

Zero-Forcing MIMO-OFDM Detection and Detectors

Field of the Invention

The present invention relates to techniques useful in Multiple Input Multiple Output -
 5 Orthogonal Frequency Division Multiplexing (MIMO-OFDM) receivers, in particular to aspects of signal detection and decoding for use in MIMO-OFDM receivers.

Background of Invention

MIMO-OFDM systems are well-known for providing space division multiplex
 10 communication. They are extensively employed in high-speed wireless communications because the combination takes advantage of multipath propagation to increase throughput, range/coverage and reliability, and provides relatively low-cost implementation particularly for high data rate communications.

15 Figure 1 is a schematic basic view of a MIMO-OFDM system, as already known. On the transmitting side T, a data signal is processed by a MIMO-OFDM transmitter 10. Separate portions or symbols of the processed signal are transmitted in parallel through multiple (N_T) spatially-separated transmitting antennas ($1, 2, \dots, N_t, \dots, N_{T-1}, N_T$), at the same carrier frequency. On the receiving side R, the signals from the transmitting
 20 antennas are received by multiple (N_R) spatially-separated receiving antennas ($1, 2, \dots, N_r, \dots, N_{R-1}, N_R$). The received signals are processed by a MIMO-OFDM receiver 20, ideally outputting the same signal as was input to the transmitter 10.

Typical processing on the transmission side T includes encoding, interleaving, mapping, modulating the signal by inverse FFT and the addition of a cyclic prefix onto each symbol prior to processing. The cyclic prefix acts as a guard space between consecutive symbols transmitted over the same transmitting antenna, to mitigate for the effects of
 5 delays in transmissions over the various channels. Pilot symbols may also be introduced to aid in channel estimation at the receiver side R.

The signal received at each receiving antenna is a combination of all the transmitted signals, with each transmitted signal received several times over with varying delays
 10 due to reflections. Sometimes the signals add up, other times they are out of phase, causing fading.

In a MIMO-OFDM based system, the signals can be expressed in the frequency domain using relatively straightforward matrix algebra. Considering a $N_T \times N_R$ MIMO-OFDM
 15 system, here N_T and N_R are the numbers of transmit and receiver antennas, respectively, the N_r received signal vector on the k -th subcarrier can be written as

$$\mathbf{y}(k) = \mathbf{H}(k)\mathbf{x}(k) + \mathbf{n}(k) \quad (1)$$

Here $\mathbf{y}(k) = (y_1(k), y_2(k), \dots, y_{N_r}(k))^T$ [O^T indicating a transposed vector or matrix as appropriate], which is the vector of received signals and $\mathbf{x}(k)$ is the N_t signal vector
 20 transmitted from the N_T transmit antennas on the k -th subcarrier, which is written as

$$\mathbf{x}(k) = (x_1(k), x_2(k), \dots, x_{N_t}(k))^T$$

The channel frequency response of the k -th subcarrier from the t -th transmit antenna to the r -th receiver antenna is denoted by $H_{r,t}(k)$, and

$$\mathbf{H}(k) = \begin{pmatrix} H_{1,1}(k) & H_{1,2}(k) & \dots & H_{1,N_t}(k) \\ H_{2,1}(k) & H_{2,2}(k) & \dots & H_{2,N_t}(k) \\ \vdots & \vdots & \dots & \vdots \\ H_{N_r,1}(k) & H_{N_r,2}(k) & \dots & H_{N_r,N_t}(k) \end{pmatrix}$$

and

$\mathbf{n}(k) = (n_1(k), n_2(k), \dots, n_{N_r}(k))^T$ is a spatially independent Gaussian noise vector for N_r receivers.

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The channel frequency responses vary with the environment between the transmitting and receiving antennas, whether with changes due to moving physical entities, e.g. doors, windows, people, cars, etc. or even due to changes in air temperature, pressure and humidity.

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Figure 2 is a schematic view of a typical receiver 20 in an MIMO-OFDM system. The different channels received at the multiple (N_r) spatially-separated receiving antennas ($1, 2, \dots, N_r, \dots, N_{r-1}, N_r$) are separately pre-processed before being demodulated. The pre-processor 22 converts from analogue to digital, provides frame and symbol synchronization and removes the cyclic prefix. Demodulation is performed separately on each received channel by FFT blocks 24. The frequency domain signals leave the FFT blocks 24 sub-carrier by sub-carrier.

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The demodulated channel signals are passed to a correction and decoding block 30. A MIMO channel estimator 32 uses the incoming signals to estimate the functions or coefficients of the different channels, using prior training and the pilot symbols

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continuously added to the transmitted signals. Channel estimation outputs from the channel estimation module 32 are utilized by a MIMO detector 34 to separate out the different transmitted symbols.

- 5 A de-mapper 38 extracts the encoded data from the carrier signals. A data de-interleaver 40 recovers the interleaved encoded data. A decoder 42 decodes the data, to output the desired data, ideally the same as the data input to the transmitting side T.

MIMO detection is applied in the MIMO detector 34 to combat frequency selective
10 fading, and to separate and recover multiple spatial stream data. MIMO detection usually involves one of linear detection, maximal likelihood detection, and BLAST detection.

Linear detection is the simplest spatial multiplexing receiver process since it requires
15 only matrix multiplication to separate the streams. It is therefore of interest in practical systems, particularly those with a large number of transmitter and receiver antennas.

Known linear detection methods include zero-forcing (ZF) linear detection and Minimal Mean Square Error (MMSE) linear detection.

- 20 A linear receiver works as follows. An $N_T \times N_R$ matrix equalizer $\mathbf{G}$ (derived from the channel estimator 32) is applied to a received frequency domain OFDM symbol to obtain an estimate of the transmitted OFDM symbol,

$$\begin{aligned}\hat{\mathbf{x}}(k) &= \mathbf{G}(k)\mathbf{y}(k) \\ &= \mathbf{G}(k)\mathbf{H}(k)\mathbf{x}(k) + \mathbf{G}(k)\mathbf{n}(k)\end{aligned}\tag{2}$$

In equation (2), $\hat{\mathbf{x}}(k)$ denotes the estimate of the transmitted symbol vector, originally derived from training but then continuously updated, and the detection matrix $\mathbf{G}(k)$ can be computed according to a zero-forcing (ZF) or minimum mean square error (MMSE) criterion.

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The detection is performed assuming that $\mathbf{G}(k)\mathbf{H}(k) = \mathbf{I}$, where $\mathbf{I}$ is an identity matrix.

Thus non-zero cancellations in the off-diagonal terms contribute to additional noise.

Unlike a maximal likelihood receiver, the minimum distance of the received constellation is the same as that of the transmit constellation. The MIMO detection,

10 however, colours the noise; thus the noise power varies as a function of the channel.

