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D3.2

Implementation of Relevant algorithms

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Abstract:

This deliverable provides a high level description of the software developed within the I-METRA project following the selection reported in D3.1 "Design, Analysis and Selection of Suitable Algorithms".

Keyword list: Adaptive Antennas, MIMO Systems, Space-Time Coding, Adaptive Modulation and Coding, HSDPA, HSUPA, Turbo Coding.

0 EXECUTIVE SUMMARY

This deliverable provides a high level description of the software developed within the I-METRA project for the HSDPA and HSUPA of UMTS following the selection reported in D3.1 "Design, Analysis and Selection of Suitable Algorithms".

With respect to the HSDPA services, only the high level description of link level simulations is presented. The transmitter follows the HS-DSCH specifications already reported in D3.1 in such a way that this deliverable is mainly focused to describe the software of the space-time block codes implemented in the transmitter as well as the reception schemes.

The herein reported HSDPA transmission techniques include the transmit antenna switching technique, four different space-time block codes based on the STTD (i.e., CSTD, Trombi, DSTTD and DABBA), the well-known V-BLAST system and the more general Linear Dispersion Codes. The channel is implemented using the MIMO channel model developed in WP2. Regarding the receiver, several schemes are described: the Interference Cancelling and Nulling technique developed for V-BLAST, the LMMSE Space-Time Equalizer technique and different suboptimal ML receivers (such as, the “Turbo algorithm”).

With respect to HSUPA, the simulation model includes full 3GPP features and has been designed for both a low-complexity UE (1 antenna UE) and a dual-antenna with MIMO capability. This approach allows a common reference point for advanced MIMO concept as well as realistic signal models and receiver implementation schemes. A standard Rake receiver structure is assumed but the model allows also the implementation of more complex receivers with interference suppression in a straight-forward manner.

DISCLAIMER

1 INTRODUCTION

[UPC]

2 HIGH SPEED DOWNLINK PACKET ACCESS (HSDPA)

2.1 HSDPA Link-Level Simulator

Figure 1 illustrates the HSDPA link-level simulator which is applicable with most of the selected MIMO transmission schemes. The simulator is implemented in CoCentric System Studio and COSSAP environments.

Transmitter

Data bits are first grouped into packets which are then channel encoded. At the same time, the packet data is stored in a memory slot for possible later retransmissions. After puncturing (rate matching) and interleaving the signal is modulated using the selected modulation scheme (QPSK or 16-QAM). The symbols are either space-time encoded using the selected space-time coding (diversity) scheme or multiplexed into several parallel streams according to the chosen layered scheme. Each antenna symbol stream is spread using the selected number of spreading codes. At his phase, common pilot signal is added to the antenna signals as well as the selected number of interfering speech or HSDPA code channels.

Channel

METRA MIMO channel model generates the desired correlation between the antennas and simulates the desired multipath fading channel. Additive white Gaussian noise (AWGN) is used to model intercell interference. It is also possible to generate actual interfering base station (BS) signal(s). This is especially important when space-time equalizing receivers are applied since they benefit from the structured interference compared to conventional receivers.

Receiver

A multiantenna receiver detects the data bits using channel estimates obtained using the common pilot channel. Only the complex channel gains are estimated while the channel delays are always assumed to be known. Estimation is performed by (moving-average) filtering the common pilot based raw channel estimates. One of the receiver algorithms described in the following sections is used. The receiver has perfect knowledge of the index of the packet it has received and, in case of a retransmitted packet, it combines the soft symbol estimates into a combiner memory slot. Chase-combining based hybrid-ARQ has been applied.

Depuncturing operation (or inverse rate matching) restores the original code frame length for the channel decoder. Cyclic redundancy check (CRC) bits in the code frame are not used but the correctness or incorrectness of the decoded packet is detected ideally. This information is applied in generation of an ARQ command to the receiver. The transmission of ARQ is not simulated explicitly but the ACK/NACK message is given to the transmitter without errors. BS transmitter does not apply any scheduling but retransmits the requested packet after a fixed, parameterized delay.

The main parameters of the HSDPA link-level simulator are:

- Channel coding rate (turbo code with puncturing or rate matching)
- Modulation method (QPSK or 16-QAM)
- Power and number of HSDPA code channels for the desired user
- Power and number of interfering HSDPA code channels
- Power and number of interfering speech users
- Common pilot channel power
- Maximum number of retransmissions per packet
- Retransmission delay

Various performance data is collected during the simulation including:

- raw (non-decoded) bit error rate
- decoded bit error rate
- data throughput (kbit/s) as a function of both average geometry parameter (G) and maximum allowable number of packet retransmissions
- residual average packet error rate (PER) as a function of both average geometry parameter (G) and maximum allowable number of retransmissions
- distribution of packet-wise signal-to-noise ratio
- PER as a function of packet-wise signal-to-noise ratio (only first transmission of each packet is included in this figure)

The last performance figure is used for generating input for system simulations.

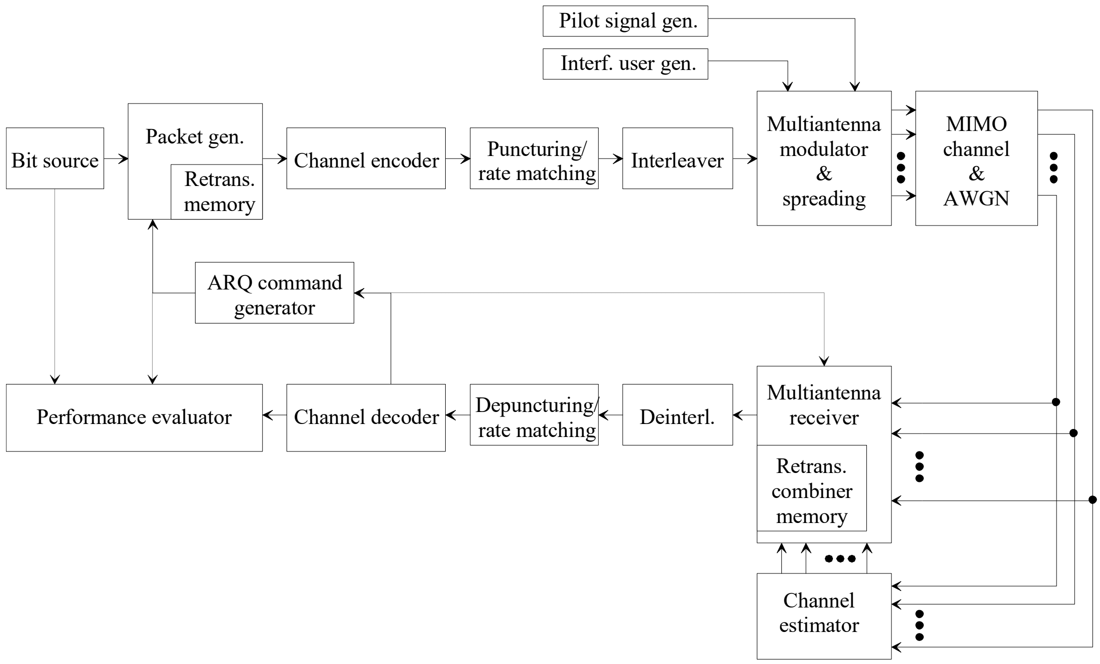


Figure 1. Structure of link-level HSDPA simulator.

2.2 Diversity MIMO Techniques

Diversity MIMO techniques utilize the multiplicity of transmit antennas to obtain transmit diversity. In HSDPA, data rate and system capacity may increase indirectly through a lower packet error rate, higher supported modulation alphabet, or higher channel coding rate.

2.2.1 STTD

In STTD transmission an orthogonal space-time block code is applied over two consecutive data symbols. STTD coding of symbols s_1 and s_2 is illustrated in Figure 2. Columns of the code matrix $C(s_1, s_2)$ represent different symbol intervals and rows different transmit antennas. The code is used as a basis for several other multiantenna transmission schemes.

$$C(s_1, s_2) = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix}$$

Figure 2. Space-Time Transmit Diversity scheme for HSDPA.

In an L-path channel and with N receive antennas, RAKE combiner decodes the STTD code as

$$\hat{s}(i) = \sum_{n=1}^{N} \sum_{l=1}^{L} \left(h_{1,n}^{*}(l) y_n(i,l) + h_{2,n}(l) y_n^{*}(i+1,l) \right) \tag{2.1}$$

$$\hat{s}(i+1) = \sum_{n=1}^{N} \sum_{l=1}^{L} \left(h_{1,n}^{*}(l) y_n(i+1,l) - h_{2,n}(l) y_n^{*}(i,l) \right) \tag{2.2}$$

where $h_{m,n}(l)$ is the channel coefficient or path l from TX antenna m to RX antenna n, $y_n(i,l)$ is the code correlator output, and $w_n(i,l)$ is the noise term. Time index i is the index of the first symbol in a space-time code word of two symbols.

In addition to the presented basic receiver algorithm, space-time equalization techniques can be used as explained in Section 2.4.1.

