



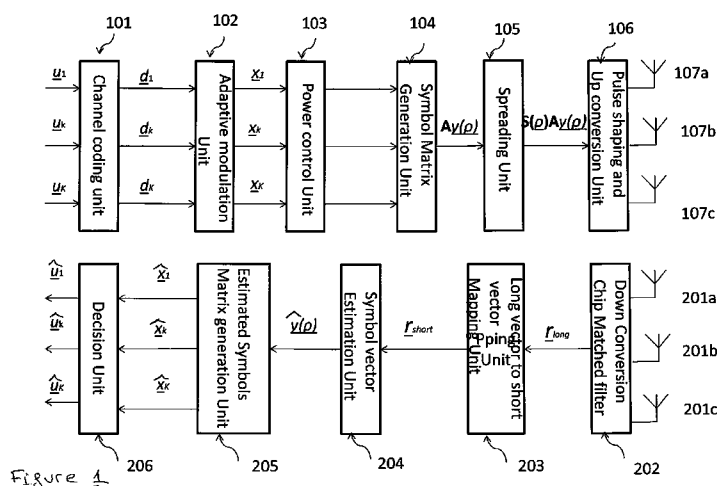
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(54) Title: DATA TRANSMISSION OPTIMISATION METHOD AND APPARATUS

(57) **Abstract:** A method for determining a transmitted signal received at a receiver in a radio transmission system is disclosed. The radio transmission system has a plurality of parallel single-input single-output or multiple-input multiple-output channels over which the signal is transmitted. The signal comprises one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis every symbol period p . The method comprises receiving a signal $r_{long}(p)$ corresponding to a signal transmitted over a channel, wherein each symbol of the transmitted signal is spread using a respective spreading sequence $S(p)$. The method also comprises determining an estimated channel characteristic H indicative of effects of the channel over which the received signal $r_{long}(p)$ has been transmitted on the transmitted signal. Furthermore, the method comprises determining the transmitted signal $r_{short}(p)$ from the received signal $r_{long}(p)$ in accordance with the estimated channel characteristic H and the spreading sequences used to spread the symbols. A method for determining data transmission rates for transmitting a signal over a radio transmission system is also disclosed. Apparatus and computer program products for performing such methods are also disclosed.

Data transmission optimisation method and apparatus

Field of Invention

The present invention relates to the field of mobile radio system data transmission. More specifically, but not exclusively, a method for determining a transmitted signal received at a receiver in a radio transmission system method is disclosed. In addition, a method for determining data transmission rates for transmitting a signal over a radio transmission system is disclosed.

Background to the Invention

Mobile radio system technologies are continuously advancing with the aim of increasing data rates. The third generation mobile radio system uses a code division multiple access (CDMA) transmission scheme and it has been extensively adopted worldwide. The third generation partnership project (3GPP) has developed the high speed down link packet access (HSDPA) system in the Release 5 specification of the Universal Mobile Telecommunications System (UMTS) as a multi-code wide-band CDMA system.

The success of the third generation wireless cellular systems is based largely on the efficient resource allocation scheme used by the HSDPA system to improve the downlink throughput. Multiple parallel channels are used to transmit data symbols. The downlink throughput optimization for the HSDPA multi-code MIMO CDMA system involves the use of the mean-square-error (MSE) minimizing receiver despreading filter coefficients. The data rate for the symbol transmitted in each channel is adjusted using a measure of the signal-to-noise ratio at the output of the MMSE despreading unit. Once the symbol rate is determined, adaptive modulation and coding methods are used to determine the type of modulation to be used over each parallel channel. The incoming data bits are then converted to the appropriate symbols for the chosen modulation type. Each symbol for each parallel channel is spread using spreading sequences and spread signals are added before being transmitted over the MIMO antennas using spatial multiplexing methods. The total throughput of the HSDPA MIMO system is given by the summation of the data rates transmitted in each parallel channel. Spatial multiplexing methods are then used to maximize the throughput in the HSDPA downlink MIMO transmission.

The number of chip sequences used to spread symbols determines the processing gain, N , for the spread sequence systems. The current standards for the HSDPA

system uses a processing gain of 16. For the processing gain of $N = 16$, the maximum number of orthogonal spreading sequences is 16. When using spatial multiplexing methods the number of spreading sequences is increased to $N \times \min(N_t, N_r)$ where N_t and N_r are the number of transmit and receive antennas. The upper bound of the spatial multiplexing gain of the HSDPA MIMO system is referred to as $\min(N_t, N_r)$. When the total throughput, R_T , for the HSDPA MIMO is normalized with respect to the $\log_2(SNR)$ at high SNR values the resultant measure $\frac{R_T}{\log_2(SNR)}$ is upper bounded by $\min(N_t, N_r)$, that is

$$\frac{R_T}{\log_2(SNR)} \leq \min(N_t, N_r).$$

The spatial multiplexing gain for the HSDPA MIMO system $\frac{R_T}{\log_2(SNR)}$ can be obtained by plotting the total rate against $\log_2(SNR)$ and measuring the slope of the capacity curve at high signal to noise ratios.

An approach was considered by 3GPP and a method was standardized to use a given fixed set size of Orthogonal Variable Spreading Factor (OVSF) spreading sequences for the processing gain of $N = 16$. To achieve the full multiplexing gain of the HSDPA MIMO system, a signature sequence set size equal to $N \times \min(N_t, N_r)$ is required. 3GPP standardized a method which increases the OVSF set size by multiplying the given set with precoding weights and then concatenating the weighted sets of spreading sequences to increase the signature sequence set size to $N \times \min(N_t, N_r)$.

Hence, a unique pre-coded spreading sequence is produced by concatenating the spreading sequences used at each antenna for each transmission symbol. At the transmitter, each transmission symbol is then spread before transmission with a different channelization spreading sequence at each MIMO antenna. These spreading sequences are known as the channelization spreading sequences. Each concatenated channelization spreading sequence is orthogonal to the remaining set of channelization spreading sequences, available at the transmitter for other transmission symbols. In addition to the channelization spreading sequences, a set of scrambling spreading sequences, specific to each downlink MIMO transceiver, is used to further scramble the spread symbols, which have already been spread using the channelization spreading sequences. The scrambling spreading sequence is

unique to each HSDPA downlink base-station and is organized to change at every symbol period.

The spreading sequence orthogonality is lost at the receiving end after transmission over frequency selective multipath channels. Therefore, it has been proposed that a linear MMSE equalizer followed by a de-spreader could be used to restore the orthogonality of the spreading sequences at each receiver and to recover the transmitted symbols after transmission over a multipath channel.

Recent developments have considered a self interference (SI) problem, which is present in linear MMSE equalizers, when operating over multipath channels. To improve the transmission throughput it is essential to reduce the large gap between the currently practical achievable rates and the theoretical upper bound for the HSDPA throughput. A receiver with an independent symbol level MMSE equalizer followed by a symbol level successive-interference-cancellation (SIC) scheme deals with the intra-cell self interference. It has been proposed that a hybrid linear equalizer/interference cancelling receiver tailored to the HSDPA standard could be utilized. Furthermore, it has been proposed that a SIC receiver with either a chip or a symbol level MMSE equalizer for the HSDPA downlink throughput optimization is used.

Instead of using a SIC based MMSE equalizer and de-spreader, if only a MMSE based equalizer and de-spreader is used at the MIMO receiver, the spatial multiplexing gain $\frac{R_T}{\log_2(SNR)}$ of the HSDPA MIMO can be much smaller than the spatial multiplexing upper bound $\min(N_t, N_r)$.

Introduction of a chip level MMSE linear equalizer followed by a de-spreader and a symbol level SIC is considered to suppress the inter-chip interference (ICI) and also all inter-stream interference. Furthermore, a channel matched filter (CMF) as a linear chip level MMSE equalizer has been shown to maximize the signal-to-noise ratio by collecting the energy at the multipath channel central tap. The chip level equalizer is used to produce an estimate of the transmitted chip sequence, which is then de-spread by one of the transmitter spreading sequences to detect one of the transmitted symbol streams. The recovered symbol is then used to remove the interference iteratively at chip level using a Successive Interference Cancellation

receiver. This work is reported in a paper entitled "Shenoy, Shakti Prasad, Irfan Ghauri, and Dirk TM Slock Optimal Precoding and MMSE receiver designs for WCDMA, In IEEE Vehicular Technology Conference, 2008, VTC Spring 2008, pp 893-897, IEEE 2008 However the throughput performance when expressed in terms of the spatial multiplexing gain $\frac{R_T}{\log_2(SNR)}$ could be significantly lower than the spatial multiplexing upper bound $\min(N_t, N_r)$ at high signal noise ratios.

The use of a receiver with the linear MMSE equalizer and a single stage SIC detector requires the joint optimization of the transmitter and receiver to improve the spatial multiplexing gain $\frac{R_T}{\log_2(SNR)}$ and hence maximize the downlink throughput. Various transmission power allocation schemes can be used for different data streams for a two stage successive interference cancellation scheme in multi-code MIMO systems. A two stage SIC detection scheme with the transmitter power optimization can improve the throughput performance for multi-code downlink transmission. However, for each iteration of the SIC, the equalizer coefficient and the power allocation calculations require an inversion of a covariance matrix for the received signal. The dimension of the covariance matrix is usually large and each of the iterative power allocation, the linear MMSE equalizer and the SIC implementations at the receiver therefore become computationally expensive. Simplifications for the inversion of large matrices has been examined to make the implementation of the linear MMSE equalizers followed by the symbol level SIC practically feasible.

It is also imperative to eliminate individual power allocations to each spreading sequence at the transmitter. It is important to use the equal energy and the equal rate allocation method for each symbol before spreading each symbol by using channelization and scrambling codes to maximize the spatial multiplexing gain of the HSDPA MIMO transceiver.

Various attempts have been made to optimize transceiver design. Usually, different optimization criteria are used when allocating powers for the multi-code downlink throughput optimization. Some techniques focus on the transceiver design optimization criteria and others concentrate on criteria for the joint rate and power allocation. One attempt to deal with these problems has been to generalise the joint rate and power adaptation methods as discussed in L. Zhao and J. Mark, " Joint rate

and power adaptation for radio resource management in uplink wideband code division multiple access systems,” IET Communications, vol. 2, no. 4, pp. 562 –572, April 2008, which generalises the joint rate and power adaptation method in three ways as follows:

1. The first criterion includes the systems which optimize the transmission power to maximize the rate for a given realization of channel gains. A typical example is L. Y. Hoon and K. S. Wu, “ Generalized joint power and rate adaptation in ds-cdma communications over fading channels,” IEEE Transactions on Vehicular Technology, vol. 57, no. 1, pp. 603 –608, Jan. 2008 which optimizes the number of symbols and the number of bits per symbol. The aim is to maximize the total rate by iteratively adjusting the transmission powers and spreading sequences whilst satisfying a target signal-to-interference-noise (SINR) ratio at each receiver.

2. The second method aims to maintain the received power at a target level, whilst maximizing the total rate by jointly optimizing the transmission power, rate and signature sequences and also the linear MMSE equalizers at the receiver. One example of such a method is S. Ulukus and A. Yener, “ Iterative transmitter and receiver optimization for cdma networks,” IEEE Transactions on Wireless Communications, vol. 3, no. 6, pp. 1879 – 1884, Nov. 2004 which jointly optimizes a set of transmission spreading sequences and receivers with linear MMSE equalizers.

3. The third method, one example of which is described in L. Zhao and J. Mark, “ Joint rate and power adaptation for radio resource management in uplink wideband code division multiple access systems,” IET Communications, vol. 2, no. 4, pp. 562 –572, April 2008, uses the average system performance as an evaluation criterion which requires the distribution of the received and the interference signal powers.

In the current HSDPA system specification, an equal energy allocation scheme is used to load each channel with either a single rate or two discrete rates. Parameters of the MMSE receivers are usually optimized using either the max-min weighted SNR criterion or the total MSE minimization criterion. Recently, an iterative power adaptation method known as the two-group resource allocation scheme has been developed as described in the paper: “The interference-reduced energy loading for multi-code HSDPA systems” , Gurcan MK, Ma I, Ab Ghani H, EURASIP Journal on Wireless Communications and Networking 2012/ 127 March 2012 for SISO HSDPA

system. This method was extended to MIMO HSDPA system in the paper: "System Value based Optimum Spreading Sequence Selection for High Speed Packet Access (HSDPA) MIMO" , Gurcan MK, Ma I, Chungtragarn A, Joudeh H, EURASIP Journal on Wireless Communications and Networking, 2013, 2013/:74. Mar 2013. Imperial Innovations Ltd have filed the following patent applications related to these two papers: UK patent applications GB1207546.1 and GB1115566.0, and International patent applications PCT/GB2013/000185 and WO2013/034875.

HSDPA systems have channel throughputs, which are usually plotted as a function of the channel signal to interference plus noise ratio. Usually HSDPA transmissions are over frequency selective multipath channels causing Inter Symbol Interference (ISI) and there are typical channels specified by the standardization organizations. Two well known specified channels are the Pedestrian A and the Pedestrian B channels.

The Pedestrian A channels cause small ISI for the HSDPA system, and the Pedestrian B channels cause severe ISI for the HSDPA system.

For the Pedestrian B channels with severe ISI the existing practical systems achieve a performance below the theoretical performance. The patents filed by Imperial Innovations Ltd referred to above have the objectives to control the number and the ordering of signature sequences to achieve a throughput upper bound close to the theoretical throughput upper bound for the HSDPA receiver for the Pedestrian B channels with severe ISI when using:

- successive interference cancellation receivers; and
- deterministic signature sequences without scrambling codes.

The paper: "System Value based Optimum Spreading Sequence Selection for High Speed Packet Access (HSDPA) MIMO" , Gurcan MK, Ma I, Chungtragarn A, Joudeh H, EURASIP Journal on Wireless Communications and Networking, 2013, 2013/:74. Mar 2013 describes a MIMO HSDPA transceivers systems, wherein these transceivers operate with the channelization codes as the signature sequences. These schemes have not used scrambling codes as part of the signature sequences.

The system disclosed in this paper achieves a much improved throughput at high signal to noise ratios by bringing the spatial multiplexing gain $\frac{R_r}{\log_2(SNR)}$ close to the spatial multiplexing upper bound $\min(N_t, N_r)$. A measurement known as the System Value measurement is used to control the rate and the power allocation when using only channelization codes as the spreading sequences. An equal SNR allocation scheme for the rate and the power allocation is considered. The System Value based allocation delivers impressive results providing throughput close to the theoretical upper bound. However, since the method requires the channel side information or the data rate for each channel to be transmitted from the receiver MIMO to the transmitter MIMO further optimisation could be possible.

Summary of Invention

In accordance with an aspect of the invention there is provided a method for determining a transmitted signal received at a receiver in a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple-output channels over which the signal is transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis every symbol period ρ . The method comprises receiving a signal $r_{long}(\rho)$ corresponding to a signal transmitted over a channel, wherein each symbol of the transmitted signal is spread using a respective spreading sequence $S(\rho)$. The method also comprises determining an estimated channel characteristic H indicative of effects of the channel over which the received signal $r_{long}(\rho)$ has been transmitted on the transmitted signal. Furthermore, the method comprises determining the transmitted signal $r_{short}(\rho)$ from the received signal $r_{long}(\rho)$ in accordance with the estimated channel characteristic H and the spreading sequences used to spread the symbols.

