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Technical Report

3rd Generation Partnership Project;
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Tx diversity solutions for multiple antennas
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Contents

Foreword

This Technical Specification has been produced by the 3rd Generation Partnership Project (3GPP).

The contents of the present document are subject to continuing work within the TSG and may change following formal TSG approval. Should the TSG modify the contents of the present document, it will be re-released by the TSG with an identifying change of release date and an increase in version number as follows:

Version x.y.z

where:

x the first digit:

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y the second digit is incremented for all changes of substance, i.e. technical enhancements, corrections, updates, etc.

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1 Scope

The study done in this TR focuses on transmit diversity methods with more than 2 antennas at the Node B and 1 antenna at the UE. The functionality introduced to support more than 2 antennas at the Node B could permit consideration of compatible operation with 2 antennas.

A proposal for a new transmit diversity method using multiple antennas may be included in the TR if at least the following requirements are fulfilled:

a) it should be shown that there is significant benefit in at least one realistic scenario, e.g. one of the (multi-)path models of table 7; for closed loop methods also feedback errors and delays need to be considered.

b) the description should be detailed enough to see how it fits into the current specification (so it is in principle clear what changes need to be done).

c) relevant R99 schemes should be used as reference cases depending on the proposal (open or closed loop, or both)

d) backwards compatibility needs to be addressed

This evaluation of different transmit diversity schemes for multiple antennas is within the scope of this TR.

2 References

The following documents contain provisions which, through reference in this text, constitute provisions of the present document.

- References are either specific (identified by date of publication, edition number, version number, etc.) or non-specific.
- For a specific reference, subsequent revisions do not apply.
- For a non-specific reference, the latest version applies.

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3 Definitions, symbols and abbreviations

3.1 Definitions

For the purposes of the present document, the [following] terms and definitions [given in ... and the following] apply.

Definition format

***<defined term>**: <definition>.*

example: text used to clarify abstract rules by applying them literally.

3.2 Symbols

For the purposes of the present document, the following symbols apply:

Symbol format

<symbol> <Explanation>

3.3 Abbreviations

For the purposes of the present document, the following abbreviations apply:

Abbreviation format

<ACRONYM> <Explanation>

4 Background and Introduction

The standardization of 3rd generation WCDMA system has been going on in 3rd Generation Partnership Project (3GPP) since the end of 1998. The 3G systems bring a promise of much higher data rates and enhanced services when compared to 2G systems. As many of the proposed services, like wireless web browsing, are expected to be downlink-intensive it was recognized from the very beginning that improvement of downlink capacity is one of the main challenges.

Performance of radio system depends on various issues but one important factor is the available diversity (time, frequency, multipath etc.). Due to wide bandwidth WCDMA systems are especially effective in exploiting the multipath diversity existing in time dispersive radio environments. If little or no multipath diversity is available the performance can degrade quite considerably. One way of improving the situation is to utilize 2 or more receive and/or transmit antennas that effectively speaking introduce additional radio paths and thereby increase the available diversity. As receiver antenna diversity is implementation wise challenging especially for low cost terminals a lot of attention have been paid to various transmit diversity solutions to be employed on radio access network side.

During 1999 a great deal of effort was put on defining transmit diversity solutions for Rel.-99 of 3GPP WCDMA specifications. As a result two open loop techniques, Space Time Transmit Diversity (STTD) and Time Switched Transmit Diversity (TSTD), and closed loop solution based on Transmit Adaptive Array (Tx AA) concept with two different modes were standardized for FDD [1,2]. For TDD, TSTD and Block STTD open loop methods can be used on SCH and P-CCPCH, respectively, and closed loop methods on DPCH [3, 4]. All the Rel.-99 Tx diversity methods assume two transmit antennas.

Already during 1999 it was recognized that further performance improvements could be possible by increasing the number of transmit antennas. Yet, it was agreed that Tx diversity for more than 2 antennas will be studied for possible inclusion to Rel.-5 of 3GPP specifications. The following chapters describe the proposed concepts, present the performance results, consider the impacts on UE and UTRAN implementation, and physical layer operation, and, finally, present issues related to backwards compatibility to Rel.-99 followed by conclusions.

Note, that the schemes included in this TR are currently for the purpose of description of new schemes only.

5 Descriptions of studied concepts

5.1 Description of the eigenbeamformer concept

With increasing the number of antenna elements by using an extension of the Release-99 TxD modes, the amount of necessary feedback is increased. When keeping the uplink bandwith the same the antenna weights cannot be adjusted fast enough to account for fast fading. Hence, for higher velocities of the UE the gain due to the additional antenna elements is low.

However there are ways to reduce the necessary feedback bandwith if the antenna channel paths are correlated. One possible concept to achieve a lower feedback bandwith is the eigenbeamformer concept which takes advantage of the correlated antenna paths. The general idea behind the eigenbeamformer is a decorrelation of the antenna signal paths to achieve a reduction in dimension of the spatial space. This enables subsequent short term processing at the UE to sufficiently mitigate fast fading.

This decorrelation is performed by exploiting the long term properties of the propagation paths based on an eigenanalysis of its long term spatial covariance matrix. The eigenvectors (in the sequel also called eigenbeams) with the largest eigenvalues (largest average UE receive power) are determined and fed back step by step to the Node B. This process takes place on the same time scale as the physical UE movement. Accordingly, the required operations in the UE as well as required feedback bits are distributed over a very large number of slots.

In addition, a short term selection between the eigenbeams is carried out at the UE to account for fast fading. This information is fed back to the Node B on (almost) every slot.

By this technique it is possible to address a larger number of antenna elements providing large beamforming gains at higher velocities.

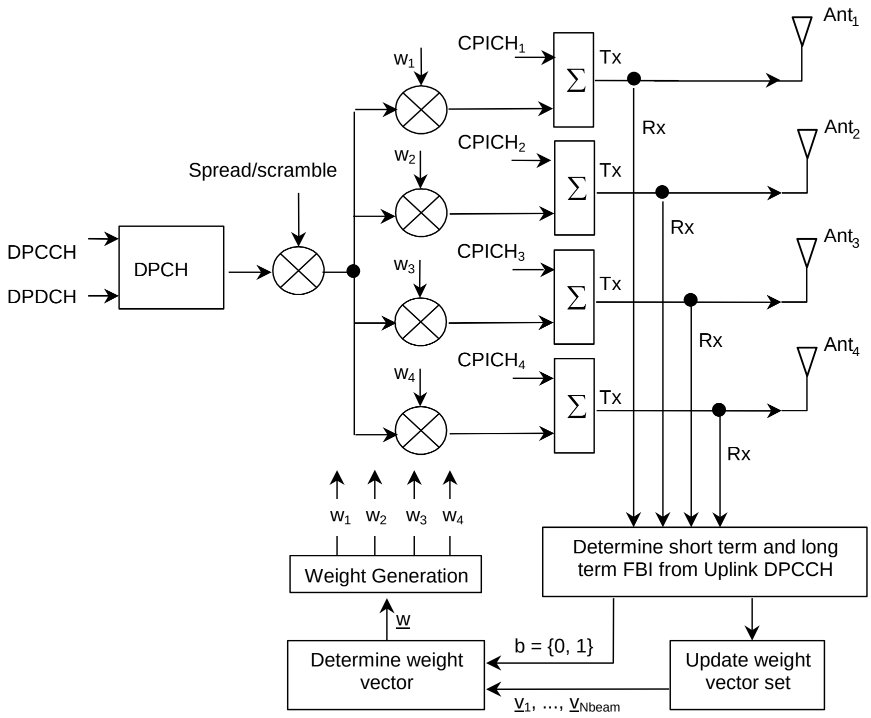

Figure 1. Generic Downlink transmitter at the Node B with M = 4 antenna elements

Figure 1 and Figure 2 show the generic architecture of the eigenbeamformer concept at the Node B and the UE. In the following sections the focus is on a system with M = 4 antenna elements and Nbeam = 2 or 4 eigenvectors. However the eigenbeamformer is easily extendable to more antenna elements.

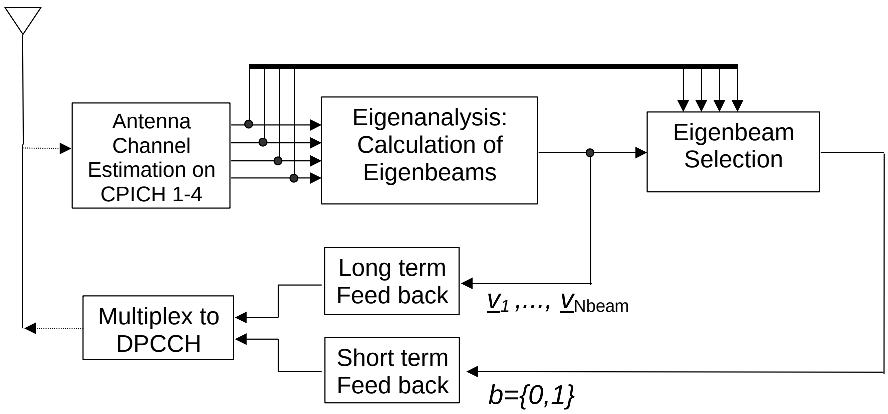

Figure 2. Generic eigenbeamformer structure at the UE for M = 4

5.1.1 Calculation of the Dominant Eigenvectors

Using orthogonal pilot sequences transmitted from the Node B antenna elements, the UE estimates the short term spatial covariance matrix averaged over the temporal taps of the channel.

$$\mathbf{R}_{\mathrm{ST}} = \sum_{n=1}^{N} \mathbf{h}_n \, \mathbf{h}_n^H \qquad (1)$$

The column vector $\mathbf{h}_n = \left(h_{n1},\ h_{n2}, \ldots, h_{nM}\right)^T$ denotes the channel vector of the n-th temporal tap. The number of taps is denoted by N; M = 4 antenna elements are assumed. The long term spatial covariance matrix is obtained by averaging the short term matrix using a forgetting factor ρ.

$$\mathbf{R}_{\mathrm{LT}}(i) = \rho\, \mathbf{R}_{\mathrm{LT}}(i-1) + (1-\rho)\, \mathbf{R}_{\mathrm{ST}}(i) \qquad (2)$$

The symbol i denotes the time index. It is sufficient to perform an update once every frame or even in larger intervals.

Decorrelation in space is achieved by an eigenanalysis of the long term spatial covariance matrix according to

$$\mathbf{R}_{\mathrm{LT}}\ \mathbf{V} = \mathbf{V}\,\mathbf{\Theta} \qquad (3)$$

The eigenvectors (eigenbeams) to be found are columns of $\mathbf{V}$. Since the matrix $\mathbf{\Theta}$ is diagonal by definition, transmission on different eigenbeams leads to uncorrelated fast fading. The diagonal entries indicate the long term UE received power of each beam.

Note that the eigenbeamformer automatically adjusts to various propagation environments (spatially correlated or uncorrelated). If the channel is spatially correlated, the channel can accurately be described by a small number of eigenbeams. If, on the other hand, the channel has a spatial correlation of zero, no long term spatial channel information can be exploited and each eigenvector addresses only one antenna element.

5.1.2 Long Term Feedback Scheme

From the set of M = 4 eigenbeams in $\mathbf{V}$, Nbeam vectors with the largest eigenvalues will be chosen to be transmitted in the long term feedback.

Each weight vector is a vector of complex numbers. The size of this vector equals the number of antenna elements (M = 4). Each complex vector element is quantized by a number of bits. There are different ways for quantization. For example, the absolute value and the phase can be quantized with 3 and 5 bits respectively. Hereby, the amount of bits can be reduced if the phase of the first vector element is set to zero. Thus, for the transmission of one eigenbeam 4□*3+3*□5 = 27 bits are necessary.

