

(19) World Intellectual Property Organization
International Bureau



(43) International Publication Date
21 September 2006 (21.09.2006)

PCT

(10) International Publication Number
WO 2006/097117 A1

(51) International Patent Classification:
H04L 1/06 (2006.01) **H04L 27/34** (2006.01)

(21) International Application Number:
PCT/EP2005/002690

(22) International Filing Date: 14 March 2005 (14.03.2005)

(25) Filing Language: English

(26) Publication Language: English

(71) Applicant (for all designated States except US): **NTT DoCoMo, Inc** [JP/JP]; 11-1 Nagata-cho 2-Chome, Chiyoda-ku, Tokyo (JP).

(72) Inventors; and

(75) Inventors/Applicants (for US only): **BAUCH, Gerhard** [DE/DE]; Barer Strasse 58, 80799 Munich (DE). **SETHURAMAN, Prasanna** [DE/DE]; Hedwig Darnsfeld Allee 7, 80637 Munich (DE).

(74) Agents: **ZINKLER, Franz** et al.; P. O. Box 246, 82043 Pullach/Munich (DE).

(81) Designated States (unless otherwise indicated, for every kind of national protection available): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NA, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SM, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

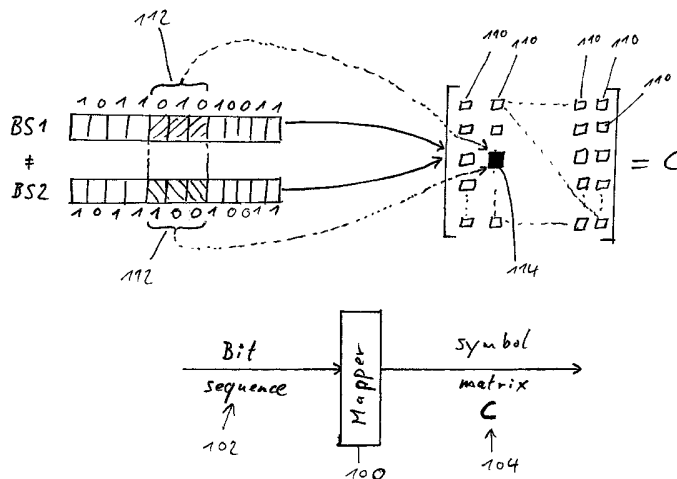
(84) Designated States (unless otherwise indicated, for every kind of regional protection available): ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IS, IT, LT, LU, MC, NL, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:

— with international search report

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: APPARATUS AND METHOD FOR PROVIDING A TRANSMIT MATRIX FOR A TRANSMITTER



(57) Abstract: An Apparatus for providing a transmit matrix (Xk) for a transmitter (400) having a plurality of transmitting points (422) comprises a mapper (100) for mapping bit sequences (BS1, BS2) to symbol matrices (C), the mapper (100) being configured to have a mapping rule in which a first bit sequence (BS1) is mapped to a symbol matrix (C) and a second bit sequence (BS2) is mapped to the same symbol matrix (C), the first and second bit sequences (BS1, BS2) being different from each other, wherein the symbol matrix (C) has a number of rows and a number of columns, the number of rows and the number of columns being larger than one. Furthermore, the apparatus comprises a processor (414, 418, 420) for processing the symbol matrix to obtain the transmit matrix (Xk), the transmit matrix (Xk) having a number of rows being equal to the number of transmitting points (422) and having a number of columns being equal to the number of columns or rows of the symbol matrix (C).

**Apparatus and Method for Providing a Transmit Matrix for a
Transmitter**

Description

5

The present invention is in the field of telecommunication and, in particular, in the field of wireless communication using a plurality of transmit antennas.

10 The usage of multiple antennas at both ends of the wireless link has been shown to provide significant capacity gains in fading environments as it is disclosed for example in G.J. Foschini and M.J. Gans, On limits of wireless communications in a fading environment when using multiple
15 antennas, *Wireless Personal Communications*, 6:311-335, 1998 or E. Telatar, Capacity of multi-antenna Gaussian channels, *European Transactions on Telecommunications (ETT)*, 10(6), November/December 1999. A variety of multiple-input multiple-output (MIMO) schemes including space-time codes
20 and spatial multiplexing have been proposed in order to exploit those capacities. However, most MIMO techniques require knowledge of the channel coefficients at the receiver. This can be obtained from a channel estimation. However, channel estimation is a more severe problem in
25 MIMO systems than in single antenna systems since more subchannels have to be estimated and the energy of pilot symbols has to be distributed over several antennas. Therefore, differential MIMO transmission schemes as proposed in

30

V. Tarokh and H. Jafarkhani, A differential detection scheme for transmit diversity, *IEEE Journal on Selected Areas in Communications*, 18(7):1169-1174, July 2000,

35 B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000, and

B. Hochwald and W. Swelden, Differential unitary space-time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000],

5 which require no channel estimation at the receiver are attractive alternatives. In general, those schemes can be viewed as differential space-time modulation methods or differential matrix modulation methods, respectively, where the constellation consists of unitary matrices. More
10 bandwidth-efficient extensions of differential matrix modulation have been proposed in

X.-G. Xia, Differentially en/decoded orthogonal space-time block codes with APSK signals, *IEEE Communications Letters*,
15 6(4):150-152, April 2002,

G. Bauch, A bandwidth-efficient scheme for non-coherent transmit diversity, in *IEEE Globecom*, December 2003,

20 G. Bauch, Differential space-time-frequency transmit diversity in OFDM, in *International Symposium on Wireless Personal Multimedia Communications (WPMC)*, October 2003,

G. Bauch, Turbo detection of differential multiple-length transmit diversity, in *3rd International Symposium on Turbo Codes and Related Topics*, pages 327-330, September 2003,
25

G. Bauch. Differential amplitude and unitary space-time modulation. In *International ITG Conference on Source and Channel Coding*, pages 135-142, January 2004,
30

G. Bauch, Higher order differential matrix modulation, in *IEEE International Symposium on Information Theory (ISIT)*, June/July 2004,
35

G. Bauch, Differential multiple-length transmit diversity, *IEEE Communications Letters*, 8(3):141-143, March 2004,

M. Tao and R.S. Cheng, Differential space-time block codes, in *IEEE Globecom Conference*, pages 1098-1102, November 2001,

5 Z. Chen, G. Zhu, J. Shen, and Y. Liu, Differential space-time block codes from amicable orthogonal designs, in *Wireless Communications and Networking Conference (WCNC)*, pages 768-772, March 2003,

10 C.-S. Hwang, S.H. Nam, J. Chung, and V. Tarokh, Differential space-time block codes using QAM constellations, in *International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC)*, September 2003,

15 C.-S. Hwang, S.H. Nam, J. Chung, and V. Tarokh, Differential space-time block codes using constant modulus constellations, *IEEE Transactions on Signal Processing*, 51(11):2955-2964, November 2003.

20 Another key technology is the application of the turbo principle for iterative detection. Turbo iterations can be performed between the differential detector and an outer FEC decoder. This requires a soft-output differential
25 detector as derived in G. Bauch, Turbo detection of differential multiple-length transmit diversity, in 3rd *International Symposium on Turbo Codes and Related Topics*, pages 327-330, September 2003 or in G. Bauch, Differential amplitude and unitary space-time modulation, in
30 *International ITG Conference on Source and Channel Coding*, pages 135-142, January 2004. The performance of this iterative detector depends on the mapping of bits to matrices. This method allows to improve the performance and to decrease the complexity of the detector.

35

Differential transmit diversity from orthogonal designs has been proposed for two transmit antennas in V. Tarokh and

H. Jafarkhani, A differential detection scheme for transmit diversity, *IEEE Journal on Selected Areas in Communications*, 18(7):1169-1174, July 2000 and generalized to more antennas in H. Jafarkhani and V. Tarokh, Multiple
5 transmit antenna differential detection from generalized orthogonal designs, *IEEE Transactions on Information Theory*, 47(6):2626-2631, September 2001. Unitary differential space-time modulations has been proposed in B. Hochwald and W. Swelden, Differential unitary space-
10 time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000 and B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000. A more bandwidth-efficient version of unitary space-time
15 modulation was presented in A. Steiner, M. Peleg, and S. Shamai, Iterative decoding of space-time differentially coded unitary matrix modulation, *IEEE Transactions on Signal Processing*, 50(10):2385-2395, October 2002. More bandwidth-efficient extensions of differential matrix
20 modulation have been proposed in

X.-G. Xia, Differentially en/decoded orthogonal space-time block codes with APSK signals, *IEEE Communications Letters*, 6(4):150-152, April 2002,

25

G. Bauch, A bandwidth-efficient scheme for non-coherent transmit diversity, in *IEEE Globecom*, December 2003,

G. Bauch, Differential space-time-frequency transmit
30 diversity in OFDM, in *International Symposium on Wireless Personal Multimedia Communications (WPMC)*, October 2003,

G. Bauch, Turbo detection of differential multiple-length transmit diversity, in *3rd International Symposium on Turbo*

Codes and Related Topics, pages 327-330, September 2003,

G. Bauch, Differential amplitude and unitary space-time modulation, in *International ITG Ccnference on Source and*
5 *Channel Coding*, pages 135-142, January 2004,

G. Bauch, Higher order differential matrix modulation, in *IEEE International Symposium on Information Theory (ISIT)*, June/July 2004,

10

G. Bauch, Differential multiple-length transmit diversity, *IEEE Communications Letters*, 8(3):141-143, March 2004,

M. Tao and R.S. Cheng, Differential space-time block codes,
15 in *IEEE Globecom Conference*, pages 1098-1102, November 2001,

Z. Chen, G. Zhu, J. Shen, and Y. Liu, Differential space-time block codes from amicable orthogonal designs, in
20 *Wireless Communications and Networking Conference (WCNC)*, pages 768-772, March 2003,

C.-S. Hwang, S.H. Nam, J. Chung, and V. Tarokh, Differential space-time block codes using QAM
25 constellations, in *International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC)*, September 2003,

C.-S. Hwang, S.H. Nam, J. Chung, and V. Tarokh,
30 Differential space-time block codes using non-constant modulus constellations, *IEEE Transactions on Signal*

Processing, 51(11):2955-2964, November 2003.

The application of the turbo principle to differential matrix modulation has been proposed in G. Bauch, Turbo
 5 detection of differential multiple-length transmit diversity, in *3rd International Symposium on Turbo Codes and Related Topics*, pages 327-330, September 2003.

A flat fading multiple-input multiple-output (MIMO) channel with n_T transmit and n_R receive antennas is considered. The
 10 channel coefficients are collected in the matrix

$$\mathbf{H}_k = \begin{bmatrix} h_k^{(11)} & \dots & h_k^{(n_T 1)} \\ \vdots & \ddots & \vdots \\ h_k^{(1 n_R)} & \dots & h_k^{(n_T n_R)} \end{bmatrix} \quad (1)$$

where $h_k^{(ij)}$ is the channel coefficient from transmit
 15 antenna i to receive antenna j at time k .

In frequency-selective environments, the flat fading channel can e.g. be accomplished by OFDM with a sufficiently large guard interval. In CDMA, the signals in each finger of a Rake
 20 receiver face a flat channel.

At the receiver, we observe

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k \quad (2)$$

where

25

$$\mathbf{X}_k = \begin{bmatrix} x_{k,1}^{(1)} & \dots & x_{k,L}^{(1)} \\ \vdots & \ddots & \vdots \\ x_{k,1}^{(n_T)} & \dots & x_{k,L}^{(n_T)} \end{bmatrix} \quad (3)$$

and

30

$$\mathbf{Y}_k = \begin{bmatrix} y_{k,1}^{(1)} & \dots & y_{k,L}^{(1)} \\ \vdots & \ddots & \vdots \\ y_{k,1}^{(n_R)} & \dots & y_{k,L}^{(n_R)} \end{bmatrix} \quad (4)$$

contain the transmitted and received symbols, respectively,
and

$$\mathbf{N}_k = \begin{bmatrix} n_{k,1}^{(1)} & \dots & n_{k,L}^{(1)} \\ \vdots & \ddots & \vdots \\ n_{k,1}^{(n_R)} & \dots & n_{k,L}^{(n_R)} \end{bmatrix} \quad (5)$$

are the noise samples, which are assumed to be independent
and Gaussian with variance

$$\sigma^2 = \frac{N_0}{2} \quad (6)$$

per real dimension.

However, it has to be mentioned that the assumption of a
flat fading multiple-input multiple-output channel is not
always fulfilled, especially not in situations in which a
high Doppler spread occurs. In such situation with a high
Doppler spread, unknown phase offsets can occur which lower
the detectability of a transmitted symbol or, in the case
of multiple-input multiple-output systems, the transmitted
symbol matrix. Especially in MIMO environments the effects
of such a Doppler spread have a severe influence as the
time interval for transmitting the information included in
a symbol matrix (or a transmit matrix respectively) is
longer than the time interval for transmitting a just one
symbol in a single antenna scenario. Thus, the effects of
phase offsets on the transmitted code or a symbol matrix
including the information to be transmitted are worse than
the effects of phase offsets influencing a transmitted
symbol in a single antenna transmission system.

Furthermore, the idea of non-unique mappings with
ambiguities, which can be resolved in an iterative process
in the detector, has been proposed for PSK modulation in
non-differential single antenna systems in T. Clevorn and

P. Vary, Iterative decoding of BICM with non-regular signal constellation sets, in *International ITC Conference on Source and Channel Coding*, pages 259-266, January 2004, or in T. Clevorn, S. Godtmann, and P. Vary, EXIT chart analysis of non-regular signal constellation sets for BICM-ID, in *International Symposium on Information Theory and its Applications (ISITA)*, pages 21-26, October 2004.