The method may further comprise determining a combined matrix Z of all spreading sequences $S(\rho)$ used to spread the symbols of the transmitted signal. The method may further comprise determining a covariance matrix C and a normalization matrix C_n in accordance with the estimated channel characteristic H , and a multiplication of the combined matrix of the spreading sequences Z and the hermitian of the combined matrix of the spreading sequences Z^H . The transmitted signal may be

determined from the received signal $r_{long}(\rho)$ by multiplying the received signal $\underline{r}_{long}(\rho)$ with the inverse of the covariance matrix $\mathbf{C}$, the hermitian of the channel matrix $\mathbf{H}$, and the hermitian of the normalization matrix $\mathbf{C}_n$.

The covariance matrix $\mathbf{C}$ may be defined as follows:

$$\mathbf{C} = \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z} \mathbf{Z}^H \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

wherein E_T is a total transmission energy, $N_t = N_s$ is a total number of MIMO transmission antennas, $N_s K$ is a total number of spreading sequences, $N^{(x)}$ is a total number of transmitted symbols used to calculate the covariance matrix $\mathbf{C}$, N_0 is a noise power spectral density, $\mathbf{I}_{N_r(N+L-1)}$ is an identity matrix, the previous symbol period channel interference matrix $\mathbf{H}_1$ is produced from the estimated channel matrix $\mathbf{H}$ using:

$$\mathbf{H}_1 = \left(\mathbf{I}_{N_r} \otimes (\mathbf{J}^T)^N \right) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

and wherein the channel matrix for the next symbol period is produced using:

$$\mathbf{H}_2 = \left(\mathbf{I}_{N_r} \otimes \mathbf{J}^N \right) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

wherein the $(N+L-1) \times (N+L-1)$ -dimensional matrix $\mathbf{J}_{N+L-1}$ is given as

$$\mathbf{J}_{N+L-1} = \begin{bmatrix} \underline{0}_{(N+L-2)}^T & 0 \\ \mathbf{I}_{(N+L-2)} & \underline{0}_{(N+L-2)} \end{bmatrix} \text{ and also the operation } \mathbf{J}^N \text{ refers to the } N^{\text{th}} \text{ power of}$$

$\mathbf{J}_{N+L-1}$ further the operator $\otimes$ corresponds to Kroneker product, $\mathbf{Z} \mathbf{Z}^H$ are transmitted signature sequence matrices and $\mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H$ are the transmitted signature sequence matrices for the current previous and next symbol periods.

The covariance matrices of the transmitted signature sequence matrices $\mathbf{Z}\mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1}\mathbf{Z}_{\rho-1}^H$ and $\mathbf{Z}_{\rho+1}\mathbf{Z}_{\rho+1}^H$ for the current, previous and next symbol periods may be determined according to the following:

$$\begin{aligned}\mathbf{Z} &= [\mathbf{S}(2), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)} + 1)] \\ \mathbf{Z}_{\rho-1} &= [\mathbf{S}(1), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)})] \\ \mathbf{Z}_{\rho+1} &= [\mathbf{S}(3), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)} + 2)]\end{aligned}$$

wherein a spreading sequence matrix $\mathbf{S}(\rho)$ of all spreading symbols for each symbol period ρ is determined according to:

$$\mathbf{S}(\rho) = \text{diag}(\underline{s}_{wsc}(\rho)) \mathbf{S}_{ch,W}.$$

wherein $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned}\mathbf{S}_{wsc} &= (\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc}) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)]\end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss}$ $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$ wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

For a Successive Interference Cancellation, SIC, receiver the signature sequence covariance matrices $\mathbf{Z}(\xi)\mathbf{Z}^H(\xi)$, $\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi)$ and $\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi)$ may be updated using:

$$\mathbf{Z}(\xi)\mathbf{Z}^H(\xi) = \mathbf{Z}(\xi)\mathbf{Z}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_k(\xi)\mathbf{S}_k^H(\xi)$$

wherein $\mathbf{S}_k(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + 1 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho \right) \right]$ and

$$\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) = \mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho-1)}(\xi)\mathbf{S}_{k,(\rho-1)}^H(\xi)$$

wherein $\mathbf{S}_{k,(\rho-1)}(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho - 1 \right) \right]$ and

$$\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) = \mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho+1)}(\xi)\mathbf{S}_{k,(\rho+1)}^H(\xi)$$

wherein $\mathbf{S}_{k,(\rho+1)}(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + 2 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho + 1 \right) \right]$ for the updated

signature sequence matrices, and $\mathbf{C}(\xi)$ and $\mathbf{C}_n(\xi)$ are the received signal

covariance matrix and the normalization matrix respectively for use in the implementation of successive interference cancellation receivers and wherein ξ is the epoch number, $N^{(x)}$ and $N_s^{(x)}$ are the number of symbols transmitted in a block and a set of the blocks respectively and wherein the covariance matrix $\mathbf{C}(\xi)$ is determined according to:

$$\mathbf{C}(\xi) = \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z}(\xi) \mathbf{Z}^H(\xi) \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1}(\xi) \mathbf{Z}_{\rho-1}^H(\xi) \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1}(\xi) \mathbf{Z}_{\rho+1}^H(\xi) \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

and the normalization matrix is determined according to:

$$\mathbf{C}_n(\xi) = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H} \right) \right) \right)^{-1}.$$

For the Successive Interference Cancellation, SIC, receiver the received long matrix $\mathbf{R}_{long}$ may be iteratively updated for the correctly detected symbol sequences in the channels updated using:

$$\begin{aligned} \mathbf{R}_{long} &= \mathbf{R}_{long} - E_k \sum_{k=k_1}^{k_{n_x}} \Phi_k \\ &= [\underline{r}_{long}(1), \dots, \underline{r}_{long}(\rho), \dots, \underline{r}_{long}(N^{(x)})] \end{aligned}$$

wherein the long matrix Φ_k for the correctly detected symbols is determined according to:

$$\Phi_k = [\underline{\phi}_k(1), \dots, \underline{\phi}_k(\rho), \dots, \underline{\phi}_k(N^{(x)})]$$

wherein the k^{th} channel's chip sequence vector $\underline{\phi}_k(\rho)$ for the symbol period ρ and spreading sequence $\underline{s}_k(\rho)$ is determined according to:

$$\underline{\phi}_k(\rho) = [\mathbf{H} \underline{s}_k(\rho) x_{k,D}(\rho) + \mathbf{H}_1 \underline{s}_k(\rho-1) x_{k,D}(\rho-1) + \mathbf{H}_2 \underline{s}_k(\rho+1) x_{k,D}(\rho+1)]$$

for successive interference cancellation receivers.

The transmitted signal $\underline{r}_{short}(\rho)$ may be determined according to the following:

$$\underline{r}_{short}(\rho) = \left(\left(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n \right)^H \underline{r}_{long}(\rho) \right)$$

wherein $\mathbf{C}^{-1}$ is an inverse of the covariance matrix, $\underline{r}_{long}(\rho)$ is the received signal, and the normalization matrix $\mathbf{C}_n$ is determined according to:

$$\mathbf{C}_n = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} \right) \right) \right)^{-1}.$$

The method may further comprise estimating a received symbol vector using the following relationship:

$$\begin{aligned} \hat{\underline{y}}(\rho) &= \mathbf{S}_{ch,W}^H \left(\text{Diag}(\underline{s}_{wsc}(\rho)) \right)^H \underline{r}_{short}(\rho) \\ &= [\hat{y}_1(\rho), \dots, \hat{y}_k(\rho), \dots, \hat{y}_{N_t K}(\rho)] \end{aligned}$$

wherein $\underline{r}_{short}(\rho)$ is the transmitted signal, $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned} \mathbf{S}_{wsc} &= \left(\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc} \right) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)] \end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$, wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

The estimated channel characteristic may be defined by:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1,1)} & \dots & \mathbf{H}^{(1,N_t)} \\ \vdots & \mathbf{H}^{(n_r,n_t)} & \vdots \\ \mathbf{H}^{(N_r,1)} & \dots & \mathbf{H}^{(N_r,N_t)} \end{bmatrix}$$

wherein the channel convolution matrix $\mathbf{H}^{(n_r,n_t)}$ between the transmit antenna n_t and the receive antenna n_r is defined by:

$$\mathbf{H}^{(n_r,n_t)} = \begin{bmatrix} h_0^{(n_r,n_t)} & 0 & 0 \\ \vdots & h_0^{(n_r,n_t)} & \vdots \\ h_{L-1}^{(n_r,n_t)} & \vdots & \ddots & 0 \\ 0 & h_{L-1}^{(n_r,n_t)} & h_0^{(n_r,n_t)} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & h_{L-1}^{(n_r,n_t)} \end{bmatrix}$$

wherein a channel impulse response vector between a transmit antenna n_t and a receive antenna n_r is defined by channel impulse response function:

$$\underline{h}^{(n_r, n_t)} = \begin{bmatrix} h_0^{(n_r, n_t)} & \dots & h_{L-1}^{(n_r, n_t)} \end{bmatrix}^T.$$

The estimated channel characteristic may be a channel impulse response function.

The received signal may be defined by:

$$\underline{r}_{long}(\rho) = \mathbf{H}\underline{z}(\rho) + \mathbf{H}_1\underline{z}(\rho-1) + \mathbf{H}_2\underline{z}(\rho-1) + \underline{n}(\rho)$$

wherein the vector $\underline{n}(\rho)$ corresponds to the received noise samples at the output of receiver chip matched filter and $\underline{z}(\rho)$ is the transmit chip vector for the current symbol period which is determined according to:

$$\underline{z}(\rho) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho) \underline{y}(\rho)$$

wherein $\underline{z}(\rho-1)$ is the transmit chip vector for the previous symbol causing intersymbol interference which is determined according to:

$$\underline{z}(\rho-1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho-1) \underline{y}(\rho-1)$$

wherein the transmitted chip vector $\underline{z}(\rho+1)$ for the next $(\rho^{th} + 1)$ symbol period is:

$$\underline{z}(\rho+1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho+1) \underline{y}(\rho+1)$$

wherein $\underline{y}(\rho) = [y_1(\rho), \dots, y_k(\rho), \dots, y_K(\rho)]^T$ is the transmitted symbols vector which contains the symbols, over the symbol period $\rho = 1, \dots, N^{(x)}$, wherein $N^{(x)}$ is the number of transmitted symbols in a transmission block wherein the entire block of transmission is represented as an $(N^{(x)} \times N_t K)$ dimensional transmit symbol matrix using the following relationship:

$$\mathbf{X} = [\underline{x}_1, \dots, \underline{x}_k, \dots, \underline{x}_{N_t K}] = [\underline{y}(1), \dots, \underline{y}(\rho), \dots, \underline{y}(N^{(x)})]^T.$$

wherein $\mathbf{X}$ is the matrix containing transmitted symbols, $\underline{x}_k$ is the vector containing symbols over channel k , $N_t K$ is the total number of spreading sequences used in the downlink MIMO system, and wherein the $(N^{(x)} \times 1)$ -dimensional symbol vector $\underline{x}_k$ is constructed using:

$$\underline{x}_k = [x_k(1), \dots, x_k(\rho), \dots, x_k(N^{(x)})]^T$$

for each channel $k = 1, \dots, N_t K$.

In accordance with a further aspect of the invention there is provided a method for determining data transmission rates for transmitting a signal over a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple output channels over which the signal is to be transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis. The method comprises receiving a signal comprising a plurality of symbols, each symbol having been spread by a spreading sequence of a plurality of spreading sequences, determining, for each symbol of the plurality of symbols, a system value λ_k indicative of a signal-to-noise ratio associated with the respective spreading sequence used to spread the symbol, and averaging the plurality of system values associated with the plurality of spreading sequences to determine an averaged system value for determining a data rate for transmission of symbols using the plurality of spreading sequences.

The method may further comprise determining a data rate to be associated with the number of signature sequences for transmission in accordance with the average system value.

The data rate for each stream may be determined by finding the discrete data rate value b_p which satisfies the following equation:

$$[\underline{\lambda}^*]_p \leq \lambda_{(n_s-1)K+1}(\xi) < [\underline{\lambda}^*]_{p+1}.$$

wherein $\lambda_{(n_s-1)K+1}(\xi)$ is the system value, which is for a streaming group number n_s

and for symbol epoch ξ and a target system value $[\underline{\lambda}^*]_p$, which is determined

according to $[\underline{\lambda}^*]_p = \frac{\Gamma(2^{b_p} - 1)}{1 - \Gamma(2^{b_p} - 1)}$ for $p = 1, \dots, P$ for a chosen discrete bit rate of b_p

wherein Γ is the gap value.

The method may further comprise estimating the symbols in the received signal in accordance with the following equation:

$$\hat{\underline{y}}(\rho) = \mathbf{S}_{ch,W}^H (\text{Diag}(\underline{s}_{wsc}(\rho)))^H \underline{r}_{short}(\rho)$$

Wherein $\underline{r}_{short}(\rho)$ is the transmitted signal, $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned} \mathbf{S}_{wsc} &= (\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc}) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)] \end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$ wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

The system value may be defined by:

$$\begin{aligned} \lambda_k &= 1 - \varepsilon_k \\ &= 1 - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D} \mathbf{Z} \mathbf{Z}^H \mathbf{D}^H \mathbf{S}_k) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_1 \mathbf{Z}_{\rho-1}^H \mathbf{Z}_{\rho-1} \mathbf{D}_1^H \mathbf{S}_k) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_2 \mathbf{Z}_{\rho+1}^H \mathbf{Z}_{\rho+1} \mathbf{D}_2^H \mathbf{S}_k) \\ &\quad - \frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B} \mathbf{B}^H) \end{aligned}$$

wherein $\mathbf{D}$ is the estimated channel characteristic converted to a square matrix

representative of intersymbol interference, which is defined as follows:

$$\mathbf{D} = \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} - \text{diag}(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H})$$

and a signature sequence matrix for channel k is defined as:

$$\mathbf{S}_k = \text{diag}(\underline{s}_{wch,k}) \mathbf{S}_{wsc} = [\underline{s}_k(1), \dots, \underline{s}_k(\rho), \dots, \underline{s}_k(N^{(x)})]$$

wherein an intersymbol matrix $\mathbf{D}_1$ for the interference from the previous symbol is determined according to:

$$\mathbf{D}_1 = (\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_1)$$

and an intersymbol interference matrix $\mathbf{D}_2$ for the interference from the next symbol is determined according to:

$$\mathbf{D}_2 = (\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_2)$$

and $\mathbf{B}$ is a noise vector despreading matrix, which is determined according to:

$$\mathbf{B} = \mathbf{S}_k^H \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1}.$$

The plurality of spreading sequences with which the average system value is associated may correspond to a first combination of receiver and antenna channels, wherein the method further comprises determining an average system value for one or more other combinations of receiver and antenna channels, and using the receiver and antenna channel having the highest system value for transmission.

Any of the different methods disclosed herein may be combined with one another, where appropriate, to form a combined process. For example, the method for determining a transmitted signal received at a receiver may be combined with the method for determining data rates, and vice versa.

Also disclosed is an apparatus arranged to perform any of the methods disclosed herein. The apparatus may be a receiver device. The apparatus may be a radio transmission system comprising one or more transmitters and one or more receivers.

Also disclosed is a computer program product arranged, in use, to instruct a computer to perform any method disclosed herein.

MMSE linear equalisers for MIMO transceivers with signature sequences constructed using scrambling and channelization codes are disclosed. Furthermore, a system model for the HSDPA MIMO transceivers which use signature sequences constructed using channelization and scrambling codes is discussed herein.