2.2.2 Circular Shifted Transmit Diversity (CSTD)

CSTD is a scheme combining orthogonal space-time block coding with antenna hopping. It can be applied with arbitrary number of TX antennas and with any space-time block code. It does not affect the data rate provided by the block code and can be detected with almost the same receiver as the underlying block code.

In CSTD transmission the data stream is first encoded with an orthogonal space-time block code. Any code and any number of TX antennas can be used, but we assume here STTD with three TX antennas for simplicity. After block coding, the encoded STTD symbols are transmitted from two antennas as shown in Figure 3. The same antennas are used to transmit the whole code block (two symbols with STTD) and the hopping is applied between two consecutive code blocks.

For channel estimation purposes separate pilot sequences are transmitted from each TX antenna. In other words, a continuous pilot signal is present in each TX antenna, even though any data might not be transmitted at certain times according to the hopping pattern. This naturally reduces the power of each CPICH channel.

The reception of CSTD is done almost similarly as with STTD. The main difference is that we now have to estimate all the channels and keep track of the hopping pattern so that right channel estimates are used at the detection phase.

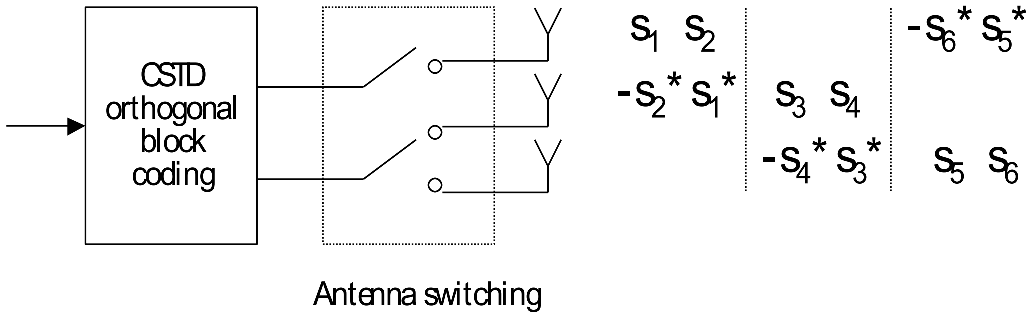


Figure 3. CSTD transmission with three transmit antennas.

2.2.3 Trombi

Trombi is an extension of STTD for more than two TX antennas, achieved by transmitting phase rotated versions of the original signal from additional antennas [Trombi]. With four TX antennas this means simply sending the first STTD diversity branch via the first and the rotated second antenna; and similarly for the second branch, as shown in Figure 4. The phase rotations are only applied for the data channel, not the CPICH channel, to allow backward compatibility to Rel.'99 users. In order to maintain orthogonality between different spreading codes the rotation angle is changed only between code blocks and at most with frequency of 256 chip intervals, corresponding to the duration of the longest spreading code. Continuous CPICH channel is transmitted from each TX antenna and the channel estimates are modified at the receiver to take into account the effects of the phase rotations.

$$C(s_1, s_2) = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix}$$

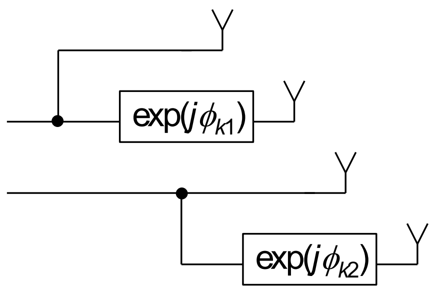


Figure 4. Trombi transmission.

2.2.4 Transmit Antenna Switching

By applying a number of transmit antennas simultaneously in an intelligent way, transmit diversity techniques offer the receiver multiple independently faded copies of the transmitted data. As a result, variation of the received signal power is smaller. However, it is clear that one of the transmit antennas is always better (in terms of total signal energy at the receiver) than the other antennas. Thus, in principle, using any of the weaker antennas for transmit diversity causes loss of signal energy.

The reasoning above suggests that, especially in slowly fading channels, transmit antenna selection based on simple feedback from terminal to base station may increase the average received signal power when compared to transmit diversity. This technique was proposed already for Release 99 of UTRA but the recent results stating that HSDPA system capacity may suffer from the use of transmit diversity makes this approach interesting.

In the applied transmit antenna switching scheme, the received power levels from two transmit antennas are estimated using antenna-specific common pilot channels (similarly to STTD). The better TX antenna is selected via one-bit feedback command transmitted to the base station. There is at least one slot delay until the feedback command affects the used TX antenna.

The receiver multipath combines the signal using both TX antenna channel estimates and selects the one that results in a higher post-combining signal power. In this way, possible feedback errors have negligible effect.

2.3 Layered MIMO Techniques

Layered MIMO techniques utilize the multiple transmit antennas by dividing the data stream into multiple parallel data streams (layers) each of which is transmitted via their own transmit antenna or transmit antenna group. A higher channel data rate can be achieved at the cost of reduced information payload per packet due to stronger required error correcting coding or lower supportable modulation order.

2.3.1 DSTTD

Double STTD is a transmission scheme that uses four TX antennas aiming at doubling the user data rate. As the underlying space-time block code is no longer orthogonal, the task of resolving the transmitted signals is left to the receiver.

In DSTTD two STTD streams are transmitted in parallel using four TX antennas [DSTTD]. The overall transmission is presented in Figure 5 using the same notation as with STTD. Continuous common pilot channel is transmitted from all four TX antennas.

Assuming for two receiver antennas and a flat fading channel, the received and despread signal over one code block can be represented as

$$\mathbf{y} = \begin{pmatrix} y_1(0) \\ y_1^*(1) \\ y_2(0) \\ y_2^*(1) \end{pmatrix} = \begin{pmatrix} h_{1,1} & -h_{2,1} & h_{3,1} & -h_{4,1} \\ h_{2,1}^* & h_{1,1}^* & h_{4,1}^* & h_{3,1}^* \\ h_{1,2} & -h_{2,2} & h_{3,2} & -h_{4,2} \\ h_{2,2}^* & h_{1,2}^* & h_{4,2}^* & h_{3,2}^* \end{pmatrix} \begin{pmatrix} s_0 \\ s_1^* \\ s_2 \\ s_3^* \end{pmatrix} + \mathbf{n} = \mathbf{Hs} + \mathbf{n}. \tag{2.3}$$

DSTTD receiver is preferably based on ML principle using the above signal model as discussed in Section 2.4.2. Both the channel coefficient matrix and noise covariance matrix are stored in memory for the most recent packets to make chase-combining possible. In case of a retransmission, the ML distance metrics are computed after updating the stored channel and noise covariance matrices. The covariance matrix has to inverted for each WCDMA slot.

$$C(s_1, s_2, s_3, s_4) = \begin{bmatrix} C(s_1, s_2) \\ C(s_3, s_4) \end{bmatrix}$$

Figure 5. DSTTD transmission.

2.3.2 V-BLAST

Vertical-BLAST (V-BLAST) is a conceptually very simple layered scheme which, similarly to DSTTD, can be understood as a rate 2 space-time block code as shown Figure 6. Only two data layers are considered here since the number of terminal antennas is assumed to be limited to two.

The channel encoding is performed over one HSDPA packet, as normally, and the interleaved code bit are mapped to symbols. The symbol stream is then serial-to-parallel converted into two (in general *M*) parallel data streams (layers) each allocated to its own transmit antenna. Each antenna also transmits its own common pilot channel.

Figure 7 shows a space-time equalization and interference cancellation based receiver for V-BLAST. First, the data layer with better signal-to-noise ratio is detected using dual-antenna LMMSE space-time equalizer and cancelled from the received signal at chip-level. The remaining data layer is detected using another space-time equalizer. Finally, the two layers are combiner into one data frame. The space-time equalizer structure presented in Section 2.4.1 is applied. It is also possible to use ML receiver as discussed in Section 2.4.2.

$$C(s_1, s_2) = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix}$$

Figure 6. V-BLAST layered scheme with two transmit antennas presented as a rate 2 space-time block code.

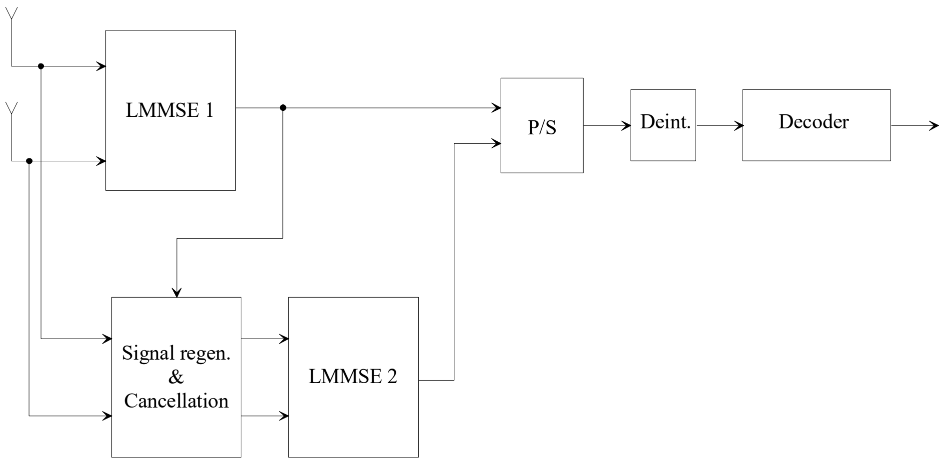


Figure 7. Receiver structure for V-BLAST layered scheme with two receive antennas.