According to the MMSE method, $\mathbf{G}(k)$ is calculated as in the following equation:

$$\mathbf{G}(k) = (\mathbf{H}^H(k)\mathbf{H}(k) + N_0 \times \mathbf{I})^{-1} \mathbf{H}^H(k) \quad (3)$$

where N_0 is noise spectral density, $(\)^H$ denotes the conjugate and transpose transform of

15 the matrix, and $(\)^{-1}$ denotes the matrix inverse. The post-detection SNR reliability

figure for a sub-stream m (spatially multiplexed transmit data stream) is given in the

following equation (4). Note that the k -th element on the diagonal of the inverse matrix

is used in the denominator.

$$SNR_m = \frac{1}{[\mathbf{H}^H(k)\mathbf{H}(k) + N_0 \mathbf{I}]_{mm}^{-1}} - 1 \quad (4)$$

20 where $[\cdot]_{mm}$ denote the element in the m -th row and m -column of the specific matrix and

$[\cdot]^{-1}$ denotes the matrix inverse. It can be seen from this that MMSE detection implies a

high complexity with matrix inverse and multi-fold matrix multiplications.

On the other hand, for a ZF (zero-forcing) receiver, $\mathbf{G}(k)$ is calculated as the pseudoinverse of $\mathbf{H}(k)$,

$$\mathbf{G}(k) = \mathbf{H}^+(k) = [\mathbf{H}^H(k)\mathbf{H}(k)]^{-1} \mathbf{H}^H(k) \quad (5)$$

5 and its post-detection SNR for a sub-stream m can be expressed as

$$SNR_m = \frac{1}{[\mathbf{H}^H(k)\mathbf{H}(k)]_{mm}^{-1}} \quad (6)$$

Comparing Eq.(3) and Eq.(4) with Eq.(5) and Eq.(6), it can be seen that ZF and MMSE detections have similar complexity. However, MMSE detection is more extensively applied in practice because it provides better detection performance, by avoiding the
10 problems of noise enhancement and matrix singularity that exist with ZF detection.

The complexity of MIMO detection is a major concern in designing MIMO-OFDM receivers, due to the huge computation involved in all kinds of MIMO detection algorithms. Linear ZF detection and MMSE detection are the simplest detection
15 methods. However, conventional ZF detection methods are seldom used in practice because MMSE detection has superior performance with comparable complexity. Moreover, conventional ZF detection faces the problems of noise enhancement and matrix singularity (which result in infeasible implementation due to zero division or non-existence of a matrix inverse).

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Another problem with OFDM systems is that some features are degraded in frequency selective fading channels. For example, some subcarriers suffer deep fades due to destructive combinations of the various reflections, while others are enhanced by

constructive addition. Coding and interleaving applied to OFDM can be a tool to average the local fadings over the whole signal bandwidth and over the time interleaving depth.

5 Moreover, MIMO-OFDM systems exploit multipath scattering to enhance detection performance. It is impossible for an Additive White Gaussian Noise (AWGN) channel to be decoupled into multiple SISO (single input single output) independent channels to increase the system capacity since these MIMO AWGN channels are spatial correlated, i.e., the channel matrix for each subcarrier is not full rank and is singular and its
10 determinant is almost zero. The richly scattering multipath decreases the correlation of the spatial correlation of MIMO fading channels, which makes it possible for MIMO to increase the capacity (the transmission within the capacity limit is feasible). Evidently, the reliability of the MIMO detection output streams also depends on the spatial property of the channel matrix.

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In an MIMO-OFDM receiver 20, the decoder 42 can be a Viterbi decoder. However, the performance of a Viterbi decoder is degraded if the reliability of received data differs from subcarrier to subcarrier, from spatial stream to spatial stream due to frequency selective fading and MIMO spatial correlation. Therefore, dynamic
20 estimation of the data reliability in each subcarrier position for all spatial streams is necessary. For a linear MIMO detection receiver it is the post-detection SNR that is the parameter of interest. The estimate of the transmitted symbol vector can be directly fed to the de-mapper 38 that produces soft bit information for the channel decoder 42 (e.g. Viterbi). However, the use of MIMO-OFDM and a frequency selective channel implies

that the reliability figures for each carrier have to be computed and combined with the soft bits before the channel decoder 42. Thus the post-detection SNR is used for this purpose.

5 **Summary of the invention**

An aim of the present invention is to provide a simplified ZF linear detection method with attractive low complexity. Additionally, or alternatively, post-detection SNR can be provided and combined with ZF detection and Viterbi decoding to alleviate, at least partially, one or more problems in ZF detection, noise enhancement and/or matrix
10 singularity.

The present invention is able to provide low cost Zero-Forcing MIMO-OFDM detection and detectors by using $\mathbf{G}(k) = \mathbf{H}^{-1}(k)$ in the zero-forcing equation $\hat{\mathbf{x}}(k) = \mathbf{G}(k)\mathbf{y}(k)$, especially for symmetric MIMO configurations. Optionally, for cases where $\det(\mathbf{H}(k))$
15 $= 0$ or is sufficiently close thereto, there is no detection and the output of the detector is zero, which may be input directly to a Viterbi decoder.

According to a first aspect of the invention, there is provided a Zero-Forcing MIMO detector for estimating symbols in a plurality of received signals on a plurality of
20 subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas. The detector is operable to estimate and output said symbols based on assuming a matrix $\mathbf{H}(k)$ to be square and not singular, where $\mathbf{H}(k)$ is a matrix of channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

The detector may further comprise means for determining $\det(\mathbf{H}'(k))$, where $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses; and means for outputting a first result, being an estimate of said symbols based on assuming the matrix $\mathbf{H}(k)$ to be square and
 5 not singular or outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.

The Zero-Forcing MIMO detector may be operable such that a detection output vector of estimated symbols $\hat{x}(k)$ for the k -th subcarrier can be represented as
 10 $\hat{x}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k)$,
 where $\mathbf{y}(k)$ is a signal vector of received signals, and $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses.

According to a second aspect of the invention, there is provided a Zero-Forcing
 15 MIMO detector for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas. The detector is operable to output a result corresponding to a detection output vector of estimated symbols $\hat{x}(k)$ for the k -th subcarrier that can be represented as

20 $\hat{x}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k)$,

where $\mathbf{y}(k)$ is a signal vector of received signals, and $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

The detector may further comprise means for determining $\det(\mathbf{H}'(k))$; and means for outputting a first result, being a result corresponding to the detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$, or for outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.

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In either detector of the first or second aspect, the output can be determined based on the determined $\det(\mathbf{H}'(k))$ relative to a threshold, with an output first result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is greater than the threshold and an output second result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is less than or equal to the threshold. Alternatively, the output first result from the MIMO detector may be if the determined $\det(\mathbf{H}'(k))$ is greater than or equal to the threshold and an output second result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is less than the threshold.