The System Value (SV) based rate allocation process may be used to allocate equal energy, equal SNR and equal rate for HSDPA MIMO downlink transmissions for a single user case. In the downlink transmission both the channelization and the scrambling codes are considered. Although the rate allocations may be based on the extended System Value calculations, known covariance matrix calculations may not be applicable to the system value calculations disclosed herein.

Covariance matrix calculations are disclosed herein to deal with spreading sequences which incorporate both channelization and scrambling codes.

A MIMO transmitter and receiver and multiple spreading sequences when operating with spreading sequences which change every symbol period when transmitting over frequency selective channels is disclosed.

A method for data transmission for a mobile radio system having a single input and single output channel or a plurality of Multiple-Input-Multiple-Output channels with a total number of N_t transmit and N_r receive antennas where N_s is the minimum of N_t and N_r over which data is transmitted, data is represented by a plurality of data symbols, data symbols being spread prior to transmission by a plurality of spreading sequences. The method comprises determining a system value λ_k for each signature sequence k of plurality of signature sequences $N_s K$, wherein the total number K of orthogonal signature sequences is the base number for spreading sequences and the system value λ_k is indicative of a signal-to-noise ratio of the associated signature sequence constructed using a combination of channelizing and scrambling sequences wherein the method determines the number K^* of signature sequences to be used for spreading the data symbols in accordance with the system values λ_k associated with a plurality of signature sequences $N_s K$ wherein the number of signature sequences selected corresponds to the determined number of signature sequences K^* and spreading the data symbols using the selected signature sequences matrix S wherein the selected number spreading sequences

are used to spread data symbols to be transmitted over a subset of transmitter antennas selected from N_t transmitter antennas if N_s is equal to N_r , alternatively the number of selected signature sequences are transmitted from N_t transmitter antennas and at the receiver the system values λ_k associated with a plurality of signature sequences $N_s K$ are used to select a group of N_s receiver antennas out of

N_r receiver antennas if N_s is equal to N_t , the signals at the output of $\begin{pmatrix} N_s \\ N_r \end{pmatrix}$ MIMO

receivers are combined to maximize the total signal to noise ratio and hence the total transmission data rate.

A number of non-deterministic channelization sequences, each with a set of scrambling sequences, may be into the detection methods, where the system values λ_k associated with a plurality of signature sequences $N_s K$ are used to produce a measurement method. This System Value (SV) based measurement method provides improvements to the HSDPA throughput without requiring any significant changes to current operational HSDPA systems. The proposed method uses the System Value based measurement method to select a sub-set of MIMO receiver and transmitter antennas to achieve throughput improvement.

A system model for the High Speed Downlink Packet Access (HSDPA) system when using spreading sequences which change at every symbol period is disclosed. This simplifies the system implementation for the HSDPA down link operations which can be described using the system model for the HSDPA downlink transmission for known systems under the assumption that $N_s = N_t$. All $N_t K$ spreading sequences are used to transmit data from N_t downlink antennas to N_r receiver antennas at a single user receiver.

Initially, the signature sequences may be constructed by combining the scrambling and the channelization codes. These signature sequences may be used to generate transmitted chip sequences for a given number of QAM symbols transmitted over KN_s parallel channels. In the MIMO system, a total of N_r MIMO receiver antennas may be considered and N may be the processing gain of the spreading sequence.

Channelization codes may be generated using Hadamard OVFS codes as specified by the 3GPP standardization organization. For a spreading sequence processing gain, N , the total number of Hadamard codes is N , the signature sequence matrix $\mathbf{S}_{ss}$ has the dimension $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}] \in \mathbb{C}^{N \times N}$. The channelization spreading sequence vector $\underline{s}_{ss,k} \in \mathbb{C}^N$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

A total of K channelization codes $\mathbf{S}_{ch,K} = [\underline{s}_{ch,1}, \dots, \underline{s}_{ch,k}, \dots, \underline{s}_{ch,K}] \in \mathbb{C}^{N \times K}$ for $K \leq N$ can be extracted from the channelization codes $\mathbf{S}_{ss}$. $\mathbf{C}$ defines the two dimensional complex space with dimension sizes given as the superscript values. The spreading sequences may be real instead of complex ones.

The number of channelization spreading sequences may be increased to $N_s K$ and the length of each spreading sequence may be increased to $N_t N$ by using the precoding matrix of $\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}] \in \mathbb{C}^{N_t \times N_s}$. The precoding vectors have the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

A total of $N_s K$ channelization codes are generated using a precoding matrix to produce an extended channelization signature sequence matrix:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}] \in \mathbb{C}^{N_t N \times N_s K}.$$

Scrambling sequences may be generated using a long binary sequence of m-sequences or any other data sequence with very low autocorrelation sidelobes. In this part of the system modeling, m-sequences of length $2^{16} - 1$ and $2^{17} - 1$ may be used to construct scrambling sequences. The m-sequence vector $\underline{m}_s = [m_s(1), \dots, m_s(l), \dots, m_s(2^{16} - 1)]^T \in \mathbb{C}^{2^{16}-1}$ elements are $m_s(l) \in \{\pm 1\}$. The

scrambling sequences are used to generate scrambling sequence matrix

$\mathbf{S}_{sc} = [\underline{s}_{sc}(1), \dots, \underline{s}_{sc}(\rho), \dots, \underline{s}_{sc}(N^{(x)} + 2)] \in \mathbb{C}^{N \times (N^{(x)} + 1)}$, where ρ symbol period epoch number and $N^{(x)}$ is the total number of symbols to be transmitted over each channelization code. Each scrambling sequence vector $\underline{s}_{sc}(\rho) = [s_{sc,1}(\rho), \dots, s_{sc,n}(\rho), \dots, s_{sc,N}(\rho)]^T$ may have complex elements constructed using the m-sequence elements:

$$s_{sc,n}(\rho) = \frac{m_s((\rho-1)N+n) + jm_s(2^{15} + (\rho-1)N+n)}{\sqrt{2}}.$$

The size of the scrambling code matrix may be increased using the following operation:

$$\begin{aligned} \mathbf{S}_{wsc} &= (\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc}) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)] \in \mathbb{C}^{N_t N \times (N^{(x)} + 2)}. \end{aligned}$$

For each symbol period a signature sequence matrix may be generated as:

$$\mathbf{S}(\rho) = \text{diag}(\underline{s}_{wsc}(\rho)) \mathbf{S}_{ch,W}.$$

The signature sequence matrix $\mathbf{S}(\rho)$ for $\rho = 1, \dots, N^{(x)}$ may be concatenated to generate the extended signature sequence matrix as:

$$\mathbf{S} = [\mathbf{S}(1), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)})] \in \mathbb{C}^{N_t N \times KN_s N^{(x)}}$$

For each channelization code a signature sequence matrix may be generated for different symbol periods as follows:

$$\mathbf{S}_k = \text{diag}(\underline{s}_{wch,k}) \mathbf{S}_{wsc} = [\underline{s}_k(1), \dots, \underline{s}_k(\rho), \dots, \underline{s}_k(N^{(x)})] \in \mathbb{C}^{N_t N \times N^{(x)}} \text{ for } k = 1, \dots, N_t K.$$

The transmitted chip sequence matrix for different symbol periods may be generated using:

$$\mathbf{Z} = [\mathbf{s}(2), \dots, \mathbf{s}(\rho), \dots, \mathbf{s}(N^{(x)} + 1)] \in \mathbb{C}^{N_t N \times N_s K N^{(x)}}.$$

and

$$\mathbf{Z}_{\rho-1} = [\mathbf{s}(1), \dots, \mathbf{s}(\rho), \dots, \mathbf{s}(N^{(x)})] \in \mathbb{C}^{N_t N \times N_s K N^{(x)}}.$$

and also

$$\mathbf{Z}_{\rho+1} = [\mathbf{s}(3), \dots, \mathbf{s}(\rho), \dots, \mathbf{s}(N^{(x)} + 2)] \in \mathbb{C}^{N_t N \times N_s K N^{(x)}}.$$

The transmitted spreading sequence matrix may be incorporated with a novel covariance matrix calculation method to obtain the inverse of the covariance matrix. The covariance matrix inversion may involve the use of transmission spreading sequence matrices $\mathbf{Z}$, $\mathbf{Z}_{\rho-1}$ and $\mathbf{Z}_{\rho+1}$ and their covariance matrices. The spreading sequence matrices may be available at the transmitter and receiver in advance of transmitting blocks of symbols. The use of preset transmission spreading sequence covariance matrix eliminates the requirement to use computationally intensive matrix calculations and inversions at the receiver each time a block of data is received.

The matrix inversion scheme disclosed herein can be incorporated with a symbol level MMSE receiver equalizer operating with short blocks of transmission symbols. The covariance matrix inversions and also the receiver equalizer designs may involve the use of link level MIMO receiver modeling. In the link level MIMO modelling of the received signals, the spread transmitted symbols may be allocated equal transmission energies with the transmission energy constraint that the total transmission energy per symbol period is E_T and each spreading sequence may be allocated equal energy such that $E_k = \frac{E_T}{N_s K}$ for $k = 1, \dots, N_s K$.

A process is disclosed herein based on the principle of utilizing a system value to determine which data rates should be allocated to increase the achievable data rates when spreading the transmitted symbols.

A symbol level equalizer may be followed by a successive interference cancellation scheme. Also, a method to use a constrained power level at the transmitter of the HSDPA downlink to adjust the transmission data rates for different groups of spreading sequences when using the proposed System Value and a measurement method and the simplified covariance matrix calculation and inversion is disclosed.

A novel method of calculating transmission data rates for different groups of transmission spreading sequences is disclosed that provides a total transmission data rate close to the theoretical upper bound.

Systems disclosed herein allocate Discrete Data Rates to maximize the total rate using System Values. Providing multiplexing versus diversity gain trade off when selecting a sub-set of receiver antennas from a large set.

Brief Description of the Drawings

Exemplary embodiments of the invention shall now be described with reference to the drawings in which:

Figure 1 provides a schematic illustration of an HSDPA MIMO transmitter and receiver arrangement; and

Figure 2 provides a schematic illustration of a Successive Interference Cancelling receiver.

Throughout the description and the drawings, like reference numerals refer to like parts.

Specific Description

Figure 1 shows a multi-code CDMA downlink system. The system has a total of N_t and N_r transmitter and receiver antennas and also $N_s K$ spreading sequences, where $N_s = N_t$ is taken for this specific example, each of which is realizable with a bit rate of b_{p_k} bits per symbol from a set of bit rates, $\{b_{p_k}\}_{p_k=1}^P$, for a given total transmission energy E_T and $p = 1, 2, \dots, P$. The basic principle of operation of this system is that by excluding the weak channels corresponding to a specific set of spreading sequences, the number of parallel transmission channels is reduced to $N_t K$ spreading sequences to transmit a symbol per channel. The data for the intended symbol for each channel is placed in an $(N_U \times 1)$ -dimensional vector $\underline{u}_k$ for $k = 1, \dots, N_t K$.

The system shall now be described in detail.

In Figure 1, a transmitter 100 receives input vectors of $\underline{u}_k$ for $k = 1, \dots, N_t K$. The input vectors represent input data to be transmitted by the transmitter 100. This input data is then input into encoding unit 101. Each of these data packets is then channel encoded by the encoding unit 101 to produce a $(B \times 1)$ -dimensional vector $\underline{d}_k$. The encoded data $\underline{d}_k$ for $k = 1, \dots, N_t K$ produced by the encoding unit 101 is then processed by the adaptive modulation and coding unit 102 to transform the encoded data into symbol vectors using a quadrature amplitude modulation scheme (QAM) with M constellations to transmit data at a rate $b = \log_2 M$ bits per symbol.

The channel encoder rate is $r_{code} = \frac{N_U}{B}$ and the realizable discrete rates are given by

$b_p = r_{code} \log_2 M$ for $p = 1, \dots, P$ where P is the number of available discrete data rates.

Data is transmitted in packets at a transmission-time-interval (TTI) and the number of symbols transmitted per packet is denoted as $N^{(x)}$ where $N^{(x)} = \frac{TTI}{NT_c}$ and N is the spreading sequence length, T_c is the chip period and NT_c is the symbol period. The modulation and coding unit 102 transforms the encoded data into the $(N^{(x)} \times 1)$ -dimensional symbol vector $\underline{x}_k = [x_k(1), \dots, x_k(\rho), \dots, x_k(N^{(x)})]^T$ for each channel $k = 1, \dots, N_t K$. The entire block of transmission can be represented as an $(N^{(x)} \times N_t K)$ dimensional transmit symbol matrix defined as:

$$\begin{aligned} \mathbf{X} &= [\underline{x}_1, \dots, \underline{x}_k, \dots, \underline{x}_{N_t K}] \\ &= [\underline{y}(1), \dots, \underline{y}(\rho), \dots, \underline{y}(N^{(x)})]^T. \end{aligned}$$

The transmitted vector $\underline{y}(\rho) = [y_1(\rho), \dots, y_k(\rho), \dots, y_K(\rho)]^T$ contains the symbols, over the symbol period $\rho = 1, \dots, N^{(x)}$, with the unit average energy

$E[y_k(\rho)y_k^*(\rho)] = 1$ for $k = 1, \dots, N_t K$ where $y_k^*(\rho)$ is the complex conjugate of the symbol $y_k(\rho)$.

The transmission symbol energies are then adjusted using the power control unit 103. Power allocation is performed on the symbols before spreading. The energies for all $N_t K$ channels are stored in an amplitude matrix $A = \text{diag}(\sqrt{E_1}, \dots, \sqrt{E_k}, \dots, \sqrt{E_{N_t K}})$ subject to the total energy E_T such that $\sum_{k=1}^K E_k = E_T$.

The energy weighted data symbols are then transformed into a transmit vector using the vector generation unit 104. After assigning energies, the amplitude weighted symbols are spread in the spreading unit 105 by a plurality of $N \times N_t K$ dimensional spreading sequences $\mathbf{S}_{n_t}(\rho) = [\underline{s}_{n_t,1}, \dots, \underline{s}_{n_t,k}, \dots, \underline{s}_{n_t,N_t K}] \in \mathbb{C}^{N \times N_t K}$, for $n_t = 1, \dots, N_t$ where $\mathbb{C}$ defines the two dimensional complex space with dimension sizes gives as the superscript values. the signature matrices are generated from $\mathbf{S}(\rho) = [\mathbf{S}_1^T(\rho), \dots, \mathbf{S}_{n_t}^T(\rho), \dots, \mathbf{S}_{N_t}^T(\rho)]^T$. At every symbol period $\rho = 1, \dots, N^{(x)}$ the length N transmit chip vector:

$$\underline{z}(\rho) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho) \underline{y}(\rho) \in \mathbb{C}^{N_t N} \text{ for } \rho = 2, \dots, (N^{(x)} + 1)$$

is generated at the input of n_t^{th} antenna unit for $n_t = 1, \dots, N_t$ where $\underline{y}(\rho) \in \mathbb{C}^{N_t K}$ is the transmitted signal symbol vector for the ρ^{th} symbol period. The transmitted chip vector for the previous $(\rho^{\text{th}} - 1)$ symbol period is:

$$\underline{z}(\rho - 1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho - 1) \underline{y}(\rho - 1) \in \mathbb{C}^{N_t N}$$

The transmitted chip vector for the next $(\rho^{\text{th}} + 1)$ symbol period is:

$$\underline{z}(\rho+1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho+1) \underline{y}(\rho+1) \in \mathbb{C}^{N_t N}$$

The spread signals are next filtered using the pulse shaping filter 106 to produce transmission signals for transmission from the MIMO transmitters 107a, 107b, ..., 107 N_T .