2.3.3 Transformed 4-Antenna Double-Rate Block Code (DABBA)

The transmission of Double ABBA (DABBA) is somewhat similar to DSTTD. Instead of transmitting two orthogonal space-time block codes in parallel, two non-orthogonal codes are transmitted [Hottinen]. This is done by applying unitary transformation to the original STTD blocks, as shown in Figure 8, resulting in less interference between parallel code blocks. The unitary transform is defined as

$$U(\alpha,\phi)=\begin{pmatrix}\mu & \nu\\ -\nu^{*} & \mu^{*}\end{pmatrix}\otimes \mathrm{I}_{N_t/2} \tag{2.4}$$

where N_t is the number of TX antennas and

$$\mu=\sqrt{\alpha} \tag{2.5}$$

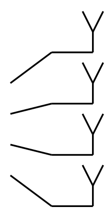
and

$$\nu=\sqrt{1-\alpha}\,\exp\left(-\,j\phi\pi\right), \tag{2.6}$$

where α and ϕ are parameters that can have different values. For example, $\alpha = 0.5$ and $\phi = 0.25$ can be used [Hottinen].

Similar model for the received signal as with DSTTD can also be formulated for DABBA. The same general receiver algorithms are applicable.

$$C(s_1,\ldots,s_8)=\begin{bmatrix}\widetilde{C}_1 & \widetilde{C}_4\\ \widetilde{C}_3 & \widetilde{C}_2\end{bmatrix}$$



$$\begin{bmatrix}\widetilde{C}_1\\ \widetilde{C}_2\end{bmatrix}=U(\alpha,\phi)\begin{bmatrix}C(s_1,s_2)\\ C(s_3,s_4)\end{bmatrix}$$

$$\begin{bmatrix}\widetilde{C}_3\\ \widetilde{C}_4\end{bmatrix}=U(\alpha,\phi)\begin{bmatrix}C(s_5,s_6)\\ C(s_7,s_8)\end{bmatrix}$$

Figure 8. DABBA transmission.

2.4 Advanced Receiver Structures

In this section, the implemented advanced receiver algorithms are defined using a general signal model which makes it straightforward to apply the algorithms to different MIMO schemes by changing the signal model to correspond to the MIMO transmission technique in question.

The applied advanced receiver structures can be roughly divided between linear space-time (ST) equalizers and non-linear maximum likelihood (ML) receivers.

2.4.1 Space-Time Equalizers

Space-time equalizers are used both in case of conventional transmission and multiantenna space-time coded transmission. They can also be applied for linear separation the interfering data layers in case of layered schemes. In most multiantenna transmission systems, however, space-time equalization has rather limited capability especially in multipath channels if the number of receive antenna does not exceed the number of transmit antennas.

ST equalization is performed using chip-level signal, i.e. prior to the despreading operation. Otherwise, due to the pseudo-random scrambling code, the signal structure would be randomized and multiple-access interference could not be suppressed by means of simple linear filtering.

The time-discrete and time-limited signal at RX antenna n can be written as

$$\mathbf{r}_n(i) = \mathbf{H}_{1,n}(i)\mathbf{d}_1(i) + \mathbf{H}_{2,n}(i)\mathbf{d}_2(i) + \ldots + \mathbf{H}_{M,n}(i)\mathbf{d}_M(i) + \mathbf{n}_n(i) \qquad (2.7)$$

where $\mathbf{d}_m(i)$ is a multiuser chip vector transmitted from TX antenna m the elements of which are formed as superposition of all users' chips (including data modulation and scrambling). Time index i refers to the multiuser chip $d_m(i)$ which is to be estimated by the space-time equalizer. $\mathbf{H}_{m,n}(i)$ is a convolutional channel matrix representing the channel from TX antenna m to RX antenna n. The matrix holds the time-discrete channel impulse responses for each chip interval, including $\mathbf{h}_{m,n}(i)$ which corresponds to $d_m(i)$. Noise-plus-interference is denoted by $\mathbf{n}_n(i)$.

In (M, N) MIMO channel, the space-time equalizer estimates the transmitted chips, $d_m(i)$, m=1, 2, …, M, using linear minimum mean-square error (LMMSE) principle, i.e.

$$\hat{d}_m(i) = \sigma_d^2 \begin{pmatrix} \mathbf{h}_{m,1}^{\mathrm{H}}(i) & \mathbf{h}_{m,2}^{\mathrm{H}}(i) & \cdots & \mathbf{h}_{m,N}^{\mathrm{H}}(i) \end{pmatrix} \mathbf{C}_{\mathbf{rr}}^{-1}(i) \begin{pmatrix} \mathbf{r}_1(i) \\ \mathbf{r}_2(i) \\ \vdots \\ \mathbf{r}_N(i) \end{pmatrix} \qquad (2.8)$$

where σ_m^2 is variance of $d_m(i)$ and the covariance matrix of the multiantenna signal is

$$\mathbf{C}_{\mathbf{rr}}(i) = \mathrm{E}\left\{ \begin{pmatrix} \mathbf{r}_1(i)\mathbf{r}_1^{\mathrm{H}}(i) & \mathbf{r}_1(i)\mathbf{r}_2^{\mathrm{H}}(i) & \cdots & \mathbf{r}_1(i)\mathbf{r}_N^{\mathrm{H}}(i) \\ \mathbf{r}_2(i)\mathbf{r}_1^{\mathrm{H}}(i) & \mathbf{r}_2(i)\mathbf{r}_2^{\mathrm{H}}(i) & \cdots & \mathbf{r}_2(i)\mathbf{r}_N^{\mathrm{H}}(i) \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{r}_N(i)\mathbf{r}_1^{\mathrm{H}}(i) & \mathbf{r}_N(i)\mathbf{r}_2^{\mathrm{H}}(i) & \cdots & \mathbf{r}_N(i)\mathbf{r}_N^{\mathrm{H}}(i) \end{pmatrix} \right\} \tag{2.9}$$

where E(·) denotes expectation. The covariance matrix is estimated as a time average. It is also possible to use adaptive techniques to implement an LMMSE equalizer.

Figure 9 shows the structure of the LMMSE space-time equalizer for the signal from transmit antenna *m* followed by a code correlator. An *M*-antenna transmission scheme requires *M* parallel space-time equalizers the outputs of which are decoupled to code correlators and followed by a proper decoder or detector for the applied MIMO transmission scheme. The processing can assume that the channel has only one multipath. This can also be seen in Figure 9 where only a single code correlator is used for detection of the symbol transmitted from antenna *m*.

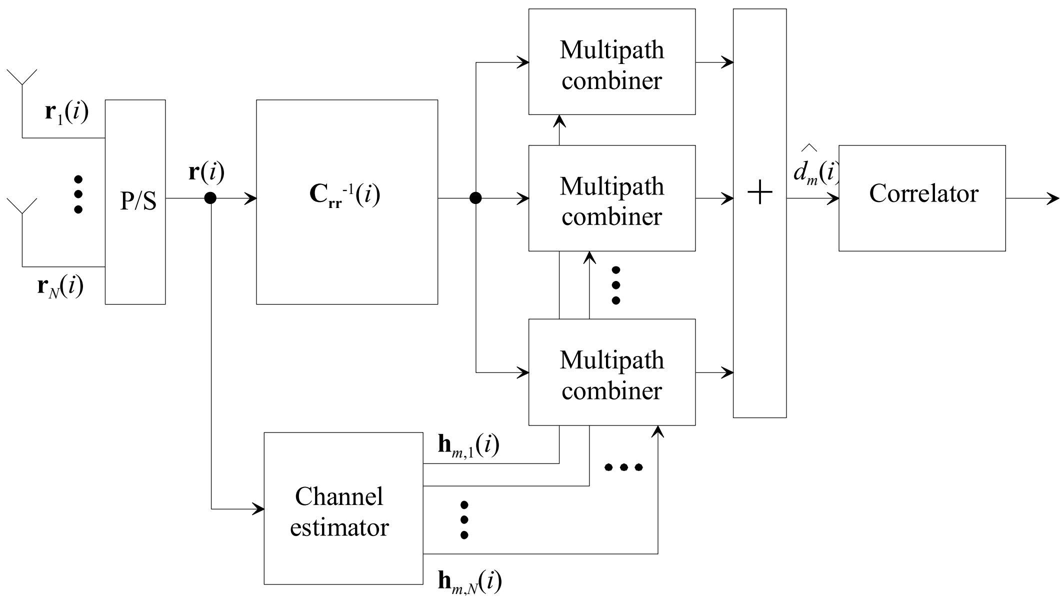


Figure 9. Basic structure of LMMSE space-time equalizer followed by a spreading code correlator.

