\boldsymbol{x}_{\mathrm{d}}(k)\boldsymbol{y}^{\mathrm{H}}(k)]\{E[\boldsymbol{y}(k)\boldsymbol{y}^{\mathrm{H}}(k)]\}^{-1}= \\ &\sigma_x^2\boldsymbol{E}_{\mathrm{d}}\boldsymbol{H}^{\mathrm{H}}(\sigma_x^2\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}+\boldsymbol{R}_n)^{-1}= \\ &\boldsymbol{E}_{\mathrm{d}}\boldsymbol{H}^{\mathrm{H}}\left(\boldsymbol{H}\boldsymbol{H}^{\mathrm{H}}+\frac{1}{\sigma_x^2}\boldsymbol{R}_n\right)^{-1}\end{aligned} \tag{6}$$

where

$$E[\boldsymbol{x}_{\mathrm{d}}(k)\boldsymbol{x}^{\mathrm{H}}(k)]=\sigma_x^2\begin{bmatrix} \boldsymbol{e}_{\mathrm{d}} & \boldsymbol{z} & \cdots & \boldsymbol{z} \\ \boldsymbol{z} & \boldsymbol{e}_{\mathrm{d}} & & \vdots \\ \vdots & & \ddots & \boldsymbol{z} \\ \boldsymbol{z} & \cdots & \boldsymbol{z} & \boldsymbol{e}_{\mathrm{d}} \end{bmatrix}=\sigma_x^2\boldsymbol{E}_{\mathrm{d}} \tag{7}$$

where $\sigma_x^2=E[x_m(k)x_m^*(k)]$ is the chip power (independent of antenna m and time k), $\boldsymbol{e}_{\mathrm{d}}=[\underbrace{0\cdots0}_{d}1\ \underbrace{0\cdots0}_{E+L-2-d}]$ is an $(E+L-1)$-dimensional unit vector, $\boldsymbol{z}$ is an $(E+L-1)$-dimensional vector of zeroes, $\boldsymbol{E}_{\mathrm{d}}$ is the M-by-$M(L+E-1)$-matrix defined in Eq.(7), and $\boldsymbol{R}_n=E[\boldsymbol{n}(k)\boldsymbol{n}^{\mathrm{H}}(k)]$ is the noise covariance matrix. If we assume that the noise is white and uncorrelated among antennas, $\boldsymbol{R}_n=\sigma_n^2\boldsymbol{I}_{PEN}$ where

$\boldsymbol{I}_j$ denotes the j-by-j identity matrix.

We can write the equalization matrix in terms of its row vectors $\boldsymbol{W}_d = \{\boldsymbol{W}_{d,1}, \cdots, \boldsymbol{W}_{d,M}\}^{\mathrm{T}}$, so that the minimum mean square error $E[\,|\boldsymbol{W}_m \boldsymbol{y}(k) - x_m(k+d)|^2]$ is minimized. We also write the channel matrix $\boldsymbol{H}$ in terms of its $M(E+L-1)$ column vectors $\boldsymbol{H} = \{\boldsymbol{H}_{1,1}, \cdots, \boldsymbol{H}_{1,d+1}, \cdots, \boldsymbol{H}_{1,E+L-1}, \cdots, \boldsymbol{H}_{M,1}, \cdots, \boldsymbol{H}_{M,d+1}, \cdots, \boldsymbol{H}_{M,E+L-1}\}$, then the equalizer output for the m-th transmit antenna can be written as

$$\begin{aligned} \boldsymbol{W}_{\mathrm{d},m}\boldsymbol{y}(k) &= \boldsymbol{W}_{\mathrm{d},m}\boldsymbol{H}_{m,d+1}x_m(k+d) + \\ &\boldsymbol{W}_{\mathrm{d},m}\sum_{i=1,i\neq m}^{M}\sum_{j=1}^{E+L-1}\boldsymbol{H}_{i,j}x_i(k+j-1) + \\ &\boldsymbol{W}_{\mathrm{d},m}\sum_{j=1,j\neq d+1}^{E+L-1}\boldsymbol{H}_{m,j}x_m(k+j-1) + \boldsymbol{W}_{\mathrm{d},m}\boldsymbol{n}(k) \end{aligned} \tag{8}$$

where the first term is the desired signal term, the second is the interference from other antennas, the third is the self-interference from the m-th antenna, and the last term is the contribution from the filtered noise. The signal to interference noise ratio (SINR) for the m-th antenna is

$$\mathrm{SINR}_m = \frac{|\boldsymbol{W}_{\mathrm{d},m}\boldsymbol{H}_{m,d+1}|^2}{\sum_{i=1,i\neq m}^{M}\sum_{j=1}^{E+L-1}|\boldsymbol{W}_{\mathrm{d},m}\boldsymbol{H}_{i,j}|^2 + \sum_{j=1,j\neq d+1}^{E+L-1}|\boldsymbol{W}_{\mathrm{d},m}\boldsymbol{H}_{m,j}|^2 + M\boldsymbol{W}_{\mathrm{d},m}\boldsymbol{R}_n\boldsymbol{W}_{\mathrm{d},m}^{\mathrm{H}}} \tag{9}$$