The signal is then transmitted from the transmitter to a receiver 200 across a channel. Since characteristics of the channel affect the received signal it is necessary to model the channel in order for the transmitted signal to be derived from the received signal. The modelling of the channel shall firstly be discussed before considering how the received signal is then dealt with.

At each TTI, the transmission procedure described above is used to transmit the pilot signals which are used in order to estimate the channel condition at each receiver to calculate the data rate $b_{p,k}$ to be transmitted over the k^{th} channel. The rate is calculated at the receiver and is feedback to the transmitter. It is assumed that the channel condition does not change for that TTI. All symbols in a block of length $N^{(x)}$ in all spread sequence channels when transmitted from the n_t^{th} transmitter antenna to the n_r^{th} receiver antenna experience the same channel condition in a multipath environment with L resolvable paths. This channel condition **[ACCEPTED]** can be represented by a channel impulse response function $\underline{h}^{(n_r, n_t)} = [h_0^{(n_r, n_t)} \dots h_{L-1}^{(n_r, n_t)}]^T$ corresponding $((N+L-1) \times N)$ -dimensional channel convolution matrix $\mathbf{H}^{(n_r, n_t)}$ can be represented as follows:

$$\mathbf{H}^{(n_r, n_t)} = \begin{bmatrix} h_0^{(n_r, n_t)} & 0 & & 0 \\ \vdots & h_0^{(n_r, n_t)} & & \vdots \\ h_{L-1}^{(n_r, n_t)} & \vdots & \ddots & 0 \\ 0 & h_{L-1}^{(n_r, n_t)} & & h_0^{(n_r, n_t)} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & & h_{L-1}^{(n_r, n_t)} \end{bmatrix}.$$

The overall $(N_r(N+L-1) \times N_t N)$ -dimensional MIMO channel convolution matrix can be formed as:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1,1)} & \dots & \mathbf{H}^{(1,N_t)} \\ \vdots & \ddots & \vdots \\ \mathbf{H}^{(N_r,1)} & \dots & \mathbf{H}^{(N_r,N_t)} \end{bmatrix} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}.$$

where L is the transmission channel impulse response length. The channel matrix for the previous symbol period is:

$$\mathbf{H}_1 = (\mathbf{I}_{N_r} \otimes (\mathbf{J}^T)^N) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

The channel matrix for the next symbol period is:

$$\mathbf{H}_2 = (\mathbf{I}_{N_r} \otimes \mathbf{J}^N) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

The $(N+L-1) \times (N+L-1)$ -dimensional matrix, i.e. the amount by which one

symbol affects the next symbol, is defined as $\mathbf{J}_{N+L-1} = \begin{bmatrix} \mathbf{0}_{(N+L-2)}^T & 0 \\ \mathbf{I}_{(N+L-2)} & \mathbf{0}_{(N+L-2)} \end{bmatrix}$. For

simplicity the subscript will be dropped from the $\mathbf{J}$ matrix notation. When the matrix $\mathbf{J}^N$ operates on a column vector, it downshifts the column by N chips filling the top of the column with N zeros. **$\mathbf{J}$ is therefore used to convert $\mathbf{H}$ to $\mathbf{H}_1$ and $\mathbf{H}_2$.**

Now that the channel has been defined we shall return to the description of Figure 1 and specifically examine the processes carried out at the receiver 200.

The transmitted pilot signal is then received at the receiver 200 by the MIMO receivers 201a, 201b, ...201 N_r . The received signal long chip vector $\underline{r}_{long}(\rho)$, which is the signal received at the receiver 200 has been affected by the channel and therefore does not directly correspond to the signal that was transmitted. It is therefore necessary to determine the signal that was transmitted by processing the received signal accordingly, as will now be discussed.

Assuming the clocks at the transmitter and receiver are fully synchronized, the received signal is then frequency down converted, filtered and sampled as chip period intervals by the chip matched filter unit 202. The received signal long chip vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ is received as:

$$\begin{aligned}\underline{r}_{long}(\rho) &= \sqrt{\frac{E_T}{N_t K}} \mathbf{H} \mathbf{S}(\rho) \underline{y}(\rho) \\ &+ \sqrt{\frac{E_T}{N_t K}} \mathbf{H}_1 \mathbf{S}(\rho-1) \underline{y}(\rho-1) \\ &+ \sqrt{\frac{E_T}{N_t K}} \mathbf{H}_2 \mathbf{S}(\rho+1) \underline{y}(\rho+1) \\ &+ \underline{n}(\rho) \\ &= \mathbf{H} \underline{z}(\rho) + \mathbf{H}_1 \underline{z}(\rho-1) + \mathbf{H}_2 \underline{z}(\rho+1) + \underline{n}(\rho)\end{aligned}$$

where $\underline{n}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ is the noise vector at the output of the receiver chip matched filter.

From the received signal $\underline{r}_{long}(\rho)$ the received long signal matrix $\mathbf{R}_{long} \in \mathbb{C}^{N_r(N+L-1) \times N^{(x)}}$ produced as follows:

$$\mathbf{R}_{long} = [\underline{r}_{long}(1), \dots, \underline{r}_{long}(\rho), \dots, \underline{r}_{long}(N^{(x)})] \in \mathbb{C}^{N_r(N+L-1) \times N^{(x)}}$$

The received long signal matrix is obtained in order to increase the processing efficiency.

Using a simplified covariance matrix inversion method and also with a minimum mean square error equalizer implementation, the signal that was transmitted can be obtained from the received long signal matrix, as shall now be described.

Firstly, the derivation of a covariance matrix used to map the long vector $\underline{r}_{long}(\rho)$ at the output of the chip matched filter unit 202 to produce the short vector $\underline{r}_{short}(\rho)$ at the output of the vector mapping unit 203 shall be considered. At the receiver chip matched filter output the long vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ has dimension $N_r(N+L-1)$ and to despread the received vector all the interferences from the

multipath MIMO channels $\mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$ must be minimized by multiplying $\mathbf{H}$ at the receiver with the vector mapping matrix $(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n)^H \in \mathbb{C}^{N_t N \times N_r(N+L-1)}$ with the dimension such that the multiplication $(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n)^H \mathbf{H} \in \mathbb{C}^{N_t N \times N_t N}$ has the dimension $N_t N \times N_t N$ and its diagonal terms are unity. Here $\mathbf{C} = E(\mathbf{r}_{long}(\rho) \mathbf{r}_{long}(\rho)^H) \in \mathbb{C}^{N_r(N+L-1) \times N_r(N+L-1)}$ is the covariance matrix of the received long vector $\mathbf{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$. The matrix $\mathbf{C}_n \in \mathbb{C}^{N_t N \times N_t N}$ is the amplitude normalization matrix. The non-diagonal terms of $(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n)^H \mathbf{H}$ identify the residual interferences between specific transmitter and receiver antennas. When multiplying the long vector $\mathbf{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ with the vector mapping matrix $(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n)^H \in \mathbb{C}^{N_t N \times N_r(N+L-1)}$ at the receiver the short $\mathbf{r}_{short}(\rho) \in \mathbb{C}^{N_t N}$ is produced using $\mathbf{r}_{short}(\rho) = (\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n)^H \mathbf{r}_{long}(\rho) \in \mathbb{C}^{N_t N}$. The resultant short vector $\mathbf{r}_{short}(\rho) \in \mathbb{C}^{N_t N}$ is the best estimate of the transmitted chip vector $\mathbf{z}(\rho) \in \mathbb{C}^{N_t N}$ produced by the MMSE equalization unit at the receiver. An estimate of the transmitted symbol is obtained at the receiver by descrambling the received $\mathbf{r}_{short}(\rho) \in \mathbb{C}^{N_t N}$ vector with the scrambling sequence and then despread it using the channelization signature sequence for a specific channel.

The received signal's covariance matrix averaged over a given number of symbols, corresponding to a given block length of the symbols observed in the demodulation process, is calculated using the following formulation:

$$\mathbf{C} = E(\mathbf{r}_{long}(\rho) \mathbf{r}_{long}(\rho)^H) \in \mathbb{C}^{N_r(N+L-1) \times N_r(N+L-1)}$$

Equation 1

$$= \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z} \mathbf{Z}^H \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

In the above equation during the averaging process the channel matrices $\mathbf{H}$, $\mathbf{H}_1$ and $\mathbf{H}_2$ are taken to be constant as the channels are considered to be stationary during one TTI period. The identity matrix $\mathbf{I}_{N_r(N+L-1)}$ is used to deal with noise

covariance matrix which has noise power spectral density of N_0 . The averaging will be taken by using the term which changes from symbol to symbol when observing the long vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$. As the only term which changes from symbol to symbol is the scrambling sequence combined with the channelization spreading sequence, the received signal vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ and its complex conjugate transpose vector $\underline{r}_{long}^H(\rho)$ will be averaged. The averaging process involves multiplying $\underline{r}_{long}(\rho)$ with $\underline{r}_{long}^H(\rho)$ resulting in the following relationship:

$$\mathbf{C} = \frac{1}{N^{(x)}} \sum_{\rho=1}^{N^{(x)}} \underline{r}_{long}(\rho) \underline{r}_{long}^H(\rho)$$

This averaging process involves the calculation of covariance matrices for every symbol period and summing it over $N^{(x)}$ symbol periods and dividing the result by $N^{(x)}$. As the matrix calculation $\underline{r}_{long}(\rho) \underline{r}_{long}^H(\rho)$ involves the use of transmitted chip vectors $\underline{z}(\rho) \underline{z}^H(\rho)$, $\underline{z}(\rho-1) \underline{z}^H(\rho-1)$ and $\underline{z}(\rho+1) \underline{z}^H(\rho+1)$ in the following form:

$$\begin{aligned} \sum_{\rho=1}^{N^{(x)}} \underline{r}_{long}(\rho) \underline{r}_{long}^H(\rho) &= \mathbf{H} \left(\sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho) \underline{z}^H(\rho) \right) \mathbf{H}^H \\ &+ \mathbf{H}_1 \left(\sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho-1) \underline{z}^H(\rho-1) \right) \mathbf{H}_1^H \\ &+ \mathbf{H}_2 \left(\sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho+1) \underline{z}^H(\rho+1) \right) \mathbf{H}_2^H \\ &+ N_o N^{(x)} \mathbf{I}_{N_r(N+L-1)} \end{aligned}$$

As the matrix summations in the above equation are expressed as

$$\sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho) \underline{z}^H(\rho) = \frac{E_T}{N_t K} \mathbf{Z} \mathbf{Z}^H \quad \text{and} \quad \sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho-1) \underline{z}^H(\rho-1) = \frac{E_T}{N_t K} \mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H \quad \text{and also}$$

$$\sum_{\rho=1}^{N^{(x)}} \underline{z}(\rho+1) \underline{z}^H(\rho+1) = \frac{E_T}{N_t K} \mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H \quad \text{the received signal covariance calculations}$$

involve the use of transmitted signature sequence covariance matrices $\mathbf{Z} \mathbf{Z}^H$ and

$\mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H$ where signature sequence $\mathbf{Z}$ and $\mathbf{Z}_{\rho-1}$ and also $\mathbf{Z}_{\rho+1}$

matrices are defined earlier. In the covariance matrix calculations the terms $\mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1}^H$ are the complex conjugate transposes of $\mathbf{Z}$ and $\mathbf{Z}_{\rho-1}$ and also $\mathbf{Z}_{\rho+1}$ matrices respectively. In equation (6) the received signal covariance matrix calculations matrices $\mathbf{Z}\mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1}\mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1}\mathbf{Z}_{\rho+1}^H$ are the most computationally heavy calculations as the number of columns $N_s KN^{(x)}$ of matrices $\mathbf{Z} \in \mathbb{C}^{N_t N \times N_s KN^{(x)}}$ and $\mathbf{Z}_{\rho-1} \in \mathbb{C}^{N_t N \times N_s KN^{(x)}}$ and $\mathbf{Z}_{\rho+1} \in \mathbb{C}^{N_t N \times N_s KN^{(x)}}$ is proportional to the number of symbols $N^{(x)}$ over which that averaging is performed. However, as both the transmitter and receiver know over which symbol period which scrambling and channelization signature sequences are used for transmission for a given number of symbol observation periods the covariances matrices $\mathbf{Z}\mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1}\mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1}\mathbf{Z}_{\rho+1}^H$ can be calculated in advance at the transmitter and receiver and may be used in conjunction with Equation 1 to calculate the covariance matrix $\mathbf{C} = \frac{1}{N^{(x)}} \sum_{\rho=1}^{N^{(x)}} \mathbf{r}_{long}(\rho) \mathbf{r}_{long}^H(\rho)$ when the channel matrices $\mathbf{H}$, $\mathbf{H}_1$ and $\mathbf{H}_2$ are estimated at the MIMO receivers using the transmitted pilot signals.

By using a given number of symbols $N_s^{(x)}$, which $1 < N_s^{(x)} \leq N^{(x)}$, the minimum mean square error equalizer can be implemented for different periods of covariance matrix averaging. There will be a total of $\frac{N^{(x)}}{N_s^{(x)}}$ different epochs with each epoch having its

covariance matrix $\mathbf{C}(\xi)$ for the epoch number $\xi = 1, \dots, \frac{N^{(x)}}{N_s^{(x)}}$. Keeping $N_s^{(x)}$ low

gives a covariance matrix which is a better representative of the signature sequence over a shorter period, however, this results in a higher computational complexity at the receiver. A compromise between the computational complexity and the averaging period of the covariance matrix calculations over a shorter number of symbols will need to be made. The most computationally heavy matrix multiplication involved in calculating the covariance matrix $\mathbf{C}$ is the calculation of matrices $\mathbf{Z}\mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1}\mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1}\mathbf{Z}_{\rho+1}^H$. If the covariance matrix averaging is undertaken for different number of symbol periods $N_s^{(x)}$, as both the transmitter and receiver know

the transmission signature sequence matrices $\mathbf{Z}$, $\mathbf{Z}_{\rho-1}$ and also $\mathbf{Z}_{\rho+1}$ at the receiver the matrix dimensions can be fixed to be $\mathbf{Z} \in \mathbb{C}^{N_t N \times N_s KN_s^{(x)}}$ and $\mathbf{Z}_{\rho-1} \in \mathbb{C}^{N_t N \times N_s KN_s^{(x)}}$ and also $\mathbf{Z}_{\rho+1} \in \mathbb{C}^{N_t N \times N_s KN_s^{(x)}}$, if the average covariance matrix is estimated using $N_s^{(x)}$ symbol periods. Once the the dimensions $\mathbb{C}^{N_t N \times N_s KN_s^{(x)}}$ are fixed for the transmission signature sequence matrices $\mathbf{Z}$, $\mathbf{Z}_{\rho-1}$ and also $\mathbf{Z}_{\rho+1}$, covariance matrices $\mathbf{Z}(\xi)\mathbf{Z}^H(\xi)$, $\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi)$ and $\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi)$ for the signature sequence matrices for the ξ^{th} observation epoch can be calculated in advance to produce the averaged covariance matrix $\mathbf{C}(\xi)$ using:

$$\mathbf{C}(\xi) = \frac{E_T}{N_t KN^{(x)}} \left(\mathbf{H}\mathbf{Z}(\xi)\mathbf{Z}^H(\xi)\mathbf{H}^H + \mathbf{H}_1\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi)\mathbf{H}_1^H + \mathbf{H}_2\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi)\mathbf{H}_2^H + \frac{N_0 N_t KN^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

Equation 2

for observation periods $\xi = 1, \dots, \frac{N^{(x)}}{N_s^{(x)}}$. It is also assumed that for the averaging process the channel matrices $\mathbf{H}$, $\mathbf{H}_1$ and $\mathbf{H}_2$ are produced at the start of every TTI period and are assumed to be stationary for the duration of TTI period.