According to a third aspect of the invention, there is provided a Zero-Forcing MIMO detector of the first or second aspect, in combination with a de-mapper and a Viterbi decoder. The MIMO detector, the de-mapper and the Viterbi decoder are arranged such that: an output first result from the MIMO detector is de-mapped by the de-mapper prior to decoding by the Viterbi decoder; and an output second result from the MIMO detector is input to the Viterbi decoder without first being de-mapped by the de-mapper.

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According to a fourth aspect of the invention, there is provided a Zero-Forcing MIMO detector for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas. The channels comprise frequency

responses for different subcarriers; and a detection output vector of estimated symbols

$\hat{x}_m(k)$, for a sub-stream m of the k -th subcarrier, can be represented as

$$\hat{x}_m(k) = \frac{\det(\mathbf{H}'(k))}{\left\| \mathbf{H}_m^p(k) \right\|^2 \text{SNR}_m(k)} \mathbf{H}_m^p(k) \mathbf{y}(k).$$

$\mathbf{y}(k)$ is a signal vector of received signals, $\mathbf{H}'(k)$ is a matrix of estimated channel

5 frequency responses for the k -th subcarrier from each of the transmitting antennas to a

second plurality of the receiving antennas; $\mathbf{H}_m^p(k)$ denotes the m -th row of $\mathbf{H}^p(k)$,

$\mathbf{H}^p(k)$ being the adjoint of $\mathbf{H}(k)$; and $\text{SNR}_m(k)$ is a SNR reliability figure for sub-

stream m for the k -th subcarrier.

10 According to a fifth aspect of the invention, there is provided a SNR calculator for use
in a MIMO receiver for receiving a plurality of signals on a plurality of subcarriers,
transmitted over a plurality of channels from a plurality of transmitting antennas to a
first plurality of receiving antennas. The SNR calculator is operable to output a SNR for
a sub-stream m

$$15 \quad \text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\left\| \mathbf{H}_m^p(k) \right\|^2}.$$

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from
each of the transmitting antennas to a second plurality of the receiving antennas; and

$\mathbf{H}_m^p(k)$ denotes the m -th row of $\mathbf{H}^p(k)$, $\mathbf{H}^p(k)$ being the adjoint of $\mathbf{H}(k)$; and

A Zero-Forcing MIMO detector of the earlier aspects may be used in combination with such a SNR calculator, where the output from the SNR calculator is an input to the MIMO detector.

- 5 A Zero-Forcing MIMO detector of the earlier aspects may be used in in combination with a SNR calculator of the fifth aspect, a de-mapper and a decoder. The MIMO detector, the SNR calculator, the de-mapper and the decoder are arranged such that: estimated outputs from the MIMO detector are input to the de-mapper for de-mapping; de-mapped outputs from the de-mapper are multiplied with the SNR outputs from the
- 10 SNR calculator; and the de-mapped outputs multiplied with the SNR output are input to the de-coder for decoding.

- According to a sixth aspect of the invention, there is provided a MIMO receiver comprising a MIMO detector of one of the above aspects, preferably also with a SNR
- 15 calculator of the fifth aspect.

- According to a seventh aspect of the invention, there is provided a method of Zero-Forcing MIMO detection for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of
- 20 transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers. The method comprises estimating and outputting said symbols based on assuming a matrix $\mathbf{H}(k)$ to be square and not singular, where $\mathbf{H}(k)$ is a matrix of channel frequency responses for the k -th

subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

According to an eighth aspect of the invention, there is provided a Zero-Forcing MIMO
 5 detection method for estimating symbols in a plurality of received signals on a plurality
 of subcarriers, transmitted over a plurality of channels from a plurality of transmitting
 antennas to a first plurality of receiving antennas, wherein the channels comprise
 frequency responses for different subcarriers. The method comprises: determining and
 outputting a result corresponding to a detection output vector of estimated symbols $\hat{x}(k)$
 10 for the k -th subcarrier that can be represented as $\hat{x}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k)$. $\mathbf{y}(k)$ is a signal
 vector of received signals, and $\mathbf{H}'(k)$ is a matrix of estimated channel frequency
 responses for the k -th subcarrier from each of the transmitting antennas to a second
 plurality of the receiving antennas.

15 According to a ninth aspect of the invention, there is provided a Zero-Forcing MIMO
 detection method for estimating symbols in a plurality of received signals on a plurality
 of subcarriers, transmitted over a plurality of channels from a plurality of transmitting
 antennas to a first plurality of receiving antennas, wherein the channels comprise
 frequency responses for different subcarriers. The method comprises: representing a
 20 detection output vector of estimated symbols $\hat{x}_m(k)$, for a sub-stream m of the k -th
 subcarrier, as

$$\hat{x}_m(k) = \frac{\det(\mathbf{H}'(k))}{\left\| \mathbf{H}'_m^p(k) \right\|^2 \text{SNR}_m(k)} \mathbf{H}'_m^p(k) \mathbf{y}(k).$$

$y(k)$ is a signal vector of received signals, $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; $H_m^P(k)$ denotes the m -th row of $H^P(k)$, $H^P(k)$ being the adjoint of $\mathbf{H}(k)$; and $\text{SNR}_m(k)$ is a SNR reliability figure for sub-stream m for the k -th subcarrier.

According to a tenth aspect of the invention, there is provided a SNR calculation method for use in a MIMO receiver for receiving a plurality of signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas. The method comprises calculating and outputting a SNR for a sub-stream m

$$\text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\|H_m^P(k)\|^2}.$$

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and $H_m^P(k)$ denotes the m -th row of $H^P(k)$, $H^P(k)$ being the adjoint of $\mathbf{H}(k)$; and

The present invention provides a ZF linear detection method which, in some circumstances at least has much lower complexity than conventional MMSE methods. This can be advantageously combined with post-detection SNR of the ZF detection.

Another aspect of the invention combines the ZF detection and post-detection SNR aided Viterbi decoding, for various modulation schemes.

The proposed combination of post-detection SNR processing and a simple ZF MIMO
5 detection processing is particularly useful, as it makes the proposed ZF method more feasible. This contrasts with the prior art, where such a ZF detection method would be generally infeasible, and explains in large part why some of the simple methods proposed herein did not occur in the prior art.

10 Preferably all the above aspects are used in or for an OFDM-MIMO system

Brief description of the drawings

The present is described herein by way of non-limitative example, with reference to the accompanying drawings, in which:-

- 15 Figure 1 is a schematic basic view of a MIMO-OFDM system, as already known;
Figure 2 is a schematic view of a typical receiver in an MIMO-OFDM system;
Figure 3 is a schematic view of a correction and decoding block according to a first embodiment of the invention;
Figure 4 is a flowchart relating to operation of the decoding block of Figure 3;
20 Figure 5 is a schematic view of a correction and decoding block according to a second embodiment of the invention;
Figure 6 is a flowchart relating to operation of the decoding block of Figure 5;
Figure 7 is a schematic view of a correction and decoding block according to a third embodiment of the invention;

- Figure 8 is a flowchart relating to operation of the decoding block of Figure 7;
 Figure 9 is a flowchart relating to operation of the MIMO detector of Figure 7;
 Figure 10 is a graph of the PER performance of ZF detections, using
 embodiments of the present invention, in Channel B;
 5 Figure 11 is a graph of the BER performance of ZF detections, using
 embodiments of the present invention, in Channel B;
 Figure 12 is a graph of the PER performance of ZF detections, using
 embodiments of the present invention, in Channel E; and
 Figure 13 is a graph of the BER performance of ZF detections, using
 10 embodiments of the present invention, in Channel E.

