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(73) Haltija - Innehavare - Holder

1• THALES , 4 rue de la Verrerie , 92190 Meudon , (FR)

(72) Keksijä - Uppfinnare - Inventor

1• PIPON, François , THALES SIX GTS France SAS 4, Avenue des Louvresses , 92622 GENNEVILLIERS CEDEX , (FR)

(74) Asiamies - Ombud - Agent

Papula Oy , P.O.Box 981 , 00101 Helsinki , (FI)

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MENETELMÄ JA JÄRJESTELMÄ ALAMOUTI-KOODATUN LOHKOISSA LÄHETETYN SIGNAALIN VASTAANOTTAMISEKSI JA TASAAMISEKSI MAHDOLLISTEN HÄIRIÖIDEN LÄSNÄOLLESSA

METHOD AND SYSTEM FOR RECEIVING AND EQUALISING A SIGNAL EMITTED WITH ALAMOUTI CODING BY BLOCKS IN THE PRESENCE OF POSSIBLE INTERFERENCE

The invention relates to an equalisation and reception method and system for a signal transmitted according to Alamouti block coding in the presence of potential interference.

It applies to MIMO (multiple input, multiple output) receive systems, in radiocommunication fields. It is for example implemented in fourth- and fifth-generation (4G and 5G) civil communication networks.

Radiocommunication systems are heavily affected by fading, which may cause variations of the received signal from a few dB to a few tens of dB over wavelength scales. To mitigate this effect, it is particularly advantageous to make use of spatial diversity on reception by using an array composed of several antennas which receive uncorrelated fading states. A large number of SIMO, single input, multiple output, processing treatments, making use of spatial diversity on reception, both in the absence and in the presence of interference, have been described in the literature, for example in patent applications FR1502720, FR1601894 and FR9514914.

Spatial diversity may also be present on transmission and numerous multiple input, single output, or MISO, and multiple input, multiple output, or MIMO, systems allowing it to be exploited by using a transmission coding scheme have been described and studied in the literature.

Among these schemes, the spatial multiplexing of independent binary trains makes it possible to simultaneously transmit, for constant power and band with respect to a reference single input, single output (SISO) system, several independent binary trains that are multiplexed spatially according to the state of the propagation channels between the transmit and receive antennas. The separation of the binary trains on reception is performed spatially and ideally requires orthogonality of the associated propagation channels. For this reason, this technique is well suited to environments with substantial interference, having many reflectors. In particular, a MIMO system with K antennas for transmission and reception allows the spatial multiplexing of K independent binary trains and provides a transmission rate that is K times higher than that of a SISO system for constant band and emitted power. The drawback of these systems is that they are adapted for transmissions with a high signal-to-noise ratio so that the optimisation of the rate is effective.

When the objective of MIMO processing is primarily to obtain an increase in range for constant emitted power or a decrease in emitted power for constant range, another coding scheme more suited to the needs has been defined in the literature: the coding scheme described by Alamouti in the reference "A simple transmit diversity technique for wireless communications", IEEE Journal on Selected areas in Communications, Vol 16, No 8, pp. 1451-1458, Oct. 1998. This scheme is particularly advantageous, since it allows full exploitation of spatial diversity on transmission, with limited computing power. It is based on two transmit antennas, but other schemes based on the same principle may also be envisaged for a larger number of antennas. The Alamouti coding scheme may be used directly on channels with flat fading for a series waveform WF or on channels with selective fading (i.e. with multipaths generating intersymbol interference ISI) with a parallel WF. The drawback of a parallel WF is that it has a ratio of the amplitude of the peak of the signal to the effective value of the signal, or PAPR, that is high, which leads to a non-negligible loss in transmit power with respect to a series WF.

In order to allow the principle of Alamouti coding to be used in the presence of selective fading for a series WF, the Alamouti block coding scheme, which is an extension of the coding scheme disclosed by Alamouti, has been described for example by S. ZHOU, G. GIANNAKIS in the reference "Single-Carrier Space-Time Block-Coded Transmissions over Frequency-Selective Fading Channels", IEEE Trans. on Info. Theory, Vol 49, No1, pp. 164-179, Jan. 2003.

In the absence of interference, the article by Zhou and Giannakis describes MISO and MIMO processing treatments that allow the reception of the signal transmitted according to Alamouti block coding.

In the presence of interference, the current solutions use a whitening method, described in the literature for SIMO processing treatments, which may also be used for MIMO processing: after estimating R, the total noise correlation matrix (noise + potential interference), the signal is filtered by the matrix $R^{-1/2}$, the inverse of the square root of R. With $\boldsymbol{x}(n)$ the vector of the signals received at the sampling time n, what is obtained at output is thus the signal $\boldsymbol{y}(n) = R^{-1/2}\boldsymbol{x}(n)$. Since the total noise correlation matrix present in $\boldsymbol{y}(n)$ is equal to the identity, the influence from the interference has been removed and MIMO demodulation in the absence of interference may be implemented. The drawback of this solution is that it is expensive in terms of computing power. One of the objectives of the present invention is to perform the reception of a signal transmitted with the Alamouti block coding scheme in the presence of potential interference with lower computing power.

Alamouti block coding consists in transmitting a series of payload symbols according to the scheme described in Figure 1. In this scheme, the symbols are transmitted in groups of two blocks of L_{DATA} symbols with the following notation:

s(i) is the ith block of symbols to be transmitted:

$$\mathbf{s}(i) = [s(i.\,L_{DATA}),\, s(i.\,L_{DATA} + 1),\, \ldots,\, s(i.\,L_{DATA} + L_{DATA} - 1)]^T,$$

where $s(n)$ is the n^{th} symbol to be transmitted, and

with the adoption of the following convention: the quantities in bold lower-case italics are vectors and the quantities in lower-case italics are scalars.