The covariance matrix of Equation 2 is then used to map the long vector $\underline{r}_{long}(\rho)$ at the output of the chip matched filter unit 202 to produce the short vector $\underline{r}_{short}(\rho)$ at the output of the vector mapping unit 203, as will now be discussed.

A minimum mean square error (MMSE) equalizer is firstly used to reduce the length $N_r(N+L-1)$ of the long received signal vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ to size a $N_t N$ vector and also to de-scramble the shortened vector elements to produce the short received vector $\underline{r}_{short}(\rho) \in \mathbb{C}^{N_t N}$. A vector mapping matrix $(\mathbf{C}^{-1}\mathbf{H}\mathbf{C}_n)^H \in \mathbb{C}^{N_t N \times N_r(N+L-1)}$ with the dimension $N_t N \times N_r(N+L-1)$ is used to shorten the vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ to size a $N_t N$ vector to produce the short received vector $\underline{r}_{short}(\rho) \in \mathbb{C}^{N_t N}$. For the receiver signal amplitude normalization the chip sequence normalization matrix $\mathbf{C}_n \in \mathbb{C}^{N_t N \times N_t N}$ is generated as follows:

$$\mathbf{C}_n = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} \right) \right) \right)^{-1} \in \mathbb{C}^{N_t N \times N_t N}$$

The shortened and de-scrambled received signal chip vector $\underline{r}_{short}(\rho) \in \mathbb{C}^{N_r N}$ is produced at the output of vector mapping unit 203 using:

$$\underline{r}_{short}(\rho) = \left(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n \right)^H \underline{r}_{long}(\rho) \in \mathbb{C}^{N_t N}$$

In a matrix form the mapping operation between the received long matrix $\mathbf{R}_{long} \in \mathbb{C}^{N_r (N+L-1) \times N^{(x)}}$ and the matrix $(\mathbf{S}_{ch,W})^*$ are used to de-scramble the short vector and produce the elements of the shortened received signal matrix $\mathbf{R}_{short} \in \mathbb{C}^{N_t N \times N^{(x)}}$ is given as

$$\begin{aligned} \mathbf{R}_{short} &= (\mathbf{S}_{ch,W})^* \left[\underline{r}_{short}(1), \dots, \underline{r}_{short}(\rho), \dots, \underline{r}_{short}(N^{(x)}) \right] \\ &= (\mathbf{S}_{ch,W})^* \left(\left(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n \right)^H \mathbf{R}_{long} \right) \end{aligned}$$

The shorted received signal vector $\underline{r}_{short}(\rho)$ is used together with the channelization codes to estimate the received symbol vector by the despreading unit 204 using the relationship

$$\begin{aligned} \underline{\hat{y}}(\rho) &= \mathbf{S}_{ch,W}^H \left(\text{Diag}(\underline{s}_{wsc}(\rho)) \right)^H \underline{r}_{short}(\rho) \\ &= \left[\hat{y}_1(\rho), \dots, \hat{y}_k(\rho), \dots, \hat{y}_{N_s K}(\rho) \right] \in \mathbb{C}^{N_s K} \end{aligned}$$

The estimated data sequences are then reorganized to produce the estimated data for each spreading sequence using the receiver mapping unit 205 and the decision unit 206.

Each of the above mentioned units of the transmitter and receiver apart from the actual MIMO transmitters 107a, 107b,, 107 N_t and receivers 201a, 201b,, 201 N_r are implemented in software.

Above, the operation of the system for determining the transmitted signal from the received signal has been discussed. Now the determination of the data rate for

transmission using a system value based method shall be discussed. The following procedure may be used in combination with the above process or with other processes.

Firstly, the system value calculation shall be defined for a system in which signature sequences are assigned on a symbol-by-symbol basis. Then, use of the System Value in order to determine which sub-set of transmit and receive antenna groups should be used when a large group of transmit and receive antennas are available shall be discussed.

It will be appreciated that by determining the System Values in the above mentioned data transmission apparatus the overall data rate achievable by the system can be improved.

The system value is a variable which is indicative of the characteristics of the channel over which data is transmitted. The system value is the normalized signal energy at the output of the despreading unit. The difference between normalized total energy of unity and the system value gives the mean square error at the output of the despreading unit. The ratio of the normalized energy, the system value, to the mean square error gives the signal to noise ratio at the output of the despreading unit.

The mean square error ε_k for the k^{th} symbol is:

$$\varepsilon_k = E\left(\left|\hat{y}_k(\rho) - y_k(\rho)\right|^2\right) = \frac{1}{\gamma_k + 1} = 1 - \lambda_k$$

where γ_k is the signal-to-noise ratio for the k^{th} symbol and $E(\bullet)$ identifies the expectation operation. The system value $\lambda_k = 1 - \varepsilon_k$ is related to the signal-to-interference-noise ratio by:

$$\gamma_k = \frac{\lambda_k}{1 - \lambda_k}$$

A modified version of the System Value optimization criterion is used to maximize the total rate when transmitting with spreading sequences which change every symbol

period. In current HSDPA standards the spreading sequence is changed every symbol period. The MMSE receivers can be configured to achieve average SNRs, which are equal at their output, when using equal energy transmission allocation based on the system value measurements. As the equal SNR scheme only requires equal rates to be allocated at the transmitter, the data rates can be estimated at the receiver when the transmission channel impulse response is available at the receiver. As groups of channels use the same transmitter data rate for each channel, the receiver only needs to report a reduced number of data rates to the transmitter by using the System Value calculations.

The System Value formulation involves the use of interference matrix for the current symbol defined as:

$$\mathbf{D} = \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} - \text{diag}(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H})$$

The channel impulse response $\mathbf{H}$ is converted to the residual channel interference matrix $\mathbf{D} \in \mathbb{C}^{N_t N \times N_t N}$, which is a square matrix for more efficient processing. In particular, as will be seen, by converting the channel impulse response $\mathbf{H}$ to matrix $\mathbf{D}$ later processing steps are optimised.

Here when the matrix $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \in \mathbb{C}^{N_t N \times N_r (N+L-1)}$ is multiplied with the channel matrix $\mathbf{H}$ the resultant matrix $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} \in \mathbb{C}^{N_t N \times N_t N}$ is a square matrix with diagonal elements equal to unity. When the diagonal terms $\text{diag}(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H})$ are subtracted from $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}$ the resultant matrix $\mathbf{D}$ identifies the Multiple Access Interference (MAI) between chip elements of the transmitted chip sequences. Hence the matrix $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \in \mathbb{C}^{N_t N \times N_r (N+L-1)}$ is multiplied with any matrix it changes the number of rows from $N_r (N+L-1)$ to $N_t N$.

When the transmission signal chip vector $\underline{z}(\rho) \in \mathbb{C}^{N_t N}$ is multiplied with the matrix $\mathbf{D} \in \mathbb{C}^{N_t N \times N_t N}$ the resultant vector $\underline{z}_{MAI}(\rho) \in \mathbb{C}^{N_t N}$ identifies the residual interference for each chip before despreading as follows:

$$\underline{z}_{MAI}(\rho) = \text{Diag}(\underline{s}_{wsc}(\rho))^H \mathbf{D} \underline{z}(\rho)$$

When the vector $\underline{z}_{MAI}(\rho)$ is de-spread using the channelization spreading sequence at the output of the k^{th} despreading unit, the error signal due to the multiple access interference is given as:

$$\xi_{MAI,k} = \underline{s}_k^H(\rho) \mathbf{D} \underline{z}(\rho)$$

Equation 3

The matrix $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \in \mathbb{C}^{N_t N \times N_r (N+L-1)}$ can also be used to change the number of rows of the intersymbol interference matrix $\mathbf{H}_1$ for the previous symbol period from $N_r(N+L-1)$ to $N_t N$ to produce the interference matrix $\mathbf{D}_1 \in \mathbb{C}^{N_t N \times N_t N}$ for the intersymbol interference from the previous symbols as follows:

$$\begin{aligned} \mathbf{D}_1 &= \left(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \left(\mathbf{I}_{N_r} \otimes (\mathbf{J}^T)^N \right) \mathbf{H} \right) \\ &= \left(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_1 \right) \in \mathbb{C}^{N_t N \times N_t N} \end{aligned}$$

When the transmission signal chip vector $\underline{z}(\rho-1) \in \mathbb{C}^{N_t N}$ is multiplied with the matrix $\mathbf{D}_1$ the resultant vector $\underline{z}_{ISI,1}(\rho) \in \mathbb{C}^{N_t N}$ identifies the residual ISI interference from the previous symbols as follows for each chip before despreading:

$$\underline{z}_{ISI,1}(\rho) = \text{Diag}(\underline{s}_{wsc}(\rho))^H \mathbf{D}_1 \underline{z}(\rho-1)$$

When the vector $\underline{z}_{ISI,1}(\rho)$ is de-spread using the channelization spreading sequence at the output of the k^{th} despreading unit, the error signal due to the previous symbol intersymbol interference is given as:

$$\xi_{ISI,1,k}(\rho) = \underline{s}_k^H(\rho) \mathbf{D}_1 \underline{z}(\rho-1)$$

Equation 4

The interference matrix for the intersymbol interference from the next symbols is also given in the following form:

$$\begin{aligned}\mathbf{D}_2 &= \left(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} (\mathbf{I}_{N_r} \otimes \mathbf{J}^N) \mathbf{H} \right) \\ &= \left(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_2 \right)\end{aligned}$$

When the transmission signal chip vector $\underline{z}(\rho+1) \in \mathbb{C}^{N_t N}$ is multiplied with the matrix $\mathbf{D}_2$ the resultant vector $\underline{z}_{ISI,2}(\rho) \in \mathbb{C}^{N_t N}$ identifies the residual interference vector for the next symbol period intersymbol interference as follows for each chip before despreading:

$$\underline{z}_{ISI,2}(\rho) = \text{Diag}(\underline{s}_{wsc}(\rho))^H \mathbf{D}_2 \underline{z}(\rho+1)$$

When the vector $\underline{z}_{ISI,2}(\rho)$ is de-spread using the channelization spreading sequence at the output of the k^{th} despreading unit, the error signal due to the next symbol intersymbol interference is given as:

$$\xi_{ISI,2,k}(\rho) = \underline{s}_k^H(\rho) \mathbf{D}_2 \underline{z}(\rho+1)$$

Equation 5

When using the matrix $\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \in \mathbb{C}^{N_t N \times N_r (N+L-1)}$ at the receiver, the noise vector $\underline{n}(\rho) \in \mathbb{C}^{N_r (N+L-1)}$ at the output of chip matched filters is also converted to the short noise vector $\underline{n}_{short}(\rho) \in \mathbb{C}^{N_t N}$ by the following operation:

$$\underline{n}_{short}(\rho) = \text{Diag}(\underline{s}_{wsc}(\rho))^H \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \underline{n}(\rho).$$

at the output of the de-scrambling unit. When the noise vector is de-spread with the channelization spreading sequence the noise signal at the output of the channelization despreading unit is given as:

$$\xi_{noise,k}(\rho) = \underline{s}_k^H(\rho) \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \underline{n}(\rho)$$

Equation 6

The error signal ξ_k for the k^{th} channelization despreading unit due to Multiple Access Interference and the Inter Symbol Interferences (ISI) from the previous and next symbols and also the noise component is obtained by adding the error signals (3), (4), (5), (6) to produce the error at the output of the de-spreading filter for the k^{th} channel as follows:

$$\begin{aligned}\xi_k &= \xi_{MAI,k} + \xi_{ISI,1,k}(\rho) + \xi_{ISI,2,k}(\rho) + \xi_{noise,k}(\rho) \\ &= \underline{s}_k^H(\rho) (\mathbf{D}\underline{z}(\rho) + \mathbf{D}_1\underline{z}(\rho-1) + \mathbf{D}_2\underline{z}(\rho+1) + \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \underline{n}(\rho))\end{aligned}$$

The Mean-Square-Error due to the MAI interference for the signal at the output of the de-spreader for the k^{th} channel is given as:

$$\varepsilon_{MAI,k} = E(\xi_{MAI,k} \xi_{MAI,k}^*) = \frac{E_T}{KN_s N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D} \mathbf{Z} \mathbf{Z}^H \mathbf{D}^H \mathbf{S}_k)$$

Equation 7

where $E(\bullet)$ refers to the expectation operation. Here the covariance matrix $\mathbf{Z}\mathbf{Z}^H$ is the signature sequence matrix $\mathbf{S}_k$ is the signature sequence matrix for channel k .

The Mean-Square-Error due to the Inter-Symbol-Interference from the previous symbol is given as:

$$\varepsilon_{ISI,1,k} = E(\xi_{ISI,1,k}(\rho) \xi_{ISI,1,k}^*(\rho)) = \frac{E_T}{KN_s N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_1 \mathbf{Z}_{\rho-1}^H \mathbf{Z}_{\rho-1} \mathbf{D}_1^H \mathbf{S}_k)$$

Equation 8

The Mean-Square-Error due to the Inter-Symbol-Interference from the next symbol is given as:

$$\varepsilon_{ISI,2,k} = E(\xi_{ISI,2,k}(\rho) \xi_{ISI,2,k}^*(\rho)) = \frac{E_T}{KN_s N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_2 \mathbf{Z}_{\rho+1}^H \mathbf{Z}_{\rho+1} \mathbf{D}_2^H \mathbf{S}_k)$$

Equation 9

The following noise vector despreading matrix $\mathbf{B} \in \mathbb{C}^{N^{(x)} \times N_r (N+L-1)}$ matrix is formulated as follows:

$$\mathbf{B} = \mathbf{S}_k^H \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \in \mathbb{C}^{N^{(x)} \times N_r (N+L-1)}$$

to calculate the Mean-Square-Error due to noise as follows:

$$\varepsilon_{noise,k} = E(\xi_{noise,k}(\rho) \xi_{noise,k}^*(\rho)) = \frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B} \mathbf{B}^H)$$

Equation 10

The normalized Mean-Square-Error due to multiple access interference, previous symbol and next symbol ISIs and also noise is obtained by adding equations (7), (8), (9) and (10 in the following form:

$$\begin{aligned} \varepsilon_k &= E(\xi_k \xi_k^*) = \varepsilon_{MAI,k} + \varepsilon_{ISI,1,k} + \varepsilon_{ISI,2,k} + \varepsilon_{noise,k} \\ &= \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D} \mathbf{Z} \mathbf{Z}^H \mathbf{D}^H \mathbf{S}_k) \\ &\quad + \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_1 \mathbf{Z}_{\rho-1}^H \mathbf{Z}_{\rho-1} \mathbf{D}_1^H \mathbf{S}_k) \\ &\quad + \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_2 \mathbf{Z}_{\rho+1}^H \mathbf{Z}_{\rho+1} \mathbf{D}_2^H \mathbf{S}_k) \\ &\quad + \frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B} \mathbf{B}^H) \end{aligned}$$

Hence the System Value, averaged over $N^{(x)}$ symbol period using spreading sequences, which change symbol by symbol basis is:

$$\begin{aligned} \lambda_k &= 1 - \varepsilon_k \\ &= 1 - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D} \mathbf{Z} \mathbf{Z}^H \mathbf{D}^H \mathbf{S}_k) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_1 \mathbf{Z}_{\rho-1}^H \mathbf{Z}_{\rho-1} \mathbf{D}_1^H \mathbf{S}_k) \end{aligned}$$

$$\begin{aligned}
& - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_2 \mathbf{Z}_{\rho+1}^H \mathbf{Z}_{\rho+1} \mathbf{D}_2^H \mathbf{S}_k) \\
& - \frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B} \mathbf{B}^H)
\end{aligned}$$

Now that it can be seen how the System Value for use in a system that applies signature sequences on a symbol-by-symbol basis has been described the determination of the data rate based on the determined System Value shall be considered.