A property of the LMMSE receiver structure in Figure 9 is that the receiver can be functionally split into two parts: (*i*) so-called prefiltering and (*ii*) conventional RAKE detection including decoding of the possibly used space-time block code. The prefiltering performs the function of the inverted $\mathbf{C}_{\mathbf{rr}}(i)$ matrix Figure 9. The inverse matrix can be shown to approach a block matrix with Toeplitz blocks. This implies that the matrix multiplication operation can be replaced by linear filtering. With multiple RX antennas this conversion from block-wise matrix operation to continuous linear filtering requires a bank of parallel linear filters, as shown in Figure 10.

The benefit of the prefilter-RAKE structure is that, regardless of the number or the origin (transmit antenna, data layer, base station etc.) of the signals to be detected, the same prefilter bank can be used prior to conventional RAKE-type receiver to convert the receiver into an LMMSE space-time equalizer.

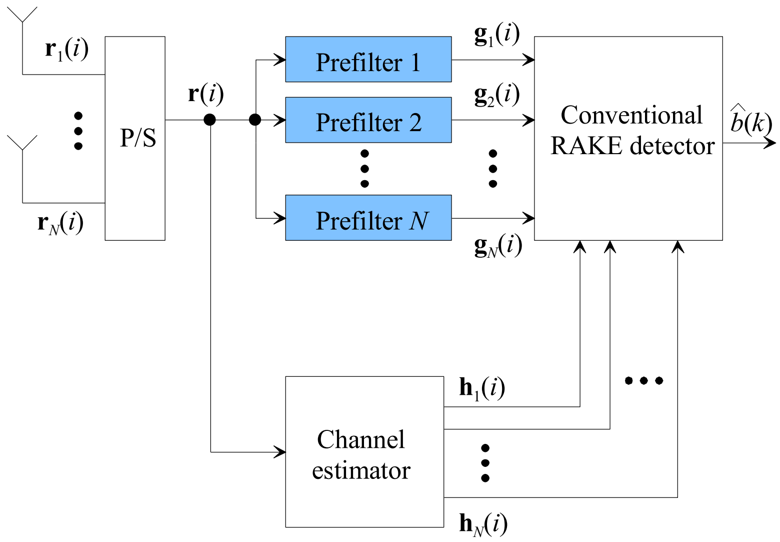


Figure 10. Prefilter-RAKE structure of LMMSE space-time equalizer.

2.4.2 Maximum Likelihood Receivers

Optimal maximum likelihood (ML) receivers cannot be applied in practice due to complexity reasons. By considering intersymbol, multiple access and intercell interference as additive noise, it is possible to formulate a suboptimal ML receiver which is useful especially with layered schemes (e.g. BLAST, DSTTD) when linear space-time equalizers do not suppress the interlayer interference effectively enough. The assumptions imply that the receiver can function in one-shot mode detecting each symbol interval or decoding each space-time code word separately.

Assuming a general L-path channel, and N receive antennas, the receiver despreads all NL signals using as many parallel code correlators. The input signal of the ML receiver is formed by the correlator outputs and can be simply written as

$$\mathbf{y}(k) = \mathbf{H}(k)\mathbf{s}(k) + \mathbf{n}(k)\,. \qquad (2.10)$$

Symbol vector $\mathbf{s}(k)$ holds the unknown symbols and $\mathbf{H}(k)$ is the channel matrix. Vector $\mathbf{n}(k)$ includes all such signal components that are considered as noise in the estimation process. (It should be noted that the above model applies directly only to the very basic V-BLAST scheme. In most cases, the effect of space-time block coding have to be incorporated into the signal model as shown e.g. in Section 2.3.1.)

The unknown symbol vector $\mathbf{s}(k)$ can be estimated using ML principle as

$$\hat{\mathbf{s}}(k) = \arg\min_{\mathbf{s}(k)} \left\{ \left(\mathbf{y}(k) - \mathbf{H}(k)\mathbf{s}(k)\right)^{\mathrm{H}} \mathbf{C}_{\mathbf{nn}}^{-1}(k) \left(\mathbf{y}(k) - \mathbf{H}(k)\mathbf{s}(k)\right) \right\}, \qquad (2.11)$$

where $\mathbf{C}_{\mathbf{nn}}$ is the noise covariance matrix. The noise covariance can either be assumed to be a unit matrix ($\mathbf{I}$) or it can be estimated. ML solution is found through exhaustive search over all possible combinations of the unknown symbols.

Since the receiver complexity grows exponentially as a function of the number of unknown parameters in vector $\mathbf{s}$, it is possible to apply the ML principle to a subset of the unknown parameters while considering the remaining parameters as additive noise (by including them in $\mathbf{C}_{\mathbf{nn}}$). A subset can be selected based on the instantaneous signal-to-noise ratios of the parameters. After detection, the subset is cancelled from $\mathbf{y}(k)$, and then the remaining unknown parameters can be estimated.

Application to Space-Time Block Coding

When ML reception is applied e.g. for DSTTD transmitted signal, a problem is that block coding cannot be presented as a linear transformation of input symbol vector $\mathbf{s}$ (due to the fact that complex conjugation is not a linear operation). In this case, the signal model have to be modified as shown in earlier sections after which the above ML solution is applicable.

Two possible ML approaches can be considered when the modified signal model is used: (*i*) to apply signal $\mathbf{y}(k)$ directly, or (*ii*) to apply signal $\mathbf{H}^{\mathrm{H}}(k)\mathbf{y}(k)$ as an input to the ML detector. The latter technique has the benefit that, by first decoding the block code, one or several interfering symbols cancel out from the signal (since some elements of $\mathbf{H}^{\mathrm{H}}(k)\mathbf{H}(k)$ equal to zero) thus simplifying the ML search. However, at the same time, the noise term becomes very coloured implying that the noise covariance matrix cannot be neglected. The latter method also fits better for coherent combining of packet retransmissions to obtain chase-combining gain.

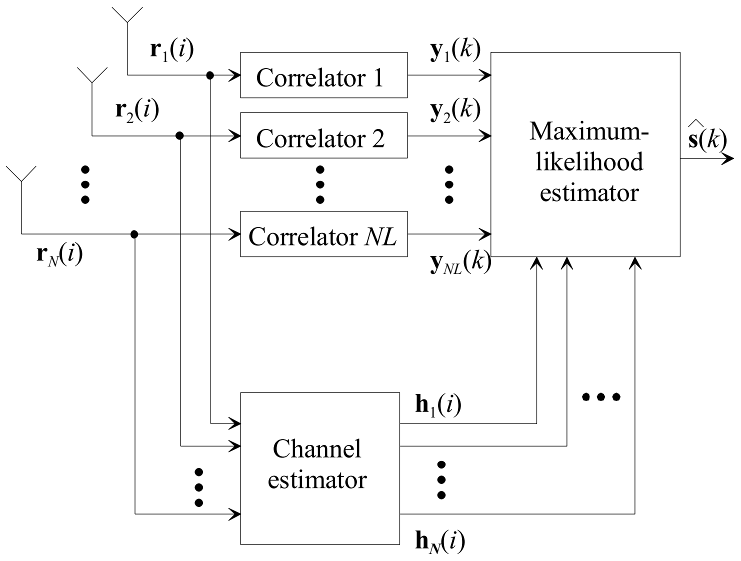


Figure 11. Maximum likelihood receiver structure.

Soft Output Generation

ML detection generates hard decisions for the symbols. Bit-wise soft outputs for the channel decoder can be generated e.g. by computing ratio

$$\frac{\left(\mathbf{y}(k)-\mathbf{H}(k)\mathbf{e}(k)\right)^{\mathrm{H}}\mathbf{C}_{\mathbf{nn}}^{-1}(k)\left(\mathbf{y}(k)-\mathbf{H}(k)\mathbf{e}(k)\right)}{\left(\mathbf{y}(k)-\mathbf{H}(k)\hat{\mathbf{s}}(k)\right)^{\mathrm{H}}\mathbf{C}_{\mathbf{nn}}^{-1}(k)\left(\mathbf{y}(k)-\mathbf{H}(k)\hat{\mathbf{s}}(k)\right)} \tag{2.12}$$

where $\mathbf{e}(k)$ is the best of all such symbol vectors which correspond to a bit vector in which the bit of the interest is other than in $\hat{\mathbf{s}}(k)$. This technique automatically takes into account the applied Gray bit-to-symbol mapping.

2.5 Combination of Linear Dispersion Codes with Turbo Space Time Decoding.

The transmitter is based on a combination of a turbo encoder, linear dispersion space time codes [Hassibi01], and multicode spreading. First, a frame of raw data bits is encoded using a standard Turbo Code with rate R. The coded bits are interleaved, and modulated using Mc bits per real symbol. These symbols are then multiplexed into the P parallel ST encoders.