P is a cyclic permutation matrix, with time reversal, which is introduced in order to make possible processing on reception. There are L_{DATA} permutation matrices, defined by the following relationships on the blocks of symbols: $P\,[a(0), \ldots, a(L_{DATA} - 1)] = [a(n), \ldots, a(0), a(L_{DATA} - 1), \ldots, a(n + 1)]$).

The use of this coding scheme entails two constraints on the design of the waveform: The stage has to contain an even number of blocks of payload data for the coding to be carried out,

The channel has to be stationary or near-stationary over two consecutive blocks, i.e., the channel varies "little" over the two blocks.

In order to be able to demodulate transmitted symbols, the receiver has to be able to perform the synchronisation and the estimation of the propagation channel between each transmit path and each receive path. To this end, training sequences, or reference REF sequences, formed of L_{REF} symbols, corresponding to sequences of symbols known on transmission and on reception, are transmitted over each transmit antenna. These sequences are denoted by $d_k(n)$, for k = 1,2 and n = 1, .., L_{REF} - 1, with k the number of the transmit antenna. Each of these sequences is preceded by its own cyclic prefix CP of size L_{CP}.

A straightforward way to synchronise would be to transmit, alternately, such a training sequence over each receive path, according to the scheme of Figure 2. However, this solution, while it prevents the training sequences REF_1, REF_2, each preceded by their cyclic prefix CP_1, CP_2 and transmitted by each of the two antennas Tx_1, Tx_2, from interfering with one another, has the drawback of decreasing the useful rate. It is preferable to simultaneously transmit two different sequences REF_1, REF_2 over the two antennas Tx_1, Tx_2 according to the scheme of Figure 3. In order to optimise the performance of the system, it is advantageous, if possible, to obtain zero autocorrelation over the length of the size of the cyclic prefix CP for the two sequences:

$$E[d_k(n)d_k(n-m)^*] = 0 \text{ for } k = 1{,}2 \text{ et } m = 0, \ldots, L_{CP}$$

and zero autocorrelation over the length of the size of the cyclic prefix between the two sequences:

$$E[d_1(n)d_2(n-m)^*] = 0 \text{ for } m = 0, \ldots, L_{CP}.$$

In one embodiment, the training sequences $d_k(n)$ transmitted over each of the two antennas are advantageously formed on the basis of a CAZAC (Constant Amplitude Zero AutoCorrelation) sequence d(n) of length L_{REF} according to the scheme of Figure 4: the sequence REF_1 transmitted over the first antenna Tx_1, denoted by $d_1(n)$, is equal to d(n) and the sequence REF_2 transmitted over the second antenna Tx_2, denoted by $d_2(n)$, is equal to the sequence d(n) cyclically shifted by a value L_c:

$$[d_2(0), \ldots, d_2(L_{REF}-1)] = [d(L_c), \ldots d(L_{REF}-1), d(0), \ldots, d(L_c-1)]$$

Throughout the rest of the description, a cyclic shift is chosen that corresponds to half of the length of the REF sequence: $L_c = L_{REF}/2$.

An object of the invention is therefore to propose an equalisation and reception method for a signal in the presence of potential interference, transmitted according to Alamouti block coding or similar coding, that is inexpensive in terms of computing power.

FR 3046311A discloses a method for receiving a signal comprising a training sequence formed of known symbols and a payload sequence. The method relates essentially to the calculation of a noise correlation matrix and of interferers with very reduced complexity in order to improve an anti-interference multipath equalisation method.

Hereinafter, the following notation will be used:

K_e: the number of transmit antennas,

K_r: the number of receive antennas,

N_{REF}: the number of reference sequences in the stage ($N_{REF} \geq 1$),

L_{CP}: the number of symbols of the cyclic prefix CP used on each reference REF sequence and each payload data DATA sequence,

L_{REF}: the number of symbols of the reference REF sequence to which must be added L_{CP} symbols for the cyclic prefix CP. The total size of the REF sequence is equal to $L_{REF} + L_{CP}$,

L_{DATA}: the number of symbols of the DATA blocks (to which must be added the cyclic prefix CP),

IR denotes a channel impulse response (frequency or time IR, depending on clarifications made each time in the description).

The invention relates to an anti-interference equalisation and reception method for a signal transmitted according to Alamouti block coding in a multipath system comprising at least one transmitter with K_e = two transmit paths and a receiver with K_r receive paths, said signal containing at least one training sequence formed of known symbols and at least one payload data sequence formed of unknown symbols to be demodulated, containing at least the following steps:

- Synchronising, in time/frequency, the signals received by the receiver,
- Estimating the frequency impulse responses of the propagation channels that are associated with each of the K_e transmit paths,
- Estimating a total noise correlation matrix R(f),
- On each receive path k_r belonging to $[1 \ldots K_r]$ performing:

a discrete Fourier transform DFT on a first type of block of payload data 2i, $x'(2i)(f)$ = DFT(x(2i)(n)) DFT, where x(2i)(n) is the signal vector received on the block 2i of length L_{DATA}, , $f = 0, \ldots, L_{DATA} - 1$ and $n = 0, \ldots, L_{DATA}-1$, and

a conjugation operation, followed by a permutation by the cyclic permutation matrix P with time reversal used for the Alamouti block coding performed on transmission and a discrete Fourier transform DFT operation on a second block of payload data 2i + 1,

$x'(2i + 1)(f) = \mathrm{DFT}(Px^*(2i + 1)(n))$, where x(2i + 1)(n) is the signal vector received on the block 2i + 1 of length L_{DATA}

where * is the conjugation operator and the matrix P is a cyclic permutation matrix, with time reversal, defined by the following relationship on a block of payload symbols $a(n)$ of length L_{DATA}:

$P\,[a(0), \ldots, a(L_{DATA} - 1)] = [a(n), \ldots, a(0), a(L_{DATA} - 1), \ldots, a(n + 1)])$, where there are L_{DATA} matrices P that are possible, each being defined by the first element a(n) of the output block of the permutation by the matrix P, the same permutation matrix being used on transit and receive.

it being understood that there are LDATA matrices P that are possible, each being defined by the first element a(n) of the output block of the permutation by the matrix P. Each of these matrices P may be used interchangeably, what matters is to use the same matrix on transmission and on reception.