However, the presented idea of non-unique mappings is limited to the non-differential single antenna system and using a non-unique mapping on PSK constellation points. This limitation on non-differential single antenna systems using PSK modulation only results in a limited improvement of performance.

It is the object of the present invention to provide a possibility to further improve the performance of a transmission system, using a plurality of transmitting points or using a plurality of receiving points. Furthermore, it is the object of the present invention to decrease the complexity of a detector for use in the above-mentioned transmission system having the plurality of transmitting points or the plurality of receiving points.

This object is achieved by an apparatus for providing transmit matrix according to claim 1, a transmitter apparatus according to claim 22, a receiver apparatus according to claim 23, a method for providing a transmit matrix according to claim 24, a transmission method according to claim 25 and a reception method according to claim 26.

The present invention provides an apparatus for providing a transmit matrix for a transmitter having a plurality of transmitting points, the apparatus comprising:

a mapper for mapping bit sequences to symbol matrices, the mapper being configured to have a mapping rule in which a first bit sequence is mapped to a symbol matrix and a second bit sequence is mapped to the same symbol matrix, the first and second bit sequences being different from each other, wherein the symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

10 a processor for processing the symbol matrix to obtain the transmit matrix, the transmit matrix having a number of rows being equal to the number of transmitting points and having a number of columns being equal to the number of columns or rows of the symbol matrix.

15

Furthermore, the present invention provides a transmitter apparatus comprising:

the apparatus for providing a transmit matrix, wherein the
20 transmit matrix defines transmit sequences to be transmitted; and

n_T transmitting points for transmitting information included in the transmit matrix, wherein the n_T
25 transmitting points are configured for simultaneously transmitting n_T values of a transmit sequence of a time instant within a time interval and for simultaneously transmitting n_T values of a further transmit sequence and a further time interval within the time interval, wherein n_T
30 denotes a number of transmitting points being greater than one.

Additionally, the present invention provides a receiver apparatus comprising:

means for receiving information via n_R receiving points in order to obtain a receiver matrix, the receiver matrix having a number of rows being equal to the number of receiving points and having a number of columns being equal to the number of columns or rows of a receiver symbol matrix, wherein n_R denotes a number of transmitting points being greater than one;

a processor for processing the receiver matrix received from the means for receiving information to obtain the receiver symbol matrix, wherein the receiver symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

a demapper for demapping bit sequences from the receiver symbol matrix according to a demapping rule in which a first bit sequence is demapped from a receiver symbol matrix and a second bit sequence is demapped from the same receiver symbol matrix, the first and second bit sequences being different from each other.

Furthermore, the present invention provides a method for providing a transmit matrix for a transmitter, the transmitter having a plurality of transmitting points, the method comprising:

mapping bit sequences to symbol matrices using a mapping rule according to which a first bit sequence is mapped to a symbol matrix and a second bit sequence is mapped to the same symbol matrix, the first and second bit sequences

being different from each other, wherein the symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

- 5 processing the symbol matrix using a processor to obtain the transmit matrix, the transmit matrix having a number of rows being equal to the number of transmitting points and having a number of columns being equal to the number of columns or rows of the symbol matrix.

10

Furthermore, the present invention provides a transmission method comprising the steps of:

- providing a transmit matrix, wherein the transmit matrix
15 defines transmit sequences to be transmitted; and

- transmitting information included in the transmit matrix via n_T transmitting points, wherein n_T values of a transmit sequence are simultaneously transmitted at the time instant
20 within a time interval and n_T values of a further transmit sequence are simultaneously transmitted at a further time interval within the time interval, wherein n_T denotes a number of transmitting points being greater than one.

- 25 Furthermore, the present invention provides a reception method comprising the steps of:

- receiving information via n_R receiving points in order to obtain a receiver matrix, the receiver matrix having a
30 number of rows being equal to the number of receiving points and having a number of columns being equal to the number of columns or rows of a receiver symbol matrix,

wherein n_R denotes a number of transmitting points being greater than one;

5 processing the receiver matrix received from the means for receiving information to obtain the receiver symbol matrix, wherein the receiver symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

10 demapping bit sequences from the receiver symbol matrix using a demapping rule according to which a first bit sequence is demapped from a receiver symbol matrix and a second bit sequence is demapped from the same receiver symbol matrix, the first and second bit sequences being
15 different from each other.

The present invention is based on the finding that an improvement of the performance of a transmission system having a plurality of transmitting points or a plurality of
20 receiving points can be achieved by a non-unique mapping in which bit sequences being different from each other are mapped to one and the same symbol matrix. This non-unique mapping provides the possibility to use a number of different symbol matrices, the number of used symbol
25 matrices being less than a number of possible bit sequences. For example, if a bit sequence has b bits, the number of possible bit sequences is 2^b . Thus, according to the invention, the number of symbol matrices used is less than 2^b due to the fact that at least two different bit
30 sequences are mapped to the same symbol matrix. The processor is can be configured to provide a change in the dimension of the symbol matrix or to provide further operations on the symbol matrix in order to obtain the transmit matrix.

35

This non-unique mapping scheme provides the advantage that an improved performance can be realized, compared to existing unique mapping schemes. This improved performance especially results from the fact that due to the smaller
5 number of possible symbol matrices, the distances between the used symbol matrices can be "artificially" enlarged by the adequate choice of a mapping scheme of bit sequences to symbol matrices and thus the detectability of a symbol matrix can be improved. This is especially important, when
10 a high phase offset, for example in an environment with a high Doppler spread, is expected. As the non-unique mapping can be compensated in a respective receiver using the turbo principle, it is now possible to improve the transmission quality via a channel having a plurality of transmit and/or
15 receiving antennas as a more secure demapping of received symbol matrices can be performed.

Furthermore, it is now possible to design the symbol matrices and the respective mapping of bit sequences to
20 these matrices individually and arbitrarily such that, for example, a maximum distance between each of the symbol matrices to be used for a transmission can be realized. Especially when an information about the channel or eigenbeams of a MIMO-channel are known, a distinct design
25 of the symbol matrices can be performed and therefore, a maximum distance of the individual symbol matrices can be achieved, resulting in an improved detectability of received symbol matrices. Thus, the inventive approach provides a further degree of freedom with respect to
30 conventional approaches. Nevertheless, it is also possible to design the symbol matrices such that they are unitary in order to allow non-coherent detection, for example.

In a preferred embodiment of the present invention, the bit
35 sequences assigned to the same symbol matrix differ in at least two bits. This improves the distinction of the individual bit sequences in the receiver, which are assigned to the same symbol matrix. However, the at least

two bits being different from each other in the bit sequences do not necessarily have to be adjacent bits in the bit sequence. This provides the possibility of assigning several bit sequences to the same symbol matrix
5 wherein a detection of the individual bit sequences is still possible. Especially in the case, in which a distinct signal constellation has superior detection characteristics over other signal constellations, a mapping of more than two bit sequences to such a signal constellation could
10 provide a significant advantage over other mapping schemes.

Furthermore, it is also possible to design the symbol matrices such that symbol matrix entries of a symbol matrix correspond to signal constellation points. On each of the
15 signal constellation points it is now possible to map a group of bits from the bit sequence to be transmitted such that the matrix consists of a number of symbol matrix entries representing signal constellation points. The non-unique mapping from bit sequences to the same symbol matrix
20 can then be realized in a splitting-off of the bit sequence to be transmitted into different groups of bits, each group of bits being mapped to one single matrix entry of the symbol matrix. The non-unique mapping can then be realized by a non-unique mapping of a group of bits to one symbol
25 matrix entry in that a group of bits of a first bit sequence, corresponding to one distinct symbol matrix entry and a group of bits of a second bit sequence are mapped to the same symbol matrix entry, wherein the groups of the first and second bit sequences are different from each
30 other and wherein the value which is assigned to the symbol matrix entry is the same for both groups of bits in the first and second bit sequences. Such a non-unique mapping can then for example be realized by a double (or more than double) assignment of bits to one constellation point of a
35 constellation diagram. It has to be mentioned that the constellation diagrams can be chosen from the diagram types of QPSK, PAM, PSK, QAM, M-PSK or M-QAM, wherein M denotes an integer value being greater than one. However, the

constellation diagram is not limited to the previously mentioned constellation diagrams. Expressed in other words, two bit sequences, which only differ in a group of bits, the group of bits corresponding to one symbol matrix entry, are mapped on the identical symbol matrix and therefore, a non-unique mapping of bit sequences to symbol matrices can be realized.

The advantage of the choice of such a non-unique mapping using non-unique constellation diagrams for one or more symbol matrix entries provide can be seen in that the construction of orthogonal or unitary symbol matrices can be simplified, thus allowing a low-complexity receiver.

Furthermore, it is possible to provide a transmit matrix comprising the information to be transmitted in a k -th time interval from the symbol matrix in that the symbol matrix is multiplied by a transmit matrix including the information to be transmitted that the $(k-1)$ -th time interval (that is the previous time interval preceding the k -th time interval). Using such a transmit matrix then provides the advantage of applying a differential matrix modulation, in which a better and more bandwidth effective transmission, or detection, can be realized. This improved transmission characteristics, based on a differential matrix modulation, increase the effect of non-unique mapping of bit sequences to symbol matrices as the differential matrix modulation also improves the transmission characteristics due to an optimization of the distances of two consecutive transmit matrices. However, especially the combination of the non-unique mapping from bit sequences to symbol matrices with the differential matrix modulation shows superior performance compared to just the differential matrix modulation or just the non-unique mapping using PSK constellation in a single antenna system when being used in time-varying or frequency-selective channels where the channel is not constant during transmission of two successive matrices and therefore in

such time-varying or frequency-selective channels a condition for non-coherent detection is violated.

Furthermore, it is also possible to multiply the transmit
5 matrix comprising the information to be transmitted within
the $(k-1)$ -th time interval with a normalization factor in
order to reduce the peak-to-average-power-ratio for a
transmission of the transmit matrix including the
information to be transmitted within the k -th time
10 interval. Thus it is possible to relax the requirements for
linearity of amplifiers. Thus, the back-off of amplifiers
can advantageously be reduced which results in a lower
battery consumption and costs.

15 The consideration of the normalization factor can be
realized by a multiplication of the symbol matrix with the
normalization factor or by a multiplication of the product
of the symbol matrix and the transmit matrix including the
information to be transmitted within the $(k-1)$ -th time
20 interval with the normalization factor. The normalization
factor can be dependent on an amplitude exponent of (a
signal being dependent from) the transmit matrix including
the information to be transmitted within the $(k-1)$ -th time
interval in order to adapt the amplitude exponent of the
25 transmit matrix including the information to be transmitted
within the k -th time interval. Furthermore, it is also
possible to map a group of bits to the normalization factor
and thus provide an additional possibility for transmitting
information. For example, a codeword can be split up in a
30 first and a second group of bits and then can be transmitted
via a mapping of the first group of bits to a symbol matrix
and a mapping of the second group of bits from the same
code word to the normalization factor. This mapping of
additional bits to the normalization factor can also be
35 done in a non-unique way, thus reducing the number of
normalization factors to be used. This, in consequence,
again results in a reduced detection complexity, when the
receiver is configured to correctly demap the non-unique

normalization factor, for example by applying the turbo principle. It has to be mentioned that the non-unique mapping of bit sequences to symbol matrices can be carried out independently from the non-unique mapping of bits to a normalization factor. This results in a possible mapping
5 scheme in which a non-mapping of bit sequences to symbol matrices is carried out when a unique or a non-unique mapping of bits to the normalization factor is realized or, on the other side, a non-unique mapping of bits to the
10 normalization factor can be combined with a unique or non-unique mapping of bit sequences to symbol matrices. This provides a further degree of freedom in designing the transmission system.

15 As a further advantage it has to be mentioned that, due to the reduced number of possible symbol matrices and/or the reduced number of possible normalization factors, the peak-to-average-power-ratio in the transmitter as well as in the receiver can be reduced, thus relaxing the requirements for
20 linearity of amplifiers. The back-off of amplifiers can be therefore reduced which results in a lower battery consumption and loss.

Further embodiments of the present invention are described
25 with respect to the following Figures, in which:

Fig. 1 shows a block diagram of an apparatus for providing symbol matrices in accordance with an embodiment of the present invention;

30

Fig. 2 shows a block diagram for a non-unique mapping of a group of bits to the same constellation point;

Fig. 3A show tables for a mapping groups of bits
35 and 3B to matrices of the symbol matrix according to an embodiment of the present invention;

Fig. 4 shows a transmitter in accordance with an embodiment of the present invention;

5 Fig. 5 shows a diagram including information for mapping bits to a normalization factor;

Fig. 6 shows a block diagram of a non-coherent soft-output decoder;

10 Fig. 7 shows a block diagram of a system for an iterative detection of differential matrix modulation;

15 Fig. 8A show simulation diagrams comparing a performance of to 8C embodiments of the invention with respect to conventional approaches.

Fig. 1 shows a block diagram of an embodiment of an apparatus for providing symbol matrices for a transmitter in which the apparatus comprises a mapper 100 for mapping a
20 bit sequence 102 onto a symbol matrix C 104. As already mentioned above, the mapping can be such that a first bit sequence BS1 and a second bit sequence BS2, which are not identical, are mapped on the identical symbol matrix C as shown in the upper part of Fig. 1. The symbol matrix C,
25 especially the symbol matrix entries 110 of the symbol matrix C can be arbitrarily filled values, such that for example the differences between the individual symbol matrices C from a set of symbol matrices have a maximum Euclidean distance from each other in a signal
30 constellation space. Especially in the case, where channel information, for example in the form of a predominant direction of eigenbeams, is already known, this provides the possibility to specifically design the symbol matrices in order to obtain an optimum detectability of a
35 transmitted symbol matrix. This, in consequence, provides the advantage that a designer of a transmission system has a very high degree of freedom in designing the individual

symbol matrices C in order to obtain the maximum possible transmission capacity.