Description

- For the particular case where $N_T = N_R$ and assuming that $\mathbf{H}(k)$ is square and not singular,
 15 the pseudoinverse of $\mathbf{H}(k)$ is an expensive way to the implementation of ZF MIMO
 detection, whose complexity is almost same as that of the MMSE MIMO detection.
 However, for such a situation, the detection matrix can be simplified as:

$$\mathbf{G}(k) = \mathbf{H}^{-1}(k)$$

Then the detection output is represented, from Eq.(2) above as,

$$\begin{aligned} \hat{\mathbf{x}}(k) &= \mathbf{G}(k)\mathbf{y}(k) \\ &= \mathbf{H}^{-1}(k)\mathbf{H}(k)\mathbf{x}(k) + \mathbf{H}^{-1}(k)\mathbf{n}(k) \\ &= \mathbf{x}(k) + \mathbf{H}^{-1}(k)\mathbf{n}(k) \end{aligned} \tag{7}$$

and the post-detection SNR for sub-stream m can be expressed as,

$$SNR_m = \frac{1}{\|\mathbf{H}_m^{-1}(k)\|^2} = \frac{\det(\mathbf{H}(k))^2}{\|\mathbf{H}_m^p(k)\|^2} \quad (8)$$

where $\mathbf{H}_m^{-1}(k)$ denotes the m -th row of $\mathbf{H}^{-1}(k)$, and $\mathbf{H}^p(k)$ is the adjoint of $\mathbf{H}(k)$, and

$$\det(\mathbf{H}(k))\mathbf{I} = \mathbf{H}(k)\mathbf{H}^p(k), \text{ since } \mathbf{H}^{-1}(k) = \frac{\mathbf{H}^p(k)}{\det(\mathbf{H}(k))}.$$

- 5 Using this approach, two matrix multiplications can be saved in MIMO detection. Thus the complexity of the MIMO detection is greatly decreased. Moreover, in the two-dimension post-detection SNR calculation, another matrix multiplication is saved. This implies a significant complexity reduction.

10 *First Embodiment*

Figure 3 is a schematic view of a correction and decoding block 130 according to a first embodiment of the invention. Figure 4 is a flowchart relating to the operation of the block 130.

- 15 The correction and decoding block 130 of Figure 3 can be used in a MIMO-OFDM system, replacing the known correction and decoding block 20 of Figure 2. Preferably, the system has a symmetric MIMO configuration, that is $N_T = N_R$.

- After the received signal streams are converted to the frequency domain by FFT
 20 transform, as in the circuit of Figure 2, they are passed to the correction and decoding block 130 of Figure 3. In the correction and decoding block 130, the transformed streams are passed to the MIMO channel estimation module 32, to estimate the channels

(step S102), and the MIMO detection module 134. Channel estimation outputs from the channel estimation module 32, as derived according to the prior art, are utilized by a MIMO detector 134 and a post-detection two-dimension SNR calculation module 136. The SNR calculator 136 calculates the SNR using a process based on applying Eq.(8) above (step S104). The MIMO detector 134 detects the different symbols using a process based on applying Eq.(7) above (step S106). The detected outputs from the MIMO detector 134 pass through the de-mapper 38 for de-mapping (step S108) and data de-interleaver 40 for de-interleaving (step S110). An SNR de-interleaver 44 recovers the sequence of the SNR from the SNR calculator 136. The recovered data from the data de-interleaver 40 and the recovered SNR from the SNR de-interleaver 44 are multiplied to provide a combined reliability (step S112) as the input to the decoder 42, preferably a Viterbi decoder, which decodes the data (step S114).

In step S106, the output of the MIMO detector 134 is

$$\hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k)$$

with $\mathbf{H}'$ being the matrix of estimates from the MIMO channel estimation module 32.

The soft-decision output of the de-mapper 38 is the signed reliability of the corresponding received bit. The two-dimension post-detection SNR for each symbol in the specific spatial stream is multiplied with this soft decision. The obtained result is the combined reliability and input to the Viterbi decoder 42.

Second Embodiment

The above embodiment of Figures 3 and 4 is based on a symmetric MIMO configuration, i.e., $N_T = N_R$, and assumes that $\mathbf{H}(k)$ is a full rank matrix. However, in

practice $\mathbf{H}(k)$ may be singular or may not be square. Especially when $\mathbf{H}(k)$ is singular, $\mathbf{H}^{-1}(k)$ does not exist. Moreover, in this case, the antenna is closely spatial correlated. Even the regular ZF MIMO detection may not applicable any more because it faces the problems of noise enhancement and matrix singularity. Such disadvantages can be at least partially overcome by way of the embodiment described below with reference to Figures 5 and 6.

If the modulation method for the transmitted symbols from the transmission side T is BPSK or QPSK or the like, then the function of a BPSK/QPSK de-mapper can be denoted as $M(\cdot)$, and its output soft reliability can be written as,

$$M(\hat{\mathbf{x}}_m(k)) = \begin{cases} \{\text{Re}(\hat{\mathbf{x}}_m(k)), \text{Im}(\hat{\mathbf{x}}_m(k))\} & QPSK \\ \text{Re}(\hat{\mathbf{x}}_m(k)) & BPSK \end{cases} \quad (9)$$

Evidently it is a linear function of the input $\hat{\mathbf{x}}_m(k)$. The post detection SNR for the k th subcarrier $SNR_m(k)$ is applicable for all output soft-decision bits of the de-mapper.