- Calculating weightings vectors $w_{ke}(f)$ that are associated with each of the transmit paths ($k_e = 1, .., Ke$):

$$\boldsymbol{w}_{\boldsymbol{k_e}}(f) = \frac{R(f)^{-1}\boldsymbol{h}_{\boldsymbol{k_e}}(f)}{1+\sum_{e=1}^{Ke}\boldsymbol{h_e}(f)^H R(f)^{-1}\boldsymbol{h_e}(f)}$$

with $R(f)^{-1}$ the inverse of the total noise correlation matrix, $h_{ke}(f)$ the vector containing the frequency impulse responses of the propagation channel on the K_r receive paths for the transmit path k_e

$f = 0, \ldots, L_{DATA} - 1$ index of the frequency,

- Performing a filtering on each frequency of the two blocks of payload data using weightings $w_{ke}(f)$ for $k_e = 1$ and $k_e = 2$

on the block 2i with:

$$y(2i)(f) = \boldsymbol{w}_1^H(f)\,\boldsymbol{x}'(2i)(f) + \boldsymbol{w}_2^T(f)\,\boldsymbol{x}'(2i+1)(f), \text{ for } f = 0, \ldots, L_{DATA} - 1,$$

$\boldsymbol{w}_1^H$ being the conjugate transpose of the weighting vector $k_e = 1$, $\boldsymbol{w}_2^T$ being the transpose of the weighting vector for $k_e = 2$,

on the block 2i + 1 with:

$$y(2i+1)(f) = \boldsymbol{w}_2^H(f)\,\boldsymbol{x}'(2i)(f) - \boldsymbol{w}_1^T(f)\,\boldsymbol{x}'(2i+1)(f), \text{ for } f = 0, \ldots, L_{DATA} - 1$$

(604), in order to estimate, respectively, the symbols corresponding to the block 2i and to the block 2i + 1,

- Performing an inverse Fourier transform DFT^{-1} operation on the signals y(2i)(f) and y(2i + 1)(f) in order to obtain an estimate of the symbols transmitted, respectively, on the blocks 2i and 2i + 1 and decode the transmitted payload symbols.

For example, two consecutive data blocks *2i* and *2i + 1* are used for coding on transmission and estimating the symbols on reception.

The method may further comprise a step of normalising the estimated symbols at the output of the inverse Fourier transform DFT^{-1} by means of a factor η defined by its estimate:

$$\hat{\eta} = \frac{1}{L_{DATA}}\sum_{f=0}^{L_{DATA}-1}\frac{\boldsymbol{h}_1(f)^H R(f)^{-1}\boldsymbol{h}_1(f)+\boldsymbol{h}_2(f)^H R(f)^{-1}\boldsymbol{h}_2(f)}{1+\boldsymbol{h}_1(f)^H R(f)^{-1}\boldsymbol{h}_1(f)+\boldsymbol{h}_2(f)^H R(f)^{-1}\boldsymbol{h}_2(f)}.$$

The step of decoding the symbols takes, for example, into account the value of the estimated signal-to-noise ratio $\widehat{SNR} = \frac{\hat{\eta}}{1-\hat{\eta}}$ obtained on each group of two blocks of symbols for the calculation of the LLR likelihood ratios used in the decoding algorithm.

The training sequences may be independent sequences on the transmit paths.

For example, the frequency impulse response of the channel that is associated with each transmit antenna k_e and with each receive antenna k_r for $f = 0, \ldots, L_{DATA} - 1$ is estimated by performing the following steps:

- Performing a discrete Fourier transform DFT of the received signal after synchronisation on each REF sequence of index i, with i = 1, ..., N_{REF}, N_{REF} being the number of reference sequences in a stage ($N_{REF} \geq 1$) for each receive antenna k_r:

$y_{kr,i}(f) = DFT[x_{kr,i}(n)]$, where $x_{kr,i}(n)$ is the signal received on the receive path k_r and the sequence REF i, for $n_0 \leq n < n_0 + L_{REF} - 1$ with n_0 the synchronisation position, the indices n corresponding to the time samples sampled at the symbol rate, with $a \leq f < L_{REF} - 1$, where L_{REF} is the number of symbols of the REF sequence,

- Precalculating the discrete Fourier transform DFT of the reference sequence transmitted over the transmit antenna k_e for the REF block of index i,

with:

$$d_{k_e,i}(f) = DFT\left[d_{k_e,i}(n)\right]$$

for $0 \leq f < L_{REF} - 1$, $d_{ke,i}(f)$ being the DFT of the reference sequence transmitted on the transmit antenna k_e for the REF sequence of index i;

- Performing a preliminary estimation of the frequency impulse response of the channel:

$$h_{k_e,k_r,i}^{prelim}(f) = \frac{y_{k_r,i}(f)}{d_{k_e,i}(f)},$$

, for $f = 0, \ldots, L_{REF} - 1$,

- Performing a preliminary estimation of the time impulse response of the channel on the basis of an inverse discrete Fourier transform, DFT^{-1}, operation for each block REF of index i, on each transmit antenna k_e and on each receive antenna k_r:

$$h_{k_e,k_r,i}^{prelim}(n) = \mathrm{DFT}^{-1}\left[h_{k_e,k_r,i}^{prelim}(f)\right],$$

for $n, f = 0, \ldots, L_{REF} - 1$,

– Decreasing the noise of the preliminary time impulse response and obtaining time impulse responses associated with each transmit path by means of the following two steps:

- Zero-forcing the time impulse responses of the channel $h_{ke,kr,i}$ over a range for which the responses do not contain any influence from the useful multipaths, aside from the size of the CP:

$$h_{k_e,k_r,i}(n) = \begin{cases} h^{prelim}_{k_e,k_r,i}(n) & \text{for} \quad n \in [0; L_{CP} + i_0] \cup [L_{REF} - i_1; L_{REF} - 1] \\ 0 & \text{for} \quad n \in [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1] \end{cases}$$

where i_0 and i_1 are two positive or zero integers, L_{CP} is the number of symbols of the cyclic prefix CP used on each reference REF sequence and each payload data DATA sequence, and the total size of the REF sequence is equal to $L_{REF} + L_{CP}$,

- For each transmit path k_e and for each receive path k_r averaging the channel time impulse responses obtained on each REF sequence over the N_{REF} sequences of the stage (in the case where $N_{REF} > 1$):

$$h_{k_e,k_r}(n) = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} h_{k_e,k_r,i}(n), \quad \text{for} \quad n \in [0; L_{REF} - 1]$$

- Determining the frequency impulse response of the channel, used to demodulate the blocks of data by means of the following two steps:

- Applying a "zero-padding" operation to the averaged time impulse responses in order to allow the interpolation of the impulse response from the length L_{REF} to the length L_{DATA} (in the case where $L_{DATA} > L_{REF}$):

$$\begin{cases} h^{ZP}_{k_e,k_r}(n) = h_{k_e,k_r}(n) \quad \text{for} \quad n = [0; L_{CP} + i_0] \\ h^{ZP}_{k_e,k_r}(n) = 0 \quad \text{for} \quad n = [L_{CP} + i_0 + 1; L_{DATA} - 1 - i_1] \\ h^{ZP}_{k_e,k_r}(n) = h_{k_e,k_r}(n - L_{DATA} + L_{REF}) \quad \text{for} \quad n = [L_{DATA} - i_1; L_{DATA} - 1] \end{cases}$$

with ZP: the zero padding (ZP),

- Estimating the impulse responses of the channel over the length L_{DATA}, denoted by $h_{ke,kr}(f)$ using the Fourier transform of the "zero-padded" time impulse responses of the channel:

$$h_{k_e,k_r}(f) = TFD\left[h^{ZP}_{k_e,k_r}(n)\right] \quad \text{for} \quad n, f = 0, \ldots, L_{DATA} - 1.$$

The total noise correlation matrix $R(f) = \hat{R}$ may be estimated by using the noise samples estimated on all of the training sequences:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H \right)$$

with the noise vector received for each sample n of each training sequence $i = 1, \ldots, N_{REF}$, estimated on the samples of the preliminary time impulse response of the channel, associated with one of the two transmit antennas corresponding to the noise samples $n \in I_b = [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1]$:

$$\boldsymbol{b}_i(n) = [b_{1,i}(n) \ldots b_{K_r,i}(n)]^T, \text{for} \quad n \in I_b = [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1]$$

with

$$b_{k_r,i}(n) = h_{1,k_r,i}^{prelim}(n).$$

For example, CAZAC sequences are used for the training sequences, the second transmit path transmitting the same sequence as the first path, after having applied a cyclic shift $L_c = L_{REF}/2$.

To estimate the frequency IR of the channel that is associated with each transmit antenna k_e and with each receive antenna k_r for $f = 0, \ldots, L_{DATA} - 1$, the following steps are performed, for example:

- Performing a discrete Fourier transform DFT of the received signal after synchronisation on each reference REF sequence associated with a block of data:

$$y_{kr,i}(f) = DFT\left[x_{kr,i}(n)\right], \text{for} \quad n_0 \leq n < n_0 + L_{REF} - 1$$

with n_0 the synchronisation position, the indices n corresponding to the time samples sampled at the symbol rate, with $0 \leq f < L_{REF} - 1$,

- Precalculating the discrete Fourier transform of the reference sequence transmitted over the first transmit antenna for the block of index i:

$$d_i(f) = DFT\left[d_i(n)\right], \text{for} \quad 0 \leq f < L_{REF} - 1:$$

- Performing a preliminary estimation of the frequency impulse response of the channel:

$$h_{k_r,i}^{prelim}(f) = \frac{y_{k_r,i}(f)}{d_i(f)}, \quad \text{for} \ f = 0, \ldots, L_{REF} - 1,$$

- Performing a preliminary estimation of the time IR of the channel on the basis of an inverse Fourier transform operation, DFT^{-1}:

$$h_{k_r,i}^{prelim}(n) = \mathrm{DFT}^{-1}\left[h_{k_r,i}^{prelim}(f)\right], \quad \text{for} \ n, f = 0, \ldots, L_{REF} - 1$$

- Separating the time impulse responses of the channel corresponding to each transmit path in the time impulse response $h_{k_r,i}^{prelim}(n)$ and zero-forcing the time impulse responses of the channel over a range for which the responses do not contain any influence from the useful multipaths, aside from the size of the cyclic prefix CP:

$$h_{1,k_r,i}(n) = \begin{cases} h_{k_r,i}^{prelim}(n) & \text{for} \quad n \in [0; L_{CP} + i_0] \\ 0 & \text{for} \quad n \in \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \\ h_{k_r,i}^{prelim}\left(n + \frac{L_{REF}}{2}\right) & \text{for} \quad n \in \left[\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1\right] \end{cases}$$

$$h_{2,k_r,i}(n) = \begin{cases} h_{k_r,i}^{prelim}\left(n + \frac{L_{REF}}{2}\right) & \text{for} \quad n \in [0; L_{CP} + i_0] \\ 0 & \text{for} \quad n \in \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \\ h_{k_r,i}^{prelim}(n) & \text{for} \quad n \in [\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1] \end{cases}$$