This number applies for the direct feedback of the eigenbeams from the UE to the Node B. Also methods with progressive refinement could be used that transmit only the difference to the previously sent vector. This could reduce the subsequent update period and an increased quantization / resolution is possible.

More advanced long term feedback concepts could be used which require less feedback bits.

To ensure good performance of the eigenbeamformer concept it is essential to add redundancy to the long term feedback. Therefore coding is used to be capable of detecting and correcting feedback errors. There are several possibilities to define such a code. One possibility described here is a simple BCH code with code rate of 54/110. For two coded eigenbeams a total of 110 bits are transmitted. The number of bits per eigenbeam roughly doubles and so the code is capable to detect and correct 9 bit errors. In the unlikely event of more than 9 bit errors in one coding block it can be assumed that the Node B detects this incorrect feedback by making a plausibility check, e.g. checking the orthogonality of the eigenbeams. In that case the previously used eigenbeams could be reused once more.

5.1.3 Short Term Feedback Scheme

A short term estimate of the UE received power is performed for each weight vector by calculating

$$P_m = \mathbf{w}_m^H \mathbf{R}_{\mathrm{ST}}^T \mathbf{w}_m = \sum_{n=1}^{N} \left| \mathbf{w}_m^T \mathbf{h}_n \right|^2 \qquad (4)$$

$$\text{where } \mathbf{w}_m = \begin{bmatrix} \mathbf{v}_1 & \mathbf{v}_2 \end{bmatrix} \begin{bmatrix} \alpha_1(m) \\ \alpha_2(m) \end{bmatrix}$$

where $[\alpha_1(m)\ \alpha_2(m)]^{\mathrm{T}} \in \{ [1\ 0]^{\mathrm{T}}, [0,1]^{\mathrm{T}} \}$ for ***beam selection scheme*** or $[\alpha_1(m)\ \alpha_2(m)]^{\mathrm{T}} \in \{ [1\ \gamma\exp(\pi/4)]^{\mathrm{T}}, [1\ \gamma\exp(3\pi/4)]^{\mathrm{T}}, [1\ \gamma\exp(-3\pi/4)]^{\mathrm{T}}, [1\ \gamma\exp(-\pi/4)]^{\mathrm{T}} \}$ for ***weighted combining scheme***, where m characterizes the degree of the combining phase, $\gamma \propto \mathrm{sqrt}(\lambda_2/\lambda_1)$, where λ_1 and λ_2 is first and second largest eigenvalue of long term covariance matrix, $\mathbf{R}_{\mathbf{LT}}$, respectably. The weight vector that results in the maximum value for the received power P_m is selected and signalled to the Node B.

For two (four) eigenbeams 2 (4) bits are transmitted to indicate the selection and progressive refine is applied to transmit each bit. The overlaying long term processing makes it possible to co-phase combine two(four) eigenbeams instead of antenna elements.

An increasing number of antenna elements can be addressed without reducing the UE velocity threshold. Note that the pilot symbols of the DPCCH may be used for eigenbeam verification similar to the closed loop modes in Release-99.

5.1.4 Format of Feedback Information

The feedback rate for the eigenbeamformer is kept at the same rate as in Release-99 and is 1500 bit/s. The long term information bits (for feedback of eigenbeams) and the short term information bits (for feedback of eigenbeam selection) are multiplexed. The following frame format for the feedback information bits is proposed:

Table 1: Multiplexing of long term / short term feedback information

Slot #	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
short term FB bits	1	1	1	1	1	1	1	1	1	1	1	1	1	1	0
long term FB bits	0	0	0	0	0	0	0	0	0	0	0	0	0	0	1

In this multiplexing format the transmission of two eigenbeams would take 2*□27=54 frames or 540 ms (see section 5.1.2). The eigenbeam selection of the previous slot is applied in the slots where no short term feedback information is received by the Node B (slot #15).

This format is confined to one radio frame. Thus, no counting over frame boundaries is necessary.

In a later extension with more than 4 antenna elements other formats could be used, e.g. using 3 long term feedback bits within one frame. This is for further study.

Table 2: Multiplexing format of long term / short term information for more than 4 antenna elements

Slot #	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
short term FB bits	1	1	1	1	0	1	1	1	1	0	1	1	1	1	0
long term FB bits	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1

Since no long term channel information is available at the Node B for a user at the start of transmission, initial weight vectors may, for instance, address only one of the antenna elements, e.g.,

$$\mathbf{v}_1 = \begin{pmatrix} 1 \\ 0 \\ 0 \\ 0 \end{pmatrix}, \quad \mathbf{v}_2 = \begin{pmatrix} 0 \\ 1 \\ 0 \\ 0 \end{pmatrix}, \text{ for M = 4 antenna elements.}$$

5.2 Basis selection scheme for > 2 Tx antennas

5.2.1 Tx antenna weights

In closed loop Tx diversity systems, the weights of transmit antennas are determined at a mobile station and fed back to the base station. These weights should result in as high SNR as possible at the mobile. The set of these weights may be viewed as a vector $\underline{w}= [w_1\ w_2\ \ldots w_i\ldots\ w_M]^T$, where w_i is a complex weight associated with the ith Tx antenna. For the maximum SNR at the mobile, the weights should maximize P below:

$$P = \underline{w}^H H^H H \underline{w}, \qquad (5)$$

when H=[$\underline{h}_1\ \underline{h}_2\ \ldots\underline{h}_i\ldots\underline{h}_M$] and M is the number of Tx antennas. The column vector $\underline{h}_i$ represents an estimated channel impulse response for the ith Tx antenna, and its vector length equals to the number of paths. The weight vector $\underline{w}$ information is periodically fed back to the base station. Note that the amount of feedback information and the implementation complexity increase with the number of Tx antennas. The efficient representation of a weight vector is desired to reduce the amount of feedback data and the implementation complexity. Furthermore, backward compatibility is desirable.

A weight vector with M elements may be represented as a linear sum of basis vectors, which span an M-dimensional space. Examples of basis vectors for 2-, 3-, 4-dimensional spaces are shown in Appendix A of[7]. Let's assume for explanation that 4 Tx antennas are used for Tx diversity. The optimal weight vector $\underline{w}_{opt}$ for this system has 4 elements and may be represented as a linear sum of four basis vectors, $\underline{B}_1$, $\underline{B}_2$, $\underline{B}_3$, $\underline{B}_4$, as follows:

$$\underline{w}_{opt} = c_1\underline{B}_1 + c_2\underline{B}_2 + c_3\underline{B}_3 + c_4\underline{B}_4 \qquad (6)$$

where $c_1,\ldots,c_4$ are complex coefficients associated with corresponding vectors. Assuming that $|c_1|>|c_2|>|c_3|>|c_4|$, $\underline{w}_{opt}$ may be approximated as

$$\underline{w}_{app\_1} \cong c_1\underline{B}_1, \qquad (7.a)$$

$$\underline{w}_{app\_2} \cong c_1\underline{B}_1 + c_2\underline{B}_2, \qquad (7.b)$$

$$\underline{w}_{app\_3} \cong c_1\underline{B}_1 + c_2\underline{B}_2 + c_3\underline{B}_3, \qquad (7.c)$$

These vectors $\underline{w}_{app\_1}$, $\underline{w}_{app\_2}$, $\underline{w}_{app\_3}$, may be viewed as the projections of $\underline{w}_{opt}$ into 1-dim, 2-dim, and 3-dim subspaces. $\underline{w}_{app\_3}$ is more accurate representation of $\underline{w}_{opt}$ than $\underline{w}_{app\_1}$ and $\underline{w}_{app\_2}$.

5.2.2 Representation of weight vectors

The conventional representation of the vector $\underline{w}_{opt}$ may require (M-1)*N_c bits, where N_c bits are required to represent each element of $\underline{w}_{opt}$. This representation indicates that the transmission of (M-1)*N_c bits at 1500Hz is required to support Tx diversity with M Tx antennas. The reason for (M-1)*N_c not M*N_c is that one of M Tx antennas may be viewed as reference and the relative weights for other antennas are required. To reduce the required number of bits, it is proposed to feedback information on the approximated vector, instead of $\underline{w}_{opt}$. The representation of the approximated vector includes the specification of basis vectors and associated coefficients. When there are M Tx antennas and the approximation is made in a S-dimensional subspace, there are ${}_M C_S$ combinations for selecting S basis vectors among M vectors and the required number of bit to specify the basis vector combination is $\lceil \log_2({}_M C_S) \rceil$.

5.2.3 Feedback protocol structure

In the simulation, the two cases for antenna selection is considered: **Case 1)** 2 antenna selection (M=4, S=2), it noted as 4C2 and **Case 2)** 3 antenna selection (M=4, S=3), it noted as 4C3. In both cases, 2bit representation for each element (phase only) is used (N_c = 2). The required number of feedback information per signalling word is: **Case 1)** 5 bits, and **Case 2)** 6bits. For detail simulation scheme, section 6.2.1 can be referred. The considered frame format of feedback information is:

(Case 1: 2 best selection among 4 basis and combine all 2 with received phase information)

Slot	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15

Phase				P_1	P_2				P_1	P_2				P_1	P_2
Selection	S_1	S_2	S_3			S_1	S_2	S_3			S_1	S_2	S_3		

S_i: Antenna selection bits
P_i: Phase difference with respect to the coefficient associated with the first basis vector

(Case 2 : 3 best selection among 4 basis and combine all 3 with received phase information)

Slot	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
Phase			P_{11}	P_{12}	P_{21}	P_{22}			P_{11}	P_{12}	P_{21}	P_{22}			P_{11}
Selection	S_1	S_2					S_1	S_2					S_1	S_2	

S_i: Antenna selection bits
P_{ij}: Phase difference with respect to the coefficient associated with the first basis vector

5.3 Closed loop mode 1 for > 2 Tx antennas: R2FNTM concept

As pointed in [6], there are much of issues that may/will degrade the "optimal" CL 4-Tx-diversity results, and in order to go forward with CL concepts, much of further analysis is needed. This section contains only a basic description (and basic results in 6.2) not trying to give the full analysis.

In the Rel.-99 closed loop mode 1 UE calculates the phase adjustment to be done at the UTRAN access point. This involves the calculation of the phase difference between the signals transmitted from the two antennas. When there are N Tx-antennas, a straight forward extension of the mode 1 is to calculate the needed phase adjustment for each N-1 antennas with respect to the reference antenna and signal those adjustments back to the UTRAN. How fast this information can be communicated to UTRAN depends naturally on the number of antennas and the feedback bit rate. This in turn will determine the sensitivity of the performance to the UE speed, i.e. the performance will deteriorate faster as a function of UE speed if the feedback bit rate is low and N large. As it is not desirable to increase the feedback bit rate from 1500 bps as defined in Rel.-99 methods trying to enhance the performance given the restrictions in feedback signaling need to be studied.

In Rel.-99 closed loop mode 1 the filtering of the phase adjustments at the UTRAN access point is specified as simple (sliding window) averaging over two consecutive values. This could be denoted as R2F2T2 where R2 denotes the number of constellation sets due to rotation, F2 denotes the filter of length 2 and T2 denotes the 2 T_x antennas. This can be generalized to R2FNTM, that is rotation constellation set per antenna remains the same as in Rel.-99, length of the phase adjustment filter is N, and there are M Tx-antennas.