Thus, the combined reliability of each bit input to a Viterbi decoder satisfies the following relationship,

$$SNR_m(k) * M(\hat{\mathbf{x}}_m(k)) = M(SNR_m(k) * \hat{\mathbf{x}}_m(k)) \quad (10)$$

Bearing in mind the property of Eq.(10), above, the embodiment of Figures 5 and 6 combines ZF MIMO detection and post-detection SNR before a de-mapper, instead of after the de-mapper,

$$\begin{aligned}
(SNR_m(k) * \hat{\mathbf{x}}_m(k)) &= \frac{\det(\mathbf{H}(k))^2}{\|\mathbf{H}_m^p(k)\|^2} (\mathbf{H}^{-1}(k) \mathbf{y}(k))_m \\
&= \frac{\det(\mathbf{H}(k))^2}{\|\mathbf{H}_m^p(k)\|^2} \frac{\mathbf{H}_m^p(k)}{\det(\mathbf{H}(k))} \mathbf{y}(k) \\
&= \frac{\det(\mathbf{H}(k))}{\|\mathbf{H}_m^p(k)\|^2} \mathbf{H}_m^p(k) \mathbf{y}(k)
\end{aligned} \tag{11}$$

Even when $\mathbf{H}^{-1}(k)$ does not exist due to $\det(\mathbf{H}(k)) = 0$, $\mathbf{H}^p(k)$ still exists. Thus ZF
MIMO detection based on applying Eq.(7) above is still applicable. Further, in the
5 scenario of $\det(\mathbf{H}(k)) \rightarrow 0$, the noise enhancement is avoided because it is removed by
higher order of $\det(\mathbf{H}(k))$ in the post-detection SNR. Thus noise enhancement and
matrix singularity can be dealt with at the same time.

Figure 5 is a schematic view of a correction and decoding block 230 according to a
10 second embodiment of the invention. Figure 6 is a flowchart relating to the operation of
the block 230.

The correction and decoding block 230 of Figure 5 can be used in a MIMO-OFDM
system, replacing the known correction and decoding block 20 of Figure 2. Preferably,
15 the system has a symmetric MIMO configuration, that is $N_T = N_R$. This is an
implementation for a MIMO-OFDM transceiver with a linear modulation scheme, such
as BPSK or QPSK.

After the received signal streams are converted to the frequency domain by FFT transform, as in the circuit of Figure 2, they are passed to the correction and decoding block 230 of Figure 5. In the correction and decoding block 230, the transformed streams are passed to the MIMO channel estimation module 32, for channel estimation (step S202) and the MIMO detection module 234. Channel estimation outputs from the channel estimation module 32, as derived according to the prior art, are utilized by a post-detection two-dimension SNR calculation module 136 and the MIMO detector 234. The SNR calculator 136 calculates the SNR using a process based on applying Eq.(8) above (step S204). The SNR from the SNR calculator 136 is also input to the MIMO detector 234.

In the MIMO detector 234 the outputs from the MIMO channel estimation module 32 are multiplied with the SNR output from the SNR calculator 136 to provide a combined reliability. The reliability combination is implemented and integrated with ZF detection. The integration is implemented based on Eq.(11) above (step S206). The output of the MIMO detector 234 is

$$\hat{\mathbf{x}}_m(k) = \frac{\det(\mathbf{H}'(k))}{\|\mathbf{H}'_m(k)\|^2 SNR_m(k)} \mathbf{H}'_m(k) \mathbf{y}(k),$$

with $\mathbf{H}'$ being the matrix of estimates from the MIMO channel estimation module 32.

The detected outputs from the MIMO detector 234 pass through a linear de-mapper 238 for de-mapping (step S208) and a data and SNR de-interleaver 240 for de-interleaving (step S210) before entering the decoder 42 for decoding (step S214).

Third Embodiment

The above-described second embodiment reduces potential problems of the first embodiment for applications of the system with a linear de-mapper. However, when the modulation method is 16QAM or 64QAM or otherwise, such that the de-mapper does not have the mentioned linear property, and $\mathbf{H}^{-1}(k)$ does not exist due to $\det(\mathbf{H}(k)) = 0$, use of Eq.(7) in simplified ZF MIMO detection is suspect. The third embodiment solves this problem and it is a general implementation method of the proposed simplified ZF detection, applicable for all symmetric MIMO configuration systems.

When a MIMO fading channel is closely spatially correlated, $\det(\mathbf{H}(k)) \rightarrow 0$ and $\mathbf{H}^{-1}(k)$ does not exist. This means that all output of the MIMO detection is unreliable and meaningless, i.e., $SNR_m(k) = 0$. However, in this case, the erasing technique in a Viterbi decoder can be utilized and $\hat{\mathbf{x}}(k) = \mathbf{G}(k)\mathbf{y}(k) = 0$ can be enforced, namely there is no operation of MIMO detection.

On the other hand, when $\det(\mathbf{H}(k)) \neq 0$, the normal simplified ZF MIMO detection using Eq.(7) above can be employed. Thus in a full scenario, the input to a Viterbi decoder in a third embodiment can be expressed as,

$$\begin{aligned}
 V_m(k) &= SNR_m(k) * M(\hat{\mathbf{x}}_m(k)) \\
 &= \begin{cases} \frac{\det(\mathbf{H}(k))^2}{\|\mathbf{H}_m^p(k)\|^2} M(\mathbf{H}_m^{-1}(k)\mathbf{y}(k)) & \det(\mathbf{H}(k)) \neq 0 \\ 0 & \det(\mathbf{H}(k)) = 0 \end{cases} \quad (12)
 \end{aligned}$$

Using this embodiment for nonlinear de-mappers, in the case of $\det(\mathbf{H}(k)) \neq 0$, the complexity is same as the first embodiment, but much lower than for conventional ZF MIMO detection. When $\det(\mathbf{H}(k)) = 0$, the processes of MIMO detection, de-mapping, and reliability combination can be eliminated, which bring a great benefit in power saving.

This implementation can be further extended to the case of $\det(\mathbf{H}(k)) \rightarrow 0$,

$$V_m(k) = SNR_m(k) * M(\hat{\mathbf{x}}_m(k))$$

$$= \begin{cases} \frac{\det(\mathbf{H}(k))^2}{\|\mathbf{H}_m^p(k)\|^2} M(\mathbf{H}_m^{-1}(k)\mathbf{y}(k)) & \det(\mathbf{H}(k)) > Threshold \\ 0 & \det(\mathbf{H}(k)) \leq Threshold \end{cases} \quad (13)$$

Alternatively, for non-zero thresholds, the estimated result can be output if $\det(\mathbf{H}(k)) \geq$ Threshold, and the zero result if $\det(\mathbf{H}(k)) < Threshold$.

With a reasonable threshold setting, this implementation further decreases the process burden with negligible performance loss.

Figure 7 is a schematic view of a correction and decoding block 330 according to a third embodiment of the invention. Figure 8 is a flowchart relating to the operation of the block 330.

The correction and decoding block 330 of Figure 7 can be used in a MIMO-OFDM system, replacing the known correction and decoding block 20 of Figure 2. Preferably, the system has a symmetric MIMO configuration, that is $N_T = N_R$. This is an

implementation for a MIMO-OFDM transceiver with a non-linear modulation scheme, such as 16QAM/64QAM (although it can also be used with a linear modulation scheme, but with some performance loss).