The following procedure is based on the principle that the number N_t of transmitter antennas is less than the number N_r of receiver antennas. The spatial multiplexing

number N_s is hence equal to N_t . At the MIMO receivers, a total of $\begin{pmatrix} N_r \\ N_s \end{pmatrix}$

transceivers can be used independently. Each MIMO receiver has a different channel matrix $\mathbf{H}$ that is uniquely associated with the MIMO transmitter and receiver antennas belonging to the sub-set of the receiver antennas selected from the full set of receiver antennas.

At a given MIMO transmitter-receiver antenna combination, the transmission scrambling and channelization spreading sequences are known to both the transmitter and receiver. Before the packets are transmitted from the transmitter to the receiver, the transmission packet duration is expressed using the number $N^{(x)}$ of

symbols transmitted in a given packet. Hence at the receiver $\frac{N^{(x)}}{N_s^{(x)}}$ different sub-

packets, with an epoch number $\xi = 1, \dots, \frac{N^{(x)}}{N_s^{(x)}}$, may be formed each with a packet

length $N_s^{(x)}$. At the receiver the signature sequence covariance matrices $\mathbf{Z}(\xi) \mathbf{Z}^H(\xi)$, $\mathbf{Z}_{\rho-1}(\xi) \mathbf{Z}_{\rho-1}^H(\xi)$ and $\mathbf{Z}_{\rho+1}(\xi) \mathbf{Z}_{\rho+1}^H(\xi)$ are calculated in advance for the signature sequence matrices and the ξ^{th} observation epoch and for the equal energy

allocation $E_k = \frac{E_T}{N_t K}$ to produce the covariance matrix $\mathbf{C}(\xi)$ averaged over a total of $N_s^{(x)}$ symbol periods using:

$$\mathbf{C}(\xi) = \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z}(\xi) \mathbf{Z}^H(\xi) \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1}(\xi) \mathbf{Z}_{\rho-1}^H(\xi) \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1}(\xi) \mathbf{Z}_{\rho+1}^H(\xi) \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

Equation 11

for observation epochs $\xi = 1, \dots, \frac{N^{(x)}}{N_s^{(x)}}$. Using the covariance matrix $\mathbf{C}(\xi)$ for epoch ξ , the interference matrices averaged over $N_s^{(x)}$ symbol periods may be calculated using:

$$\mathbf{C}_n(\xi) = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H} \right) \right) \right)^{-1}$$

Equation 12

$$\mathbf{D}(\xi) = \mathbf{C}_n^H(\xi) \mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H} - \text{diag} \left(\mathbf{C}_n^H(\xi) \mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H} \right)$$

and

$$\mathbf{D}_1(\xi) = \left(\mathbf{C}_n^H(\xi) \mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H}_1 \right)$$

and also

$$\mathbf{D}_2(\xi) = \left(\mathbf{C}_n^H(\xi) \mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H}_2 \right)$$

and also the noise matrix

$$\mathbf{B}(\xi) = \mathbf{S}_k^H(\xi) \mathbf{C}_n^H(\xi) \mathbf{H}^H \mathbf{C}^{-1}(\xi)$$

The System Value for the ξ^{th} epoch and the k^{th} channel is then given by:

$$\lambda_k(\xi) = 1 - \varepsilon_k(\xi)$$

Equation 13

$$\begin{aligned} &= 1 - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace} \left(\mathbf{S}_k^H(\xi) \mathbf{D}(\xi) \mathbf{Z}(\xi) \mathbf{Z}^H(\xi) \mathbf{D}^H(\xi) \mathbf{S}_k(\xi) \right) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace} \left(\mathbf{S}_k^H(\xi) \mathbf{D}_1(\xi) \mathbf{Z}_{\rho-1}^H(\xi) \mathbf{Z}_{\rho-1}(\xi) \mathbf{D}_1^H(\xi) \mathbf{S}_k(\xi) \right) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace} \left(\mathbf{S}_k^H(\xi) \mathbf{D}_2(\xi) \mathbf{Z}_{\rho+1}^H(\xi) \mathbf{Z}_{\rho+1}(\xi) \mathbf{D}_2^H(\xi) \mathbf{S}_k(\xi) \right) \end{aligned}$$

$$-\frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B}(\xi) \mathbf{B}^H(\xi))$$

If the number $N_s^{(x)}$ of symbols is a small number, the System Value $\lambda_k(\xi)$ for each channel for epoch ξ will be almost identical for each group of K signature sequences out of N_t different groups of streams. Hence, the streams will have the following System Values.

$$\lambda_k(\xi) = \begin{cases} \lambda_1(\xi) & \text{for } 1 \leq k \leq K \text{ and the stream number } n_s = 1 \\ \lambda_{K+1}(\xi) & \text{for } K+1 \leq k \leq 2K \text{ and the stream number } n_s = 2 \\ \vdots & \vdots \\ \lambda_{(n_s-1)K+1}(\xi) & \text{for } (n_s-1)K+1 \leq k \leq n_s K \text{ and the stream number } n_s \\ \vdots & \vdots \\ \lambda_{(N_t-1)K+1}(\xi) & \text{for } (N_t-1)K+1 \leq k \leq N_t K \text{ and the stream number } n_s = N_t \end{cases}$$

On this basis, the discrete data rate for each stream is allocated as follows. For the set of bit rates $\{b_p\}_{p=1}^P$, the corresponding target system value is determined according to the following equation:

$$[\underline{\lambda}^*]_p = \frac{\Gamma(2^{b_p} - 1)}{1 - \Gamma(2^{b_p} - 1)} \text{ for } p = 1, \dots, P$$

For the stream group number n_s find $b_p(n_s)$ that satisfies $[\underline{\lambda}^*]_p \leq \lambda_{(n_s-1)K+1}(\xi) < [\underline{\lambda}^*]_{p+1}$. This determines the symbol rate allocation for stream group n_s . In the receiver implementation, the stream with the largest system value $\lambda_{(n_s-1)K+1}(\xi) - [\underline{\lambda}^*]_p$ is selected. Symbol rates are allocated by finding the b_p value that satisfies the equation $[\underline{\lambda}^*]_p \leq \lambda_{(n_s-1)K+1}(\xi) < [\underline{\lambda}^*]_{p+1}$.

$$\underline{\hat{y}}(\rho) = \mathbf{S}_{ch,W}^H (\text{Diag}(\underline{s}_{wsc}(\rho)))^H \underline{r}_{short}(\rho)$$

Equation 14

for the symbols received during epoch ξ . From the estimated signal's vector $\underline{\hat{y}}(\rho)$, the received signal matrix $\hat{\mathbf{X}}$ is generated as follows:

$$\hat{\mathbf{X}} = [\hat{\underline{x}}_1, \dots, \hat{\underline{x}}_k, \dots, \hat{\underline{x}}_K] = [\hat{y}(1), \dots, \hat{y}(\rho), \dots, \hat{y}(\rho)]^T$$

Equation 15

From the estimated signals matrix $\hat{\mathbf{X}}$ the decision matrix $\mathbf{X}_D$ channel decoders is generated in the following form:

$$\mathbf{X}_D = [\underline{x}_{1,D}, \dots, \underline{x}_{k,D}, \dots, \underline{x}_{K,D}]$$

Equation 16

and

$$\underline{x}_{k,D} = [x_{k,D}(1), \dots, x_{k,D}(\rho), \dots, x_{k,D}(N^{(x)})]^T$$

The system values can be used to implement successive interference cancellation receiver as shall now be described.

In the successive interference cancellation receiver implementation, the estimated symbols for the stream group with the largest value of $\lambda_{(n_s-1)K+1}(\xi) - [\underline{\lambda}^*]_p$ are considered. The channel decoders are used to decide which symbols are detected correctly for epochs $\xi = 1, \dots, \frac{N^{(x)}}{N_s}$. If up to n_x channels are detected correctly in the stream, the channel numbers $k_1, \dots, k_{n_x}$ are identified for the channels which have all the symbols detected correctly for $\rho = 1, \dots, N^{(x)}$. The signature sequence covariance matrices $\mathbf{Z}(\xi)\mathbf{Z}^H(\xi)$, $\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi)$ are updated and $\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi)$ using:

$$\mathbf{Z}(\xi)\mathbf{Z}^H(\xi) = \mathbf{Z}(\xi)\mathbf{Z}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_k(\xi)\mathbf{S}_k^H(\xi)$$

$$\text{where } \mathbf{S}_k(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} + 1 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} + \rho \right) \right] \text{ and}$$

$$\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) = \mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho-1)}(\xi)\mathbf{S}_{k,(\rho-1)}^H(\xi)$$

where $\mathbf{S}_{k,(\rho-1)}(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} + \rho-1 \right) \right]$ and

$$\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) = \mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho+1)}(\xi)\mathbf{S}_{k,(\rho+1)}^H(\xi)$$

where $\mathbf{S}_{k,(\rho+1)} = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} + 2 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s} + \rho+1 \right) \right]$ for the updated signature sequence matrices, the received signal covariance matrix $\mathbf{C}(\xi)$ and $\mathbf{C}_n(\xi)$ are calculated using Equations 11 and 12. The received long matrix $\mathbf{R}_{long}$ is iteratively updated for the correctly detected symbol sequences in the channels updated using:

$$\mathbf{R}_{long} = \mathbf{R}_{long} - E_k \sum_{k=k_1}^{k_{n_x}} \Phi_k$$

Equation 16

$$= \left[\underline{r}_{long}(1), \dots, \underline{r}_{long}(\rho), \dots, \underline{r}_{long}(N^{(x)}) \right]$$

where

$$\Phi_k = \left[\underline{\phi}_k(1), \dots, \underline{\phi}_k(\rho), \dots, \underline{\phi}_k(N^{(x)}) \right]$$

Equation 17

with

$$\underline{\phi}_k(\rho) = \left[\mathbf{H}_{\underline{s}_k}(\rho)x_{k,D}(\rho) + \mathbf{H}_{1\underline{s}_k}(\rho-1)x_{k,D}(\rho-1) + \mathbf{H}_{2\underline{s}_k}(\rho+1)x_{k,D}(\rho+1) \right]$$

after updating the receiver long matrix using Equation 16 the vector for the estimated symbols are generated using Equation 14 The successive interference cancellation is iteratively run for the remaining data streams in the order with the highest remaining values of $\lambda_{(n_s-1)K+1}(\xi) - [\underline{\lambda}^*]_p$.

A Successive Interference Cancellation (SIC) receiver shall now be described with reference to Figure 2. The SIC receiver uses the System Value discrete rate allocation scheme just described. In Figure 2, the signal transmitted from MIMO transmitter, as shown in Figure 1, is received at the receiver 300 by the MIMO antennas 301a, 301b,301 N_r . Assuming the clocks at the transmitter and receiver are fully synchronized, the received signal is then frequency down converted, filtered and sampled at chip period intervals by the chip matched filter unit 302. The received signal long chip vector $\underline{r}_{long}(\rho) \in \mathbb{C}^{N_r(N+L-1)}$ is concatenated to produce the long received signal matrix $\mathbf{R}_{long}$ by the concatenation unit 303. The successive interference cancellation matrix handling unit 304, processes the long received matrix $\mathbf{R}_{long}$ to remove signals iteratively using Equation 16 and the signal removal matrix Φ_k given in Equation 17. The generation of the signal removal matrix Φ_k is handled by units 304, 305 and 307. The estimated signal vector generator unit 305 handles the generation of $\underline{y}(\rho)$ using Equation 14 and produces the estimated signals matrix $\hat{\mathbf{X}}$ using Equation 15. The decision unit 307 produces the decisions matrix $\mathbf{X}_D$ given in Equation 16 using the estimated symbols matrix $\hat{\mathbf{X}}$. The decisions matrix $\mathbf{X}_D$ is passed to the unit 304 to produce the removal matrix Φ_k using Equation 17 and hence the long received matrix $\mathbf{R}_{long}$ for the iterative SIC operations. The estimated symbols matrix $\hat{\mathbf{X}}$ is also passed from unit 305 to produce the estimated symbols vector $\underline{y}(\rho)$. The binary data unit 308 takes the symbol decision matrix $\mathbf{X}_D$ from unit 307 and produces the $(N_U \times 1)$ -dimensional binary decisions vector $\underline{u}_{k,D}$.

In alternative arrangement, the system values given in Equation 13 are used to choose a subset of MIMO receiver antennas with the subset cardinality is equal to N_s if the number of receiver antennas is N_r and $N_s = N_t$. In other words, if there are more transmit antennas than receiver antennas then the antennas to be used, i.e. those with the best channels, can be selected for use in transmitting to the receiver.

The system values are used to determine which subset of the receiver antennas may be used over the epoch ξ given that the number of bits to be transmitted over each

channelization code in a stream group is already determined using the rate allocation algorithm described above. The idea of using different set of receiver antennas is based on the fact that System Value obtained over a given large number of symbols $N_s^{(x)}$ will be constant for a given stream. However the System Value will vary from channelization code to channelization code and also from symbol period to symbol period over a short number of symbols $N_s^{(x)}$. If the system value is low for a channelization code for a given stream for a given subset of MIMO receiver antennas, the system value for the same channelization code and epoch for a different subset of MIMO receiver antennas. Hence, the System Value calculations described above may be used to identify how different subset of MIMO receiver antennas can be combined at the receiver.

The various methods described above may be implemented by a computer program. The computer program may include computer code arranged to instruct a computer to perform the functions of one or more of the various methods described above. The computer program and/or the code for performing such methods may be provided to an apparatus, such as a computer, on a computer readable medium or computer program product. The computer readable medium could be, for example, an electronic, magnetic, optical, electromagnetic, infrared, or semiconductor system, or a propagation medium for data transmission, for example for downloading the code over the Internet. Alternatively, the computer readable medium could take the form of a physical computer readable medium such as semiconductor or solid state memory, magnetic tape, a removable computer diskette, a random access memory (RAM), a read-only memory (ROM), a rigid magnetic disc, and an optical disk, such as a CD-ROM, CD-R/W or DVD.

An apparatus such as a computer may be configured in accordance with such code to perform one or more processes in accordance with the various methods discussed herein. Such an apparatus may take the form of a data processing system. Such a data processing system may be a distributed system. For example, such a data processing system may be distributed across a network.

Claims:

1. A method for determining a transmitted signal received at a receiver in a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple-output channels over which the signal is transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis every symbol period ρ , the method comprising:

receiving a signal $r_{long}(\rho)$ corresponding to a signal transmitted over a channel, wherein each symbol of the transmitted signal is spread using a respective spreading sequence $S(\rho)$;

determining an estimated channel characteristic H indicative of effects of the channel over which the received signal $r_{long}(\rho)$ has been transmitted on the transmitted signal; and

determining the transmitted signal $r_{short}(\rho)$ from the received signal $r_{long}(\rho)$ in accordance with the estimated channel characteristic H and the spreading sequences used to spread the symbols.