In order to maintain the same rate of feedback as in the Rel.-99 mode 1, the weight of only one antenna is fed back in one slot. Hence, the actual memory of the filter is (M-1)(N-1) slots. In case of 4 Tx-antennas (M=4), let Z_1, Z_2, Z_3 be the feedback for antennas 2,3,4 respectively, with antenna 1 being the reference antenna. Let N = 4 (R2F4T4). It is clear that in this example case the current antenna weight relies on feedback weights sent 9 slots in the past as illustrated in theFigure 3.

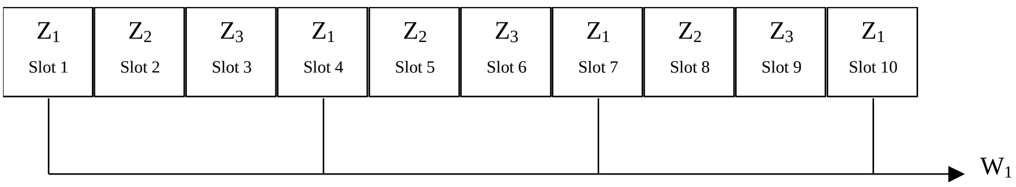

Figure 3. Filtering of the feedback commands

The filter itself could be of various types but a simple solution is to use the same averaging equation as in the Rel.-99 mode 1. End of frame adjustments can also be made as in the case of Rel. –99 mode 1. Some of the combinations fed to the filter lead to a filtered output of zero. In those cases, either the previously used antenna weight can be reused, or the next closest weight constellation point can be used.

5.4 Tx diversity scheme with beamforming feature

It is desirable that the closed loop multiple antenna transmit diversity/beamforming scheme can support a variety of antenna configurations and beamforming algorithms. To achieve it efficiently, the higher layer signaling information about Tx antenna configurations of Node B to UE is necessary. As the spatial correlation largely depends on the Tx antenna configurations in most cases, it will greatly help UEs to determine how appropriately the diversity and beamforming be combined, e.g. the number of beams and the feedback frame format.

Spatial correlation property depends on both the transmit antenna configuration and the radio propagation environment. The latter is unpredictable and performance depends on how feedback scheme matches the channel. For example, in strong spatially correlated channels, frequent update of short-term diversity weights is not efficient. In spatially uncorrelated channels, feedback bits for long-term beamforming weights are useless. The former, selection of antenna configuration, is one of the design criteria for cellular operators. It is rather easy to control spatial correlation by choosing appropriate antenna configuration.

In theory, employing multiple transmit antenna elements can achieve both diversity gain and beamforming gain. If the antennas are placed far away from each other, maximum diversity gain can be achieved but, due to the grating lobe problem, the achievable beamforming gain is limited. On the other hand, if the inter-element spacing in the antenna array is small, maximum beamforming gain can be obtained but the diversity gain will be limited as signals from different antenna elements are highly correlated.

5.4.1 Description of our solution

According to the transmit antenna configuration of the Node B, UE calculate short-term diversity weights and long-term beamforming weights. The hierarchical weighting is defined as shown in Figure 4.

Consider an M sub-arrays configuration in which each sub-array consists of $K=N/M$ elements. Firstly, UE finds an M-dimensional short-term diversity weight vector $\underline{w}_D$, which maximize

$$P_D = \underline{w}_D{}^H H_D{}^H H_D \underline{w}_D \qquad (8)$$

with $H_D = [\underline{h}_1, \underline{h}_{K+1}, \cdots, \underline{h}_{(M-1)K+1}]$

where, $\underline{h}_{(m-1)K+1} \quad (m = 1 \cdots M)$ is the channel response vector, which represents the m-th sub-array.

Secondly, UE finds a K-dimensional beamforming weight vector $\underline{w}_{B,m}$ for each sub-array which maximize

$$P_{B,m} = \underline{w}_{B,m}{}^H H_{B,m}{}^H H_{B,m} \underline{w}_{B,m} \qquad (9)$$

with $H_{B,m} = [\underline{h}_{(m-1)K+1}, \underline{h}_{(m-1)K+2}, \cdots, \underline{h}_{(m-1)K+K}]$

where, $\underline{h}_{(m-1)K+k} \quad (k = 1 \cdots K)$ is the channel response vector of the k-th element in the m-th sub-array.

Then, short-term diversity weights $D_{1,m}$ and long-term beamforming weights $B_{m,k}$ for the hierarchical weighting are calculated from $\underline{w}_D$ and $\underline{w}_{B,m}$ as follows.

$$D_{1,m} = \frac{w_D(m)}{w_D(1)} \quad (m = 1 \cdots M) \tag{10}$$

$$B_{m,k} = \frac{w_{B,m}(k)}{w_{B,m}(1)} \quad (m = 1 \cdots M, \; k = 1 \cdots K) \tag{11}$$

$\{D_{1,m}\}$ corresponds to the M-branch Tx diversity weights and $\{B_{m,k}\}$ are beamformer weights for the m-th antenna group. The feedback frequency for $\{D_{1,m}\}$ is much higher than that for $\{B_{m,k}\}$ to suit for fast fading environment.

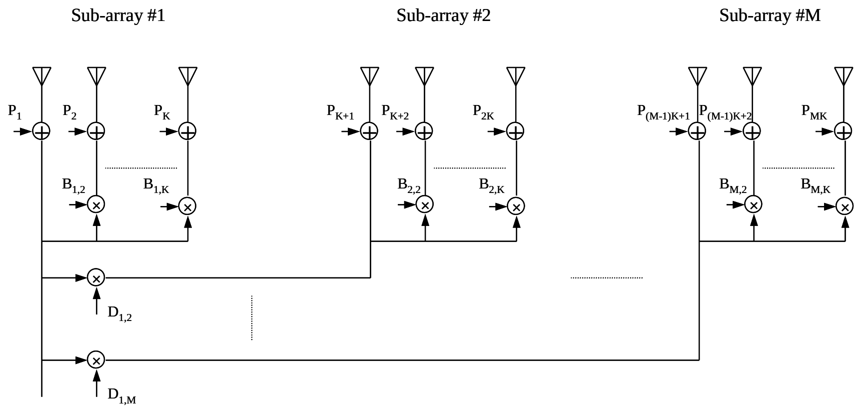

Figure 4. Hierarchical weighting for the transmit antenna diversity/beamforming

5.4.2 Example format of feedback information

The following frame formats for the feedback information bits are desirable. Format 1 is for Tx diversity and it allows employing the scheme without beamforming features. Format 2 is for beamforming with small inter-element spacing less than spatial correlation length. All of 15 bits/frame for beamformer weights can be used for accurate control of the beam. Format 3 is for combination of diversity and beamforming, which is suitable for the sub-array antenna configuration. Single beamformer weight is quantized by 3 bits and fed back in a frame to Node B. In case of two-sub-array system, the beamformer weight of sub-array #1 is transmitted in the first frame, then the beamformer weight of sub-array #2 is transmitted in the second frame. The number of antenna elements in each sub-array can be increased, as long as the feedback delay is permissible for updating the beamformer weights.

Table 3. Multiplexing format 1 of feedback information for Tx diversity

Slot #	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
FB bits for D	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
FB bits for B	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0

Table 4. Multiplexing format 2 of feedback information for beamforming

Slot #	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
FB bits for D	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
FB bits for B	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1

Table 5. Multiplexing format 3 of feedback information for combination of diversity and beamforming

Slot #	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
FB bits for D	1	1	1	1	0	1	1	1	1	0	1	1	1	1	0
FB bits for B	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1

5.5 Closed loop transmit diversity mode 2 with reduced states for 4 elements

Similar to Rel-99 closed loop Tx diversity systems, the weights of transmit antennas are determined at a mobile station and fed back to the base station. The set of weights at the UTRAN-AP $\underline{w}=[w_1\ w_2\ \ldots w_i \ldots\ w_M]^T$, where w_i is a complex weight associated with the ith Tx antenna are chosen to maximize P below:

$$P=\underline{w}^H H^H H\underline{w}, \tag{12}$$

where $H=[\underline{h}_1\ \underline{h}_2\ \ldots\underline{h}_i\ldots\underline{h}_M]$ and M is the number of Tx antennas. The column vector $\underline{h}_i$ represents an estimated channel impulse response for the ith Tx antenna, and its vector length equals to the number of paths.

Two extensions of the Rel'99 closed loop transmit diversity mode 2 for the case of 4 elements are considered with 4 phases and 8 phases states per element as defined in Table x1 and Table x2 respectively.

Table x1. FSM of modified closed loop mode 2 (4p0g) signalling message per element

FSM	Phase difference between antennas
00	$\pi/4$
01	$3\pi/4$
11	$-3\pi/4$
10	$-\pi/4$

Table x2. FSM of modified closed loop mode 2 (8p0g) signalling message per element

FSM	Phase difference between antennas
000	0
001	$\pi/4$
010	$3\pi/4$
011	$\pi/2$
100	$-\pi/4$
101	$-\pi/2$
110	π
111	$-3\pi/4$

Therefore 6 bits and 9 bits feedback per slot are needed for the update of the antenna coefficients for 4p0g and 8p0g respectively . If we consider **1 bit feedback per slot**, Progressive refinement (as described for Rel 99 Closed loop transmit diversity Mode 2) is used to update the antenna coefficients.

5.5.1 Format of Feedback Information

The uplink feedback information signalling is similar to the Rel-99 mode 2, that is using progressive refinement both at the UE and UTRAN-AP. The only change is the number of states, at the UE, to be compared for maximising $P=\underline{w}^H H^H H\underline{w}$, is increased due to 4 elements instead of 2 (Rel 99). The FSM (Tables x1 and x2) corresponding to coefficient of the antenna 2, 3 and 4 are sent successively.

5.6 Description of the 4-Tx-STTD diversity scheme

Starting point in the design of the proposed 4-Tx-STTD has been an observation that open loop (OL) schemas perform much better at high velocities than at stabile channels since of better interleaving caused by velocity. Complexity increase is also required to be minimize compared to 2-Tx-STTD, as well as full compatibility with rel'99 structures (e.g. interleavers).

4-tx-STTD utilizes Alamouti's space-time block code: encoding by 4-Tx-STTD means to transmit the first STTD diversity branch ($s_1, s_2, \ldots$) via the first and the phase rotated replica via the second antenna. The second STTD diversity branch is transmitted in an analog manner from the antennas 3 and 4.The general encoding scheme for 4-tx-STTD is presented in the Figure 5~~Figure 5Figure 5~~.

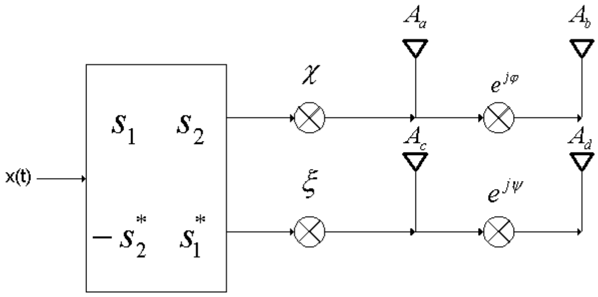

Figure 5. Encoding for 4-Tx-STTD.

Pseudo-antennas can be defined e.g. as $A_a := A_1 + A_2$, $A_b := A_3 + A_4$, $A_c := A_1 - A_2$ and $A_d := A_3 - A_4$ if power balancing is required.

Phase hopping by using 8-level quantization gives the required performance. The rotation phases are picked from a short look-up-table, keeping the same phases over the period of at least 2 symbols.