5 After the received signal streams are converted to the frequency domain by FFT transform, as in the circuit of Figure 2, they are passed to the correction and decoding block 330 of Figure 7. In the correction and decoding block 330, the transformed streams are passed to the MIMO channel estimation module 32, for channel estimation (step S302) and the MIMO detection module 334. Channel estimation outputs from the
10 channel estimation module 32, as derived according to the prior art are utilized by a post-detection two-dimension SNR calculation module 136 and the MIMO detector 334. The SNR calculator 136 calculates the SNR using a process based on applying Eq.(8) above (step S304). The MIMO detector 334 detects and outputs the different symbols or other results, as described below (step S306).

15 In effect, the MIMO detector 334 has two possible alternative outputs and outputs the appropriate one (step S306). The first output passes from the MIMO detector 334 through a de-mapper 338 (generally non-linear for this embodiment), for de-mapping (step S308), and data de-interleaver 40, for de-interleaving (step S310). An SNR de-
20 interleaver 44 recovers the sequence of the SNR from the SNR calculator 136. The recovered data from the data de-interleaver 40 and the recovered SNR from the SNR de-interleaver 44 are multiplied to provide a combined reliability (step S312) as the input to a Viterbi decoder 342. The second output passes straight from the MIMO detector 334

to the Viterbi decoder 342. The input data to the Viterbi decoder 342 is decoded (step S314).

Figure 9 is a flowchart relating to the operation of the MIMO detector 334, that is a more detailed operation of step S306 of Figure 8. To decide which output is used, the MIMO detector 334 determines $\det(\mathbf{H}(k))$ (step S350) and compares the result against a threshold (step S352). The threshold may be zero or close thereto. When $\det(\mathbf{H}(k))$ is above the relevant erase threshold, then processing occurs in the same manner as according to the first embodiment. That is the MIMO detector 334 detects the different symbols using a process based on applying Eq.(7) above (step S354). The output of the MIMO detector 334 would be

$$\hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k)$$

with $\mathbf{H}'$ being the matrix of estimates from the MIMO channel estimation module 32.

The results of this detection are provided as the first output as mentioned above, passing to the de-mapper 338.

On the other hand, when $\det(\mathbf{H}(k)) = 0$ or is below or is equal to the relevant erase threshold, if set, then MIMO detection, de-mapping, de-interleaving and the combined reliability calculation are bypassed (or possibly disabled) (step S360), and zeros are directly input (step S362) to the Viterbi decoder 342. An input of zeros means the higher puncturing rate for the codec, namely an erasing effect, pending state. The MIMO detection is performed symbol by symbol, and with streams in parallel. Thus the number of zeros input to the decoder can easily be predetermined, depending on the modulation scheme and number of streams.

In the embodiment of Figure 7, multiplication of the data and SNR occurs after de-interleaving. This is generally for the case where the demapper function is not linear (for example, 16QAM/64QAM). However, where the demapper function is linear (for example, BPSK/QPSK), multiplication before and after de-interleaving is the same. Thus it can be performed before de-interleaving, as in the embodiment of Figure 5, to save separate de-interleaving for SNR.

In the above described embodiments, the SNR has been used, in effect, as channel state information (CSI) by the decoder. However, other sources of CSI can be used, for instance fading path delay spread, path numbers, etc.

In an OFDM system with a frequency-selective fading channel, the SNR also varies subcarrier by subcarrier. Embodiments make use of the SNR for a specific subcarrier in a specific data stream after MIMO detection.

The codes used in the above described embodiments are exemplified by Viterbi codes, decoded by a Viterbi decoder. However, other types of codes (and decoder) can also be used, for example, but not limited to: all other FEC codec, including all convolutional codes, Turbo codes, and LDPC codes.

Whilst the rationale behind Eq.(7) and Eq.(8) above, as used in all three embodiments, is that the MIMO system has a symmetric configuration (i.e. $N_T = N_R$, for instance in 2*2 MIMO and 4*4 MIMO), that does not mean the present invention can only be used

in symmetric configurations. The invention can be used with asymmetric configurations, where the number of transmitted data streams is not **[Inventor – please confirm this “not” is correct]** the same as the number of the receiving antennas, e.g., 3*2, or 4*2 MIMO configurations with two spatial data streams, or others, where cyclic
 5 delay diversity (CDD) or transmitter beamforming technique is used for spatial expansion.

Whilst the above-described embodiments are all based on MIMO-OFDM systems, the present invention also extends to other narrow-band, frequency non-selective fading
 10 MIMO systems.

Simulation and Verification

The above described embodiments were simulated and verified. The receivers were
 15 simulated as used in a 2*2 MIMO-OFDM based WLAN system platform. In the simulation, two spatial streams were transmitted over a MIMO fading channel with a 20MHz frequency band, utilizing the spatial division multiplexing (SDM) technique. Convolutional codes with different puncturing rates were adopted. The detailed system specification is referred to the WLAN 802.11n standard draft, “EWC HT PHY
 20 Specification”, *Enhanced Wireless Consortium Publication*. V1.27, December 23rd, 2005. Type B and type E MIMO fading channel models, as described in “IEEE P802.11 Wireless LANs TGN Channel Models”, *doc.: IEEE 802.11-03/940r4*, May 2004, were employed in the system simulation.

The antenna configuration in the simulation was a linear array and the distance between adjacent two antennas was a half carrier wavelength. In the considered system, perfect synchronization was assumed. Least square MIMO channel estimation without smoothing was adopted in the receiver for the Channel Estimator 32. Moreover, the
 5 decoder 42, 342 used Viterbi decoding with trace-back length of 128.

In the simulation, the simplified ZF detection based on the use of Eq.(7) and its implementation variations were verified.

10 The first embodiment, under the condition that the channel matrix was not singular had the same performance as the conventional ZF detection method.

The second embodiment, under the condition that the de-mapper function has a linear property had same performance as the conventional ZF detection method.

15

The third embodiment is the general implementation method of the proposed simplified ZF detection. It is applicable for all symmetric MIMO detection MIMO-OFDM systems with linear or nonlinear modulation schemes. Its performance varied with the threshold of the implementation. The simulations were performed in various conditions to
 20 demonstrate this, including different modulation schemes, BPSK and 64QAM, Channel B and channel E, the erase threshold being 0, 0.1, and 0.01. The simulation results are shown in Figures 10 to 13. Figures 10 and 11 show the PER and BER performance, respectively, in channel B and Figures 12 and 13 show the PER and BER performance, respectively, in channel E. In all the Figures 10 to 13, simplified ZF detection, based on

Eq.(7), with different erase thresholds were adopted for BPSK and 64QAM. From these simulation results it can be seen that the performance for BSPK is robust for these threshold settings. When 64QAM is considered, the performances were almost same with the threshold of 0 and 0.01. However, when the threshold increased further to 0.1, there was obvious performance degradation for 64QAM.

Further, a comparison is now made between the number of calculations made during use of various embodiments of the present invention, relative to those made during operation of conventional ZF or MMSE.