In another approach, the symbol matrix entries 110 of the
5 symbol matrix C can be considered to be constellation
points (or constellation spaces) to which a group of bits
from the bit sequences BS1 or BS2 are mapped to. In the
upper part of Fig. 1 this is shown by the mapping of the
group of three bits 112 which were both mapped to the same
10 symbol matrix entry 114 in the symbol matrix C. A non-
unique mapping can then be carried out in that the groups
112 of the bit sequences BS1 and BS2, which are different
from each other are mapped such that the symbol matrix
entry 114 is assigned the identical value.

15 Such a mapping is shown in more detail in Fig. 2. Herein a
mapper 200 is configured to map, for example, a group of
three bits onto a 4-QAM signal constellation point. Due to
the fact that the group of three bits would result in 8 (= 2^3)
20 different bit sequences, the non-unique mapping can be
carried out, such that for example the bit sequence 010 and
100 are mapped onto the same signal constellation point
202, which has for example a complex value of $1+i$.
Therefore, both bit sequences 010 and 100 are mapped on the
25 signal constellation point $1+i$. It has to be mentioned that
the assignment of bit sequences to the individual signal
points, such as the signal point 202 can be done
arbitrarily, but, for an optimum detection, the bit
sequences, assigned to the identical signal constellation
30 point 202 should differ in (at least) 2 bits.

Referring back to Fig. 1, it has to be mentioned that, if
the bit sequences BS1 and BS2 differ only in the bit group
112, the resulting symbol matrix C will be identical such
35 that a non-unique mapping of the bit sequences BS1 and BS2
to the symbol matrix C can be carried out by the mapper 100
on a matrix scale.

Differential unitary space-time modulation was introduced simultaneously in Differential Unitary Space-Time Modulation according to B. Hochwald and W. Swelden, Differential unitary space-time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000 and B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000. A group of $b_1 = \log_2(M_1)$ data bits $\mathbf{u}_k^{(1)} = u_{k,1}, \dots, u_{k,b_1}, u_{k,t} \in \{+1, -1\}$, is mapped on a $L \times L$ info matrix $\mathbf{C}_k$. The $n_T \times L$ transmit matrix $\mathbf{X}_k$ is determined by $\mathbf{C}_k$ and the previously transmitted matrix $\mathbf{C}_{k-1} = \mathbf{C}(\mathbf{u}_k^{(1)})$ according to the differential encoding rule

$$\mathbf{X}_k = \mathbf{X}_{k-1} \mathbf{C}_k \quad (7)$$

In order to allow non-coherent detection, $\mathbf{C}_k$ must be unitary, i.e.

$$\mathbf{C}_k \mathbf{C}_k^H = \mathbf{I}_L \quad (8)$$

where $\mathbf{C}^H$ denotes the conjugate transpose of $\mathbf{C}$ and $\mathbf{I}_L$ is the $L \times L$ unity matrix. A unitary reference matrix $\mathbf{X}_0$ has to be transmitted first. All transmit matrices $\mathbf{X}_k$ are unitary. Plugging (7) in (2) yields

$$\mathbf{Y}_k = \mathbf{Y}_{k-1} \mathbf{C}_k + \mathbf{N}_{k-1} \mathbf{C}_k + \mathbf{N}_k \quad (9)$$

This describes the transmission of the info matrix $\mathbf{C}_k$ over an equivalent channel with L transmit and n_R receive antennas, channel coefficients $\tilde{\mathbf{H}} = \mathbf{Y}_{k-1}$ and - since $\mathbf{C}_k$ is unitary - a noise variance of $\tilde{\sigma}^2 = 2\sigma^2$ per real dimension at each receive antenna.

Throughout the application the following proposals for the choice of the space-time constellation $\mathbf{C}$ are considered:

a) Differential Unitary Space-Time Modulation according to B. Hochwald and W. Swelden, Differential unitary space-time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000.

5

The constellations proposed in Differential Unitary Space-Time Modulation according to B. Hochwald and W. Swelden, Differential unitary space-time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000, consist of diagonal matrices

$$\mathbf{c}(m) = \begin{bmatrix} e^{j2\pi d_1 m/M_1} & 0 & 0 \\ 0 & \ddots & 0 \\ 0 & 0 & e^{j2\pi d_{n_T} m/M_1} \end{bmatrix}, m = 0, \dots, M_1 - 1 \quad (10)$$

The diagonal entries are M_1 -PSK constellation elements which are determined by the values of d_1 . For example reference is made to B. Hochwald and W. Swelden, Differential unitary space-time modulation, *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000. Given the reference matrix $\mathbf{X}_0 = \mathbf{I}_{n_T}$, only one antenna is active at a time.

b) Differential Space-Time Modulation according to B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000.

The constellations proposed in B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000 are determined by the group consisting of all distinct products of powers of certain matrices $G_1, \dots, G_m$. For examples reference is made to B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000. Due to the group property, the transmit matrices $\mathbf{X}_k$ are also elements of the group. The symbols transmitted from each antenna are again PSK constellation elements. For $n_T = 2$ transmit antennas, the reference matrix is given by

$$\mathbf{x}_0 = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix}. \quad (11)$$

Fig. 3A and 3B show tables of examples of matrices $G_1, \dots, G_m$ for $n_T = 2$ transmit antennas wherein the matrices $G_1, \dots, G_m$ denote a set of possible symbol matrices for mapping a group of four bits onto these symbol matrices. As could be seen in Fig. 3B the natural assignment of groups of bits to the matrices $G_1, \dots, G_m$ can be carried out chronologically or, in order to obtain an improved bit error rate, the assignment of groups of bits to the symbol matrices $G_1, \dots, G_m$, can also be carried out according to the values in the column named "optimum". In order to provide a non-unique mapping, the assignment of groups of four bits can be done according to values denoted in columns named "H7" of Fig. 3A or the column named "H8" of Fig. 3B. Expressed in other words, a mapper being configured for mapping according to the mapping scheme H7, as shown in Fig. 3A, is not allowed to use the symbol matrices G_2, G_4, G_{8-10} and G_{14-16} . However, the matrix G_3 is assigned to five different groups of bits (1111, 1100, 0000, 1001, 1010), the symbol matrices G_6 (1110, 0001, 0010) and G_7 (1101, 1000, 1011) are assigned to three different groups of bits wherein the symbol matrix G_5 (0100, 0111) is assigned two different groups of bits. Analogously, a mapper being configured for mapping according to the mapping scheme H8 is not allowed to use the symbol matrices $G_1, G_3, G_5, G_7, G_{11-12}$ and G_{15-16} . However, the matrix G_9 is assigned to six different groups of four bits (0011, 0000, 0101, 0110, 1001, 1100) wherein the matrix G_{10} is assigned to four different groups of four bits (0001, 0100, 1000, 1101). Therefore, the mapping according to one of the mapping schemes H7 and H8 is not unique, as at least one symbol matrix G is assigned to at least two groups of four bits, being different from each other. However, it has to be mentioned that the mapping schemes H7 and H8 are only exemplary mapping schemes, alternative non-uniform mapping schemes can also be applied.

c) Orthogonal designs as given for coherent space-time block codes in V. Tarokh, H. Jafarkhani, and A.R. Calderbank, Space-time block codes from orthogonal designs, *IEEE Transactions on Information Theory*, 45(5):1456-1467, June 1999 or O. Tirkkonen and A. Hottinen, Complex space-time block codes for four TX antennas, in *IEEE GLOBECOM*, pages 1005-1009, November/December 2000 are defined as matrices with orthogonal columns. Unitary info matrices $\mathbf{C}_k$ can be obtained from orthogonal designs with PSK symbols $c_{k,l}$, $l = 1, \dots, K$. For $n_T = 2, K = 2$ and $n_T = 4, K = 3$, orthogonal designs are given by S. Alamouti, A simple transmitter diversity technique for wireless communications, *IEEE Journal on Selected Areas of Communications, Special Issue on Signal Processing for Wireless Communications*, 16(8):1451-1458, 1998 or O. Tirkkonen and A. Hottinen, Complex space-time block codes for four TX antennas, in *IEEE GLOBECOM*, pages 1005-1009, November/December 2000:

$$\mathbf{C}_k = \begin{bmatrix} c_{k,1} & c_{k,2} \\ -c_{k,2}^* & c_{k,1}^* \end{bmatrix}, \quad \mathbf{C}_k = \begin{bmatrix} c_{k,1} & c_{k,2} & c_{k,3} & 0 \\ -c_{k,2}^* & c_{k,1}^* & 0 & c_{k,3} \\ c_{k,3}^* & 0 & -c_{k,1}^* & c_{k,2} \\ 0 & c_{k,3}^* & -c_{k,2}^* & -c_{k,1} \end{bmatrix} \quad (12)$$

The reference matrix can be chosen as an orthogonal design with arbitrary PSK symbols. The orthogonality enables the application of a low-complexity receiver. However, the transmitted symbols after differential encoding are no PSK constellation elements any more, i.e. the constellation is expanded.

d) Another, more bandwidth-efficient proposal for unitary space-time modulation is given in A. Steiner, M. Peleg, and S. Shamai, Iterative decoding of space-time differentially coded unitary matrix modulation, *IEEE Transactions on Signal Processing*, 50(10):2385-2395, October 2002. E.g. for $L = 2$ a mapping of info bits on

three PSK symbols $c_{k,1}, \dots, c_{k,3}$ is chosen and then remaining entry of $\mathbf{c}_k$ such that $\mathbf{C}_k$ becomes unitary, i.e.

$$\mathbf{C}_k = \begin{bmatrix} c_{k,1} & c_{k,2} \\ c_{k,3} & -\frac{c_{k,1}^* c_{k,3}}{c_{k,2}^*} \end{bmatrix} \quad (13)$$

As with orthogonal designs, the constellation of the transmit symbols is expanded. Furthermore, the prize paid for higher bandwidth efficiency compared to orthogonal designs is a more complex receiver.

Alternatively, the symbol matrix elements c_k, m can be other signal constellation symbols than PSK symbols, such as, for example, QPSK, QAM, PAM, M-QAM or M-PSK, M denoting an integer value being greater than 1.

e) For higher bandwidth-efficiency, a transmission of more bits per matrix is proposed, i.e. the cardinality M_1 of the unitary space-time modulation constellation is increased. However, the receiver complexity will increase exponentially with the number of bits per matrix. In G. Bauch, Differential amplitude and unitary space-time modulation, in *International ITG Ccnference on Source and Channel Coding*, pages 135-142, January 2004 or G. Bauch, Higher order differential matrix modulation, in *IEEE International Symposium on Information Theory (ISIT)*, June/July 2004, a general scheme for differential amplitude and unitary matrix modulation is proposed which can be applied as an extension of any differential unitary space-time modulation scheme. The transmit matrices are not unitary any more but satisfy

$$\mathbf{X}_k \mathbf{X}_k^H = a^{z_k} \mathbf{I}_{n_T} \quad (14)$$

where a^{z_k} can take the discrete real values $a^{z_k} \in \{1, a, a^2, \dots, a^{M_2-1}\}$. An embodiment of the differential transmitter is depicted in Figure 4.

Fig. 4 shows a block diagram of a transmitter using a differential amplitude and linearity matrix modulation. The transmitter 400 comprises a mapper 100 for mapping a group u_k of M_1 bits onto a symbol matrix C_k , the symbol matrix C including the information to be transmitted in the k -th time interval. Furthermore, the transmitter 400 includes a mapper 410 for mapping a group $u_k^{(2)}$ of M_2 on an amplification or normalization factor 412. Furthermore, the mapper 410 for mapping the bits $u_k^{(2)}$ to the normalization factor 412 is configured to use a knowledge of a transmit sequence X_{k-1} , including information to be transmitted within the $(k-1)$ -th time interval. Furthermore, the transmitter 400 comprises a first multiplier 414 for multiplying the transmit matrix X_{k-1} comprising information to be transmitted within the $(k-1)$ -th time interval with the normalization factor 412 in order to obtain a multiplication factor 416. Furthermore, the transmitter 400 comprises a second multiplier 418 for multiplying the symbol matrix C_k with the multiplication factor 416 in order to obtain the transmit matrix X_k comprising information to be transmitted within the k -th time interval. Additionally, the transmitter 400 includes a means 420 to transmit the information included in the transmitting matrix X_k using the transmission points 422.

The input bits for the transmit matrix X_k are grouped in two sets

$$\mathbf{u}_k^{(1)} = [u_{k,1}, \dots, u_{k,b_1}]^T \quad (15)$$

$$\mathbf{u}_k^{(2)} = [u_{k,b_1+1}, \dots, u_{k,b_1+b_2}]^T \quad (16)$$

and

$$b_1 = \log_2(M_1) \text{ und } b_2 = \log_2(M_2) \quad (17)$$

bits, respectively. The first b_1 bits are mapped on a unitary matrix $\mathbf{C}_k$ as described in the previous section. The last b_2 bits determine the amplitude difference of the transmit matrix $\mathbf{X}_k$ compared to the previously transmitted matrix $\mathbf{X}_{k-1}$. More precisely, the transmit matrix is obtained from the differential encoding

$$\mathbf{X}_k = \sqrt{a^{q_k}} \mathbf{X}_{k-1} \mathbf{C}_k \quad (18)$$

where a is a real constant and $q_k \in \{-M_2 + 1, -M_2 + 2, \dots, -1, 0, 1, \dots, M_2 - 1\}$. Using the square-root is convenient for the description of the receiver. Depending on the b_2 last input bits, the amplitude a^{q_k} as defined in (14) is cyclically increased compared to the previously transmitted matrix by a factor of $1, a, a^2, \dots, \text{or } a^{M_2-1}$. Fig. 5 gives an example for Gray mapping of input bits to the amplitude difference exponent q_k for $M_2 = 4$.