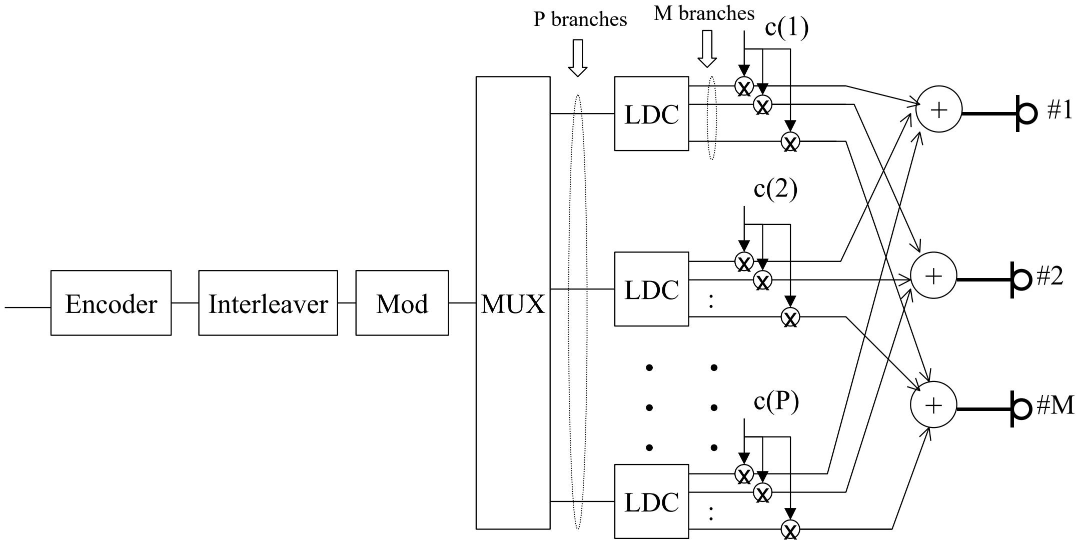


Figure 12. – Transmitter structure.

Each LDC ST encoder brakes its symbol sequence into B blocks of 2Q symbols. Each block will be transmitted through T time slots. Before transmission, the output of the encoders are spreaded using Hadamard codes.

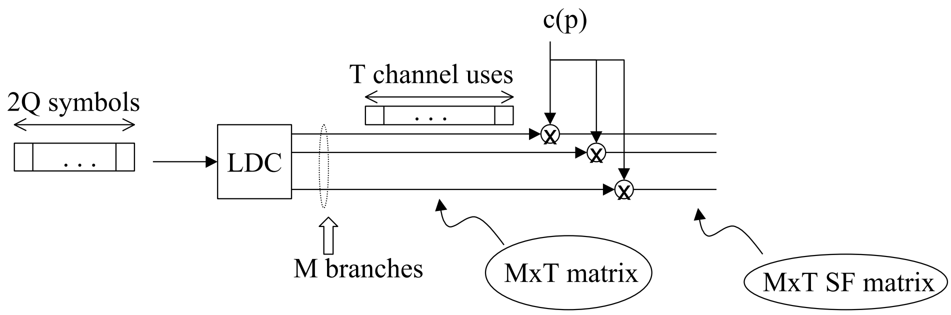


Figure 13. – Space time encoder and spreader.

The overall output of the transmitter's M antennas at the i'th chip can be expressed as a linear function of all the symbols in the associated block:

$$\mathbf{x}(i) = \left(\mathbf{c}^T\left(c(i)\right) \otimes \mathbf{D}\left(t(i)\right)\right)\mathbf{s}\left(b(i)\right) \qquad i = 1,\ldots, SF \cdot T \cdot B \qquad (2.13)$$

where $\mathbf{c}(c)$ is a matrix representing the spreading codes, $\mathbf{D}(t)$ is a matrix representing the LDC structure, $\otimes$ denotes the Kronecker product, $\mathbf{s}(b)$ is a vector representing the $2QP$ transmitted symbols (of all the parallel encoders) at the b'th block, and we define the indexes $c(i)$, $t(i)$ and $b(i)$ follow the following structure:

$$\underbrace{\underbrace{c=1\cdots c=SF}_{t=1}\ \cdots\ \underbrace{c=1\cdots c=SF}_{t=T}}_{b=1}\ \cdots\ \underbrace{\underbrace{c=1\cdots c=SF}_{t=1}\ \cdots\ \underbrace{c=1\cdots c=SF}_{t=T}}_{b=B}$$

The optimal receiver for the previous problem formulation is impractical. It involves a Viterbi Algorithm of the effective super-trellis associated with the concatenated encoders, and the frequency selective channel. The complexity of such a solution is too high for implementation. We therefore resort to a suboptimal approach, usually referred to as the ``Turbo Algorithm''. The use of interleavers decouples the demodulation stage and the decoding stage. This allows for iterative exchange of soft information between them:

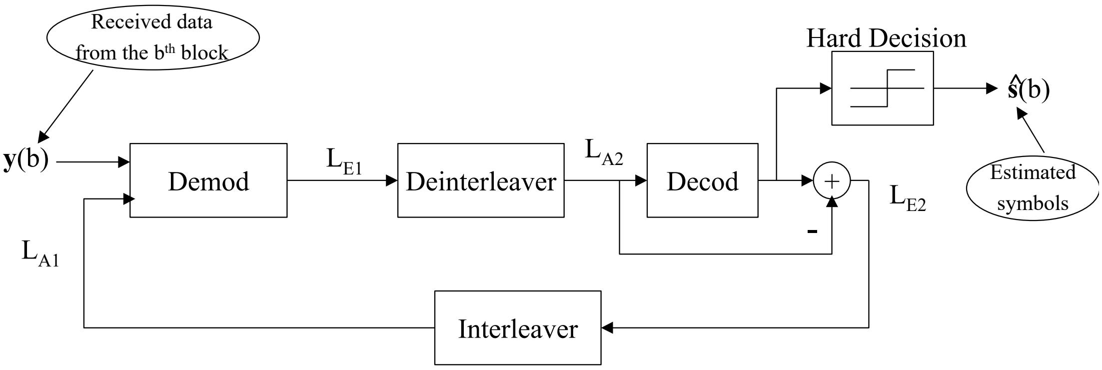


Figure 14. – Iterative receiver.

The soft information is expressed using log likelihood ratios (LLR). The input to each stage is the a priori LLRs denoted by $L_A(d)$. Using them, it calculates the a posteriori LLRs, denoted by $L_D(d)$, which are supposedly ``better''. The difference between these LLRs is usually referred to as extrinsic LLR and is denoted by $L_E(d)$. The extrinsic LLRs of each stage are used as the a priori LLRs for the other stage. This procedure is repeated in an iterative fashion.

$$\underbrace{\log\frac{P(d=1/\mathbf{y})}{P(d=-1/\mathbf{y})}}_{L_D(d)}=\underbrace{\log\frac{P(d=1)}{P(d=-1)}}_{L_A(d)}+\underbrace{\log\frac{P(\mathbf{y}/d=1)}{P(\mathbf{y}/d=-1)}}_{L_E(d)} \tag{2.14}$$

The decoder uses the well known BCJR algorithm [Pollara96]. In brevity, this algorithm calculates the a posteriori LLR of each bit as a function of the a priori LLR of all the other bits and the correlations between them induced by the encoder's trellis. This algorithm is well known, therefore we leave its details to references.

The demodulation algorithm calculates the extrinsic LLR of each bit as a function of the a priori LLR of all the other bits and the received samples. This algorithm is based on the soft input soft output, linear multiuser detector, e.g., [Wang99].

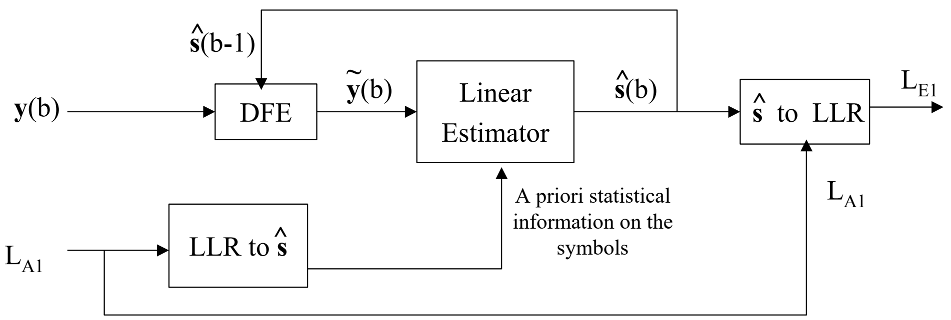


Figure 15. – Soft input soft output demodulator.

The demodulator is basically a weighted least squares linear estimator. The estimator approximates the received vector as a linear function of the unknown symbols, and then uses a priori statistical information on these symbols to derive the linear estimator. The estimates are then converted to extrinsic LLRs.