The MMSE equalizer $\boldsymbol{W}_{\mathrm{d}}$ simultaneously accounts for three types of interference with respect to a desired data stream corresponding to a given code and transmit antenna, the self-interference (intersymbol interference) due to the multipath delay spread, multipath interference from data streams spread by other codes, and spatial interference from data streams sharing the same code but transmitted from other antennas.

Because the MMSE equalizer is only dependent on the power of the chip sequences and not their actual values, $\boldsymbol{W}_{\mathrm{d}}$ is independent of the time index k. Hence it needs to be recomputed at the rate of significant channel variations. Using an alternative MMSE detector design which depends on the spreading sequences, the equalizer taps would have to be updated whenever either the spreading codes change or the channel changes significantly. In systems with long spreading codes, the spreading codes change every symbol, therefore the equalizers would have to be computed each symbol interval. The computations would be in an enormous computational burden.

3 Computer Simulation

We performed link level simulations and measured the frame error rate (FER) for a system equipped with 2 transmits and 2 receive antennas. We consider two transport block formats, one with QPSK modulation and the other with 16QAM modulation, each with 5 code channels. The channel simulated is a typical urban channel profile ITU Pedestrian A profile with vehicle speed 3 km/h.

The chip-level MMSE equalizer uses 64 taps and 2 times oversampling. The data power is assumed to be 80% of the total transmitted power. In Fig.2, the performance of the rake receiver is interference limited, even for QPSK constellations. The advanced receiver provides a performance improvement of over 6 dB at 10% FER. For 16QAM modulation in Fig.3, the conventional receiver's FER is pegged to unity no matter how high the E_{b}/N_0 is. In contrast, the advanced receiver performs well without an error floor as low as 1% FER.

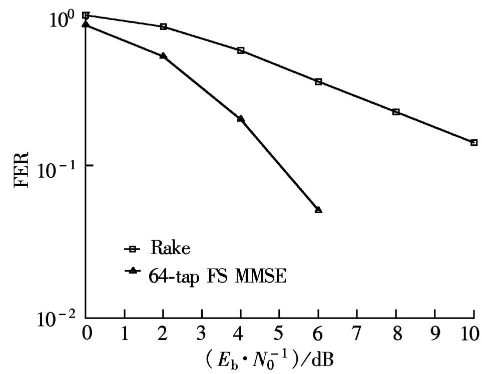

Fig.2 QPSK modulation link level performance

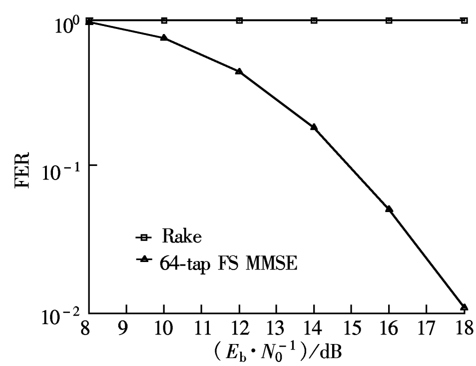

Fig.3 16QAM modulation link level performance

4 Conclusion

We have shown that the performance of the conventional receiver architecture for MIMO HSDPA results in an error floor in a frequency selective typical urban channel. A chip-level fractional-sample MMSE (FS-MMSE) space-time equalizer architecture is proposed which can effectively mitigate the effects of interference caused by the frequency selective fading channel and co-channel transmission. Performance results indicate that the MMSE equalizer followed by despreading and decoding not only removes the error floor but also provides performance improvements due to multipath diversity over the flat channel.

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MIMO HSDPA 系统中的 chip 级空时均衡接收方案

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摘要: 提出了一种频率选择性衰落环境下多输入多输出高速下行分组接入(MIMO HSDPA)系统中 chip 级的空时均衡接收方案, 以消除多天线和多码扩频传输时产生的严重的码间干扰和共信道间干扰; 并推导出分数采样时的空时均衡接收机以进一步提高接收机性能, 然后给出分析和接收信干噪比来设计最优权重. ITU Pedestrian A 信道环境下的仿真结果验证了本文的分析和接收机优越的性能.

关键词: 多输入多输出; chip 级; 干扰; 最小均方误差; 权系数; 空时均衡

中图分类号: TN929. 533