In general, the amplitude exponent is given by

$$q_k = z_{k-1} - M_2 \lfloor (z_{k-1} + d_k) / M_2 \rfloor \quad (19)$$

where $\lfloor \cdot \rfloor$ is the floor function. The input bits $\mathbf{u}_k^{(2)}$ are mapped on an integer $d_k \in \{0, 1, \dots, M_2 - 1\}$ and z_{k-1} denotes the amplitude exponent of the previously transmitted matrix $\mathbf{X}_{k-1}$ which is determined by

$$z_k = z_{k-1} + q_{k-1} \quad (20)$$

with the arbitrary choice $z_0 = q_0 = 0$. Fig. 5 shows an example for assigning the amplitude difference exponent q_k , for $M_2 = 4$ and Gray mapping.

In this section, the non-coherent soft-output detector is described which was derived in G. Bauch, Differential

amplitude and unitary space-time modulation, in *International ITG Conference on Source and Channel Coding*, pages 135-142, January 2004 and G. Bauch, Higher order differential matrix modulation, in *IEEE International Symposium on Information Theory (ISIT)*, June/July 2004 for differential matrix modulation. The detector which is illustrated in Figure 6 is described for differential amplitude and unitary matrix modulation according to a previous section.

10

It can be used for all other unitary matrix modulation schemes mentioned above by setting $M_2 = 1$. A simple non-coherent receiver takes into account two successively received matrices

15

$$\mathbf{Y}_{k-1} = \mathbf{H}_{k-1} \mathbf{X}_{k-1} + \mathbf{N}_{k-1} \quad (21)$$

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k = \sqrt{a^{q_k}} \mathbf{H}_k \mathbf{X}_{k-1} \mathbf{C}_k + \mathbf{N}_k \quad (22)$$

where the channel is assumed to be constant during transmission of two matrices, i.e.

$$\mathbf{H}_{k-1} = \mathbf{H}_k \quad (23)$$

Plugging $\mathbf{Y}_{k-1} = \mathbf{H}_{k-1} \mathbf{X}_{k-1} + \mathbf{N}_{k-1}$ in

25 $\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k = \sqrt{a^{q_k}} \mathbf{H}_k \mathbf{X}_{k-1} \mathbf{C}_k + \mathbf{N}_k$ yields

$$\begin{aligned} \mathbf{Y}_k &= \sqrt{a^{q_k}} \mathbf{Y}_{k-1} \mathbf{C}_k - \sqrt{a^{q_k}} \mathbf{N}_{k-1} \mathbf{C}_k + \mathbf{N}_k \\ &= \sqrt{a^{q_k}} \mathbf{Y}_{k-1} \mathbf{C}_k + \bar{\mathbf{N}}_k \end{aligned} \quad (24)$$

30 This describes the transmission of the info matrix $\mathbf{C}_k$ over an equivalent channel with L transmit and n_R receive antennas, channel coefficients $\tilde{\mathbf{H}} = \sqrt{a^{q_k}} \mathbf{Y}_{k-1}$ and - since $\mathbf{C}_k$ is unitary - additive white Gaussian noise with variance

35

$$\tilde{\sigma}^2 = \sigma^2(a^{q_k} + 1) \quad (25)$$

per real dimension at each receive antenna.

The non-coherent detection of the sets $\mathbf{u}_k^{(1)}$ and $\mathbf{u}_k^{(2)}$ can be completely separated.

5

From (24) it can be seen that the amplitude modulation takes effect only as a scaling factor of the resulting channel matrix $\sqrt{a^{q_k}} \mathbf{Y}_{k-1}$. Therefore, for a hard decision, the maximum-likelihood detector derived in B. Hochwald and
 10 W. Swelden, Differential unitary space-time modulation. *IEEE Transactions on Communications*, 48(12):2041-2052, December 2000 and B.L. Hughes, Differential space-time modulation, *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000 can be applied, i.e.

15

$$\hat{\mathbf{c}}_k = \arg \min \left\{ \left\| \mathbf{y}_k - \sqrt{a^{q_k}} \mathbf{Y}_{k-1} \mathbf{c}_k \right\|^2 \right\} \quad (26)$$

$$= \arg \max \left\{ \text{trace} \left\{ \text{Re} \left\{ \mathbf{c}_k \mathbf{Y}_k^H \mathbf{Y}_{k-1} \right\} \right\} \right\} \quad (27)$$

20 The bits $\mathbf{u}_k^{(i)}$ are obtained by demapping from $\hat{\mathbf{c}}_k$.

For a soft-output detection we use (24), (26) and (25) in order to compute the logarithmic probability

$$\begin{aligned} \log p(\mathbf{c}_k | \mathbf{y}_k, \mathbf{Y}_{k-1}) &= \text{const} + \frac{2\sqrt{a^{q_k}}}{2\tilde{\sigma}^2} \text{trace} \left\{ \text{Re} \left\{ \mathbf{c}_k \mathbf{Y}_k^H \mathbf{Y}_{k-1} \right\} \right\} + \frac{1}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \\ &= \text{const} + \frac{\sqrt{a^{q_k}}}{\sigma^2(a^{q_k} + 1)} \text{trace} \left\{ \text{Re} \left\{ \mathbf{c}_k \mathbf{Y}_k^H \mathbf{Y}_{k-1} \right\} \right\} + \frac{1}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \end{aligned} \quad (28)$$

25

where

$$\mathbf{L}_a^n(\mathbf{u}_k^{(1)}) = \left[L_a^n(\mathbf{u}_{k,1}), \dots, L_a^n(\mathbf{u}_{k,b_1}) \right]^T \quad (29)$$

is a vector containing the bit a priori
 30 information

$$L_a^n(u_{k,t}) = \frac{P(u_{k,t} = +1)}{P(u_{k,t} = -1)} \quad (30)$$

If no a priori information is available, we have $\mathbf{L}_a^n(\mathbf{u}_k^{(1)}) = 0$.
 5 With the notation

$$L(\hat{u}_{k,t}) = L(u_{k,t} | \mathbf{y}_k \mathbf{y}_{k-1}) \quad (31)$$

the APP log-likelihood ratios are given by
 10

$$\begin{aligned} L(\hat{u}_{k,t}) &= \frac{p(u_{k,t} = +1 | \mathbf{y}_{k-1}, \mathbf{y}_k)}{p(u_{k,t} = -1 | \mathbf{y}_{k-1}, \mathbf{y}_k)} \\ &= \frac{\sum_{\substack{\mathbf{C}_k \\ u_{k,t}=+1}} p(\mathbf{C}_k | \mathbf{y}_k \mathbf{y}_{k-1})}{\sum_{\substack{\mathbf{C}_k \\ u_{k,t}=-1}} p(\mathbf{C}_k | \mathbf{y}_k \mathbf{y}_{k-1})} \end{aligned} \quad (32)$$

where the sum in the nominator is taken over all info
 20 matrices $\mathbf{C}_k$ which are associated with $u_{k,t} = +1$ and the sum
 in the denominator is taken over all info matrices $\mathbf{C}_k$
 associated with $u_{k,t} = -1$. Using (28) and the max-log
 approximation $\ln[e^{\delta_1} + e^{\delta_2}] \approx \max\{\delta_1, \delta_2\}$ we obtain

$$\begin{aligned} L(\hat{u}_{k,t}) &\approx \max_{\substack{\mathbf{C}_k \\ u_{k,t}=+1}} \left\{ \frac{\sqrt{a^{q_k}}}{\sigma^2(a^{q_k} + 1)} \text{trace} \{ \text{Re} \{ \mathbf{C}_k \mathbf{y}_k^H \mathbf{y}_{k-1} \} \} + \frac{1}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \right\} \\ &- \max_{\substack{\mathbf{C}_k \\ u_{k,t}=-1}} \left\{ \frac{\sqrt{a^{q_k}}}{\sigma^2(a^{q_k} + 1)} \text{trace} \{ \text{Re} \{ \mathbf{C}_k \mathbf{y}_k^H \mathbf{y}_{k-1} \} \} + \frac{1}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \right\}, t = 1, \dots, b_1 \end{aligned} \quad (33)$$

The above equation (33) still requires knowledge of a^{q_k} and
 of the noise variance σ^2 . Given the definition (40), the
 35 approximation

$$a^{q_k} \approx |v_k| \quad (34)$$

can be carried out, which is valid for reasonably high SNR. In order to avoid estimation of the noise variance, (33) is simply multiplied by σ^2 and (35) is computed

$$\begin{aligned} \sigma^2 L(\hat{u}_{k,t}) \approx & \max_{\substack{c_k \\ u_{k,t}=+1}} \left\{ \frac{\sqrt{|v_k|}}{|v_k| + 1} \text{trace} \left\{ \text{Re} \{ \mathbf{C}_k \mathbf{y}_k^H \mathbf{y}_{k-1} \} \right\} + \frac{\sigma^2}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \right\} \\ & - \max_{\substack{c_k \\ u_{k,t}=-1}} \left\{ \frac{\sqrt{|v_k|}}{|v_k| + 1} \text{trace} \left\{ \text{Re} \{ \mathbf{C}_k \mathbf{y}_k^H \mathbf{y}_{k-1} \} \right\} + \frac{\sigma^2}{2} \mathbf{u}_k^{(1)T} \mathbf{L}_a^n(\mathbf{u}_k^{(1)}) \right\}, t = 1, \dots, b_1 \end{aligned} \quad (35)$$

If the noise variance is constant over a frame, which is a reasonable assumption, all log-likelihood ratios are scaled by the constant factor σ^2 . This has no effect on the hard output of an outer Viterbi or Max-Log-type APP decoder. However, the APP log-likelihood ratios of the outer decoder will also be scaled by the same factor. Even in a turbo scheme this will cause no degradation as long as only Max-Log components are applied, since the fed back a priori info is scaled as required in (35). However, if a-priori information which is gained outside the turbo scheme is used G. Bauch and V. Franz, A comparison of soft-in/soft-out algorithms for "turbo-detection", in *International Conference on Telecommunications (ICT)*, June 1998 the noise variance for correct weighting is needed to know.

If $\mathbf{C}_k$ is an orthogonal design as described in a previous section, the detector can be simplified since decoupled expressions for the PSK symbols $c_{k,l}, l = 1, \dots, K$ can be obtained. Hence, in order to find the most likely transmitted data, only $KM_1^{1/K}$ comparisons are needed instead of M_1 . Starting from (22), a standard space-time block code diversity combiner can be applied as disclosed in V. Tarokh, H. Jafarkhani, and A.R. Calderbank, Space-time block codes

from orthogonal designs, *IEEE Transactions on Information Theory*, 45(5):1456--1467, June 1999 or in G. Bauch, J. Hagenauer, and N. Seshadri, Turbo processing in transmit antenna diversity systems, *Annals of Telecommunications*,
 5 Special Issue: Turbo codes - a widespreading technique, 56(7-8):455--471, August 2001, where the matrix $\mathbf{H}_k$ of the channel coefficients is replaced by $\mathbf{Y}_{k-1}$. This yields an equivalent system

$$10 \quad \hat{y}_{k,l} = \sqrt{a^{q_k}} \text{trace} \{ \mathbf{y}_{k-1} \mathbf{y}_{k-1}^H \} c_{k,l} + \hat{n}_{k,l}, \quad l = 1, \dots, K \quad (36)$$

of K single-input single-output (SISO) channels with noise variance

$$15 \quad \hat{\sigma}^2 = \sigma^2 (a^{q_k} + 1) \text{trace} \{ \mathbf{y}_{k-1} \mathbf{y}_{k-1}^H \}. \quad (37)$$

With the approximation (34), equation (38) is obtained

$$\log p(c_{k,l} | \hat{y}_{k,l}, \mathbf{y}_{k-1}) = \text{const} + \frac{\sqrt{|v_k|}}{(|v_k| + 1)\sigma^2} \text{Re} \{ \hat{y}_{k,l}^* c_{k,l} \} \quad (38)$$

20 and

$$\begin{aligned} \sigma^2 L(\hat{u}_{k,t}) \approx & \max_{\substack{c_{k,l} \\ u_{k,t}=+1}} \left\{ \frac{\sqrt{|v_k|}}{|v_k| + 1} \text{Re} \{ \hat{y}_{k,l}^* c_{k,l} \} + \frac{\sigma^2}{2} \mathbf{u}_{k,l}^{(1)T} \mathbf{L}_a^n(\mathbf{u}_{k,l}^{(1)}) \right\} \\ & - \max_{\substack{c_{k,l} \\ u_{k,t}=-1}} \left\{ \frac{\sqrt{|v_k|}}{|v_k| + 1} \text{Re} \{ \hat{y}_{k,l}^* c_{k,l} \} + \frac{\sigma^2}{2} \mathbf{u}_{k,l}^{(1)T} \mathbf{L}_a^n(\mathbf{u}_{k,l}^{(1)}) \right\}, \quad t = 1, \dots, b_1, l = 1, \dots, K \end{aligned} \quad (39)$$

25 From (24) we obtain

$$v_k = \frac{\text{trace} \{ \mathbf{Y}_k \mathbf{Y}_k^H \}}{\text{trace} \{ \mathbf{Y}_{k-1} \mathbf{Y}_{k-1}^H \}} = a^{q_k} + \bar{\mathbf{N}}_k \quad (40)$$

30 where

$$\bar{\mathbf{N}}_k = \frac{\text{trace}\{\sqrt{a_k^q} \mathbf{Y}_{k-1} \mathbf{C}_k \tilde{\mathbf{N}}_k^H + \sqrt{a_k^q} \tilde{\mathbf{N}}_k \mathbf{C}_k^H \mathbf{Y}_{k-1}^H + \tilde{\mathbf{N}}_k \tilde{\mathbf{N}}_k^H\}}{\text{trace}\{\mathbf{Y}_{k-1} \mathbf{Y}_{k-1}^H\}}. \quad (41)$$

A hard decision on q_k is obtained from

5

$$\hat{q}_k = \arg \min \{v_k - a^{q_k}\} \quad (42)$$

The bits $\mathbf{u}_k^{(2)}$ are obtained by demapping from $\hat{q}_k$.