- For each transmit path k_e and for each receive path k_r, averaging the channel time impulse responses obtained on each REF sequence over the N_{REF} sequences of the stage (in the case where $N_{REF} > 1$):

$$h_{k_e,k_r}(n) = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} h_{k_e,k_r,i}(n), \quad \text{for} \quad n \in [0; L_{REF}/2 - 1]$$

- Determining the frequency impulse response of the channel, used to demodulate the blocks of data by means of the following two steps:

- Applying a "zero-padding" operation to the averaged time impulse responses in order to allow the interpolation of the impulse response from the length $L_{REF}/2$ to the length L_{DATA} (in the case where $L_{DATA} > L_{REF}/2$):

$$\begin{cases} h_{k_e,k_r}^{ZP}(n) = h_{k_e,k_r}(n) & \text{for} \quad n = [0; L_{CP} + i_0] \\ h_{k_e,k_r}^{ZP}(n) = 0 & \text{for} \quad n = [L_{CP} + i_0 + 1; L_{DATA} - 1 - i_1] \\ h_{k_e,k_r}^{ZP}(n) = h_{k_e,k_r}(n - L_{DATA} + L_{REF}/2) & \text{for} \quad n = [L_{DATA} - i_1; L_{DATA} - 1] \end{cases}$$

with ZP: the zero padding (ZP),

where i_0 and i_1 are two positive or zero integers,

- Estimating the impulse responses of the channel over the length L_{DATA}, denoted by $h_{ke,kr}(f)$ using the Fourier transform of the "zero-padded" time impulse responses of the channel,

$$h_{k_e,k_r}(f) = DFT\left[h_{k_e,k_r}^{ZP}(n)\right] \text{ for } \quad n, f = 0, \ldots, L_{DATA} - 1.$$

For example, the total noise correlation matrix R(f) = R is estimated by using the noise samples estimated on all of the training sequences:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H\right)$$

With the noise vector received for each sample n of each training sequence i = 1, ..., N_{REF}, estimated on the samples of the preliminary time impulse response of the channel corresponding to the noise samples for $n \in I_b = \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \cup \left[\frac{L_{REF}}{2} + L_{CP} + i_0 + 1; L_{REF} - 1 - i_1\right]$

$$\boldsymbol{b}_i(n) = [b_{1,i}(n) \ldots b_{K,i}(n)]^T,$$

with

$$b_{k_r,i}(n) = h_{k_r,i}^{prelim}(n).$$

The invention also relates to an anti-interference equalisation and reception system for a signal transmitted according to Alamouti block coding in a multipath system comprising at least one transmitter with K_e = two receive paths and a receiver with K_r receive paths, said signal containing at least one training sequence formed of known symbols and at least one payload data sequence formed of unknown symbols to be demodulated, characterised in that the receiver contains a processor connected to the receive paths and configured to perform the steps of the method according to the invention. The system may be a 4G communication system.

Other features, details and advantages of the invention will become apparent from reading the description, which is with reference to the appended drawings, which are given by way of non-limiting example and which show, respectively:

[Fig.1] A reminder of the Alamouti block coding scheme,

[Fig.2] An example of reference sequences transmitted alternately over two antennas,

[Fig.3] An example of reference sequences transmitted simultaneously over two antennas,

[Fig.4] An illustration of the formation of the reference sequences transmitted over the two antennas on the basis of a CAZAC sequence,

[Fig.5] An exemplary architecture of a system for the implementation of the invention,

[Fig.6] A diagram of the processing of the signals on reception according to the invention,

[Fig.7] A figure illustrating steps 3 and 4, in the general case, for the construction of channel time IRs with resetting and "zero padding", for each transmit path k_e and each receive path k_r,

[Fig.8] A figure illustrating the construction of the channel time IRs in the case where the reference sequences transmitted by the two transmit antennas are formed on the basis of a CAZAC sequence according to the scheme of Figure 4,

[Fig.9] A figure illustrating, for a system comprising two transmit paths, the construction of the channel time IRs with "zero padding" in the case where the reference sequences transmitted by the two transmit antennas are formed on the basis of a CAZAC sequence according to the scheme of Figure 4.

The following example is given by way of entirely non-limiting illustration for a MIMO system comprising a transmitter 51 comprising at least two transmit antennas K_e referenced 511, 512 in the figure, and a receiver 52 comprising K_r receive antennas represented by a block 521 for the sake of simplicity. The receiver is equipped with a processor 523 that receives the information received at the antennas and is configured to carry out the steps of the method detailed hereinafter. The transmitted frame is formed, on each stage, of a succession of blocks of payload data and of reference sequences, denoted hereinafter by DATA and REF, respectively.

In order to prevent interference between blocks, a cyclic prefix CP, the length L_{CP} of which is greater than or equal to the expected size of the propagation channel, is added to each data DATA or reference REF sequence block transmitted, such that, on each block of symbols, the intersymbol interference ISI comes only from the symbols belonging to the block itself.

The method according to the invention relates to the processing of the information of the signals at the receiver which receives a signal transmitted according to Alamouti block coding and which, in the presence of potential interference, will perform, on each frequency, a twofold recombination:

The first to take advantage of the spatial diversity on reception and to remove the influence of the potential interference,

The second to take advantage of the spatial diversity on transmission.

The steps of processing the signals which will be described are carried out by the processor of the receiver.

The first recombination involves an adaptive spatial filter for each frequency, and it is therefore based on calculating the total noise correlation matrix $R(f)$ (noise of the receiver and of the interference) at each frequency. Since this matrix is difficult to estimate on each frequency, the method according to the invention implements the total noise correlation matrix R ("average" correlation matrix over all of the frequencies of the useful band), which may be estimated as described hereinafter. The receiver thus implemented is optimal when the total noise is temporally white, i.e., $R(f) = R$.