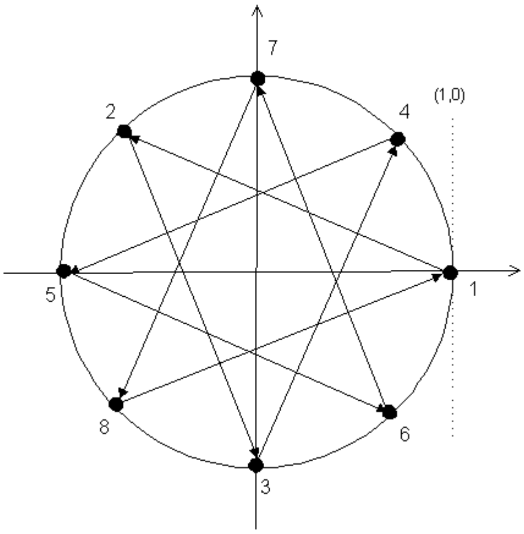

Figure 6 One optimal configuration of phase offsets

One optimal configuration is obtained when the periodic phase shift pattern of shifts in degrees of {0, 135, 270, 45, 180, 315, 90, 225} on pseudo antenna 2 and respectively {180, 315, 90, 225, 0, 135, 270, 45} on pseudo antenna 4 is applied.

It is required that $\varphi = \psi + \pi$, and therefore it is sufficient to have only one look-up table for the both phase rotated channels.

Decoding procedure for 4-Tx-STTD is similar to that of 2-Tx-STTD except that in the decoding of the STTD-branches estimates of rotated channels

$$\begin{cases} h_1 + \exp(j\varphi)h_2 \\ h_3 + \exp(j\psi)h_4 \end{cases} \tag{13}$$

are used, where h_i is the channel estimation of the i'th physical antenna. Gain factors χ and ξ are adjusted depending on channel estimation quality.

5.6.1 Channel estimation for the 4-Tx-STTD concept

Depending on the CpiCH transmission scheme used we distinguish two different cases how the channel estimation for 4-tx-STTD can be done.

<u>Channel estimation for symmetric pilots:</u>

In this case the common pilot power is evenly distributed between primary and secondary CPiCH, and all the four channels are estimated from common pilots (i.e., from P-CpiCH and S-CpiCH). In this case $\chi = \xi = 1$.

<u>Channel estimation for non-symmetric pilots:</u>

Non-symmetric channel estimation approach is used in the case that rel'99 backward compatibility reasons force us to avoid the use of S-CpiCH or it's power is at remarkably lower level than that of P-CpiCH. In this case the first STTD branch transmitting non-transformed signal is estimated from P-CpiCH and the additional channels transmitting the transformed signal are estimated from the dedicated pilots. This is illustrated in the Figure 7~~Figure 7Figure 7~~. In order to improve the quality of the dedicated channel estimates of the additional channels additional pilot power offset may be

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Formatted: Font: Times New Roman, Not Bold, English (United Kingdom) used. If we are forced to use non-symmetric channel estimation, then from the performance point of view it is beneficial to set $\chi > \xi$.

The non-symmetric pilot arrangement minimizes the effect of channel estimation errors and guarantees that 4-tx-STTD always outperforms one antenna transmission ($\xi = 0$). The scheme is also well power balanced.

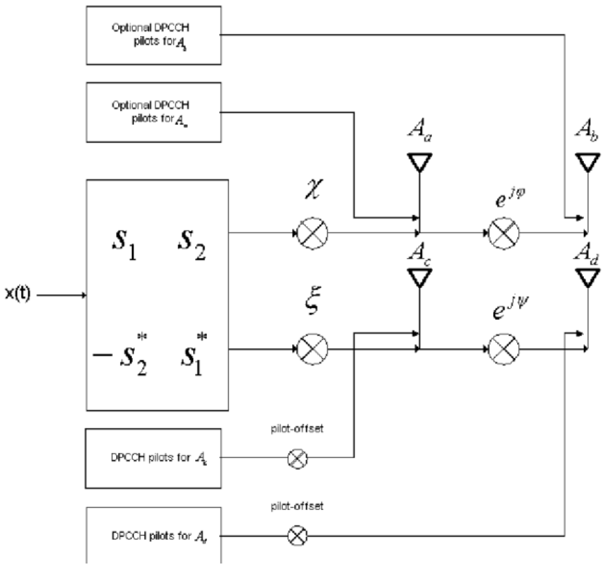


Figure 7. Using dedicated pilots for channel estimation of the second diversity branch.

5.6.2 Phase rotation frequency of the 4-Tx-STTD concept

One issue that should be considered when the 4-Tx-STTD is used is the phase rotation frequency. Two different options exists. First option is to use fixed phase rotation frequency *in time* meaning that the rotation phases are updated once in a certain time period. Another option is to use fixed phase rotation frequency *in symbols* meaning that a certain number of consecutive 4-Tx-STTD encoded data symbols in phase rotated channels are rotated by using the same rotating phase.

5.7 STTD-OTD scheme

The STTD-OTD technique for a symbol sequence S_1, S_2, S_3 and S_4 is given below [18]:

Antenna 1: S_1 | S_1 | S_2 | S_2 |

Antenna 2: $-S_2^*$ | $-S_2^*$ | S_1^* | S_1^* |

Antenna 3: S_3 | $-S_3$ | S_4 | $-S_4$ |

Antenna 4: $-S_4^*$ | S_4^* | S_3^* | $-S_3^*$ |

A block diagram of the transmitter chain is shown in Figure 8 below. A symbol level scrambling to randomize the fading antenna pattern for STTD-OTD is done before STTD-OTD encoding. The four possible ways that can be employed to scramble the symbol sequence S_1, S_2, S_3, S_4 are given below:

(b_0,b_1)

(0, 0): S_1, S_2, S_3, S_4

(0, 1): S_2, S_1, S_4, S_3

(1, 0): S_3, S_4, S_1, S_2

(1, 1): S_4, S_3, S_2, S_1

The value of the bits b_0, b_1 determines the scrambling of the symbols S_1, S_2, S_3, S_4. The bits b_0, b_1 can be derived from the long code for the base station. This symbol level scrambling is employed before the STTD-OTD encoding is done.

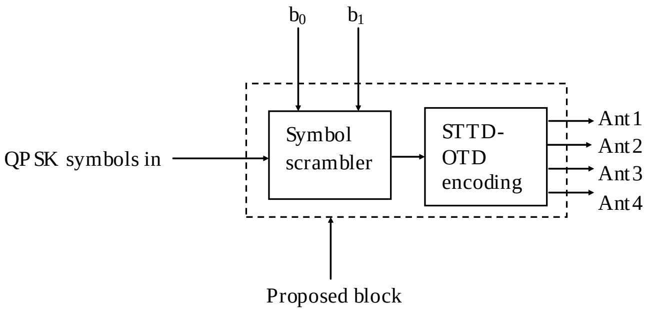


Figure 8. Proposed symbol level scrambling with STTD-OTD at the transmitter is shown.

5.8 Closed-Loop STTD with multiple antennas

The Closed-Loop (CL) STTD with multiple antennas is a concept valid for any even number of transmit antennas. Here it will be described for the case of 4 antennas (CL-4-Tx-STTD). The transmitter of CL-4-Tx-STTD is depicted in Figure 9~~Figure 9~~~~Figure 9~~ [26]-[28].

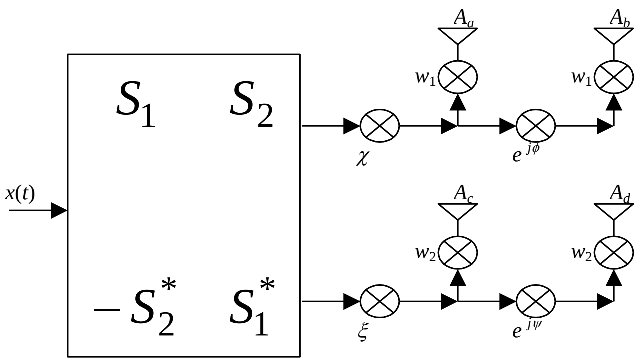


Figure 9. CL-4-Tx-STTD transmitter.

The received signal during two symbol periods at the receiver front end with one antenna element can be represented as

$$r_1 = w_1 \alpha S_1 - w_2 \beta S_2^* + n_1$$

$$r_2 = w_1 \alpha S_2 + w_2 \beta S_1^* + n_2 \qquad (14)$$

where S_1 and S_2 are the data symbols, and n_1 and n_2 are complex Gaussian noise samples. The parameters α and β are defined as follows,

$$\alpha = \sum_{k=1}^{K} \alpha_k = \sum_{k=1}^{K} \left(h_{1k} + h_{2k} e^{j\phi} \right)$$

$$\beta = \sum_{k=1}^{K} \beta_k = \sum_{k=1}^{K} \left(h_{3k} + h_{4k} e^{j\psi} \right), \tag{15}$$

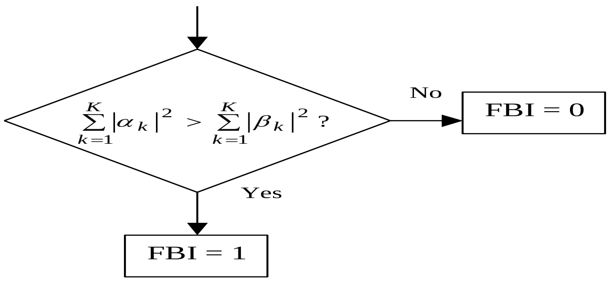
where h_{1k}, h_{2k}, h_{3k} and h_{4k} are the complex amplitude coefficients of the propagation paths from 4 transmit antennas, having the same propagation delay. There are K propagation paths from each transmit antenna. The parameters ϕ and ψ are the rotation phases. It is required that $\phi = \psi + \pi$, so one of the rotation phases is picked up from a look-up table given in Section 5.6, while the other is calculated accordingly. The rotation phase should be kept constant over the period of at least 2 symbols.

The symbols S_1 and S_2 can be decoded from r_1 and r_2 by using the receiver consisting of an ordinary STTD decoder, having α_k replaced by $\alpha_k w_1$, and β_k replaced by $\beta_k w_2$.

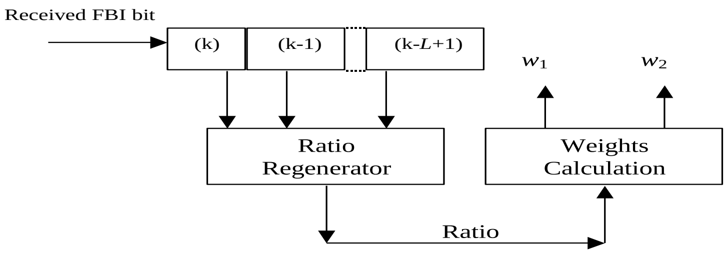
The weights w_1 and w_2 are defined as [27]

$$w_1 = \frac{1}{\sqrt{1 + \left(\sum_{k=1}^{K} |\beta_k|^2 / \sum_{k=1}^{K} |\alpha_k|^2 \right)^2}}, \quad w_2 = \frac{1}{\sqrt{1 + \left(\sum_{k=1}^{K} |\alpha_k|^2 / \sum_{k=1}^{K} |\beta_k|^2 \right)^2}}. \tag{16}$$

The feedback quantization method, consisting of FBI bit generation (encoding) in the UE and decoding in the Node B, is shown in Figure 10.



(a) FBI encoding



(b) FBI decoding

Figure 10. Predictive Feedback Quantization method.

The one-bit FBI encoding of power ratio of two propagation channels is performed similarly as for selection transmit diversity: if the power of the first propagation channel is greater than or equal to the power of the second channel, FBI bit is set to 1, otherwise FBI bit is set to 0. The FBI bit is then transmitted to Node B.