- 10 The noise $\bar{\mathbf{N}}_k$ is not Gaussian. For computation of a soft-output, the logarithm of (40) is taken and (8), (21), (22), (23) are used and obtain

$$\begin{aligned} \log v_k &= \log \frac{\text{trace}\{\mathbf{Y}_k \mathbf{Y}_k^H\}}{\text{trace}\{\mathbf{Y}_{k-1} \mathbf{Y}_{k-1}^H\}} \\ &= \log \{ \text{trace}\{a^{q_k} \mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H + 2 \text{Re}\{\mathbf{H}_k \mathbf{X}_k \mathbf{N}_k^H\} + \mathbf{N}_k \mathbf{N}_k^H\} \} \\ 15 \quad &- \log \{ \text{trace}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H + 2 \text{Re}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{N}_{k-1}^H\} + \mathbf{N}_{k-1} \mathbf{N}_{k-1}^H\} \} \end{aligned} \quad (44)$$

The noise terms $\mathbf{N}_k \mathbf{N}_k^H$ and $\mathbf{N}_{k-1} \mathbf{N}_{k-1}^H$ are neglected and are approximated (43) by the first two terms of the Taylor series $f(a+b) = f(a) + \frac{b}{11} f'(a) \dots$ which yields

20

$$\begin{aligned} \log v_k &\approx \log [a^{q_k} \text{trace}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H\}] + \frac{\text{trace}\{2 \text{Re}\{\mathbf{H}_k \mathbf{X}_k \mathbf{N}_k^H\}\}}{a^{q_k} \text{trace}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H\}} \\ &- \log [\text{trace}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H\}] - \frac{\text{trace}\{2 \text{Re}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{N}_{k-1}^H\}\}}{\text{trace}\{\mathbf{H}_k \mathbf{X}_{k-1} \mathbf{X}_{k-1}^H \mathbf{H}_k^H\}} \\ &= \log a^{q_k} + \tilde{n}_k \end{aligned} \quad (45)$$

where $\tilde{n}_k$ is real white Gaussian noise with variance

25

$$\tilde{\sigma}^2 = \frac{1 + a^{q_k}}{a^{q_k}} \frac{4\sigma^2}{a^{q_k} \text{trace}\{\mathbf{H}_k \mathbf{H}_k^H\}} \quad (46)$$

With the approximations (34) and

$$a^{z_{k-1}} \text{trace}\{\mathbf{H}_k \mathbf{H}_k^H\} \approx \text{trace}\{\mathbf{y}_{k-1} \mathbf{y}_{k-1}^H\} \quad (47)$$

we obtain

$$\begin{aligned} \log p(q_k | \mathbf{y}_{k-1}, \mathbf{y}_k) &= \text{const} + \frac{1}{2\sigma^2} \left| \log v_k - \log a^{q_k} \right|^2 \\ &\approx \text{const} + \frac{|v_k| \text{trace}\{\mathbf{y}_{k-1} \mathbf{y}_{k-1}^H\}}{8\sigma^2(1 + |v_k|)} \left| \log \frac{v_k}{a^{q_k}} \right|^2 \end{aligned}$$

As in (39), we use the max-log approximation of the APP log-likelihood ratios and multiply by σ^2 which yields

$$\begin{aligned} \sigma^2 L(\hat{u}_{k,t}) &\approx \max_{\substack{q_k \\ u_{k,t} = +1}} \left\{ \frac{|v_k| \text{trace}\{\mathbf{y}_{k-1} \mathbf{y}_{k-1}^H\}}{8(1 + |v_k|)} \left| \log \frac{v_k}{a^{q_k}} \right|^2 + \frac{\sigma^2}{2} \mathbf{u}_k^{(2)T} \mathbf{I}_a^n(\mathbf{u}_k^{(2)}) \right\} \\ &- \max_{\substack{q_k \\ u_{k,t} = -1}} \left\{ \frac{|v_k| \text{trace}\{\mathbf{y}_{k-1} \mathbf{y}_{k-1}^H\}}{8(1 + |v_k|)} \left| \log \frac{v_k}{a^{q_k}} \right|^2 + \frac{\sigma^2}{2} \mathbf{u}_k^{(2)T} \mathbf{I}_a^n(\mathbf{u}_k^{(2)}) \right\}, t = b_1, \dots, b_1 \end{aligned} \quad (49)$$

where

$$\mathbf{I}_a^n(\mathbf{u}_k^{(2)}) = \left[L_a^n(u_{k,b_1}), \dots, L_a^n(u_{k,b_1+b_2}) \right]^T \quad (50)$$

Note that, as desired, the log-likelihood ratios in (49) are scaled by the same (unknown) factor σ^2 as the log-likelihood ratios in (35) and (39), respectively.

The principle of iterative ("turbo") detection is illustrated in Figure 7. Bit interleaved coded modulation is considered with iterative ("turbo") detection. The data is encoded by the encoder of a forward error control (= FEC) code, e.g. by a binary convolutional code. The code bits $u_{k,l}$ are bit-interleaved according to the interleaving rule Π and mapped on entries of a matrix $\mathbf{C}_k$ out of the set $\mathcal{C}$ of possible matrices $\mathbf{C}_k$. The transmit matrix $\mathbf{X}_k$ is obtained by differential matrix encoding based on the data matrix $\mathbf{C}_k$ and the previously transmitted matrix $\mathbf{X}_{k-1}$.

At the receiver, a non-coherent differential detector computes log likelihood ratios $L^n(\hat{u}_{k,l})$ as described above. From the a-posteriori log-likelihood ratios $L^n(\hat{u}_{k,l})$, we
 5 compute the extrinsic information $L_e^n(\hat{u}_{k,l})$ by bit-wise subtraction of the a-priori log-likelihood ratios $L_a^n(\hat{u}_{k,l})$ which are input to the non-coherent detector. In the first detection step, $L_a^n(\hat{u}_{k,l}) = 0$ results. The extrinsic L-values are passed to the FEC decoder. The FEC decoder, which can
 10 be based on e.g. the BCJR algorithm as disclosed in L.R. Bahl, J. Cocke, F. Jelinek and J. Raviv, Optimal decoding of linear codes for minimizing symbol error rate, IEEE Transactions on Information Theory, IT-20:284-287, March 1974 or the Viterbi algorithm, computes improved a-
 15 posteriori L-values $L^d(\hat{u}_{k,l})$ about the code bits taking into account the code constraints. Again, extrinsic information $L_e^d(\hat{u}_{k,l})$ is computed which is fed back to the non coherent detector. The non-coherent detector delivers new extrinsic L-values $L_e^n(\hat{u}_{k,l})$ based on the stored received matrices $\mathbf{Y}_k$ and
 20 $\mathbf{Y}_{k-1}$ and the fed back L-values as an additional a-priori information $L_a^n(u_{k,l}) = L_e^d(\hat{u}_{k,l})$. Performing some iterations can significantly improve the error rate.

However, similarly as has been shown in S. ten Brink,
 25 J. Speidel, and R.-H. Yam, Iterative demapping for QPSK modulation, *Electronic Letters*, pages 1459-1460, July 1998, A. Chindapol and A. Ritcey, Design, analysis and performance evaluation for BICM-ID with square QAM constellations in Rayleigh fading channels, *IEEE Journal*
 30 *on Selected Areas in Communications*, 19(5):944-957, May 2001, F. Schreckenbach, N. Görtz, J. Hagenauer and G. Bauch, Optimized symbol mappings for bit-interleaved coded modulation with iterative decoding, *IEEE Communications Letters*, 7(12):593-595, December 2003 and
 35 F. Schreckenbach, N. Görtz, J. Hagenauer, and G. Bauch, Optimized symbol mappings for bit-interleaved coded modulation with iterative decoding, *IEEE Communications*

Letters, 7(12):593-595, December 2003 for iterative demapping and decoding in non-differential single antenna systems, the particular mapping of bits to matrices $\mathbf{C}_k$ has a great impact on gains which can be achieved by turbo iterations.

A particularly interesting mapping scheme for non-differential PSK modulation in single antenna systems was presented in T. Clevorn and P. Vary, Iterative decoding of BICM with non-regular signal constellation sets, in *International ITC Conference on Source and Channel Coding*, pages 259-266, January 2004 and T. Clevorn, S. Godtmann, and P. Vary, EXIT chart analysis of non-regular signal constellation sets for BICM-ID, in *International Symposium on Information Theory and its Applications (ISITA)*, pages 21-26, October 2004. Here, a non-unique mapping of bits to constellation points of the modulation scheme is used, i.e. b bits are mapped on a constellation with M different points where $M < 2^b$. For example, a 6-PSK constellation is used for transmission of 3 bit/symbol. Since a unique mapping of 3 bit/symbol would require a constellation with $2^3 = 8$ different points, ambiguities occur. Those ambiguities are resolved in the iterative process: In the first detection step, the differential detector delivers log-likelihood ratios $L^n(\hat{u}_{k,l}) = 0$ for those bits with ambiguities, i.e. no information is passed to the decoder for these bits. Consequently, the decoder sees a punctured code. If the mapping is carefully designed, the decoder will deliver extrinsic information $L_e^d(\hat{u}_{k,l}) \neq 0$ about all code bits which is fed back to the differential detector. Hence, in the following iteration, the demapper can use input information on all code bits and delivers improved L-values to the decoder.

35

The advantage of those exemplary mappings is a lower error floor on the expense of convergence at slightly higher SNR

compared to unique mappings. The bit sequences which are assigned to the same constellation point should differ in at least two bits. Otherwise, ambiguities possibly cannot be resolved based on fed back extrinsic information. This is because extrinsic information means that information is available for the other bits but not for the bit on which the current decision is taken. If two bit sequences which are assigned to the same constellation point differ in the last bit, the extrinsic information for the decision on the last bit would be the same for both bit sequences and, hence, no gain would be obtained by turbo iterations.

The discussion in T. Clevorn and P. Vary, Iterative decoding of BICM with non-regular signal constellation sets, in *International ITC Conference on Source and Channel Coding*, pages 259-266, January 2004 and T. Clevorn, S. Godtmann, and P. Vary, EXIT chart analysis of non-regular signal constellation sets for BICM-ID, in *International Symposium on Information Theory and its Applications (ISITA)*, pages 21-26, October 2004, is limited to signal constellations for non-differential transmission in single antenna systems where all points are located on a circle, i.e. PSK type constellations, with $b \leq 3$ bit symbol. Furthermore, not more than two bit sequences are mapped to a particular constellation point.

The presented proposal is to generalize the conventional idea especially to MIMO-Systems as several significant improvements can be expected, which are not achievable by just applying the conventional approaches using distinct signal constellation points to the symbol matrix of a MIMO-System. Furthermore, an even higher performance improvement can be realized by using differential matrix modulation. A bit sequence of b bits is mapped to an info matrix $\mathbf{C}_k$ where the number of possible info matrices $\mathbf{C}_k$ is $M < 2^b$. This means that the mapping of bits to matrices is not

unique. The ambiguities are resolved by the iterative detector.

In case of differential unitary matrix modulation based on
5 orthogonal designs with PSK modulation as described in a
previous Section, e.g. the non-unique mappings which were
introduced in T. Clevorn and P. Vary, Iterative decoding of
BICM with non-regular signal constellation sets, in
10 *International ITC Conference on Source and Channel
Coding*, pages 259-266, January 2004 or T. Clevorn,
S. Godtmann, and P. Vary, EXIT chart analysis of non-
regular signal constellation sets for BICM-ID, in
*International Symposium on Information Theory and its
Applications (ISITA)*, pages 21-26, October 2004 can be
15 used for the mapping of bits to PSK symbols which then
constitute the entries of the unitary matrix $\mathbf{C}_k$.

For the differential amplitude and unitary matrix
modulation scheme which was described in a previous
20 section, non-unique mappings can be applied in both the
mapping of bits $\mathbf{u}_k^{(1)}$ in set 1 to a unitary matrix $\mathbf{C}_k$ and
for the mapping of bits $\mathbf{u}_k^{(2)}$ in set 2 to the amplitude
modulation exponent q_k .

25 To prove the performance of the invention, simulations for
 $n_T = 2$ transmit antennas and $n_R = 1$ receive antenna are
simulated in a spatially uncorrelated Rayleigh fading
channel which varies in time according to a Jakes spectrum
with normalized maximum Doppler frequencies
30 $f_d T_s = 0,005, \dots, 0,02$. BER results are depicted in Figures 8A
to 8C for non-unique differential unitary matrix modulation
based on orthogonal designs with transmission of
8 bit/matrix.

35 The matrix modulation based on 16-PSK modulation is
compared with Gray mapping, 16-PSK with turbo optimized
mapping (optimized) and non-unique mapping (non regular).