It is also possible to take into account the correlation matrices R(f) estimated on each frequency f, $\widetilde{R(f)}$ or per block of frequencies.

One possibility for estimating the value of R is described in patent FR1502720 by adapting it for the multiple input, multiple output, MIMO, processing treatment, as described below.

Figure 6 illustrates the anti-interference MIMO processing treatment implemented by the method according to the invention.

The example is given for two consecutive DATA blocks, the block 2i, 61, and the block 2i + 1, 62, of length L_{DATA}. It is also possible to consider two non-consecutive, but close, data blocks, such that the channel does not vary or varies little over the two blocks.

In a prior step, the method will carry out a step of MIMO synchronisation in the presence of potential interference, according to the methods described for example in patent FR1501813. At the output of this step, the signal is sampled at the symbol rate.

Next, in a first step detailed below, the frequency IRs of the propagation channels $h_{1,kr}(f)$ and $h_{2,kr}(f)$ that are associated with the two transmit paths (antenna 511 and antenna 512) are estimated independently on each receive path (antennas 521) $k_r = 1, ..., K_r$, with $f = 0, ..., L_{DATA} - 1$. The channel vectors that are associated with the two transmit paths, containing the frequency IRs obtained on the different receive paths, are denoted by:

$$\boldsymbol{h}_{k_e}(f) = \left[h_{k_e,1}(f), \dots, h_{k_e,K_r}(f)\right]^T \text{ for } \quad k_e = 1{,}2 \quad (1)$$

The total noise correlation matrix R is also estimated in this step detailed below.

In a second step, on each receive path k_r, the Alamouti block recombination preprocessing treatment is performed on the signals from the two DATA blocks, according to the method described for example in the aforementioned document by Zhou and Giannakis:

on the first block 2i: a discrete Fourier transform, DFT, operation 63:

$$x_k(2i)(n), \text{ for } \quad n = 0, \dots, L_{DATA} - 1 \xrightarrow{DFT} x'_k(2i)(f), \text{ for } \quad f = 0, \dots, L_{DATA} - 1 \qquad (2)$$

on the second block 2i + 1: a conjugation operation, followed by a permutation by the matrix P used for the Alamouti block coding performed on transmission 64 and a discrete Fourier transform DFT operation 66:

$$Px_k^*(2i+1)(n), \text{ for } \quad n = 0, \dots, L_{DATA} - 1 \xrightarrow{DFT} x'_k(2i+1)(f) \quad (3)$$

for $f = 0, ..., L_{DATA} - 1$.

The third step, 60, consists in calculating weighting vectors $w_{ke}(f)$ that are associated with each of the transmit antennas ($k_e = 1$ and $k_e = 2$), 601, 602, which will be used to perform the Alamouti block recombination with anti-interference on each frequency of the two DATA blocks. The value of the weighting coefficient is expressed as:

$$\boldsymbol{w}_{k_e}(f) = \frac{R(f)^{-1}\boldsymbol{h}_{k_e}(f)}{1+\boldsymbol{h}_1(f)^H R(f)^{-1}\boldsymbol{h}_1(f)+\boldsymbol{h}_2(f)^H R(f)^{-1}\boldsymbol{h}_2(f)} \qquad (4)$$

The exponent $[^H]$ corresponds to the conjugate transpose.

$\boldsymbol{h}_{ke}(f)$ is a vector that comprises all of the channel estimates performed on the different receive paths, for the transmitter of index $\boldsymbol{k_e}$ with $k_e = 1$ and $k_e = 2$, according to equation (1) $\boldsymbol{h}_{ke}(f) = [h_{ke,1}(f), ..., h_{ke,Kr}(f)]^T$.

The estimate of the signal-to-noise ratio SNR at the output of the Alamouti block recombination performed at the end of the fourth and fifth step is obtained on the basis of the following formulas:

$$\hat{\eta} = \frac{1}{L_{DATA}} \sum_{f=0}^{L_{DATA}-1} \frac{h_1(f)^H R(f)^{-1} h_1(f) + h_2(f)^H R(f)^{-1} h_2(f)}{1 + h_1(f)^H R(f)^{-1} h_1(f) + h_2(f)^H R(f)^{-1} h_2(f)} \quad (5)$$

$$\widehat{SNR} = \frac{\hat{\eta}}{1-\hat{\eta}} \ (6)$$

In a fourth step, the method will perform an Alamouti block recombination on each frequency of the two DATA blocks using the weightings calculated in the preceding step:

603 - on the first block 2i:

$$y(2i)(f) = w_1^H(f)\, x'(2i)(f) + w_2^T(f)\, x'(2i+1)(f), \text{ for}$$

$$f = 0, \ldots, L_{DATA} - 1 \ (7).$$

This recombination makes it possible to estimate the symbols corresponding to the block 2i

604 - on the second block 2i + 1:

$$y(2i+1)(f) = w_2^H(f)\, x'(2i)(f) - w_1^T(f)\, x'(2i+1)(f), \text{ for}$$

$$f = 0, \ldots, L_{DATA} - 1 \quad (8)$$

This recombination makes it possible to estimate the symbols corresponding to the block 2i + 1.

Thus obtained, in the frequency domain, on each receive block, are the symbols transmitted over one of the two blocks, by having put back into phase the different contributions corresponding to the block in question by means of a twofold recombination of MRC (Maximum Ratio Combining) type.

The first recombination performed with the weightings $w_{ke}(f)$ makes it possible to put back into phase the symbols received over each receive antenna, to thus make use of the spatial diversity on reception, and to remove the influence of potential interference.

The second recombination, which consists in summing or subtracting the contributions from the filtering by the weightings $w_{ke}(f)$, makes it possible to put back into phase the symbols from each transmit antenna and thus to make use of the spatial diversity on transmission.