The received FBI decoder at the NodeB consists of a delay-line buffer for storing a number of most recently received FBI bits, and the ratio regenerator. The ratio regenerator regenerates the power ratio $\sum_{k=1}^{K} |\alpha_k|^2 / \sum_{k=1}^{K} |\beta_k|^2$. The ratio regenerator is a look-up table. The three different ratio regenerators are given in Figure 11~~Figure 11Figure 11~~, corresponding to the decoding lengths L equal to 1, 2 and 3 FBI bits. Once the power ratio is regenerated (Ratio=$10^{\text{Ratio(dB)}/10}$), the weights w_1 and w_2 are calculated according to the equation (16).

Mapping Table for *L* = 1

FBI(k)	Ratio (dB)
1	6
0	-6

Mapping Table for *L* = 2

FBI(k)	FBI(k-1)	Ratio (dB)
1	1	6
1	0	0
0	1	0
0	0	-6

Mapping Table for *L* = 3

FBI(k)	FBI(k-1)	FBI(k-2)	Ratio(dB)
0	0	0	-6
0	0	1	-4
0	1	0	0
0	1	1	-2
1	0	0	2
1	0	1	0
1	1	0	4
1	1	1	6

Figure 11. Power ratio regenerators for 1, 2, and 3 consecutive FBI bits.

5.8.1 Antenna verification algorithm

The outcome of antenna verification algorithm is the estimate $R_q^{(est)}$ of the quantized power ratio R_q used in the Node B's FBI decoder. The FBI decoder of length 1 will be assumed. Thus in each slot there are two hypothesises to choose among in the UE. The hypothesis H_0 corresponds to the FBI=0, while the hypothesis H_1 corresponds to FBI=1.

The channel coefficient of i-th path estimated in the UE from the CPICH from transmit antenna 1 is denoted by $\alpha_i^{(p)}$ and from antenna 2 by $\beta_i^{(p)}$. The channel coefficient of i-th path estimated in the UE from the DPCCH from transmit antenna 1 is denoted by $\alpha_i^{(d)}$ and from antenna 2 by $\beta_i^{(d)}$.

The hypothesis H_0, i.e. $R_q^{(est)}=R_0=4$, should be chosen if the following inequality is satisfied

$$\frac{\gamma(1+R_0)}{\sqrt{1+R_0^2}}\sum_{i=1}^{K}\left[\frac{\left|\alpha_i^{(p)}\right|^2}{\sigma_i^2}-\frac{\left|\beta_i^{(p)}\right|^2}{\sigma_{K+i}^2}\right]-2\sum_{i=1}^{K}\mathrm{Re}\left\{\frac{\alpha_i^{(d)}\cdot\alpha_i^{(p)^*}}{\sigma_i^2}-\frac{\beta_i^{(d)}\cdot\beta_i^{(p)^*}}{\sigma_{K+i}^2}\right\}>\frac{\sqrt{1+R_0^2}}{\gamma(1-R_0)}\ln\frac{P(H_1)}{P(H_0)},$$

where $\gamma^2=\mathrm{SNR}_{\text{dpcch pilot}}/\mathrm{SNR}_{\text{cpich}}$.

The a priori probabilities of both hypothesises in each slot can be determined on the basis of the previously issued FBI bit from the UE (taking also into account the FBI delay from the UE to the NodeB). If it is the FBI=0 that has been previously issued by the UE and used by the NodeB in the current slot after the delay known to the UE, then the a priori probability of the hypotheses H_0 is $P(H_0)=1-p_{fbe}$, while $P(H_1)=p_{fbe}$, where p_{fbe} is the feedback error rate. Typically the feedback error rate is assumed to be 4%, so an upper bound (e.g. of 10%) can be used. If there is no exact knowledge of the feedback delay, then the a priori probabilities of both hypothesises have to be assumed equal.

5.9 4Tx OL-CL Diversity

It is desirable that Tx diversity solutions shall be applicable to a variety of different antenna constellations. The 4Tx open loop – closed loop (OL-CL) diversity scheme [34] is suitable for application with sub-array antenna arrangements, in which case a combined spatial/polarisation diversity and beamforming gain is achievable. Possible 4Tx sub-array antenna arrangements include two spatially separated 2-element linear-polarised arrays or an array using two cross-polarised elements. The elements of a sub-array are assumed to have a narrow spacing, e.g. half a wavelength.

The 4Tx OL-CL scheme is further characterised by the following design considerations:

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- The 4Tx OL-CL scheme combines the existing R99/4/5 2Tx diversity solutions.
- Closed loop feedback mechanisms are applied in order to provide beamforming gain. Due to small antenna spacing, closed loop beamforming is comparatively insensitive to UE movement.
- A spatial or polarisation diversity gain is achievable in downlink by means of the STTD component, provided the sub-arrays are well separated in position or polarisation orientation, respectively.

The Node B transmit part of the 4Tx OL-CL diversity scheme is depicted in Figure 12. The symbol stream to be transmitted is split into two STTD encoded sub-streams. These two sub-streams are transmitted via different 2Tx sub-arrays, each using 2Tx Closed Loop Mode 1 (i.e. a complex-valued weighting factor is used per 2Tx sub-array).

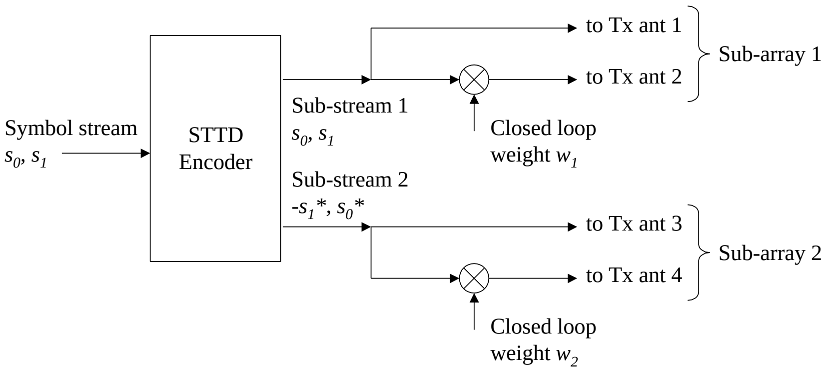

Figure 12. Node B transmit part of 4Tx OL-CL diversity scheme.

We use a notation as indicated in Figure 12 to describe the UE receiver algorithms for the 1Rx case. For the sake of convenience, four one-path fading channels characterised by the complex-valued coefficients h_1, h_2, h_3, and h_4 are assumed. It should be noted that the 4Tx OL-CL diversity scheme is applicable also in multipath case. In response to the two transmitted symbols s_0 and s_1, the UE receives the two symbols

$$r_0 = (h_1 + h_2 w_1) s_0 - (h_3 + h_4 w_2) s_1 *,$$

and

$$r_1 = (h_1 + h_2 w_1) s_1 + (h_3 + h_4 w_2) s_0 *,$$

where w_1 and w_2 denote the complex-valued antenna weights for 2Tx closed loop diversity for sub-array 1 and 2, respectively, and s* denotes the conjugate complex of the symbol s. UE performs channel estimation for all the four propagation channels, e.g., by means of a CPICH transmission scheme for >2Tx antennas as described in §6. Using knowledge of the channel coefficients h_1, h_2, h_3 and h_4, and knowledge of the antenna weights w_1 and w_2, UE applies its R99/4/5 channel weighting algorithm to compute the effective channel coefficients $\tilde{h}_1 = (h_1 + h_2 w_1)$ and $\tilde{h}_2 = (h_3 + h_4 w_2)$ seen from sub-array 1 and sub-array 2, respectively. UE can now apply the R99/4/5 STTD decoding algorithm in order to obtain estimates $\hat{s}_0$ and $\hat{s}_1$ of the transmitted symbols s_0 and s_1 according to

$$\hat{s}_0 = \tilde{h}_1 * r_0 + \tilde{h}_2 r_1 *,$$

and

$$\hat{s}_1 = -\tilde{h}_2 r_0 * + \tilde{h}_1 * r_1 .$$

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To determine optimum antenna weights w_1 and w_2, UE applies its R99/4/5 feedback information computation algorithm such as to separately maximise $P_1 = w_1^H H_1^H H_1 w_1$ and $P_2 = w_2^H H_2^H H_2 w_2$, where $H_1 = [h_1 \; h_2]$, $H_2 = [h_3 \; h_4]$, $w_1 = [1 \; w_1]^T$, and $w_2 = [1 \; w_2]^T$.

The building blocks of a UE receiver are sketched in Figure 13 for 4Tx OL-CL diversity transmission. As can be seen, the UE can fully reuse its R99/4/5 receiver algorithms.

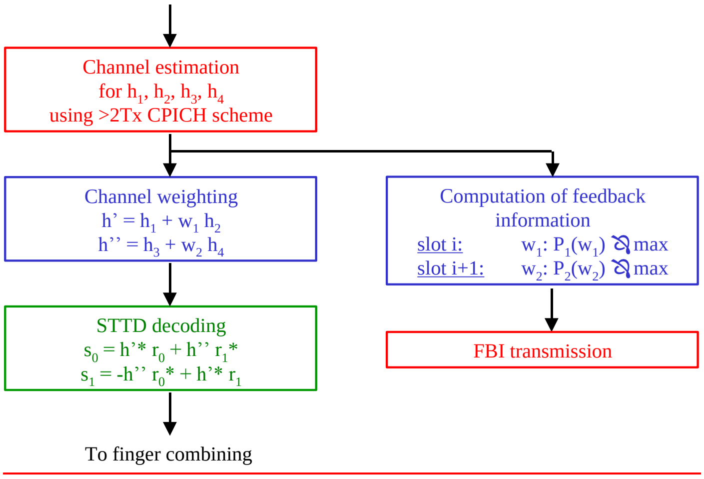

Figure 13. Building blocks of a UE receiver in case of a 4Tx OL-CL diversity transmission. Blue and green colours indicate that R99/4/5 algorithms can be reused, whereas red colour indicates an algorithm not used in R99/4/5.

An FBI transmission scheme for use with 4Tx OL-CL diversity is proposed in Figure 14. The capacity for FBI on the UL DPCCH is limited to 1 bit per slot, and thus a FBI transmission scheme is used in which the bits to represent the weights w_1 and w_2 of the two sub-arrays are transmitted in time multiplex.

UL slot #	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14
w_1 for sub-array 1	I_1	-	Q_1	-	I_1	-	Q_1	-	I_1	-	Q_1	-	I_1	-	Q_1
w_2 for sub-array 2	-	I_2	-	Q_2	-	I_2	-	Q_2	-	I_2	-	Q_2	-	I_2	-

Figure 14. Possible FBI transmission scheme for 4Tx OL-CL diversity scheme. Per antenna weight w, a bit pair (I,Q) is used for coding the phase of w as with R99/4/5 2Tx CL Mode 1, i.e., I/Q = +1 if real/imaginary part of w > 0, 0 else. The bit I/Q transmitted in slot #i is updated in the same slot prior to transmission.

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6 New CPICH Transmission Schemes

6.1 New CPICH transmission scheme for > 2 Tx antennas with reduced PAPR and power balancing

In Release 99 specification, transmit diversity using 2 antennas is included. Currently, various transmit diversity schemes using 4 transmit antenna are considered for possible extension to Release 2000. However, the most important thing that should be solved first is the pilot reference channel for 4-antenna.

In this document, a new CPICH transmission scheme is propose for 4-antenna transmit diversity. The most important thing that should be kept in mind when proposing new CPICH transmission scheme for 4-antenna transmit diversity is the backward compatibility. Here, the backward compatibility means that the Release 99 UE should be able to demodulate the whole physical channels (dedicated or common physical channel) without any change in receiver structure.