2. The method according to claim 1, wherein the method further comprises:

determining a combined matrix Z of all spreading sequences $S(\rho)$ used to spread the symbols of the transmitted signal; and

determining a covariance matrix C and a normalization matrix C_n in accordance with:

the estimated channel characteristic H ; and

a multiplication of the combined matrix of the spreading sequences Z and the hermitian of the combined matrix of the spreading sequences Z^H ;

wherein

the transmitted signal is determined from the received signal $r_{long}(\rho)$ by multiplying the received signal $r_{long}(\rho)$ with:

the inverse of the covariance matrix C ;

the hermitian of the channel matrix H ; and

the hermitian of the normalization matrix C_n .

3. The method according to claim 1 or claim 2, wherein the covariance matrix $\mathbf{C}$ is defined as follows:

$$\mathbf{C} = \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z} \mathbf{Z}^H \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

wherein E_T is a total transmission energy, $N_t = N_s$ is a total number of MIMO transmission antennas, $N_s K$ is a total number of spreading sequences, $N^{(x)}$ is a total number of transmitted symbols used to calculate the covariance matrix $\mathbf{C}$, N_0 is a noise power spectral density, $\mathbf{I}_{N_r(N+L-1)}$ is an identity matrix, the previous symbol period channel interference matrix $\mathbf{H}_1$ is produced from the estimated channel matrix $\mathbf{H}$ using:

$$\mathbf{H}_1 = \left(\mathbf{I}_{N_r} \otimes (\mathbf{J}^T)^N \right) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

and wherein the channel matrix for the next symbol period is produced using:

$$\mathbf{H}_2 = \left(\mathbf{I}_{N_r} \otimes \mathbf{J}^N \right) \mathbf{H} \in \mathbb{C}^{N_r(N+L-1) \times N_t N}$$

wherein the $(N+L-1) \times (N+L-1)$ -dimensional matrix $\mathbf{J}_{N+L-1}$ is given as

$$\mathbf{J}_{N+L-1} = \begin{bmatrix} \underline{0}_{(N+L-2)}^T & 0 \\ \mathbf{I}_{(N+L-2)} & \underline{0}_{(N+L-2)} \end{bmatrix} \text{ and also the operation } \mathbf{J}^N \text{ refers to the } N^{\text{th}} \text{ power of}$$

$\mathbf{J}_{N+L-1}$ further the operator $\otimes$ corresponds to Kroneker product, $\mathbf{Z} \mathbf{Z}^H$ are transmitted signature sequence matrices and $\mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H$ and also $\mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H$ are the transmitted signature sequence matrices for the current previous and next symbol periods.

4. The method according claim 3, wherein the covariance matrices of the transmitted signature sequence matrices $\mathbf{Z} \mathbf{Z}^H$ and $\mathbf{Z}_{\rho-1} \mathbf{Z}_{\rho-1}^H$ and $\mathbf{Z}_{\rho+1} \mathbf{Z}_{\rho+1}^H$ for the current, previous and next symbol periods are determined according to the following:

$$\begin{aligned}\mathbf{Z} &= [\mathbf{S}(2), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)} + 1)] \\ \mathbf{Z}_{\rho-1} &= [\mathbf{S}(1), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)})] \\ \mathbf{Z}_{\rho+1} &= [\mathbf{S}(3), \dots, \mathbf{S}(\rho), \dots, \mathbf{S}(N^{(x)} + 2)]\end{aligned}$$

wherein a spreading sequence matrix $\mathbf{S}(\rho)$ of all spreading symbols for each symbol period ρ is determined according to:

$$\mathbf{S}(\rho) = \text{diag}(\underline{s}_{wsc}(\rho)) \mathbf{S}_{ch,W}.$$

wherein $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned}\mathbf{S}_{wsc} &= (\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc}) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)]\end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss}$ $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$ wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

5. The method according to claim 4, wherein, for a Successive Interference Cancellation, SIC, receiver the signature sequence covariance matrices $\mathbf{Z}(\xi)\mathbf{Z}^H(\xi)$, $\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi)$ and $\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi)$ are updated using:

$$\mathbf{Z}(\xi)\mathbf{Z}^H(\xi) = \mathbf{Z}(\xi)\mathbf{Z}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_k(\xi)\mathbf{S}_k^H(\xi)$$

wherein $\mathbf{S}_k(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + 1 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho \right) \right]$ and

$$\mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) = \mathbf{Z}_{\rho-1}(\xi)\mathbf{Z}_{\rho-1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho-1)}(\xi)\mathbf{S}_{k,(\rho-1)}^H(\xi)$$

wherein $\mathbf{S}_{k,(\rho-1)}(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho - 1 \right) \right]$ and

$$\mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) = \mathbf{Z}_{\rho+1}(\xi)\mathbf{Z}_{\rho+1}^H(\xi) - \sum_{k=k_1}^{k_{n_x}} \mathbf{S}_{k,(\rho+1)}(\xi)\mathbf{S}_{k,(\rho+1)}^H(\xi)$$

wherein $\mathbf{S}_{k,(\rho+1)}(\xi) = \left[\underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + 2 \right), \dots, \underline{s}_k \left((\xi-1) \frac{N^{(x)}}{N_s^{(x)}} + \rho + 1 \right) \right]$ for the updated

signature sequence matrices, and $\mathbf{C}(\xi)$ and $\mathbf{C}_n(\xi)$ are the received signal covariance matrix and the normalization matrix respectively for use in the implementation of successive interference cancellation receivers and wherein ξ is the epoch number, $N^{(x)}$ and $N_s^{(x)}$ are the number of symbols transmitted in a block and a

set of the blocks respectively and wherein the covariance matrix $\mathbf{C}(\xi)$ is determined according to:

$$\mathbf{C}(\xi) = \frac{E_T}{N_t K N^{(x)}} \left(\mathbf{H} \mathbf{Z}(\xi) \mathbf{Z}^H(\xi) \mathbf{H}^H + \mathbf{H}_1 \mathbf{Z}_{\rho-1}(\xi) \mathbf{Z}_{\rho-1}^H(\xi) \mathbf{H}_1^H + \mathbf{H}_2 \mathbf{Z}_{\rho+1}(\xi) \mathbf{Z}_{\rho+1}^H(\xi) \mathbf{H}_2^H + \frac{N_0 N_t K N^{(x)}}{E_T} \mathbf{I}_{N_r(N+L-1)} \right)$$

and the normalization matrix is determined according to:

$$\mathbf{C}_n(\xi) = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1}(\xi) \mathbf{H} \right) \right) \right)^{-1}$$

6. The method according to claim 5, wherein, for the Successive Interference Cancellation, SIC, receiver the received long matrix $\mathbf{R}_{long}$ is iteratively updated for the correctly detected symbol sequences in the channels updated using:

$$\begin{aligned} \mathbf{R}_{long} &= \mathbf{R}_{long} - E_k \sum_{k=k_1}^{k_{n_x}} \Phi_k \\ &= \left[\underline{r}_{long}(1), \dots, \underline{r}_{long}(\rho), \dots, \underline{r}_{long}(N^{(x)}) \right] \end{aligned}$$

wherein the long matrix Φ_k for the correctly detected symbols is determined according to:

$$\Phi_k = \left[\underline{\phi}_k(1), \dots, \underline{\phi}_k(\rho), \dots, \underline{\phi}_k(N^{(x)}) \right]$$

wherein the k^{th} channel's chip sequence vector $\underline{\phi}_k(\rho)$ for the symbol period ρ and spreading sequence $\underline{s}_k(\rho)$ is determined according to:

$$\underline{\phi}_k(\rho) = \left[\mathbf{H} \underline{s}_k(\rho) x_{k,D}(\rho) + \mathbf{H}_1 \underline{s}_k(\rho-1) x_{k,D}(\rho-1) + \mathbf{H}_2 \underline{s}_k(\rho+1) x_{k,D}(\rho+1) \right]$$

for successive interference cancellation receivers.

7. The method according to any one of claims 2 to 6, wherein the transmitted signal $\underline{r}_{short}(\rho)$ is determined according to the following:

$$\underline{r}_{short}(\rho) = \left(\left(\mathbf{C}^{-1} \mathbf{H} \mathbf{C}_n \right)^H \underline{r}_{long}(\rho) \right)$$

Wherein $\mathbf{C}^{-1}$ is an inverse of the covariance matrix, $\underline{r}_{long}(\rho)$ is the received signal, and the normalization matrix $\mathbf{C}_n$ is determined according to:

$$\mathbf{C}_n = \left(\text{Diag} \left(\text{Diag} \left(\mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} \right) \right) \right)^{-1}.$$

8. The method according to claim 7, further comprising estimating a received symbol vector using the following relationship:

$$\begin{aligned} \hat{\underline{y}}(\rho) &= \mathbf{S}_{ch,W}^H \left(\text{Diag}(\underline{s}_{wsc}(\rho)) \right)^H \underline{r}_{short}(\rho) \\ &= [\hat{y}_1(\rho), \dots, \hat{y}_k(\rho), \dots, \hat{y}_{N_t K}(\rho)] \end{aligned}$$

Wherein $\underline{r}_{short}(\rho)$ is the transmitted signal, $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned} \mathbf{S}_{wsc} &= \left(\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc} \right) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)] \end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$, wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

9. The method according to any preceding claim, wherein the estimated channel characteristic is defined by:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1,1)} & \dots & \mathbf{H}^{(1,N_t)} \\ \vdots & \mathbf{H}^{(n_r,n_t)} & \vdots \\ \mathbf{H}^{(N_r,1)} & \dots & \mathbf{H}^{(N_r,N_t)} \end{bmatrix}$$

wherein the channel convolution matrix $\mathbf{H}^{(n_r,n_t)}$ between the transmit antenna n_t and the receive antenna n_r is defined by:

$$\mathbf{H}^{(n_r,n_t)} = \begin{bmatrix} h_0^{(n_r,n_t)} & 0 & 0 \\ \vdots & h_0^{(n_r,n_t)} & \vdots \\ h_{L-1}^{(n_r,n_t)} & \vdots & \ddots & 0 \\ 0 & h_{L-1}^{(n_r,n_t)} & h_0^{(n_r,n_t)} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & h_{L-1}^{(n_r,n_t)} \end{bmatrix}$$

wherein a channel impulse response vector between a transmit antenna n_t and a receive antenna n_r is defined by channel impulse response function:

$$\underline{h}^{(n_r, n_t)} = \begin{bmatrix} h_0^{(n_r, n_t)} & \dots & h_{L-1}^{(n_r, n_t)} \end{bmatrix}^T.$$

10. The method according to any preceding claim, wherein the estimated channel characteristic is a channel impulse response function.

11. The method according to any preceding claim, wherein the received signal is

defined by:

$$\underline{r}_{long}(\rho) = \mathbf{H}\underline{z}(\rho) + \mathbf{H}_1\underline{z}(\rho-1) + \mathbf{H}_2\underline{z}(\rho-1) + \underline{n}(\rho)$$

wherein the vector $\underline{n}(\rho)$ corresponds to the received noise samples at the output of receiver chip matched filter and $\underline{z}(\rho)$ is the transmit chip vector for the current symbol period which is determined according to:

$$\underline{z}(\rho) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho) \underline{y}(\rho)$$

wherein $\underline{z}(\rho-1)$ is the transmit chip vector for the previous symbol causing intersymbol interference which is determined according to:

$$\underline{z}(\rho-1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho-1) \underline{y}(\rho-1)$$

wherein the transmitted chip vector $\underline{z}(\rho+1)$ for the next $(\rho^{th} + 1)$ symbol period is:

$$\underline{z}(\rho+1) = \sqrt{\frac{E_T}{N_t K}} \mathbf{S}(\rho+1) \underline{y}(\rho+1)$$

wherein $\underline{y}(\rho) = [y_1(\rho), \dots, y_k(\rho), \dots, y_K(\rho)]^T$ is the transmitted symbols vector which contains the symbols, over the symbol period $\rho = 1, \dots, N^{(x)}$, wherein $N^{(x)}$ is the

number of transmitted symbols in a transmission block wherein the entire block of transmission is represented as an $(N^{(x)} \times N_t K)$ dimensional transmit symbol matrix using the following relationship:

$$\mathbf{X} = [\underline{x}_1, \dots, \underline{x}_k, \dots, \underline{x}_{N_t K}] = [\underline{y}(1), \dots, \underline{y}(\rho), \dots, \underline{y}(N^{(x)})]^T.$$

wherein $\mathbf{X}$ is the matrix containing transmitted symbols, $\underline{x}_k$ is the vector containing symbols over channel k , $N_t K$ is the total number of spreading sequences used in the downlink MIMO system, and wherein the $(N^{(x)} \times 1)$ -dimensional symbol vector $\underline{x}_k$ is constructed using:

$$\underline{x}_k = [x_k(1), \dots, x_k(\rho), \dots, x_k(N^{(x)})]^T$$

for each channel $k = 1, \dots, N_t K$.

12. A method for determining data transmission rates for transmitting a signal over a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple output channels over which the signal is to be transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis, the method comprising:

receiving a signal comprising a plurality of symbols, each symbol having been spread by a spreading sequence of a plurality of spreading sequences;

determining, for each symbol of the plurality of symbols, a system value λ_k indicative of a signal-to-noise ratio associated with the respective spreading sequence used to spread the symbol; and

averaging the plurality of system values associated with the plurality of spreading sequences to determine an averaged system value for determining a data rate for transmission of symbols using the plurality of spreading sequences.

13. The method according to claim 12, further comprising:

determining a data rate to be associated with the number of signature sequences for transmission in accordance with the average system value.

14. The method according to claim 13, wherein the data rate for each stream is determined by finding the discrete data rate value b_p which satisfies the following equation:

$$[\underline{\lambda}^*]_p \leq \lambda_{(n_s-1)K+1}(\xi) < [\underline{\lambda}^*]_{p+1}.$$

wherein $\lambda_{(n_s-1)K+1}(\xi)$ is the system value, which is for a streaming group number n_s

and for symbol epoch ξ and a target system value $[\underline{\lambda}^*]_p$, which is determined

according to $[\underline{\lambda}^*]_p = \frac{\Gamma(2^{b_p} - 1)}{1 - \Gamma(2^{b_p} - 1)}$ for $p = 1, \dots, P$ for a chosen discrete bit rate of b_p

wherein Γ is the gap value.