At the output of the fourth step, thus obtained is the frequency-equalised signal corresponding to the two processed data blocks. The fifth and sixth steps are identical to the steps in an SC-FDE (Single Carrier Frequency Domain Equalisation) receiver which are known to a person skilled in the art.

In a fifth step, an inverse discrete Fourier transform, DFT^{-1}, 65, 68 is applied to obtain the (undecided) equalised symbols and normalisation by the factor η calculated in formula (5) is performed.

The sixth step 69 consists in de-interleaving/decoding the transmitted symbols, for example by taking into account the signal-to-noise ratio SNR, estimated according to formula (6), that is obtained on each group of two blocks of symbols for the calculation of the LLRs (Logarithm of Likelihood Ratio) which are known to a person skilled in the art.

The first step of the method according to the invention consists in estimating the frequency IRs of the propagation channels that are associated with the two transmit paths and the total noise correlation matrix R. These estimations may be performed for example according to two embodiments according to the training sequences used in the waveform. The common principle used for these two embodiments is that of firstly estimating the channel in the frequency domain on the REF sequences, then of returning to the time domain in order to denoise and perform the interpolation (from L_{REF}, length of the REF blocks to L_{DATA}, length of the DATA blocks) of the channel estimate, then of returning to the frequency domain to obtain the frequency IRs used in the method according to the invention.

The objective of the algorithm is to estimate the frequency IR of the multipath propagation channel associated with each transmit antenna k_e:

$$\boldsymbol{h}_{k_e}(f) = [h_{k_e,1}(f) \ldots h_{k_e,K_r}(f)]^T, \quad \text{for} \quad f = 0, \ldots, L_{DATA}\text{-}1 \quad \text{and}$$

$$k_e = 1 \text{ and } 2, \text{ with } k_r \geq 1, k_r = 1, \ldots, K_r. \qquad (9)$$

The two DATA blocks on which the Alamouti block coding has been performed are demodulated on the basis of the same estimate of $\boldsymbol{h}_{ke}(f)$.

According to a first embodiment of the invention, there is no deduction of the sequences transmitted over the two antennas from one another, they are independent, and the algorithm is run according to the following steps:

Step 1 – Perform a discrete Fourier transform DFT of the received signal after synchronisation on each REF sequence of index i, with i = 1, …, N_{REF}, for each receive antenna k_r:

$$y_{k_r,i}(f) = DFT\left[x_{k_r,i}(n)\right], \ (10),$$

for $n_0 \leq n < n_0 + L_{REF} - 1$ with n_0 the synchronisation position, the indices n corresponding to the time samples sampled at the symbol rate.

Step 2 – Preliminary estimation of the channel frequency IR on each transmit antenna k_e and on each receive antenna k_r:

$$h^{prelim}_{k_e,k_r,i}(f) = \frac{y_{k_r,i}(f)}{d_{k_e,i}(f)} \ (11)$$

for $f = 0, \ldots, L_{REF} - 1$, with $d_{ke,i}(f)$ DFT of the reference sequence transmitted over the transmit antenna k_e for the REF block of index i,

The discrete Fourier transform DFT of the reference sequence is precalculated:

$$d_{k_e,i}(f) = DFT\left[d_{k_e,i}(n)\right], \quad \text{for} \quad 0 \leq f < L_{REF} - 1 \qquad (12)$$

Step 3 – Preliminary estimation of the time IR of the channel on the basis of an inverse discrete Fourier transform, DFT^{-1}, operation, on each transmit antenna k_e and on each receive antenna k_r:

$$h_{k_e,k_r,i}^{prelim}(n) = \mathrm{DFT}^{-1}\left[h_{k_e,k_r,i}^{prelim}(f)\right], \quad \text{for} \quad n, f = 0, \ldots, L_{REF} - 1, \; (13) \; (\text{Fig.7}).$$

Step 4 – Decrease the noise of the preliminary time IR and obtain time IRs associated with each transmit path by means of the steps described below.

In order to improve the quality of the channel estimate, the time IRs of the channel $h_{ke,kr,i}$ are zero-forced over a range for which the responses are assumed not to be affected by the multipath channel: only the elements of the time IR over a length at least equal to the length of the cyclic prefix CP are retained.

$$h_{k_e,k_r,i}(n) = \begin{cases} h_{k_e,k_r,i}^{prelim}(n) & \text{for} \quad n \in [0; L_{CP} + i_0] \cup [L_{REF} - i_1; L_{REF} - 1] \\ 0 & \text{for} \quad n \in [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1] \end{cases} \qquad (14)$$

Where i_0 and i_1 are two positive or zero integers making it possible to retain a larger portion of the time IR of the channel than the portion corresponding to the cyclic prefix (to manage edge effects) and satisfying the following conditions:

$0 \leq i_0 \leq L_{REF} - L_{CP} - 1$ and $0 \leq i_1 \leq L_{REF} - L_{CP} - 2 - i_0$

- if $i_0 = L_{REF} - L_{CP} - 1$, this amounts to retaining all of the elements of the time IR (and i_1 is not used),
- if $i_1 = 0$, no sample belongs to the interval $[L_{REF} - i_1; L_{REF} - 1])$.

Step 5 – Averaging the "denoised" time IRs if N_{REF} is strictly greater than one. This step is not performed if N_{REF} is equal to 1.

For each transmit path k_e and for each receive path k_r, the channel time IRs obtained on each REF sequence are averaged over the N_{REF} sequences of the stage:

$$h_{k_e,k_r}(n) = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} h_{k_e,k_r,i}(n), \quad \text{for} \quad n \in [0; L_{REF} - 1] \qquad (15)$$

In the following steps, the method will estimate the frequency IR of the multipath channel over the length L_{DATA}.