In section 6.1.1, the transmission scheme of CPICH for 4-antenna transmit diversity is described and the backward compatibility is proved. In section 6.1.2, corresponding common physical channel transmission scheme with 4 transmit antenna to satisfy the backward compatibility is proposed. And the corresponding transmission schemes of dedicated physical channel with 2-antenna and 4-antenna transmit diversity UE are described in section 6.1.3~~6.1.3~~~~6.1.3~~.

Let's distinguish UEs by their diversity mode as following.

2-ant diversity UE: UE in 2-antenna diversity mode (open/closed)
4-ant diversity UE: UE in 4-antenna diversity mode (open?/closed?)

6.1.1 CPICH Transmission Scheme

6.1.1.1 CPICH Transmission Scheme

If UTRAN supports 4 transmit diversity (open or closed loop) for dedicated channel to UE in the cell, then it should provide 3 additional diversity pilot channels as well as primary CPICH. However, since the CPICH is a common physical channel it also should be received by all UEs with different diversity mode, and thus one and only one CPICH transmission scheme should be used. Each UE should estimate the channel(s) as many as the number of transmit antenna, since each antenna has its own path. That is, common CPICH transmission scheme for 4 transmit antenna shall have the property that it must be recognised as one, two, or 4 pilot channels to 2-ant diversity UE and 4-ant diversity UE, respectively. Figure 15~~Figure 15~~~~Figure 12~~ shows the proposed CPICH transmission scheme for 4 transmit antenna which satisfies the property. The main characteristics of the proposed CPICH transmission scheme are:

- using two OVSF codes (C_{OVSF1} and C_{OVSF2})
- same pilot pattern as Release 99' 2-ant CPICH (AA and A-A/ –AA)
- different control of pilot channel gain for 2-ant or 4-ant diversity reception
- backward compatible with Release 99
- reduce PAPR by distributing physical channels to 4 antenna

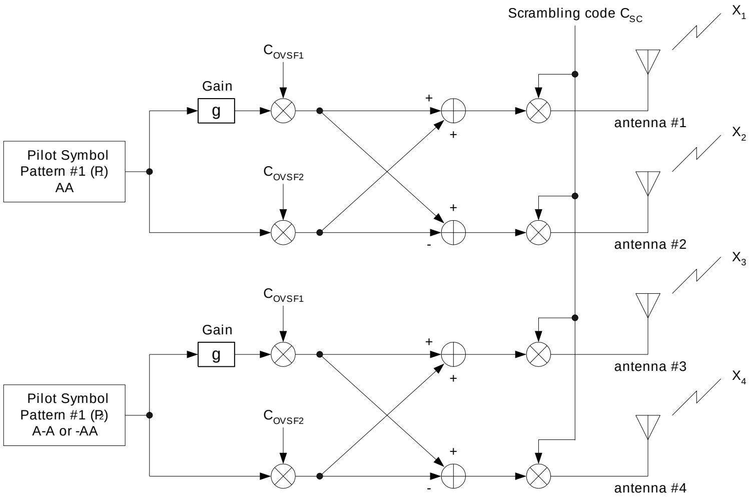

Figure 15~~15~~~~12~~. Proposed CPICH transmission scheme for 4 antenna transmit diversity

The CPICH signal from each antenna at the receiver side is given by the following equations. The timing index and background noise is ignored for simplicity.

$$
\begin{aligned}
X_1 &= \mathrm{P}_1 \times (g \cdot C_{OVSF1} + C_{OVSF2}) \times C_{SC} \times h_1 \\
X_2 &= \mathrm{P}_1 \times (g \cdot C_{OVSF1} - C_{OVSF2}) \times C_{SC} \times h_2 \\
X_3 &= \mathrm{P}_2 \times (g \cdot C_{OVSF1} + C_{OVSF2}) \times C_{SC} \times h_3 \\
X_4 &= \mathrm{P}_2 \times (g \cdot C_{OVSF1} - C_{OVSF2}) \times C_{SC} \times h_4
\end{aligned}
\qquad (17)
$$

where P_1 (=AA) and P_2 (=A-A or –AA) are the two pilot patterns defined for 2-ant CPICH in Release 99, and C_{SC} is the primary scrambling code. In Eq. (17), C_{OVSF1} and C_{OVSF2} are two OVSF codes where C_{OVSF1} is $C_{ch,256,0}$ and C_{OVSF2} is one additional OVSF code. h_1, h_2, h_3, h_4 are the channel coefficients for each antenna path. It is worth noting that the parameter *g* in Eq. (17) is the gain factor to discriminate the received pilot power for 2-ant and 4-ant diversity UE. By varying the gain *g*, the received pilot strength can be controlled to 2-ant diversity UE and 4-ant diversity UE. For 2-ant diversity UE only C_{OVSF1} is used and thus the second term in Eq. (17) is removed. On the other hand, for 4-ant diversity UE, both C_{OVSF1} and C_{OVSF2} will be used and it can discriminate 4 different antenna paths. Detail receiver structure of different diversity UE are described in next section.

6.1.1.2 Receiver Structure of CPICH

6.1.1.2.1 Receiver structure of 4-ant diversity UE

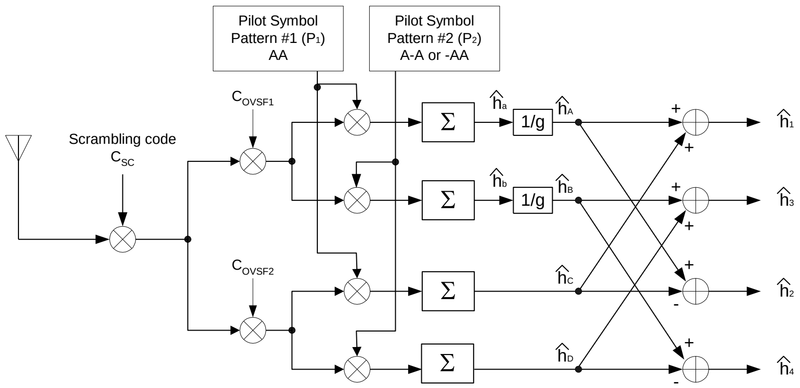

Figure 16~~1613~~. Receiver structure of 4-ant diversity UE

Figure 16~~Figure 16Figure 13~~ shows how the 4-ant diversity UE can receive and estimate the 4 channels. In Figure 16~~Figure 16Figure 13~~, $\hat{h}_a, \hat{h}_b$ denote the channel estimation of $h_a = g(h_1 + h_2)$, $h_b = g(h_3 + h_4)$, respectively. Similarly, $\hat{h}_A, \hat{h}_B, \hat{h}_C, \hat{h}_D$ denote the estimation of $h_A = h_1 + h_2$, $h_B = h_3 + h_4$, $h_C = h_1 - h_2$, $h_D = h_3 - h_4$, respectively. Note that these channel estimation pairs $\{\hat{h}_a, \hat{h}_b\}$, $\{\hat{h}_A, \hat{h}_B, \hat{h}_C, \hat{h}_D\}$, or $\{\hat{h}_1, \hat{h}_2, \hat{h}_3, \hat{h}_4\}$ can be used to compensate the common or dedicated physical channels.

6.1.1.2.2 Receiver structure of 2-ant diversity UE

Figure 17~~Figure 17Figure 14~~ is the CPICH receiver structure of 2-ant diversity UE and it can also be used with the proposed CPICH transmission scheme without any change. Note that the channel estimation value with the receiver is exactly same as the output of the first branch output in Figure 16~~Figure 16Figure 13~~. That is, $h_a = g(h_1 + h_2)$ and $h_b = g(h_3 + h_4)$. Consequently, there's no change required to Release 99 UE in terms of pilot channel estimation. The only difference is the effective channel through which the pilot signal is transmitted. In order for the proposed CPICH transmission scheme to become fully backward compatible , UTRAN also should modify the transmission structure of common/dedicated physical channels with considering the channel estimation outputs of 2-ant diversity UE, and 4-ant diversity UE. The detail of the common/dedicated physical channel transmission structure is given in section 6.1.2~~6.1.26.1.2~~ and 6.1.3~~6.1.36.1.3~~.

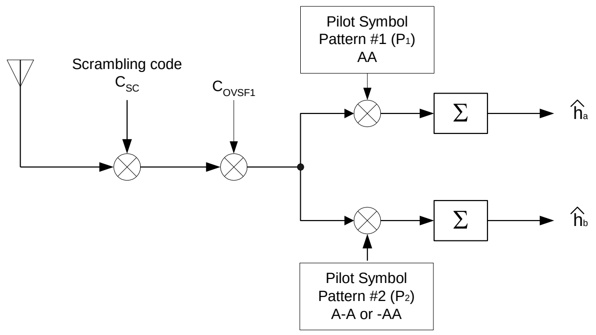

Figure 17~~1714~~. Receiver structure of 2-ant diversity UE

6.1.1.2.3 Summary of channel estimation outputs

Table 6. shows the summary of demodulation parameters and the channel estimation output according to the UE diversity mode. In Table 6 the related physical channel implies the physical channel that utilises the corresponding channel estimation output during demodulation. The main idea of the Table 6 is that the transmission structure of the related physical channel should be designed carefully with considering the corresponding channel estimation output.

Table 6. Summary of demodulation parameters and channel estimation output

Rx parameters / UE mode	Pilot Channel				Related physical channel
	scrambling code	channelisation code	pilot pattern	channel estimation output	
2-ant diversity	C_{SC}	$C_{OVSF1} = C_{ch,256,0}$	AA A-A/-AA	$h_a = g(h_1 + h_2)$ $h_b = g(h_3 + h_4)$	Common CH Dedicated CH
4-ant diversity	C_{SC}	$C_{OVSF1} = C_{ch,256,0}$ $C_{OVSF2} = C_{ch,256,1}$	AA A-A/-AA	$h_a = g(h_1 + h_2)$ $h_b = g(h_3 + h_4)$	Common CH
				h_1, h_2, h_3, h_4	Dedicated CH

6.1.2 Common Physical Channel Tx Scheme

6.1.2.1 Common Physical Channel Tx Scheme

Common physical channel should be transmitted with one and only one transmission scheme. However, each UE should receive the common physical channel as their transmit diversity mode. Figure 18~~Figure 18Figure 15~~ is the proposed common physical channel transmission scheme where the original symbols (S_1, S_2) are transmitted to antenna 1 and 2, while the STTD encoded symbols ($-S_2^*$, S_1^*) are transmitted to antenna 3 and 4. Backward compatibility of this scheme can be easily proved and shown in section 6.1.2.2~~6.1.2.26.1.2.2~~.

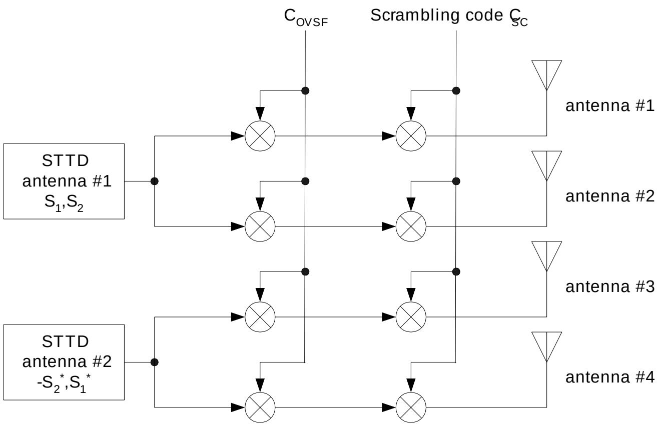

Figure 18~~1815~~. Common Physical Channel Transmission

6.1.2.2 Receiver Structure of Common Physical Channel Tx Scheme

If the common physical channel transmission scheme in Figure 18~~Figure 18Figure 15~~ is used, 2-ant and 4-ant diversity UE can receive the signals with the conventional STTD decoder as shown in Figure 19~~Figure 19Figure 16~~.