10

Estimates of the required multiplications for the proposed methods are based on the following assumptions:

- Complex matrix multiplication $A_{L \times M} \times B_{M \times N}$ requires $L \times M \times N$ complex scalar multiplications;
- 15 • The matrix inversion operation of matrix $A_{N \times N}$ takes $O(N^3)$ multiplications.
- Each complex multiplication consists of 4 real ones, and each real multiplication takes one operation to finish
- The adjoint calculation of a 2×2 matrix involves no complex multiplication.
- The matrix inversion of a 2×2 matrix involves 6 complex multiplications.

20

Assuming a 2×2 MIMO system, the corresponding results (including the SNR calculation) based on these assumptions are given in Table I. From the comparisons, it can be seen that the complexity of ZF detection for symmetric MIMO configuration,

using the present invention, is much lower than for conventional ZF detection and MMSE detection. The number of multiplications involved in the simplified ZF detection for linear de-mappers are less half that for conventional ZF/MMSE detections. The complexity of the general ZF detection of the third embodiment depends on the

5 erasion rate of the employed ZF detection, $p\%$.

Table I, complexity comparison

Detection method	Real multiplications
Conventional ZF/MMSE	104 (excluding the combined reliability calculation)
The invented ZF detection for linear de-mapper (second embodiment)	44 (including the combined reliability calculation)
The invented general ZF detection (third embodiment)	$48 \cdot p\%$ (excluding the combined reliability calculation)

Embodiments of the invention may be implemented, inter alia, using just hardware, with

10 circuits dedicated for the required processing, by software or by a combination of hardware and software modules, even if the hardware is just that of a conventional computer system.

A module, and in particular the module's functionality, can be implemented in either

15 hardware or software. In the software sense, a module is a process, program, or portion

thereof, that usually performs a particular function or related functions. In the hardware sense, a module is a functional hardware unit designed for use with other components or modules. For example, a module may be implemented using discrete electronic components, or it can form a portion of an entire electronic circuit such as an
5 application specific integrated circuit (ASIC). Numerous other possibilities exist.

Other embodiments of the invention will be apparent to those skilled in the art from consideration of the specification and practice of the invention. Furthermore, certain terminology has been used for the purposes of descriptive clarity, and not to limit the
10 present invention. The embodiments and preferred features described above should be considered exemplary, with the invention being defined by the appended claims and/or as described.

Claims

1. A Zero-Forcing MIMO detector for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers; and the detector is operable to estimate and output said symbols based on assuming a matrix $\mathbf{H}(k)$ to be square and not singular, where $\mathbf{H}(k)$ is a matrix of channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

2. A Zero-Forcing MIMO detector according to claim 1 further comprising:
means for determining $\det(\mathbf{H}'(k))$, where $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses; and
means for outputting a first result, being an estimate of said symbols based on assuming the matrix $\mathbf{H}(k)$ to be square and not singular or outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.

3. A Zero-Forcing MIMO detector according to claim 1 or 2, operable such that a detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$ for the k -th subcarrier can be represented as

$$\hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k),$$

where

$\mathbf{y}(k)$ is a signal vector of received signals, and

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses.

4. A Zero-Forcing MIMO detector for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers; and the detector is operable to output a result corresponding to a detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$ for the k -th subcarrier that can be represented as
- $$\hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k),$$

10 where

$\mathbf{y}(k)$ is a signal vector of received signals, and

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

15

5. A Zero-Forcing MIMO detector according to claim 4 further comprising:
means for determining $\det(\mathbf{H}'(k))$; and
means for outputting a first result, being a result corresponding to the detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$, or for outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.
- 20

6. A Zero-Forcing MIMO detector according to claim 2 or 5, wherein the output is determined based on the determined $\det(\mathbf{H}'(k))$ relative to a threshold, with an output first result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is greater than the

threshold and an output second result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is less than or equal to the threshold.

7. A Zero-Forcing MIMO detector according to claim 6, wherein the threshold = 0.

5

8. A Zero-Forcing MIMO detector according to claim 2 or 5, wherein the output is determined based on the determined $\det(\mathbf{H}'(k))$ relative to a threshold, with an output first result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is greater than or equal to the threshold and an output second result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is less than the threshold.

10

9. A Zero-Forcing MIMO detector according to any one of claims 2 and 5 to 8, in combination with a de-mapper and a Viterbi decoder, wherein the MIMO detector, the de-mapper and the Viterbi decoder are arranged such that:

15 an output first result from the MIMO detector is de-mapped by the de-mapper prior to decoding by the Viterbi decoder; and

an output second result from the MIMO detector is input to the Viterbi decoder without first being de-mapped by the de-mapper.

20 10. A Zero-Forcing MIMO detector for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers; and

a detection output vector of estimated symbols $\hat{x}_m(k)$, for a sub-stream m of the k -th subcarrier, can be represented as

$$\hat{x}_m(k) = \frac{\det(\mathbf{H}'(k))}{\left\| \mathbf{H}_m^p(k) \right\|^2 \text{SNR}_m(k)} \mathbf{H}_m^p(k) \mathbf{y}(k),$$

where

5 $\mathbf{y}(k)$ is a signal vector of received signals,

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas;

$\mathbf{H}_m^p(k)$ denotes the m -th row of $\mathbf{H}^p(k)$, $\mathbf{H}^p(k)$ being the adjoint of $\mathbf{H}(k)$;

10 and

$\text{SNR}_m(k)$ is a SNR reliability figure for sub-stream m for the k -th subcarrier.

11. A Zero-Forcing MIMO detector according to claim 10 in combination with a linear de-mapper for de-mapping the output from the MIMO detector.

15

12. A Zero-Forcing MIMO detector according to claim 11 further in combination with a decoder, wherein the output from the linear de-mapper is input to the decoder.

13. A Zero-Forcing MIMO detector according to claim 10, 11 or 12 in combination

20 with a SNR calculator for use in a MIMO receiver for receiving the plurality of signals, wherein

the SNR calculator is operable to output a SNR for a sub-stream m

$$\text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\|\mathbf{H}'_m^p(k)\|^2}$$

where

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

$\mathbf{H}'_m^p(k)$ denotes the m -th row of $\mathbf{H}'^p(k)$, $\mathbf{H}'^p(k)$ being the adjoint of $\mathbf{H}(k)$; and

wherein the output from the SNR calculator is an input to the MIMO detector.

14. A Zero-Forcing MIMO detector according to any one of claims 1 to 8, in combination with a SNR calculator for use in a MIMO receiver for receiving the plurality of signals, a de-mapper and a decoder, wherein

the SNR calculator is operable to output a SNR for a sub-stream m

$$\text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\|\mathbf{H}'_m^p(k)\|^2}$$

where

15 $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

$\mathbf{H}'_m^p(k)$ denotes the m -th row of $\mathbf{H}'^p(k)$, $\mathbf{H}'^p(k)$ being the adjoint of $\mathbf{H}(k)$;

and wherein the MIMO detector, the SNR calculator, the de-mapper and the decoder are arranged such that:

estimated outputs from the MIMO detector are input to the de-mapper for de-mapping;

de-mapped outputs from the de-mapper are multiplied with the SNR outputs from the SNR calculator; and

5 the de-mapped outputs multiplied with the SNR output are input to the de-coder for decoding.