The performance advantage of the non-unique mapping is obvious. Particularly, the gain increases with increasing maximum Doppler frequency. This is due to the fact that the assumption of a constant channel during transmission of two successive matrices does hold the less the higher the Doppler spread. This causes an unknown phase offset. A constellation with less PSK points as used in the non-unique scheme is more robust to a phase offset than PSK with unique mapping and, hence, more constellation points with smaller Euclidean distance. The Figures 8A to 8C show especially:

Figure 8A: A Simulation of non-unique differential unitary matrix modulation based on orthogonal designs with 8 bits/matrix. Time-varying fading channel ($f_d T_s = 0.005$),

Figure 8B: A Simulation of non-unique differential unitary matrix modulation based on orthogonal designs with 8 bits/matrix. Time-varying fading channel ($f_d T_s = 0.01$), and

Figure 8C: A Simulation of non-unique differential unitary matrix modulation based on orthogonal designs with 8 bits/matrix. Time-varying fading channel ($f_d T_s = 0.02$).

Furthermore, it is obvious for a person skilled in the art that the detector as shown in the lower part of Fig. 7 must be provided with a mapping scheme equivalent to the mapping scheme for mapping bit sequences to symbol matrices in the transmitter in order to be able to demap the respective bit sequences from a received symbol matrix.

Especially, a differential matrix modulation with forward error control coding is considered herewith. A differential matrix modulation where the number of possible transmit matrices is smaller than the number of possible bit sequences is proposed in the present invention. The ambiguities are resolved in an iterative ("turbo") detector. The mapping of bit sequences to matrices has to be carefully chosen in order to obtain gains in the iterative scheme. Particular mappings are also proposed herein.

Moreover, depending on certain implementation requirements of the inventive method or the inventive methods can be implemented in hardware or in software. The implementation can be performed using a digital storing medium, in particular a disc or a CD having electronically readable control signals stored thereon, which can operate with a programmable computer system such that the inventive methods are performed. Generally, the present invention is, therefore, a computer program product with a program code stored on a machine-readable carrier, the program code being configured for performing the inventive methods, when the computer program product runs on a computer. In other words, the inventive methods are, therefore, a computer program having a program code for performing the inventive methods, wherein the computer program runs on a computer.

Claims

1. Apparatus for providing a transmit matrix (X_k) for a transmitter (400) having a plurality of transmitting points (422), the apparatus comprising:
- 5 a mapper (100) for mapping bit sequences (BS1, BS2) to symbol matrices (C), the mapper (100) being configured to have a mapping rule in which a first bit sequence (BS1) is mapped to a symbol matrix (C) and a second
- 10 bit sequence (BS2) is mapped to the same symbol matrix (C), the first and second bit sequences (BS1, BS2) being different from each other, wherein the symbol matrix (C) has a number of rows and a number of
- 15 columns, the number of rows and the number of columns being larger than one; and
- a processor (414, 418, 420) for processing the symbol matrix to obtain the transmit matrix (X_k), the
- 20 transmit matrix (X_k) having a number of rows being equal to the number of transmitting points (422) and having a number of columns being equal to the number of columns or rows of the symbol matrix (C).
- 25 2. Apparatus according to claim 1, wherein first bit sequence (BS1) and the second bit sequence (BS2) differ in at least two bits.
3. Apparatus according to claim 1 or 2, wherein the
- 30 symbol matrix (C) is unitary.
4. Apparatus according to one of claims 1 to 3, wherein the symbol matrix (C) comprises symbol matrix entries

(110) representing signal space constellation points
(202) according to a mapping scheme.

- 5 5. Apparatus according to claim 4, wherein the mapping
 scheme comprises signal space constellation points
 (202) having different phases.
6. Apparatus according to claim 4 or 5, wherein the
10 mapping scheme comprises signal space constellation
 points (202) having different amplitudes.
7. Apparatus according to one of claims 4 to 6, wherein
 the mapping scheme is one of the following mapping
 schemes: QPSK, PAM, QAM, PSK, M-QAM, M-PSK, M being a
15 number of equal or greater than one.
8. Apparatus according to one of claims 4 to 6, wherein
 the mapper is configured to a map a group of bits
 (112) of the first bit sequence (BS1) to a non-unique
20 symbol matrix entry (114) and to a map a group of bits
 (112) of the second bit sequence (BS2) to the same
 non-unique symbol matrix entry (114), the position of
 group of bits (112) in the first and second bit
 sequences corresponding to each other and the group of
25 bits (112) in the first and second bit sequence being
 different from each other.
9. Apparatus according to claim 8, wherein the groups of
 bits (112) comprises at least three bits.
- 30 10. Apparatus according to one of claims 1 to 9, wherein
 the symbol matrices (C) are diagonal matrices.

11. Apparatus according to one of claims 1 to 9, wherein the mapper (100) is configured to provide a symbol matrix

$$C = \begin{bmatrix} c_1 & c_2 \\ -c_2^* & c_1^* \end{bmatrix},$$

a symbol matrix

$$C = \begin{bmatrix} c_1 & c_2 \\ -c_3 & -\frac{c_1^* \cdot c_3}{c_2^*} \end{bmatrix}$$

or a symbol matrix

$$C = \begin{bmatrix} c_1 & c_2 & c_3 & 0 \\ -c_2^* & c_1^* & 0 & c_3 \\ c_3^* & 0 & -c_1^* & c_2 \\ 0 & c_3^* & -c_2^* & -c_1 \end{bmatrix},$$

- 10 wherein c_1 , c_2 and c_3 are signal constellation points of a signal space and c_1^* , c_2^* , c_3^* are complex conjugates of c_1 , c_2 and c_3 , respectively.

12. Apparatus according to one of claims 1 to 3, wherein the mapper (100) is configured to map the bit sequences 1111, 1100, 0000, 1001 and 1010 to the symbol matrix

$$C = \begin{pmatrix} -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \end{pmatrix};$$

- 20 wherein the mapper (100) is configured to map the bit sequences 0100 and 0111 to the symbol matrix

$$C = \begin{pmatrix} -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix};$$

wherein the mapper (100) is configured to map the bit sequences 1110, 0001 and 0010 to the symbol matrix

$$C = \begin{pmatrix} -i & 0 \\ 0 & i \end{pmatrix};$$

5 and wherein the mapper (100) is configured to map the bit sequence 1101, 1000 and 1011 to the symbol matrix

$$C = \begin{pmatrix} \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix}.$$

13. Apparatus according to one of claims 1 to 3, wherein the mapper (100) is configured to map the bit sequences 0011, 0000, 0101, 0110, 1001 and 1100 to the symbol matrix

$$C = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix};$$

15 and wherein the mapper is configured to map the bit sequences 0001, 0100, 1000 and 1101 to the symbol matrix

$$C = \begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}.$$

14. Apparatus according to one of claims 1 to 13, wherein the mapper (100) is configured to map a third bit sequence to the same symbol matrix (C), the third bit sequence being different from the first bit sequence (BS1) and being different from the second bit sequence (BS2).

25 15. Apparatus according to one of claims 1 to 14, wherein the transmit matrix is to be transmitted within a k-th time interval,

wherein the processor (414, 418, 420) comprises a multiplier (418, 414) for multiplying the symbol matrix (C) with a normalization factor (412) or for multiplying the symbol matrix (C) with a transmit matrix (X_{k-1}) to be transmitted within the (k-1)-th time interval or for multiplying the symbol matrix (C) with a product of the normalization factor (416) and the transmit matrix (X_{k-1}) to be transmitted within the (k-1)-th time interval in order to obtain the transmit matrix (X_k).

16. Apparatus according to claim 15, wherein the multiplier comprises a further mapper (410) for mapping an additional bit sequence ($u_k^{(2)}$) to the normalization factor (412).

17. Apparatus according to claim 16, wherein the first or second bit sequence (BS1, BS2) forms a codeword when combined with the additional bit sequence ($u_k^{(2)}$).

18. Apparatus according to claim 16 or 17, wherein the further mapper (410) is configured to map a first and second additional bit sequence ($u_k^{(2)}$) to the same normalization factor (412), the first and second additional bit sequence ($u_k^{(2)}$) being different from each other.

19. Apparatus according to one of claims 16 to 18, wherein the further mapper (410) is configured to map the additional bit sequence ($u_k^{(2)}$) to the normalization factor (412) on the basis of an amplitude (z_{k-1}) of a signal being derived from the transmit matrix (X_{k-1}) to be transmitted within the (k-1)-th time interval.

20. Apparatus according to claim 19, wherein the further mapper (410) is configured to provide a value of $\sqrt{a^{q_k}}$ for the normalization factor (412), wherein a is a real constant and q_k is given by

$$q_k = z_{k-1} - M_2 \lfloor (z_{k-1} + d_k) / M_2 \rfloor,$$

wherein z_{k-1} denotes an amplitude exponent of a signal being derived from the transmit matrix X_{k-1} to be transmitted within the $(k-1)$ -th time interval, wherein $\lfloor \cdot \rfloor$ denoted the floor function, M_2 denotes a maximum possible number of different additional bit sequences ($u_k^{(2)}$) each having a number of $\log_2(M_2)$ bits and wherein d_k denotes a number according to a mapping scheme being used for mapping the additional bit sequences ($u_k^{(2)}$) to the normalization factor (412).

21. Apparatus according to claims 15 to 20, wherein the multiplier (414, 418) is configured to multiply the normalization factor (412) and the transmit matrix (X_{k-1}) to be transmitted within the $(k-1)$ -th time interval in order to obtain the transmit matrix (X_k) to be transmitted in the k -th time interval.

22. Transmitter apparatus comprising:

the apparatus for providing a transmit matrix (X_k) according to one of claims 1 to 21, wherein the transmit matrix (X_k) defines transmit sequences to be transmitted; and

n_T transmitting points for transmitting information included in the transmit matrix (X_k), wherein the n_T transmitting points are configured for simultaneously transmitting n_T values of a transmit sequence of a time instant within a time interval and for simultaneously transmitting n_T values of a further transmit sequence and a further time interval within the time interval, wherein n_T denotes a number of transmitting points being greater than 1.

10

23. Receiver apparatus comprising:

means for receiving information via n_R receiving points in order to obtain a receiver matrix, the receiver matrix having a number of rows being equal to the number of receiving points and having a number of columns being equal to the number of columns or rows of a receiver symbol matrix, wherein n_R denotes a number of transmitting points being greater than 1;

15

20

a processor for processing the receiver matrix received from the means for receiving information to obtain the receiver symbol matrix, wherein the receiver symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

25

a demapper for demapping bit sequences from the receiver symbol matrix according to a demapping rule in which a first bit sequence is demapped from a receiver symbol matrix and a second bit sequence is demapped from the same receiver symbol matrix, the

30

first and second bit sequences being different from each other.

24. Method for providing a transmit matrix (X_k) for a transmitter (400), the transmitter (400) having a plurality of transmitting points (422), the method comprising:

mapping bit sequences (BS1, BS2) to symbol matrices (C) using a mapping rule according to which a first bit sequence (BS1) is mapped to a symbol matrix (C) and a second bit sequence (BS2) is mapped to the same symbol matrix (C), the first and second bit sequences (BS1, BS2) being different from each other, wherein the symbol matrix (C) has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

processing the symbol matrix (C) using a processor (414, 418, 420) to obtain the transmit matrix (X_k), the transmit matrix (X_k) having a number of rows being equal to the number of transmitting points (422) and having a number of columns being equal to the number of columns or rows of the symbol matrix (C).

25

25. Transmission method comprising the steps of:

providing a transmit matrix (X_k) according to claim 24, wherein the transmit matrix (X_k) defines transmit sequences to be transmitted; and

30

transmitting information included in the transmit matrix (X_k) via n_T transmitting points, wherein n_T

values of a transmit sequence are simultaneously transmitted at the time instant within a time interval and n_T values of a further transmit sequence are simultaneously transmitted at a further time interval within the time interval, wherein n_T denotes a number of transmitting points being greater than 1.

26. Reception method comprising the steps of:

receiving information via n_R receiving points in order to obtain a receiver matrix, the receiver matrix having a number of rows being equal to the number of receiving points and having a number of columns being equal to the number of columns or rows of a receiver symbol matrix, wherein n_R denotes a number of transmitting points being greater than 1;

processing the receiver matrix received from the means for receiving information to obtain the receiver symbol matrix, wherein the receiver symbol matrix has a number of rows and a number of columns, the number of rows and the number of columns being larger than one; and

demapping bit sequences from the receiver symbol matrix using a demapping rule according to which a first bit sequence is demapped from a receiver symbol matrix and a second bit sequence is demapped from the same receiver symbol matrix, the first and second bit sequences being different from each other.

27. Computer program for performing at least one of the methods according to one of claims 24 to 26, when the computer program runs on a computer.