The linear estimator uses the following approximation for the received vector structure:

$$\mathbf{y}(b) \approx \mathbf{G}(b)\mathbf{s}(b) + \mathbf{w}(b) \quad b = 1 \cdots B \tag{2.15}$$

with

$$\mathbf{y}(b) = \begin{bmatrix} \mathrm{Re}\{\bar{\mathbf{y}}(b)\} \\ \mathrm{Im}\{\bar{\mathbf{y}}(b)\} \end{bmatrix}, \quad \mathbf{G}_b = \begin{bmatrix} \mathrm{Re}\{\mathbf{H}(b)\mathbf{D}\} \\ \mathrm{Im}\{\mathbf{H}(b)\mathbf{D}\} \end{bmatrix}, \quad \mathbf{w}(b) = \begin{bmatrix} \mathrm{Re}\{\bar{\mathbf{w}}(b)\} \\ \mathrm{Im}\{\bar{\mathbf{w}}(b)\} \end{bmatrix},$$

and

$$\bar{\mathbf{y}}(b) = \left[\mathbf{y}^T\left((b-1)SF \cdot T + 1\right) \quad \cdots \quad \mathbf{y}^T\left(b \cdot SF \cdot T\right)\right]^T$$

$$\mathbf{H}(b) = \begin{bmatrix} \mathbf{H}^{(0)}\left((b-1)SF \cdot T + 1\right) & 0 & \cdots & 0 \\ \mathbf{H}^{(1)}\left((b-1)SF \cdot T + 2\right) & \mathbf{H}^{(0)}\left((b-1)SF \cdot T + 2\right) & 0 & \vdots \\ 0 & \ddots & \ddots & 0 \\ 0 & \mathbf{H}^{(L-1)}\left(b \cdot SF \cdot T\right) & \cdots & \mathbf{H}^{(0)}\left(b \cdot SF \cdot T\right) \end{bmatrix}$$

$$\mathbf{D} = \left[\left(\mathbf{c}^T(1) \otimes \mathbf{D}(1)\right)^T \quad \cdots \quad \left(\mathbf{c}^T(c) \otimes \mathbf{D}(t)\right)^T \quad \cdots \quad \left(\mathbf{c}^T(SF) \otimes \mathbf{D}(T)\right)^T\right]^T$$

$$\bar{\mathbf{w}}(b) = \left[\mathbf{w}^T\left((b-1)SF \cdot T + 1\right) \quad \cdots \quad \mathbf{w}^T\left(b \cdot SF \cdot T\right)\right]^T .$$

Due to the frequency selective nature of the channel, this structure is only an approximation. However, as will be explained in the sequel, using the DFE block validates this approximation.

Using the well known matrix inversion lemma, e.g., [Kay93], it is possible to show that the weighted least squares estimate for the unknown symbols is (for simplicity we omit the block indexes):

$$\begin{aligned}
&\hat{\mathbf{s}} = \mathbf{\Sigma}\mathbf{G}^H\mathbf{W}\left(\mathbf{y}-\mathbf{E}_S\right)+\mathbf{E}_S \\
&\mathbf{W} = \left(\mathbf{G}\mathbf{C}_{SS}\mathbf{G}^H+\sigma^2\mathbf{I}\right)^{-1} \\
&\mathbf{\Sigma} = diag\left(1\big/\left[\mathbf{G}^H\mathbf{W}\mathbf{G}\right]_{11} \quad \cdots \quad 1\big/\left[\mathbf{G}^H\mathbf{W}\mathbf{G}\right]_{KK}\right)
\end{aligned} \tag{2.16}$$

where $\mathbf{E}_S$ and $\mathbf{C}_{SS}$ are the a priori expectation and covariance of the unknown symbols. These a priori symbols statistics can be easily derived from the a priori bits LLRs:

$$P\left(d_i=\beta\right)=\frac{e^{\beta L_A(d_i)/2}}{e^{-L_A(d_i)/2}+e^{+L_A(d_i)/2}} \quad \Rightarrow \quad \begin{aligned} E\left[s_k\right] &= \sum_{\mathbf{d}}\left[s(\mathbf{d})\prod_i P(d_i)\right] \\ E\left[s_k^{\,2}\right] &= \sum_{\mathbf{d}}\left[s(\mathbf{d})^2\prod_i P(d_i)\right]\end{aligned} \tag{2.17}$$

where $\mathbf{d}$ denotes the vector of all the bits associated with the symbol $s(\mathbf{d})$. Using these moments it is easy to express $\mathbf{E}_S$ and $\mathbf{C}_{SS}$:

$$\mathbf{E}_S = \begin{bmatrix} E[s_1] \\ \vdots \\ E[s_K] \end{bmatrix} \quad \mathbf{C}_{SS} = \begin{bmatrix} E\left[s_1^{\,2}\right]-E^2[s_1] & 0 & 0 \\ 0 & \ddots & 0 \\ 0 & 0 & E\left[s_K^{\,2}\right]-E^2[s_K] \end{bmatrix} \tag{2.18}$$

Therefore, the demodulator derives the a priori statistics, and then applies the weighted least squares estimators. Finally, the last stage of the demodulator transforms the symbols estimates into bits LLRs using the following formula:

$$\begin{aligned}
L_D\left(d_j\right) = &\log\left\{\sum_{\mathbf{d}_{\neq j}}\exp\left[-\frac{1}{2\sigma_{\hat{s}_k}^2}\left(\hat{s}_k - s\left(d_j=+1,\mathbf{d}_{\neq j}\right)\right)^2+\tfrac{1}{2}\mathbf{d}^T\mathbf{L}_A\right]\right\} \\
-&\log\left\{\sum_{\mathbf{d}_{\neq j}}\exp\left[-\frac{1}{2\sigma_{\hat{s}_k}^2}\left(\hat{s}_k - s\left(d_j=-1,\mathbf{d}_{\neq j}\right)\right)^2+\tfrac{1}{2}\mathbf{d}^T\mathbf{L}_A\right]\right\}
\end{aligned} \tag{2.19}$$

where in the last equation, $\mathbf{d}_{\neq j}$ denotes the vector of all the associated bits excluding the current bit d_j.

As previously explained, the receiver assumes a block channel structure. In order to validate this assumption, we propose the use of a decision feedback equalizer (DFE), which will mitigate the inter block interference. Thus, before applying the estimator for each block, the DFE tries to eliminate the interference of the previous block:

$$\tilde{\mathbf{y}}\left(\left(b-1\right)B+i\right)=\mathbf{y}\left(\left(b-1\right)B+i\right)-\sum_{l=i}^{L-1}\mathbf{H}^{(l)}\left(\left(b-1\right)B+i-l\right)\hat{\mathbf{x}}\left(\left(b-1\right)B+i-l\right) \quad (2.20)$$

for $b=2\cdots B$ and $i=1\cdots L$, where $\hat{\mathbf{x}}$ is obtained from equation (**¡Error! No se encuentra el origen de la referencia.**) by substituting the symbols with their previous estimates.

3 HIGH SPEED UPLINK PACKET ACCESS (HSUPA)

3.1 Introduction

In the following we consider the simulation models for both a low-complexity UE (1-antenna UE) and a dual-antenna UE with MIMO capability. The single antenna case serves as a reference case with "standard" 1-code transmission but multicode transmission is also considered. In addition, a multi-antenna UE with single Tx-chain may be interesting from viewpoint of implementation complexity and cost. This approach allows antenna switching in an open loop mode and antenna selection using feedback from Node B.

After considering the results from the initial study and taking into account the complexity issues of the Node B, the following cases were selected for detailed simulation study. This approach was chosen also because it represents a straight-forward evolution path from the UE capabilities of the current 3GPP specification. Therefore, in this study, we assume that the signal structure of Uplink Common Packet Access using the Random Access Channel (RACH) is employed. It must be noted, however, that concept proposals for uplink high-speed packet access has not been considered yet in 3GPP. For example, it is not possible to obtain uplink channel state information at Node B because no reference channel exists fro uplink. Moreover, there are no feedback channels from Node B to UE to carry the channel state information.

3.2 Simulation model

3.2.1 3GPP compatibility

The simulation model has been designed so that it model includes full 3GPP features. This approach allows a common reference point for advanced MIMO concept as well as realistic signal models and receiver implementation schemes. In a first stage a standard Rake receiver structure is assumed but the model allows also the implementation of more complex receivers with interference suppression in a straight-forward manner.

3.2.2 High-level modelling of uplink high-speed data with MIMO approach

The detailed simulation model includes features such as:

- packet mode data with ARQ
- channel coding (convolutional/turbo) with different code rates
- interleaving
- rate matching / puncturing
- fast power control
- realistic interference modelling (64 kb/s data users)

- realistic pilot power
- realistic channel estimation
- realistic SIR estimation
- adjustable receiver complexity
- 3GPP compatible radio channel models for micro and macro cells
- MIMO radio channel model (Metra, I-Metra) with adjustable spatial/correlation characteristics and Doppler spread

Figure 16 depicts a high-level description of the simulation model. The model includes a MIMO radio channel model from the I-METRA project which is described in detail in deliverable D2.