15. A Zero-Forcing MIMO detector according to claim 9 in combination with a SNR calculator for use in a MIMO receiver for receiving the plurality of signals ,

10 wherein

the SNR calculator is operable to output a SNR for a sub-stream m

$$\text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\|\mathbf{H}_m^p(k)\|^2}$$

where

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th

15 subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

$\mathbf{H}_m^p(k)$ denotes the m -th row of $\mathbf{H}^p(k)$, $\mathbf{H}^p(k)$ being the adjoint of $\mathbf{H}(k)$;

and wherein the de-mapped outputs from the de-mapper are multiplied with the SNR outputs from the SNR calculator prior to decoding by the Viterbi decoder.

20

16. A Zero-Forcing MIMO detector according to any preceding claim, wherein the second plurality of receiving antennas comprises the same number of receiving antennas as the number of transmitting antennas.
- 5 17. A Zero-Forcing MIMO detector according to claim 16, wherein the first plurality of receiving antennas comprises the same number of receiving antennas as the number of transmitting antennas.
18. A MIMO detector according to any preceding claim, being a MIMO-OFDM
10 detector.
19. A MIMO receiver comprising a MIMO detector according to any preceding claim.
- 15 20. A MIMO receiver according to claim 19, further comprising a SNR calculator, wherein
the SNR calculator is operable to output a SNR for a sub-stream m
- $$\text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\|\mathbf{H}'_m^p(k)\|^2}$$
- where
- 20 $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

$H_m^P(k)$ denotes the m -th row of $H^P(k)$, $H^P(k)$ being the adjoint of $\mathbf{H}(k)$.

21. A MIMO receiver comprising a SNR calculator for receiving a plurality of signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the SNR calculator is operable to output a SNR for a sub-stream m

$$\text{SNR}_m = \frac{\det(H'(k))^2}{\|H_m^P(k)\|^2}$$

where

- $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

$H_m^P(k)$ denotes the m -th row of $H^P(k)$, $H^P(k)$ being the adjoint of $\mathbf{H}(k)$;

22. A MIMO receiver according to any one of claims 19 to 21, further comprising a MIMO channel estimator.

23. A MIMO receiver according to claim 22, when dependent on at least one of claims 3, 4 10, wherein the output of the channel estimator corresponds to the contents of matrix $\mathbf{H}'(k)$.

24. A method of Zero-Forcing MIMO detection for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers, the
 5 method comprising

estimating and outputting said symbols based on assuming a matrix $\mathbf{H}(k)$ to be square and not singular, where $\mathbf{H}(k)$ is a matrix of channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

10

25. A Zero-Forcing MIMO detection method according to claim 24, further comprising:

determining $\det(\mathbf{H}'(k))$, where $\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses; and

15 outputting a first result, being an estimate of said symbols based on assuming the matrix $\mathbf{H}(k)$ to be square and not singular or outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.

26. A Zero-Forcing MIMO detection method according to claim 24 or 25, wherein a
 20 detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$ for the k -th subcarrier can be represented as

$$\hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k),$$

where

$\mathbf{y}(k)$ is a signal vector of received signals, and

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses.

27. A Zero-Forcing MIMO detection method for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers, the method comprising:

determining and outputting a result corresponding to a detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$ for the k -th subcarrier that can be represented as

$$10 \quad \hat{\mathbf{x}}(k) = \mathbf{H}'^{-1}(k)\mathbf{y}(k),$$

where

$\mathbf{y}(k)$ is a signal vector of received signals, and

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas.

28. A Zero-Forcing MIMO detection method according to claim 27, further comprising:

determining $\det(\mathbf{H}'(k))$; and

20 outputting a first result, being a result corresponding to the detection output vector of estimated symbols $\hat{\mathbf{x}}(k)$, or for outputting a second result, being a zero signal, based on the determined $\det(\mathbf{H}'(k))$.

29. A Zero-Forcing MIMO detection method according to claim 25 or 28, wherein the output is determined based on the determined $\det(\mathbf{H}'(k))$ relative to a threshold, with an output first result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is greater than the threshold and an output second result from the MIMO detector if the
- 5 determined $\det(\mathbf{H}'(k))$ is less than or equal to the threshold.
30. A Zero-Forcing MIMO detection method according to claim 29, wherein the threshold = 0.
- 10 31. A Zero-Forcing MIMO detection method according to claim 25 or 28, wherein the output is determined based on the determined $\det(\mathbf{H}'(k))$ relative to a threshold, with an output first result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is greater than or equal to the threshold and an output second result from the MIMO detector if the determined $\det(\mathbf{H}'(k))$ is less than the threshold.
- 15 32. A Zero-Forcing MIMO detection method for estimating symbols in a plurality of received signals on a plurality of subcarriers, transmitted over a plurality of channels from a plurality of transmitting antennas to a first plurality of receiving antennas, wherein the channels comprise frequency responses for different subcarriers, the
- 20 method comprising:
- representing a detection output vector of estimated symbols $\hat{\mathbf{x}}_m(k)$, for a sub-stream m of the k -th subcarrier, as

$$\hat{x}_m(k) = \frac{\det(\mathbf{H}'(k))}{\left\| \mathbf{H}'_m^p(k) \right\|^2 \text{SNR}_m(k)} \mathbf{H}'_m^p(k) y(k),$$

where

$y(k)$ is a signal vector of received signals,

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th

5 subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas;

$\mathbf{H}'_m^p(k)$ denotes the m -th row of $\mathbf{H}'^p(k)$, $\mathbf{H}'^p(k)$ being the adjoint of $\mathbf{H}(k)$;

and

$\text{SNR}_m(k)$ is a SNR reliability figure for sub-stream m for the k -th subcarrier.

10

33. A MIMO detection method according to any one of claims 24 to 32 in combination with an SNR calculation method for use in a MIMO receiver for receiving the plurality of signals, the SNR calculation method comprising calculating and outputting a SNR for a sub-stream m

$$15 \quad \text{SNR}_m = \frac{\det(\mathbf{H}'(k))^2}{\left\| \mathbf{H}'_m^p(k) \right\|^2}$$

where

$\mathbf{H}'(k)$ is a matrix of estimated channel frequency responses for the k -th

subcarrier from each of the transmitting antennas to a second plurality of the receiving antennas; and

20 $\mathbf{H}'_m^p(k)$ denotes the m -th row of $\mathbf{H}'^p(k)$, $\mathbf{H}'^p(k)$ being the adjoint of $\mathbf{H}(k)$.

34. A MIMO detection method according to any one of claims 24 to 33, being a MIMO-OFDM detection method.