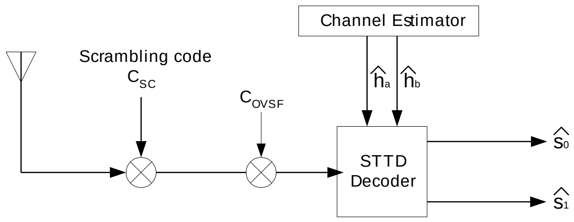

Figure 19~~19~~~~16~~. Receiver Structure for Common Physical Channel Transmission Scheme 1

The received signal after multiplication of OVSF in Figure 19~~Figure 19Figure 16~~ is given by

$$\begin{aligned} r_{t1} &= S_1(h_1+h_2) - S_2^*(h_3+h_4) = S_1 h_a - S_2^* h_b \\ r_{t2} &= S_2(h_1+h_2) + S_1^*(h_3+h_4) = S_2 h_a + S_1^* h_b \end{aligned} \quad (18)$$

where $t1, t2$ denote the time unit. Since the channel estimation provided by 2-antenna CPICH receiver are $\hat{h}_a$ and $\hat{h}_b$ (See Table 6), conventional STTD receiver can be used without any change.

6.1.3 Dedicated Physical Channel Tx Scheme

In case of a dedicated physical channel, transmission scheme should be different for each UE according to the diversity mode. However, the transmission scheme should be carefully designed with considering the available channel estimation output as given in Table 6.

6.1.3.1 Dedicated Physical Channel Tx Scheme for R99/R4 UE

For R99/R4 UE which does not know the usage of 4 ant transmission in the Node-B, the available channel estimation is $h_a = g(h_1+h_2)$, $h_b = g(h_3+h_4)$ (See Table 6). It means that the antenna 1 and antenna 2 should transmit one signal, and antenna 3 and antenna 4 should transmit the other signal. Based on the above constraint, Figure 20~~Figure 20Figure 17~~ and Figure 21~~Figure 21Figure 18~~ show the proposed transmission scheme for dedicated physical channel to 2-ant STTD and closed loop transmit diversity UE, respectively. By distributing 2-antenna signals to 4-antenna, such situation that the power of R99/R4 UE's concentrate on two antenna can be avoided. Definitely, it reduces the PAPR (Peak to Average Power Ratio)

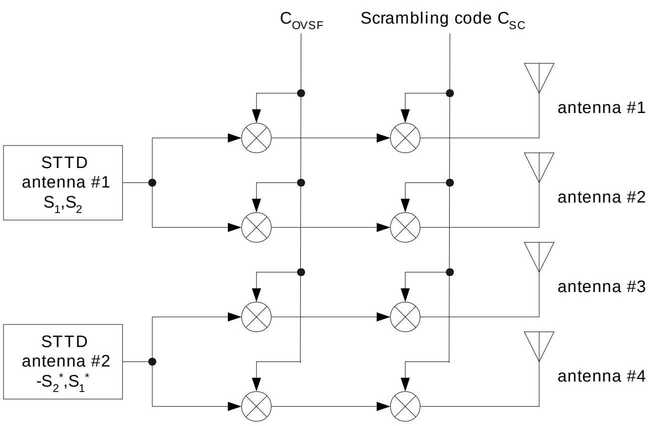

Figure 20~~17~~. Dedicated physical channel transmission scheme for R99/R4 STTD diversity UE

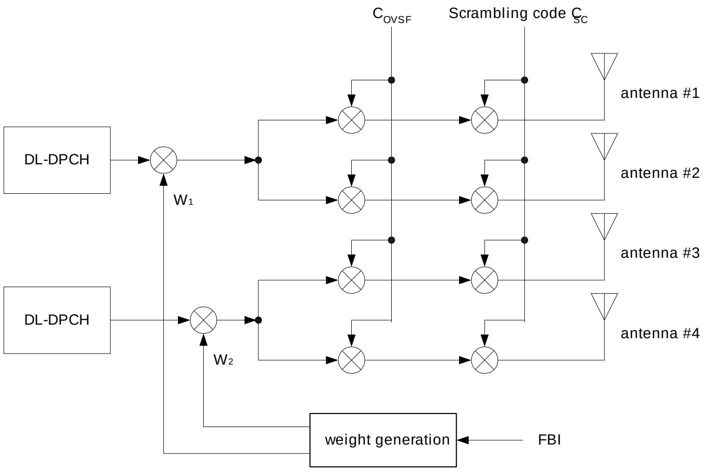

Figure 21~~18~~. Dedicated physical channel transmission scheme for R99/R4 closed loop transmit diversity UE

6.1.3.2 Dedicated Physical Channel Transmission Scheme for beyond R99/R4 UE

If a dedicated physical channel is transmitted to a beyond R99/R4 diversity UE which assumes knowing the usage of 4 ant transmission in the Node-B, e.g. R5 and/or R6, the transmission scheme should be designed with considering the available channel estimations output to the UE. With the proposed CPICH transmission scheme, the available channel estimation output of the beyond R99/R5 UE are $\hat{h}_1, \hat{h}_2, \hat{h}_3, \hat{h}_4$ (See Table 6). Currently, there is no accepted 4-antenna open/closed loop transmit diversity scheme but the proposed CPICH transmission scheme can be used with any kind of open/closed loop diversity proposal.

6.2 New pilot scheme for 4-Tx-antennas

To keep the specification as simple and effective as possible we propose not to extend Tx diversity for 4-antennas to the all channels – instead we suggest to keep all the general measurements as indicated in Rel-99/4 specification. Only channels that could arguably benefit from 4-antenna Tx-diversity (e.g. DPCH and DSCH) are proposed to be extended. P-CPiCH is would remain basically unchanged and could be used as a reference for measurements in Rel-99/4 UEs as well.

It seems evident that different 4-Tx-diversity concept state different requirements for pilots. One important requirement is the possibility for good channel estimations for additional channels. We propose to utilize dedicated pilots for the additional channels, and recommend to justify slot formats that are favorable for each concept, including possibility to utilize power offsets for dedicated pilots.

The use of S-CPiCH will add the overall power loss of a cell when reserving acceptable P-CPiCH level, hence its use is recommended only for the situations when no Rel'99 terminals are in the cell coverage area or when a proposed scheme offers big enough gain to compensate for S-CPiCH energy.

6.2.1 Description of the proposed pilot transmission scheme

This section contains the description of the non-symmetric orthogonal pilots. Let a pseudo-antenna configuration be defined by

$$\begin{cases} A_a := A_1 + A_2 \\ A_b := A_3 + A_4 \\ A_c := A_1 - A_2 \\ A_d := A_3 - A_4 \end{cases} \tag{19}$$

where $A_1,...,A_4$ are the physical antennas; another configuration would be e.g. $A_a := A_1$, $A_b := A_2$, $A_c := A_3$ and $A_d := A_4$.

Then primary common pilots are spread over all antennas as shown in Figure 22~~Figure 22~~~~Figure 19~~ - hence P-CPiCH separates two orthogonal channels via four physical antennas such that P-CPiCH power is evenly distributed over them. Therefore, 2-Tx-antenna schemes (for Rel-99/4) utilize two sum channels (see [17]) (i.e., the corresponding dedicated channels are transmitted over four antennas, the first diversity branch via antennas 1&2, and the second via antennas 3&4). This arrangement guarantees good power balance.

In order to guarantee Rel-99/4 UE's full functionality (as long they are in the market), P-CPiCH energy should be adjusted as for 2-Tx-cells. All general measurements from common channels (e.g. SHO evaluation, idle mode cell re-selection and synchronization as indicated in [10]) are done using only the proposed P-CPiCH scheme – so there is no need to extend all the channels for 4-Tx-diversity schemas and the corresponding Rel-99/4 functionality can be maintained in later releases.

Dedicated pilots are proposed to be utilized in order to separate two additional channels, possibly with the help of S-CPiCH (e.g. according to the operators choice and to the amount of rel'99 UE's in the cell area; S-CPiCH energy may be time variant) with adjustable gain factor g as presented in Figure 23~~Figure 23~~~~Figure 20~~($g = g(t)$ is possibly time varying). Note that specifying $g = f$ we get the pilot extension scheme already described above. Dedicated pilots can be powered by 0-6 dB offsets (as allowed already in Rel-99/4 specification) and short pilot symbol intervals (e.g. two symbols) will not be allowed – each dedicated pilot arrangement is designed based on the corresponding Tx-diversity schema.

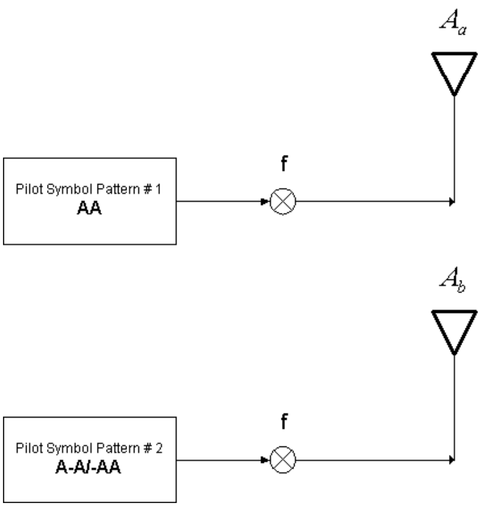

Figure 22~~22~~~~19~~. Transmission of P-CPiCH via pseudo-antennas A_a, A_b.

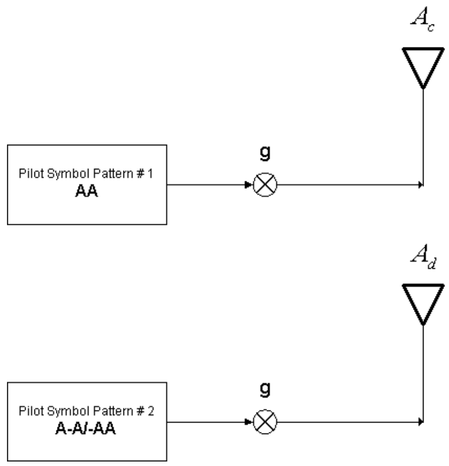

Figure 23~~2320~~. Transmission of S-CPiCH via two pseudo-antennas $A_{c,} A_d$ with adjustable gain factor *g*.

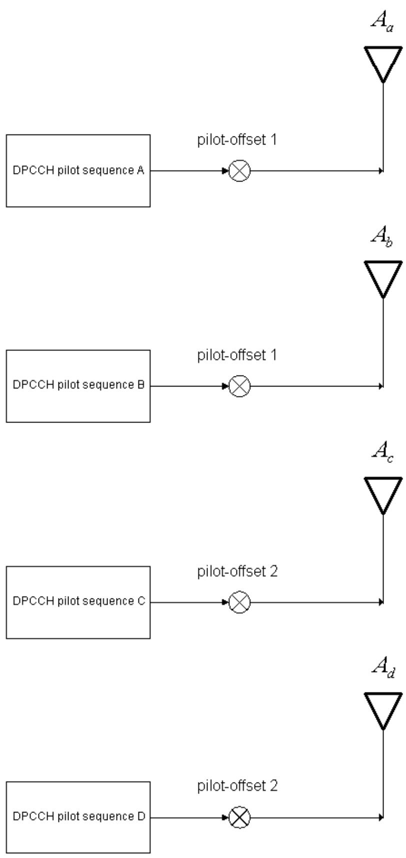

Figure 24~~24~~~~21~~. Transmission of DPCCH pilots via two or four pseudo-antennas $A_a, A_b, A_{c,} A_d$ with adjustable pilot offsets. Four orthogonal Hadamard sequences are used as pilot symbols.