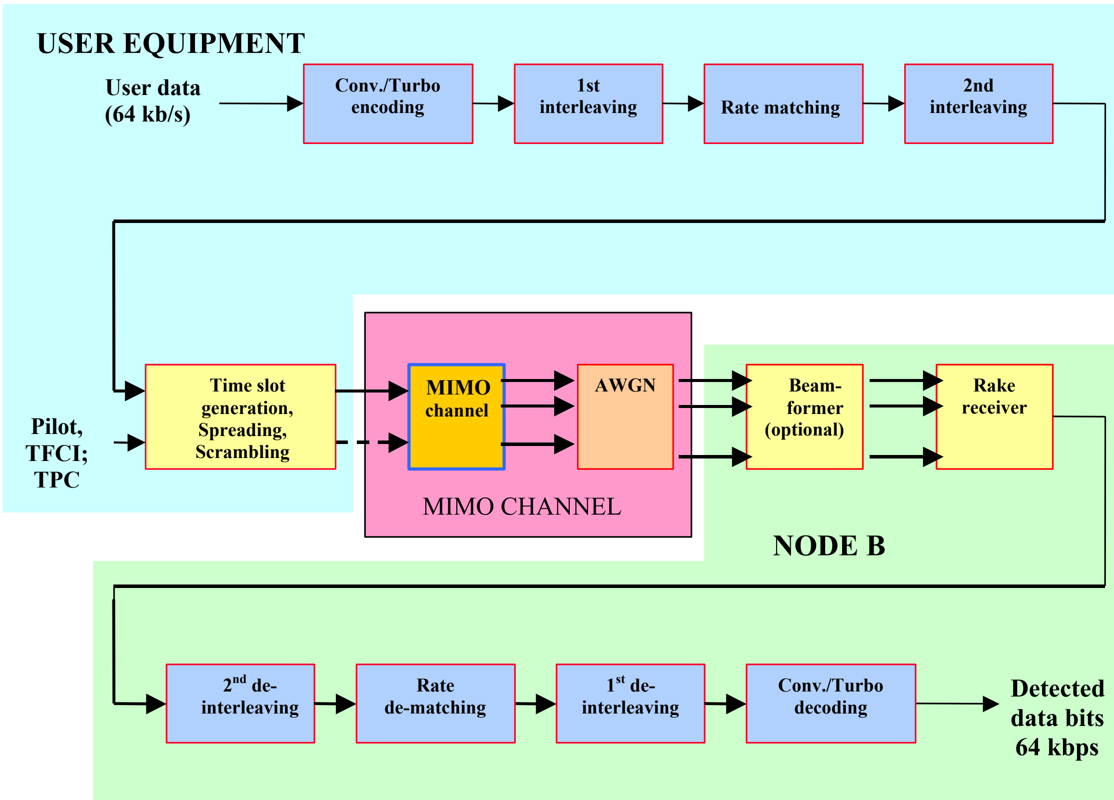


Figure 16. Simulation model for uplink MIMO studies.

In simulations a maximum of two transmit antennas are assumed for the UE. This choice is based simply in strict implementation requirements of the terminal as well as the limited size of the handsets. Node B receiver can accommodate typically two or four diversity antennas or an antenna array of 4-8 antenna elements. As mentioned before standard Rake receiver is employed. Packet data rate of 64 kbit/s with spreading factor of 16 is assumed with 10% BLER target and 10 ms interleaving. Both inner and outer loop power control is applied with PC signaling errors of 4%. These choices support the 3GPP compatibility assumption so that they correspond to the characteristics of the uplink common packet access channel. Interference is modelled as AWGN. Channel estimation assumes perfect timing, amplitude and phase of the channel impulse response are estimated from Dedicated Physical Control Channel (DPCCH). Signal-to-Interference ratio, SIR, is also estimated from the DPCCH. In reception the signal is sampled at rate of one sample per chip.

3.2.3 Signal structure

Figure 17 illustrates the basic uplink I/Q multiplexed WCDMA traffic channel signal structure which has been used in the preliminary simulations. The data bits are transmitted on the I branch and the control channel bits on the Q branch. The duration of the slot is 10/15 ms (0.667 ms).

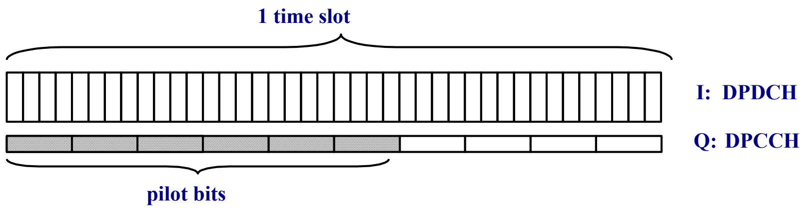


Figure 17. Signal structure for uplink MIMO studies.

The frame structure of the message part of uplink common packet access channel is depicted in Figure 18. The uplink CPCH can be modified for HSUPA for example by adopting higher order modulation combined with stronger coding and HARQ.

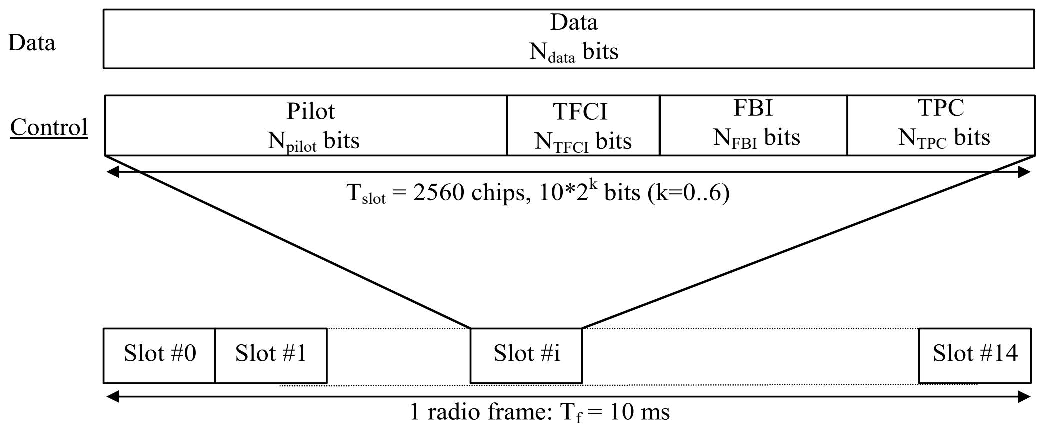


Figure 18. Frame structure for uplink PCPCH.

Table 1 describes in more detail the simulation parameters. These parameters have been obtained from the 3GPP specification TS 25.141 which covers the Node B

Conformance Testing and they confirm realistic performance comparison between standard 3GPP uplink common packet channel and proposed uplink MIMO schemes.

Parameter		Dedicated channel for DTCH / DCCH	Unit
DPDCH	Information bit rate	64 / 0	kbps
	Physical channel	240	kbps
	Spreading factor	16	
	Repetition rate	21	%
	Interleaving	10	ms
	Channel coding	Turbo code, R=1/3	
	# of DPDCHs	1	
DPCCH	Dedicated pilot	6	bit/slot
	Power control	2	bit/slot
	TFCI	2	bit/slot
	Spreading factor	256	
Power ratio of DPCCH/DPDCH		-5.46	dB

Table 1: Simulation parameters. DPDCH (Dedicated Physical Data Channel), DPCCH (Dedicated Physical Control Channel), DTCH (Dedicated Traffic Channel), DCCH (Dedicated Control Channel).

3.2.4 Transmitter structure

Figure 19 depicts the transmitter block diagram in which the panel on top represents the SIMO (Single-Input, Multiple-Output) approach with single transmit antenna at the UE and the panel on bottom represents the MIMO approach. In SIMO case QPSK symbols are transmitted from a single antenna while the MIMO approach allows two separate (or space-time coded) signals to be transmitted from two UE antennas. In MIMO case of this study the same user bits are allocated to different TX antennas so that effective bit rate is the same. In other words, the UE transmits QPSK-symbols via two antennas and the two antenna signals use different channelization (Walsh) codes). The performance of the considered MIMO scheme should be the same as with STTD (used in DL) since the STTD scheme does not give any coding gain. The merit of STTD is to orthogonalise two transmit signals using only a single Walsh code. In asynchronous uplink of WCDMA there is no code limitation which allows the use of two times more Walsh codes. The benefit of the transmit scheme under study is that no space-time (STTD) encoder/decoder structures are needed which makes the implementation less complex. The total transmit power level is kept the same as compared to the reference case of single antenna transmission.