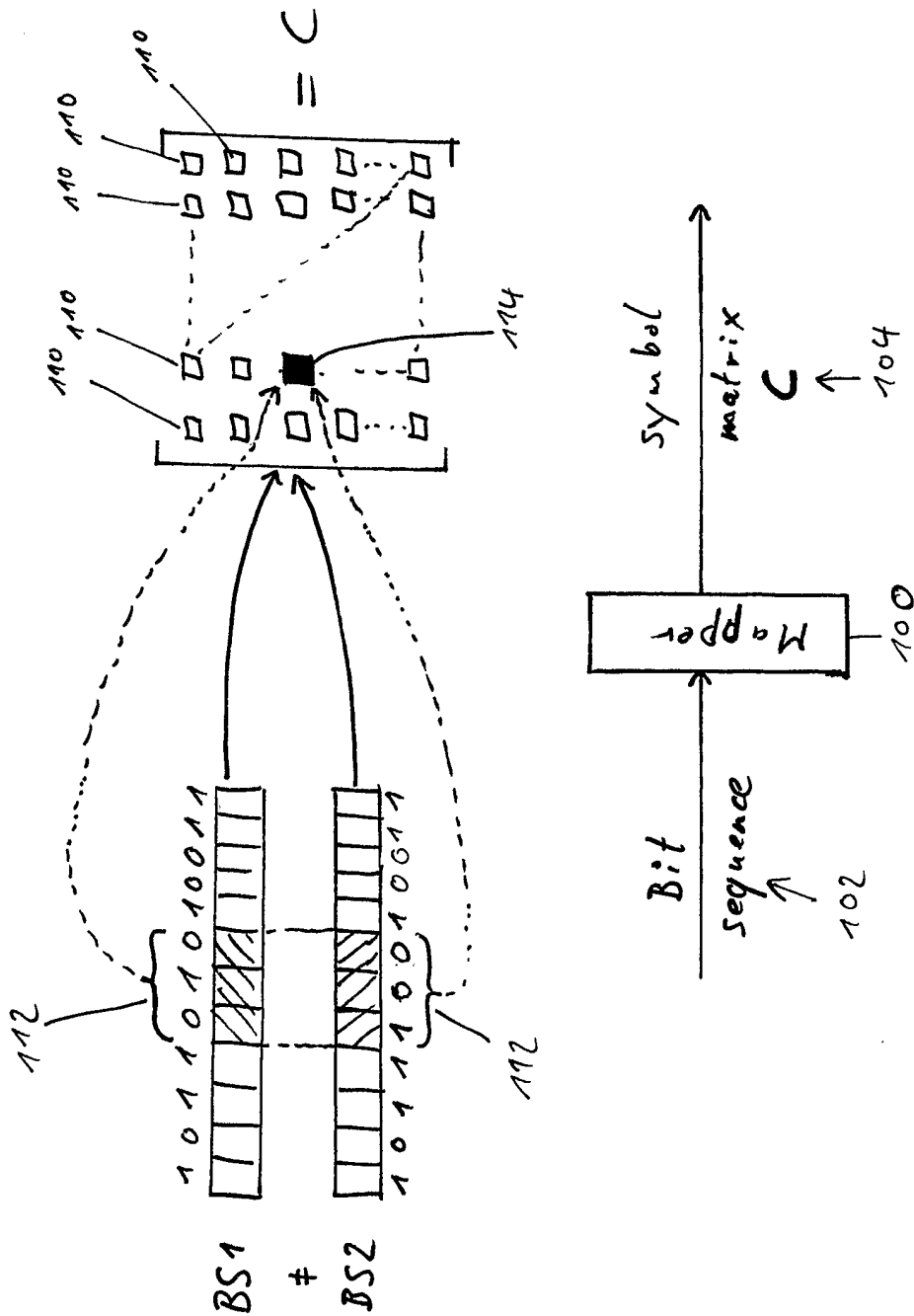


Fig. 1

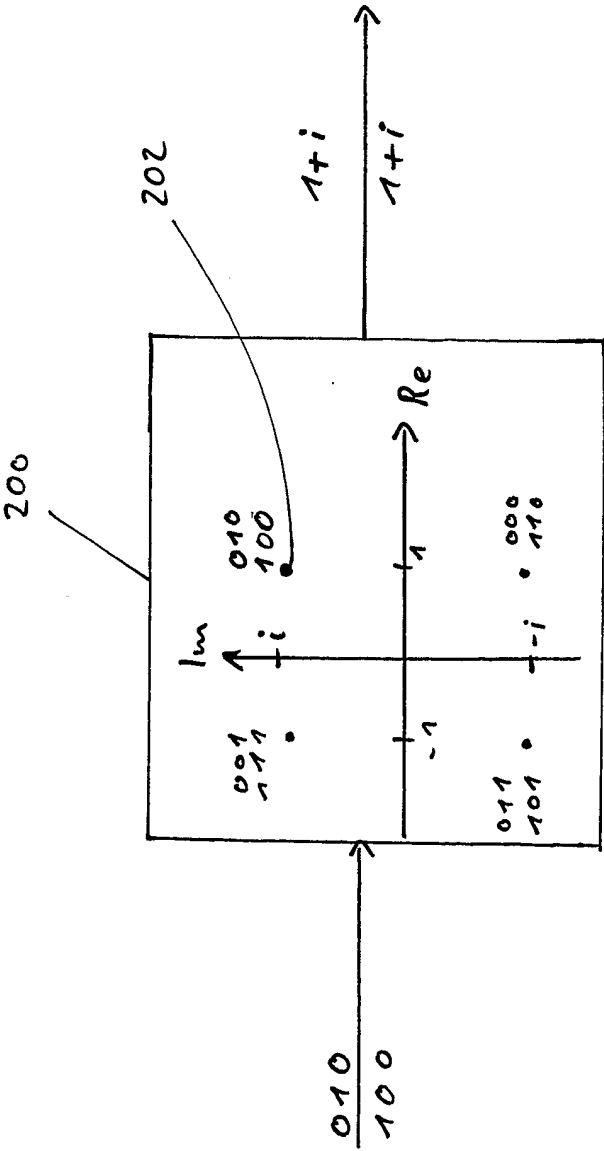


Fig. 2

	Hughes Constellation	H7
$G_1 =$	$\begin{pmatrix} \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \\ 0 & \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \end{pmatrix}$	0110
$G_2 =$	$\begin{pmatrix} i & 0 \\ 0 & -i \end{pmatrix}$	-
$G_3 =$	$\begin{pmatrix} -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \end{pmatrix}$	1111, 1100, 0000, 1001, 1010
$G_4 =$	$\begin{pmatrix} -1 & 0 \\ 0 & -1 \end{pmatrix}$	-
$G_5 =$	$\begin{pmatrix} -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix}$	0100, 0111
$G_6 =$	$\begin{pmatrix} -i & 0 \\ 0 & i \end{pmatrix}$	1110, 0001, 0010
$G_7 =$	$\begin{pmatrix} \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix}$	1101, 1000, 1011
$G_8 =$	$\begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}$	-
$G_9 =$	$\begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}$	-
$G_{10} =$	$\begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}$	-
$G_{11} =$	$\begin{pmatrix} 0 & -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	0011
$G_{12} =$	$\begin{pmatrix} 0 & \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \\ -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	-
$G_{13} =$	$\begin{pmatrix} 0 & -i \\ -i & 0 \end{pmatrix}$	0101
$G_{14} =$	$\begin{pmatrix} 0 & i \\ i & 0 \end{pmatrix}$	-
$G_{15} =$	$\begin{pmatrix} 0 & \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \\ -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	-
$G_{16} =$	$\begin{pmatrix} 0 & -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	-

Fig. 3 A

	Hughes Constellation	Optimum	Natural	H8
$G_1 =$	$\begin{pmatrix} \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \\ 0 & \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \end{pmatrix}$	1100	0000	-
$G_2 =$	$\begin{pmatrix} i & 0 \\ 0 & -i \end{pmatrix}$	1011	0001	0111
$G_3 =$	$\begin{pmatrix} -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \end{pmatrix}$	0111	0010	-
$G_4 =$	$\begin{pmatrix} -1 & 0 \\ 0 & -1 \end{pmatrix}$	1110	0011	1111
$G_5 =$	$\begin{pmatrix} -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix}$	0100	0100	-
$G_6 =$	$\begin{pmatrix} -i & 0 \\ 0 & i \end{pmatrix}$	0011	0101	0010
$G_7 =$	$\begin{pmatrix} \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \\ 0 & \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \end{pmatrix}$	1111	0110	-
$G_8 =$	$\begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}$	0110	0111	1110
$G_9 =$	$\begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}$	1001	1000	0011, 0000, 0101, 0110, 1001, 1100
$G_{10} =$	$\begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}$	0001	1001	0001, 0100, 1000, 1101
$G_{11} =$	$\begin{pmatrix} 0 & -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	0101	1010	-
$G_{12} =$	$\begin{pmatrix} 0 & \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \\ -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	1101	1011	-
$G_{13} =$	$\begin{pmatrix} 0 & -i \\ -i & 0 \end{pmatrix}$	0010	1100	1011
$G_{14} =$	$\begin{pmatrix} 0 & i \\ i & 0 \end{pmatrix}$	1010	1101	1010
$G_{15} =$	$\begin{pmatrix} 0 & \frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} \\ -\frac{1}{\sqrt{2}} - \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	1000	1110	-
$G_{16} =$	$\begin{pmatrix} 0 & -\frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} + \frac{i}{\sqrt{2}} & 0 \end{pmatrix}$	0000	1111	-

Fig. 3B

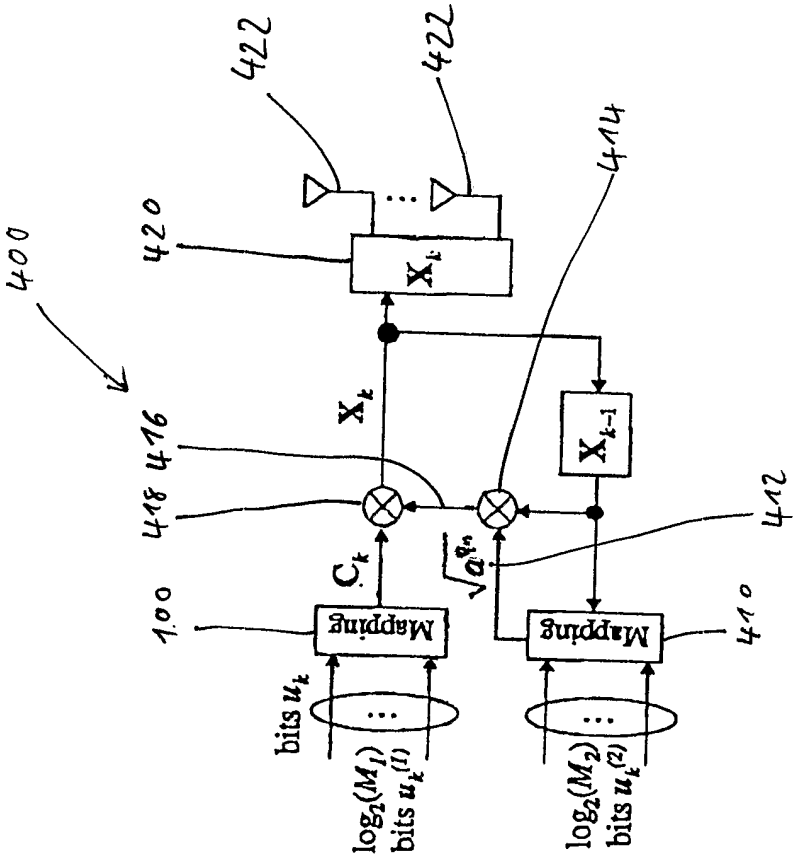


Fig. 4

z_{k-1}	$ X_{k-1}X_{k-1}^H _1$	$u_{k,b_1+1}, u_{k,b_1+2} d_k$					
		00 0	01 1	11 2	10 3		
0	1	0	1	2	3		
1	a	0	1	2	-1		
2	a^2	0	1	-2	-1		
3	a^3	0	-3	-2	-1		