6.3 New CPICH Transmission scheme for > 2 Tx antennas

In Release 99 specification, transmit diversity using 2 antennas is included. For an extension of transmit diversity schemes using more than 2 transmit antennas the definition of a new CPICH transmission scheme is essential. The UE must be enabled to perform channel estimation for each antenna. In this section, a new CPICH transmission scheme is described for 4-antenna transmit diversity. The shown principles can be also applied to more than 4 antennas.

The following principles apply:

- definition of the orthogonal pilots for CPICH using OVSF codes
- same pilot patterns (modulation with AA and A-A/-AA) as in Release 99 are used
- definition of unequal CPICH power ratio between CPICH1&2 and CPICH3&4
- possibility to utilize only dedicated pilots

A detailed example of the new CPICH transmission scheme for four antennas that also has the advantage of reduced PAPR and balancing the power of the antennas at the Node B is described in section 6.1.

6.3.1 Definition of new orthogonal pilot sequences

The definition of new orthogonal pilot sequences is described in this section. The extension to four antennas is shown inFigure 25~~Figure 25~~~~Figure 22~~. One can see the modulated CPICH for two antennas as specified in Rel99/4, which is denoted here as CPICH1 and CPICH2. Hereby the two different pilot symbol patterns are spread using channelisation code $C_{ch,256,0}$.

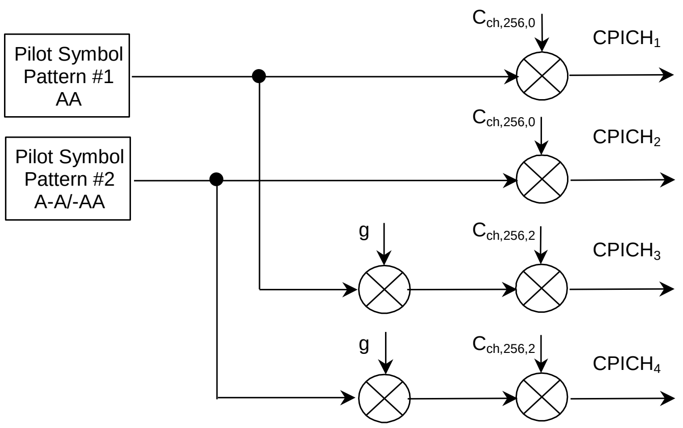

Figure 25~~25~~~~22~~ Orthogonal pilots for 4 antennas

When using four antennas, two additional orthogonal pilot sequences are defined. The same pilot symbol patterns as in Rel99/4 spread by a different channelisation code are used. To keep the complexity at the UE simple a channelisation code as close as possible in the code tree to Cch,256,0 is desired. However, code Cch,256,1 is already reserved for the Primary CCPCH, so the next available one is Cch,256,2. This code is therefore used for spreading of CPICH3 and CPICH4. The resulting structure can be seen in Figure 25~~Figure 25~~~~Figure 22~~.

Note that scrambling is done with the same scrambling code for all channels and is not shown in Figure 25~~Figure 25~~~~Figure 22~~.

For the Rel 99/4 DPCH it shall be possible to use CPICH1 and CPICH 2 as a phase reference. If CPICH is used as a phase reference for a DPCH using four antennas it shall be possible for the UE to use CPICH 1, 2, 3 and 4.

All common channels (other than pilot) are using CPICH1 and CPICH2 as phase reference.

It is still possible for the UTRAN to use non-standardised beamforming techniques. Hereby a S-CPICH or dedicated pilot bits can be used as phase reference for a DPCH.

The need for a further extension to more than 4 antennas is ffs.

6.3.2 Unequal power setting for CPICH

The total power of the CPICH 1,2,..,4 can be configured by the network. The power of CPICH3&4 can be adjusted relative to the power of CPICH1&2 by the gain factor g with $0 \leq g \leq 1$. It is ffs whether good performance of TX diversity techniques requires that the value of g is signalled to the UE.

Examples:

If g = 1 the power on CPICH 3&4 is the same as the power on CPICH 1&2 and therefore the CPICH power is equally distributed with ration 50:50.

If g = 0.25 a power ratio between CPICH 1&2 and CPICH 3&4 of 80:20 is obtained.

The case of unequal power allocation is also covered in the simulations (section 7.1.1).

7 Performance

7.1 Link level simulation assumptions

7.1.1. Simulation assumptions

Table 1 and Table 2 list the simulation parameters that should be used in the Tx diversity simulations for DPCH for R99 and HS-channels respectively. It is recommended that each proponent have the same receiver related to each algorithm of channel estimation and SIR estimation so as to make fair comparisons without disputes. Table 3 represents one of recommendation of receiver algorithms. Also corresponding antenna verification algorithm to be used for evaluations should be described for each closed loop transmit diversity scheme. If antenna verification algorithm is not described for a given scheme, the scheme will be evaluated without antenna verification algorithm.

Table 1. Recommended simulation parameters for Tx diversity simulations (DPCH for R99).

Bit Rate	12.2 kbps
Chip Rate	3.84 Mcps
Convolutional code rate	1/3
Carrier frequency	2 GHz
PC rate	1500 Hz
PC delay	1 slot*
PC error rate	4 %
PC Step Size	1 dB per antenna
Channel model(s) and UE velocities	As in SCM link level channel model
CL TxD feedback rate	1 bit / slot
CL TxD feedback delay	1 slot
CL TxD feedback error rate	4 %
TTI	20 ms
Downlink DPCH slot format	#10 or #11
Number of RAKE fingers	Equal to number of taps in the channel model
Target FER/BlkER	1 %

Geometry (G)	-3, 0 and 6 dB
CPICH Power	-10 dB total
Performance measure	$T_x E_c/I_{or}$

* Power control cmd will be applied after 1 slot as in Annex B.1 of 25.214 (V.5.4.0).

Table 2. Recommended simulation parameters for Tx diversity simulations (HS-channels for HSDPA)

Bit Rate	As defined in HS-channels
Chip Rate	3.84 Mcps
MCS	1/4, 1/2, 3/4 with QPSK 1/2, 3/4 with 16QAM
Max no. of iterations for Turbo Coder	8
Metric for Turbo Coder	Max Log Map
Turbo Interleaver	Random
Carrier frequency	2 GHz
Power control	OFF
Channel model(s) and UE velocities	As in SCM link level channel model
CL TxD feedback rate	1 bit/ slot
CL TxD feedback delay	1 slot
CL TxD feedback bit error rate	4 %
TTI	2 ms
Number of RAKE fingers	Equal to number of taps in the channel model
Target FER/BlkER	1 %
Geometry (G)	-3, 0 and 6 dB
CPICH Power	-10 dB total
Performance measure	$T_x E_c/I_{or}$
CQI feedback delay	3 slots , 6 slots
CQI feedback error	0 %

Table 3. Recommended Receiver Setup

Channel Estimation	**From Dedicated Pilots for Antenna Verification**	Simple Average over dedicated pilots of each slot
	From CPICH for demodulation of DPCH	Simple Average over common pilot in each slot
SIR Estimation	**From Dedicated Pilots for Power Control**	Summed estimates from each finger
	From both Dedicated and common pilots for Antenna Verification	1-order filtering for each finger [33]

7.1.2. Link Level Channel Model

In this section, link level channel model parameters are described in Table 4 considering one antenna element at UE.

Table 4. Spatial channel model parameters for multiple transmit antennas

Model		Case I		Case II		Case III		Case IV	
PDP		Modified Pedestrian A		Vehicular A		Pedestrian B		Single Path	
# of Paths		1) 4+1 (LOS on, K = 6dB) 2) 4 (LOS off)		6		6		1	
Relative Path Power (dB)	Delay (ns)	1) 0.0 2) -Inf	0	0,0	0	0.0	0	0	0
		1) -6.51 2) 0.0	0	-1.0	310	-0.9	200		
		1) -16.21 2) -9.7	110	-9.0	710	-4.9	800		
		1) -25.71 2) –19.2	190	-10.0	1090	-8.0	1200		
		1) -29.31 2) -22.8	410	-15.0	1730	-7.8	2300		
				-20.0	2510	-23.9	3700		
Speed (km/h)		1) 3 2) 30, 120		3, 30, 120		3, 30, 120		3	
UE	Topology	Reference 0.5λ		Reference 0.5λ		Reference 0.5λ		N/A	
UE	PAS	1) LOS on: Fixed AoA for LOS component, remaining power has 360 degree uniform PAS. 2) LOS off: PAS with a Lapacian distribution, RMS angle spread of 35 degrees per path		RMS angle spread of 35 degrees per path with a Lapacian distribution Or 360 degree uniform PAS.		RMS angle spread of 35 degrees per path with a Lapacian distribution		N/A	
UE	DoT (degrees)	0		22.5		-22.5		N/A	
UE	AoA (degrees)	22.5 (LOS component) 67.5 (all other paths)		67.5 (all paths)		22.5 (odd numbered paths), -67.5 (even numbered paths)		N/A	
Node B	Topology	Reference: ULA with 0.5λ-spacing or 4λ-spacing or 10λ-spacing						N/A	

Model		Case I	Case II	Case III	Case IV
	PAS	Lapacian distribution with RMS angle spread of 2 degrees or 5 degrees, per path depending on AoA/AoD			N/A
	AoD/AoA (degrees)	50^o for 2^o RMS angle spread per path 20^o for 5^o RMS angle spread per path			N/A

NOTE: The Angle of Arrival (AoA) and Power azimuth spectrum (PAS) may not be applicable to the UE because it only has 1 antenna.

8 Impacts to UE and UTRAN implementation

9 Impacts to physical layer operation

10 Backwards compatibility to R99/R4

10.1 Eigenbeamformer concept

With the eigenbeamformer no backward compatibility problem is identified for Release-99.

10.2 Basis selection scheme for > 2 Tx antennas

10.3 New CPICH Transmission scheme for > 2 Tx antennas

This proposed scheme satisfies the backward compatibility with Release 99 2-ant diversity UE. The proper common/dedicated physical channel transmission scheme is also considered to be used with the proposed CPICH transmission scheme. With only one additional channelisation code, the proposed CPICH transmission scheme can be used as diversity pilot for 4 transmit antenna.

10.4 4-Tx-STTD concept

Pilot backward (rel'99) compatibility problem (meaning that decreasing of P-CPiCH transmit power for rel'99 UE's) can fully be avoided e.g. by utilizing the dedicated pilots for the two additional channels. Indeed, decoding of 4-Tx-STTD is requiring only two channel estimates for rel'99 STTD-decoder.

10.5 Closed loop mode 1 for > 2 Tx antennas

As Rel.-99 closed loop mode 1 is one instance (i.e. R2F2T2) of R2FNTM solutions there are no special backwards compatibility problems.

11 Evaluations of the decribed schemes

[tbd]

History

Document history		
Date	**Version**	**Comment**
	0.1.0	First draft
Nov. 22, 2001	0.1.1	Revised version
Feb. 21, 2002	0.2.1	Merged version
Feb. 22, 2002	1.0.0	Updated version to informative present in the plenary
Nov. 11, 2002	1.0.1	Two contributions are included
Mar. 26, 2003	1.0.2	Scope and a contribution are included
April 3, 2003	1.1.0	The clean version of 1.0.2 (e-mail)
August 29, 2003	1.1.1	Based on #32's decesion, two contributions are added.
September 27, 2003	1.2.0	WG1 approved version of V1.1.1
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