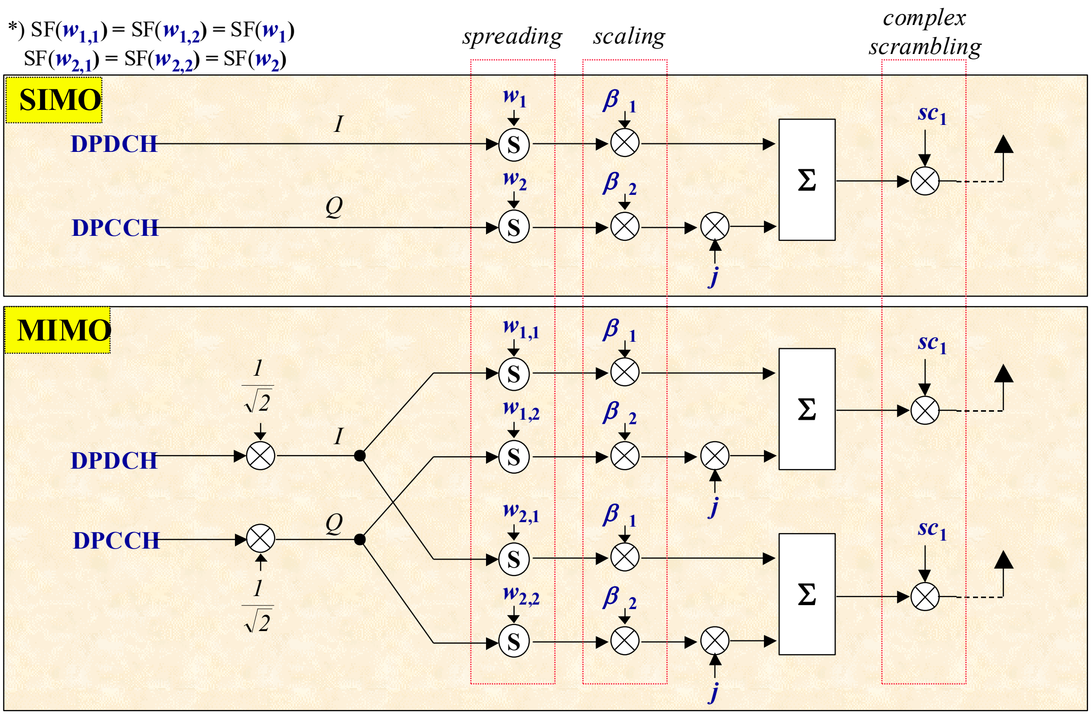


Figure 19. **UE transmitter structure for uplink SIMO (top, reference case) and MIMO (bottom) studies. In SIMO case BPSK symbols are transmitted from a single antenna while in MIMO case the same bit is transmitted simultaneously from two antennas (data rate is the same in both cases).**

The transmitter structure allows also parallel data transmission from two Tx antennas so that different data bits are transmitted from different antennas at the same time interval. It is also possible to transmit space-time coded signals.

3.2.5 Receiver structure

Receiver structure is shown in Figure 20. The receiver represents the 2D Rake receiver structure in which delay estimation, channel estimation and SIR estimation is based on pilot symbols of the received UE signal. The desired signal is coherently combined over the propagation paths and over the antennas. Thus each antenna is allocated a number of Rake fingers. The total number of the fingers can be varied for the desired user depending on the multipath characteristics of the radio channel and/or the allowed receiver complexity. In MIMO case, the different bits from different transmit antennas are finally combined to a QPSK symbol.

The receiver structure allows also the use of more advanced receiver algorithms such as optimum combining with interference suppression.

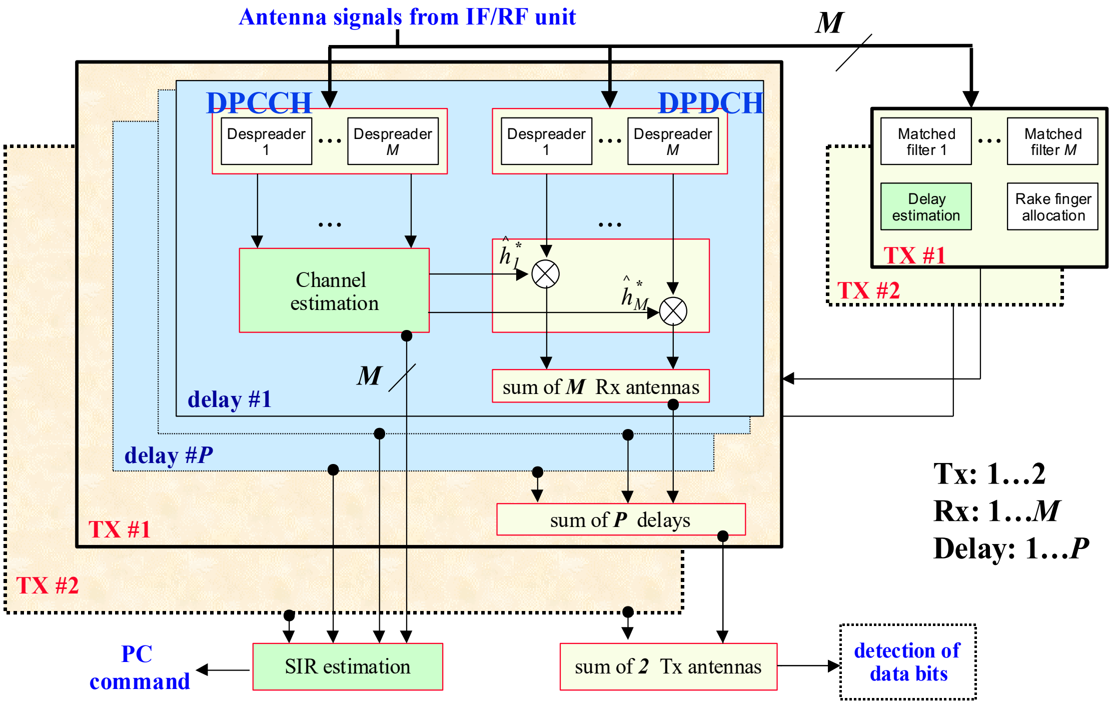


Figure 20. Receiver structure of Node B in case of SIMO (TX #1 applied) and MIMO (TX#1 and TX#2 applied).

3.2.6 MIMO channel properties

The following temporal properties will be studied:

- 1-path Rayleigh, UE velocity 3 km/h
- 1-path Rayleigh, UE velocity 50 km/h
- modified ITU Pedestrian A, UE velocity 3 km/h
  - two taps
  - tap powers: [0 -12.7] dB
- modified ITU Vehicular A, UE velocity 50 km/h
  - six taps
  - tap powers: [0 -1.92 -7.31 -10.39 -10.89 -17.31] dB

The spatial properties which define the correlation coefficients between the UE and Node B antennas are assumed to be the following:

- UE: uncorrelated antennas with angular spread of 360 degrees

- Node B: correlation varies in large range from uncorrelated antennas to fully correlated antennas (in case of beamforming)

- antenna setup (Tx-antennas x Rx-antennas): 1x1, 1x2, 2x1, 2x2, 1x4, 2x4

3.2.7 Performance measure

The link level performance can be measured in simulation as the required receive and transmit Eb/No (energy per bit / total noise power density) that gives the required Quality of Service, QoS (which is 10% BLER). Transmit (Tx) Eb/No refers to the required transmit power (battery consumption) of the UE and corresponds to the amount of inter-cell interference that the given UE generates. Therefore this figure affects also to the total capacity of the network. Required receive (Rx) Eb/No per Rx antenna corresponds to the amount of intra-cell interference that the other users experience from the given UE. Thus this figure has a direct effect to the uplink capacity of the particular cell, i.e. max. number of users that can be supported in one sector of a Node B.

3.2.8 Simulation cases

It is assumed here that 64 kb/s packet mode data is the desired service, and the capacity increase is estimated as throughput improvement in terms of maximum number of 64 kb/s channels. Thus, the desired UE can employ a maximum data rate of MxZx64 kb/s, in which M denotes the number of TX antennas and Z the number of multicodes. The following uplink MIMO concepts will be studied in performance evaluation:

<u>Case 1 (reference):</u>

- Single antenna transmission in UE
  - Reference case: 1x2 (2-antenna diversity reception at Node B)
  - Another reference: 1x4

<u>Case 2 (2xM MIMO):</u>

- 2-antenna Tx diversity
  - Antenna switching
  - Higher modulation schemes
  - ARQ option
  - Different Walsh codes from different antennas
  - 2 or 4 receive antennas in Node B

<u>Case 3 (MIMO with link level adaptation):</u>

- STTD and/or 1-antenna transmission (continuous transmission) compared to adaptive STTD and/or single antenna transmission (Tx: ON/OFF)
  - No measurement channel for UL quality

- Known channel state
- Adaptivity on the link level: transmission during favorable channel state
- Antenna switching w/wo ARQ
- Scheduling not considered

Case 4 (tentative):

- Multi-code via 1 Tx-antenna vs. multi-codes via 2 Tx antennas
  - Reference case: 1x2, 1x4
  - Evaluating uplink peak data rate with MIMO approach

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