15. The method according to claim 13 or claim 14, further comprising estimating the symbols in the received signal in accordance with the following equation:

$$\hat{\underline{y}}(\rho) = \mathbf{S}_{ch,W}^H (\text{Diag}(\underline{s}_{wsc}(\rho)))^H \underline{r}_{short}(\rho)$$

Wherein $\underline{r}_{short}(\rho)$ is the transmitted signal, $\mathbf{S}_{ch,W}$ is an extended channelization matrix and $\underline{s}_{wsc}(\rho)$ is a scrambling sequence vector for symbol period ρ wherein the scrambling sequence vector $\underline{s}_{wsc}(\rho)$ is determined according to:

$$\begin{aligned} \mathbf{S}_{wsc} &= (\text{diag}(\mathbf{I}_{N_t}) \otimes \mathbf{S}_{sc}) \\ &= [\underline{s}_{wsc}(1), \dots, \underline{s}_{wsc}(\rho), \dots, \underline{s}_{wsc}(N^{(x)} + 2)] \end{aligned}$$

and wherein the extended channelization matrix $\mathbf{S}_{ch,W}$ is determined according to:

$$\mathbf{S}_{ch,W} = (\mathbf{W} \otimes \mathbf{S}_{ch,K}) = [\underline{s}_{wch,1}, \dots, \underline{s}_{wch,k}, \dots, \underline{s}_{wch,K}]$$

wherein a total of K channelization spreading sequences $\mathbf{S}_{ch,K}$ are generated from a total of N Hadamard codes with spreading sequence processing gain N , and a signature sequence matrix $\mathbf{S}_{ss} = [\underline{s}_{ss,1}, \dots, \underline{s}_{ss,k}, \dots, \underline{s}_{ss,N}]$ wherein a channelization spreading sequence vector $\underline{s}_{ss,k}$ associated with the signature sequence matrix $\mathbf{S}_{ss}$ has the property:

$$\underline{s}_{ss,i}^H \underline{s}_{ss,j} = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

and wherein a transmitter precoding matrix $\mathbf{W}$ is determined according to:

$$\mathbf{W} = [\underline{w}_1, \dots, \underline{w}_{n_t}, \dots, \underline{w}_{N_s}]$$

wherein the precoding vectors $\underline{w}_{n_t}$ for $n_t = 1, \dots, N_s$ with the property:

$$\underline{w}_i^H \underline{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases}$$

are the beam steering vectors used at the transmitter antennas.

16. The method according to any one of claims 12 to 15, wherein the system value is defined by:

$$\begin{aligned} \lambda_k &= 1 - \varepsilon_k \\ &= 1 - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D} \mathbf{Z} \mathbf{Z}^H \mathbf{D}^H \mathbf{S}_k) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_1 \mathbf{Z}_{\rho-1}^H \mathbf{Z}_{\rho-1} \mathbf{D}_1^H \mathbf{S}_k) \\ &\quad - \frac{E_T}{N_t K N^{(x)} N^{(x)}} \text{Trace}(\mathbf{S}_k^H \mathbf{D}_2 \mathbf{Z}_{\rho+1}^H \mathbf{Z}_{\rho+1} \mathbf{D}_2^H \mathbf{S}_k) \\ &\quad - \frac{N_0}{N^{(x)}} \text{Trace}(\mathbf{B} \mathbf{B}^H) \end{aligned}$$

wherein $\mathbf{D}$ is the estimated channel characteristic converted to a square matrix

representative of intersymbol interference, which is defined as follows:

$$\mathbf{D} = \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H} - \text{diag}(\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H})$$

and a signature sequence matrix for channel k is defined as:

$$\mathbf{S}_k = \text{diag}(\underline{s}_{wch,k}) \mathbf{S}_{wsc} = [\underline{s}_k(1), \dots, \underline{s}_k(\rho), \dots, \underline{s}_k(N^{(x)})]$$

wherein an intersymbol matrix $\mathbf{D}_1$ for the interference from the previous symbol is determined according to:

$$\mathbf{D}_1 = (\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_1)$$

and an intersymbol interference matrix $\mathbf{D}_2$ for the interference from the next symbol is determined according to:

$$\mathbf{D}_2 = (\mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1} \mathbf{H}_2)$$

and $\mathbf{B}$ is a noise vector despreading matrix, which is determined according to:

$$\mathbf{B} = \mathbf{S}_k^H \mathbf{C}_n^H \mathbf{H}^H \mathbf{C}^{-1}.$$

17. The method according to any one of claims 12 to 16, wherein the plurality of spreading sequences with which the average system value is associated corresponds to a first combination of receiver and antenna channels, wherein the method further comprises:

determining an average system value for one or more other combinations of receiver and antenna channels; and

using the receiver and antenna channel having the highest system value for transmission.

18. The method according to any one of claims 1 to 11, further comprising the method of any one of claims 12 to 17.
19. Apparatus arranged to perform the method according to any preceding claim.
20. The apparatus according to claim 19, wherein the apparatus is a receiver device.
21. The apparatus according to claim 19, wherein the apparatus is a radio transmission system comprising one or more transmitters and one or more receivers.
22. A computer program product arranged, in use, to instruct a computer to perform the method according to any one of claims 1 to 18.

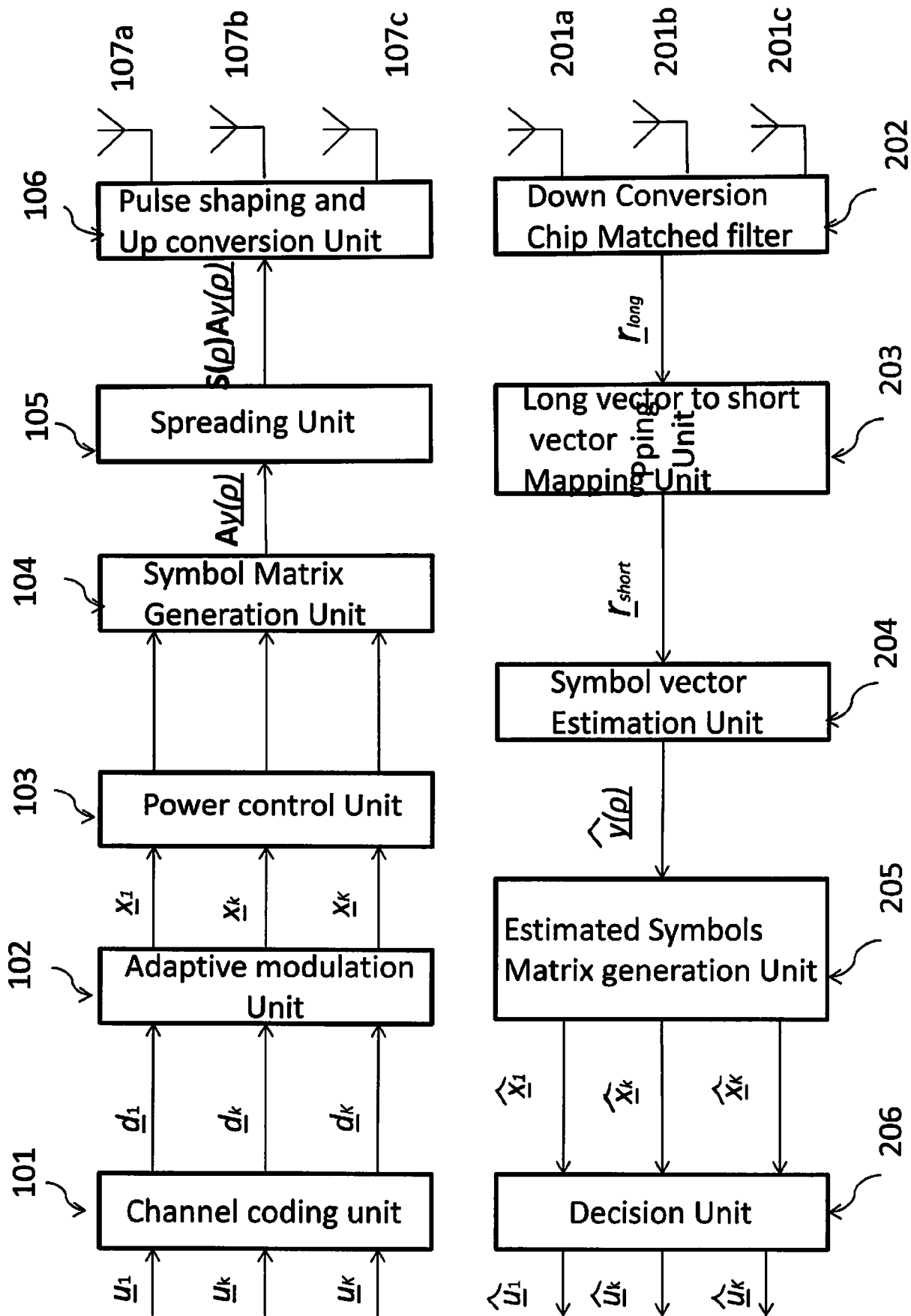
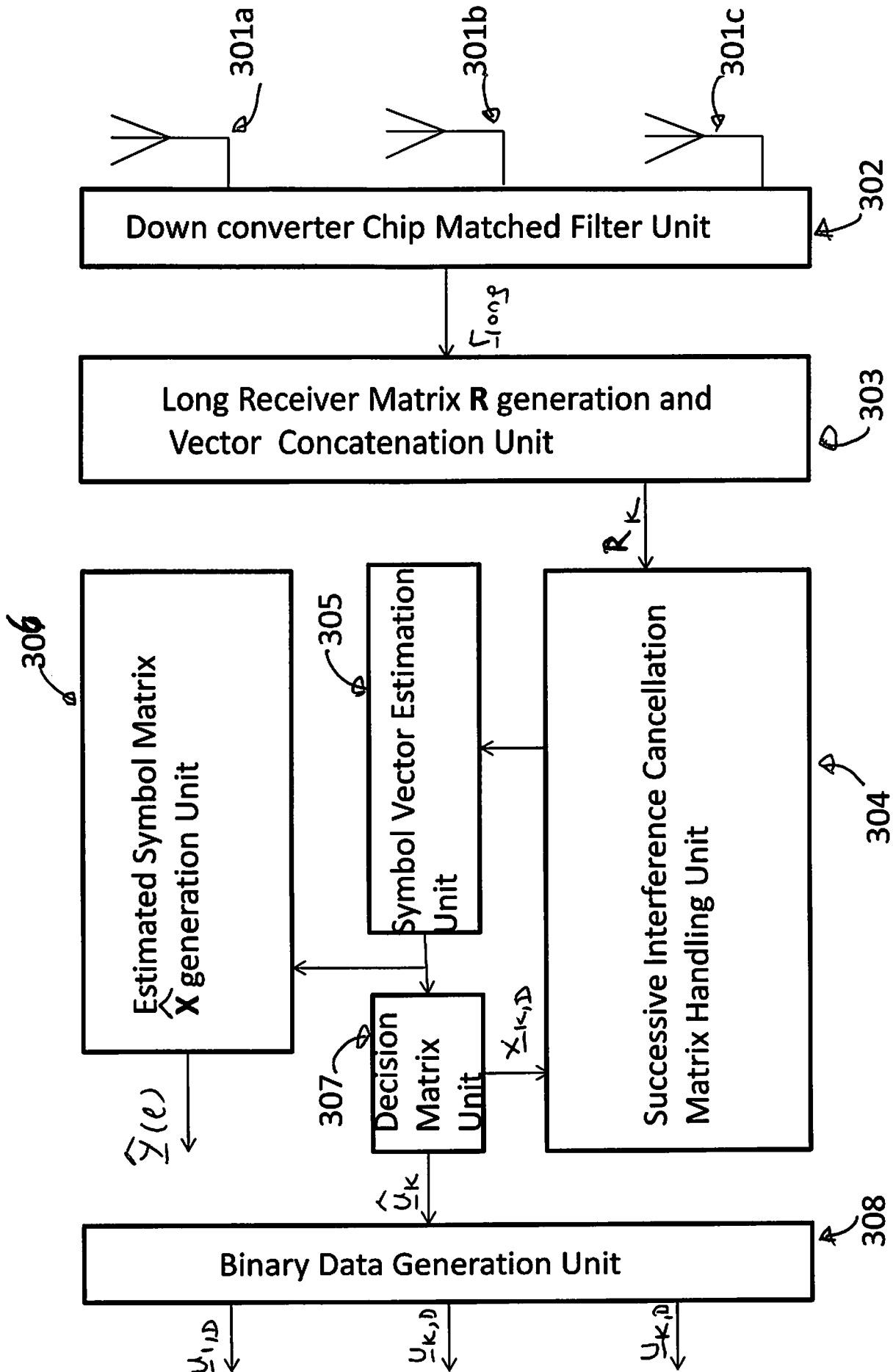


Figure 1



INTERNATIONAL SEARCH REPORT

International application No

PCT/GB2014/051937

A. CLASSIFICATION OF SUBJECT MATTER
 INV. H04B1/7105 H04B7/04
 ADD.

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)

H04B 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 practicable, search terms used)

EPO-Internal, WPI Data

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 2006/256843 A1 (GRANT STEPHEN J [US] ET AL) 16 November 2006 (2006-11-16)	1
A	paragraph [0021] paragraph [0024] paragraph [0026] paragraph [0029] paragraph [0032] paragraph [0043] paragraph [0045]	2-11, 18-22
A	----- EP 1 821 445 A1 (SIEMENS SPA ITALIANA [IT]) 22 August 2007 (2007-08-22) paragraph [0001] paragraph [0002] - paragraph [0004] paragraph [0008] paragraph [0018] paragraph [0023] paragraph [0029] - paragraph [0032] -----	1-11, 18-22



Further documents are listed in the continuation of Box C.



See patent family 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 application or patent 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

2 October 2014

Date of mailing of the international search report

28/11/2014

Name and mailing address of the ISA/

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INTERNATIONAL SEARCH REPORT

Information on patent family members

International application No

PCT/GB2014/051937

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US 2006256843 A1	16-11-2006	EP 1894311 A2	05-03-2008
		US 2006256843 A1	16-11-2006
		WO 2006132592 A2	14-12-2006

EP 1821445 A1	22-08-2007	CN 101056286 A	17-10-2007
		EP 1821445 A1	22-08-2007
		EP 1994667 A1	26-11-2008
		US 2010067366 A1	18-03-2010
		WO 2007093379 A1	23-08-2007

INTERNATIONAL SEARCH REPORT

International application No.
PCT/GB2014/051937

Box No. II Observations where certain claims were found unsearchable (Continuation of item 2 of first sheet)

This international search report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:
2. ☐ Claims Nos.:
because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:
3. ☐ Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box No. III Observations where unity of invention is lacking (Continuation of Item 3 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

see additional sheet

1. ☐ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.
2. ☐ As all searchable claims could be searched without effort justifying an additional fees, this Authority did not invite payment of additional fees.
3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:
4. ☒ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

1-11(completely); 18-22(partially)

Remark on Protest

- ☐ The additional search fees were accompanied by the applicant's protest and, where applicable, the payment of a protest fee.
- ☐ The additional search fees were accompanied by the applicant's protest but the applicable protest fee was not paid within the time limit specified in the invitation.
- ☐ No protest accompanied the payment of additional search fees.

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

This International Searching Authority found multiple (groups of) inventions in this international application, as follows:

1. claims: 1-11(completely); 18-22(partially)

A method for determining a transmitted signal received at a receiver in a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple-output channels over which the signal is transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis every symbol period p , the method comprising:
receiving a signal corresponding to a signal transmitted over a channel;
determining an estimated channel characteristic H ; and
determining the transmitted signal from the received signal in accordance with the estimated channel characteristic H and the spreading sequences used to spread the symbols.

2. claims: 12-17(completely); 18-22(partially)

A method for determining data transmission rates for transmitting a signal over a radio transmission system having a plurality of parallel single-input single-output or multiple-input multiple output channels over which the signal is to be transmitted, the signal comprising one or more symbols, the symbols being spread prior to transmission by a plurality of spreading sequences on a symbol-by-symbol basis, the method comprising:
receiving a signal comprising a plurality of symbols;
determining, for each symbol of the plurality of symbols, a system value indicative of a signal-to-noise ratio associated with the respective spreading sequence used to spread the symbol; and
averaging the plurality of system values associated with the plurality of spreading sequences to determine an averaged system value for determining a data rate for transmission of symbols using the plurality of spreading sequences.