Fig. 5

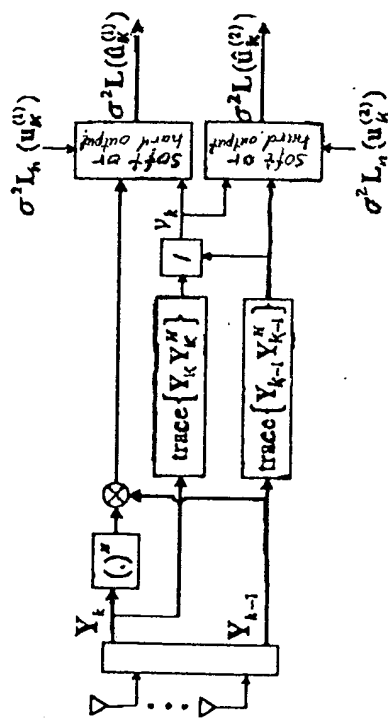


Fig. 6

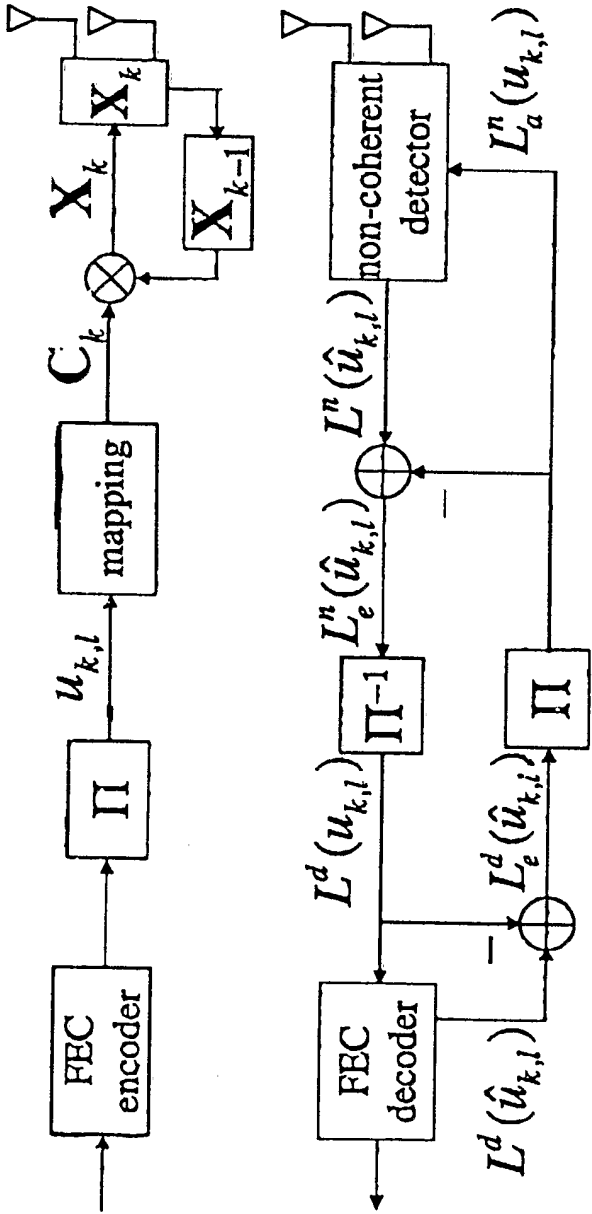


Fig. 7

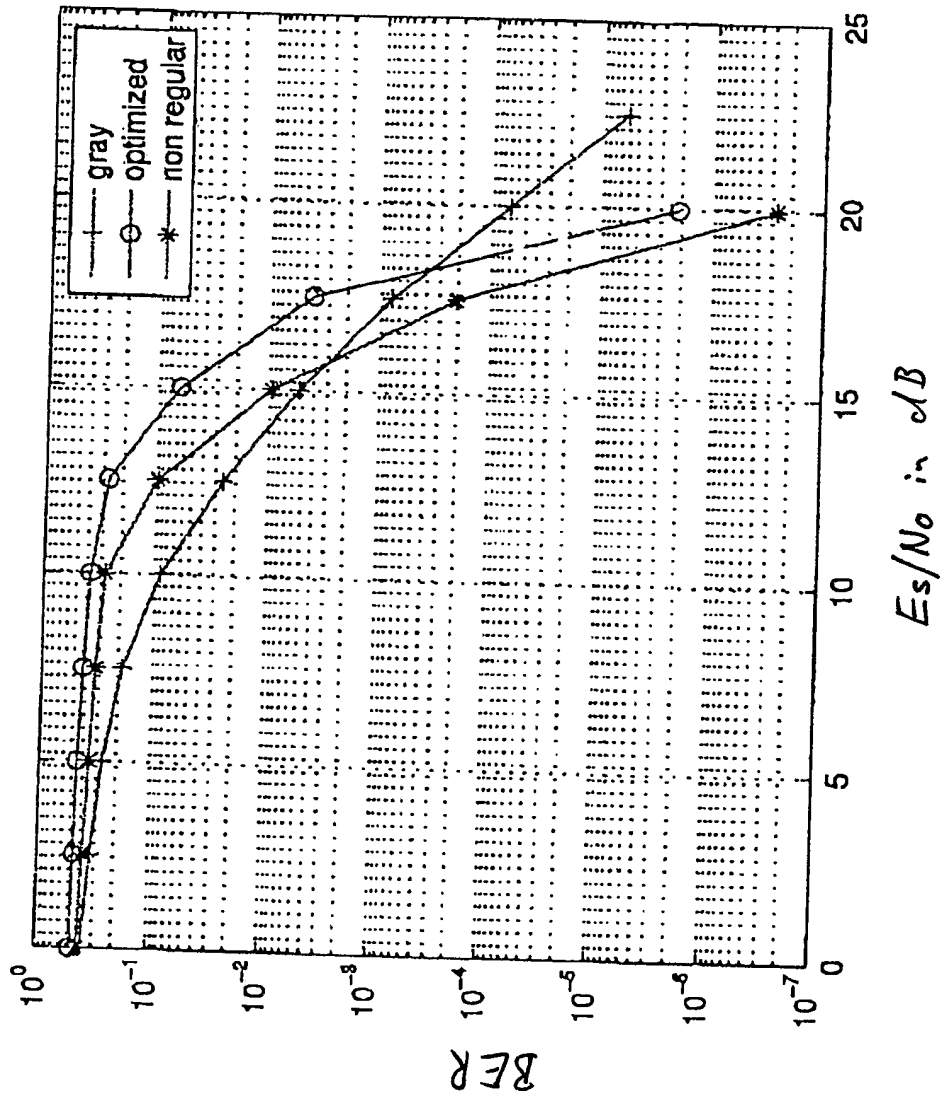


Fig. 8A

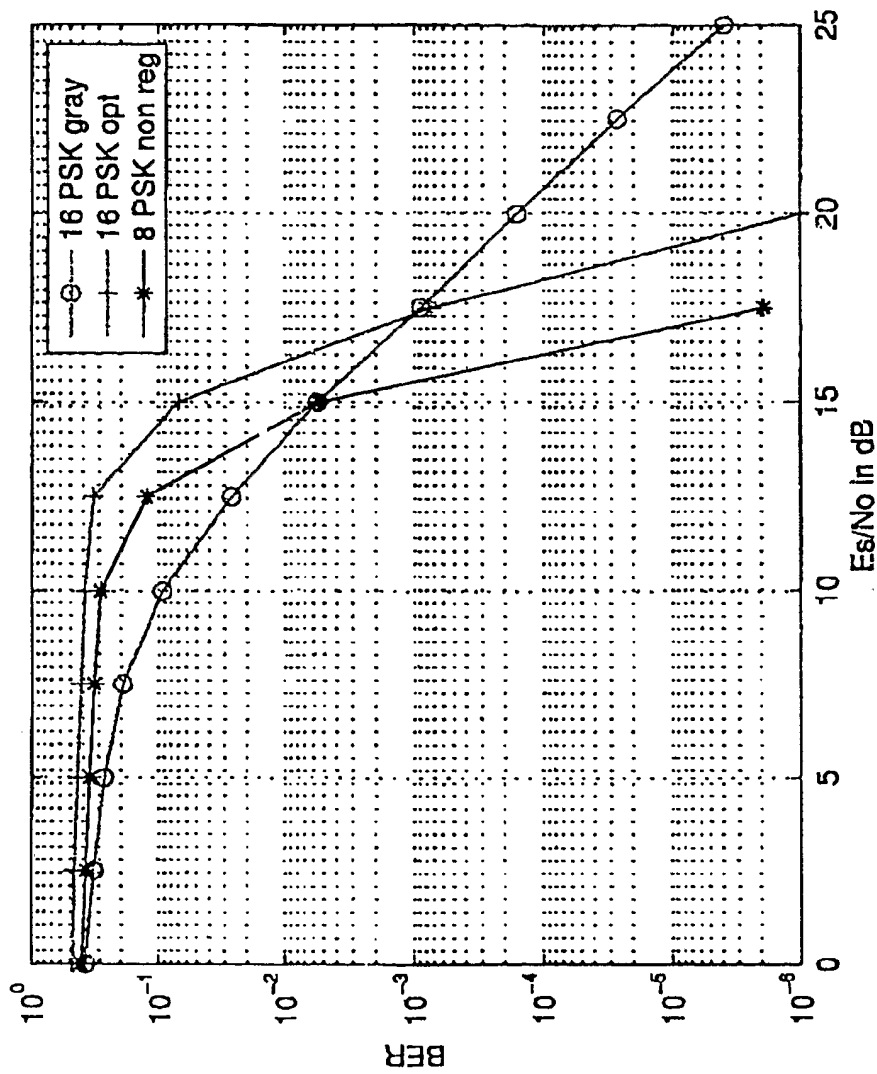


Fig. 8B

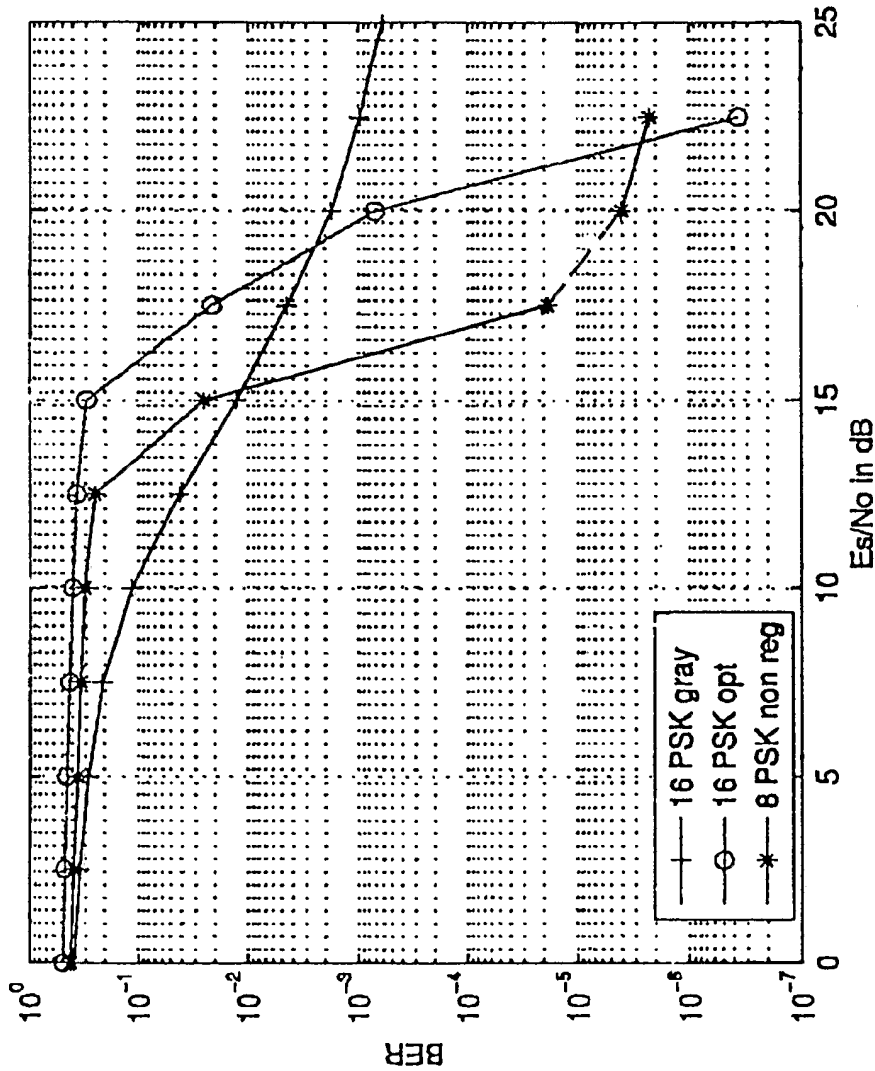


Fig. 8C

INTERNATIONAL SEARCH REPORT

International Application No
/EP2005/002690A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/06 H04L27/34

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, IBM-TDB, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	LAMPE, L.H.-J. ET AL.: "Bit-interleaved coded differential space-time modulation" COMMUNICATIONS, IEEE TRANSACTIONS ON, vol. 50, no. 9, September 2002 (2002-09), pages 1429-1439, XP011071002 page 1430, left-hand column, paragraph 2 - right-hand column, paragraph 3; figure 1	1-27
A	LUTZ H.-J. LAMPE ET AL.: "Coded differential space--Time modulation for flat fading channels" IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS, vol. 2, no. 3, May 2003 (2003-05), pages 582-590, XP002350796 page 582, right-hand column, paragraph 4 page 584, right-hand column, paragraph 2; figure 1b	1-27



Further documents are listed in the continuation of box C.



Patent family members are listed in annex.

* Special categories of cited documents:

- *A* document defining the general state of the art which is not considered to be of particular relevance
- *E* earlier document but published on or after the international filing date
- *L* document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
- *O* document referring to an oral disclosure, use, exhibition or other means
- *P* document published prior to the international filing date but later than the priority date claimed

- *T* later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
- *X* document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
- *Y* document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.
- *&* document member of the same patent family

Date of the actual completion of the international search

25 October 2005

Date of mailing of the international search report

08/11/2005

Name and mailing address of the ISA

European Patent Office, P.B. 5818 Patentlaan 2
NL - 2280 HV Rijswijk
Tel. (+31-70) 340-2040, Tx. 31 651 epo nl,
Fax: (+31-70) 340-3016

Authorized officer

Agudo Cortada, E

INTERNATIONAL SEARCH REPORT

International Application No.
PCT/EP2005/002690

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	A.STEINER: "Iterative decoding of space-time differentially coded unitary matrix modulation" IEEE TRANS. SIGNAL PROCESSING, vol. 50, October 2002 (2002-10), pages 2385-2395, XP002350797 cited in the application page 2386, right-hand column, paragraph 4 - last paragraph	1-27
Y	CLEVORN T ET AL: "ITERATIVE DECODING OF BICM WITH NON-REGULAR SIGNAL CONSTELLATION SETS" ITG FACHBERICHTE, VDE VERLAG, BERLIN, DE, no. 181, 2003, pages 259-266, XP001203424 ISSN: 0932-6022 cited in the application sections II and IV	1-27
Y	BAUCH G A: "Higher order differential matrix modulation" INFORMATION THEORY, 2004. ISIT 2004. PROCEEDINGS. INTERNATIONAL SYMPOSIUM ON CHICAGO, ILLINOIS, USA JUNE 27-JULY 2, 2004, PISCATAWAY, NJ, USA, IEEE, 27 June 2004 (2004-06-27), pages 189-189, XP010749900 ISBN: 0-7803-8280-3 cited in the application abstract section I	15-21
A	CLEVORN T ET AL: "PSK versus QAM for iterative decoding of bit-interleaved coded modulation" GLOBAL TELECOMMUNICATIONS CONFERENCE, 2004. GLOBECOM '04. IEEE DALLAS, TX, USA 29 NOV.-3 DEC., 2004, PISCATAWAY, NJ, USA, IEEE, vol. 1, 29 November 2004 (2004-11-29), pages 341-345, XP010758907 ISBN: 0-7803-8794-5 section II	6-9