On the basis of the estimate of the channel time IR obtained at the end of the fifth step on each REF sequence of the stage, the frequency IR of the channel used to demodulate the DATA blocks is obtained on the basis of the following two steps, which are performed for each transmit antenna k_e, with $k_e = 1,2$, and each receive antenna k_r, with $k_r = 1, \ldots, K_r$.

Step 6 - Interpolation of the channel time IR over the length L_{DATA} if $L_{DATA} > L_{REF}$.

A "zero-padding" operation illustrated in Figure 7 is applied to the averaged time IRs in order to allow the interpolation of the IRs from the length L_{REF} to the length L_{DATA}:

$$\begin{cases} h_{k_e,k_r}^{ZP}(n) = h_{k_e,k_r}(n) & \text{for} \quad n = [0; L_{CP} + i_0] \\ h_{k_e,k_r}^{ZP}(n) = 0 & \text{for} \quad n = [L_{CP} + i_0 + 1; L_{DATA} - 1 - i_1] \\ h_{k_e,k_r}^{ZP}(n) = h_{k_e,k_r}(n - L_{DATA} + L_{REF}) & \text{for} \quad n = [L_{DATA} - i_1; L_{DATA} - 1] \end{cases} \tag{16}$$

With ZP: the zero padding (ZP)

Step 7 - Estimation of the channel frequency IRs over the length L_{DATA}, denoted by $h_{ke,kr}(f)$ is then performed using the DFT of the "zero-padded" time IRs of the channel:

$$h_{k_e,k_r}(f) = DFT\left[h_{k_e,k_r}^{ZP}(n)\right] \text{ for } \quad n, f = 0, \dots, L_{DATA} - 1. \tag{17}$$

Step 8: In order to estimate the total noise correlation matrix R, firstly, for each receive antenna k_r and on each REF sequence i, the noise samples $b_{kr,i}(n)$ are initialised on the basis of the samples of the channel time IR $h_{1,k_r,i}^{prelim}(n)$ estimated for one of the two transmit antennas (the first for example as described in the rest of this document):

$$b_{k_r,i}(n) = h_{1,k_r,i}^{prelim}(n) \tag{18}$$

for

$$n \in I_b = [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1]$$

with:

$$i = 1, \dots, N_{REF}$$
$$k_r = 1, \dots, K_r$$

Let $\boldsymbol{b}_i(n)$ denote the noise vector received on each sequence i

$$\boldsymbol{b}_i(n) = [b_{1,i}(n) \dots b_{K_r,i}(n)]^T, \text{ for } \quad n \in I_b \tag{19}$$

The correlation matrix $\hat{R}$ is then estimated by using the noise samples estimated on all of the training sequences:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H \right). \tag{20}$$

According to a second embodiment of the invention, the training sequences allowing the channel estimation are formed on the basis of a CAZAC sequence and transmitted over the two transmit antennas according to the scheme of Figure 4. The channel estimation thus exploits the deduction of the sequence transmitted over the second transmit antenna from the sequence transmitted over the first antenna via a cyclic shift of L_c symbols in order to decrease the computing power of steps 2 to 4 by making use of the properties of the DFT, step 1 remaining unchanged. In the rest of the description, $L_c = L_{REF}/2$ is chosen.

To estimate the frequency IR on each transmit path k_e, with $k_e = 1, ..., K_e$ and on each receive path k_r, with $k_r = 1, ..., K_r$, and on each training sequence i, with i = 1, ..., N_{REF}, the method consists in proceeding to estimate the time IR of the channel on each receive path k_r and on each training sequence i in steps 2 to 3. At the end of step 3, the time IRs associated with the two transmit paths $k_e = 1, 2$ are present on the channel time IR thus estimated and step 4 makes it possible to "separate" the two channel time IRs.

Step 2 – Preliminary estimation of the channel frequency IR on each receive antenna k_r:

$$h_{k_r,i}^{prelim}(f) = \frac{y_{k_r,i}(f)}{d_i(f)}, \text{ for } \quad f = 0, \ldots, L_{REF} - 1,$$

with $d_i(f)$ the DFT of the reference sequence transmitted over the first transmit antenna for the REF block of index i (22) (Figure 8).

At this point, unlike the first embodiment of the channel estimation step, this channel estimate takes into account the two transmit antennas.

The discrete Fourier transform DFT of the reference sequence is precalculated:

$$d_i(f) = TFD[d_i(n)]\,, \text{ for } \quad 0 \leq f < L_{REF} - 1 \qquad (23)$$

Step 3 – Preliminary estimation of the time IR of the channel on the basis of an inverse discrete Fourier transform, DFT^{-1}, operation, on each receive antenna k_r:

$$h_{k_r,i}^{prelim}(n) = \mathrm{DFT}^{-1}\left[h_{k_r,i}^{prelim}(f)\right], \text{ for } \; n, f = 0, \ldots, L_{REF} - 1 \qquad (24)$$

At this point, unlike the first embodiment of the channel estimation step, this channel estimate also takes into account the two transmit antennas. The following step allows the algorithm to perform the channel estimation for each transmit path.

Step 4 – Decrease the noise of the preliminary time IR and obtain time IRs associated with each transmit path by means of the steps described below.

In order to improve the quality of the channel estimate, the time IRs of the channel $h_{kr,i}$ are zero-forced over a range for which the responses are assumed not to be affected by the multipath channel, figure 9: only the elements of the time IR over a length at least equal to the length of the cyclic prefix CP are retained. It is also necessary to take into account the fact that the two transmit antennas transmit the same sequence, to within a cyclic shift of value $\frac{L_{REF}}{2}$: the first portion of the time IR obtained therefore corresponds to the first transmit antenna and the second portion to the second transmit antenna:

$$h_{1,k_r,i}(n) = \begin{cases} h_{k_r,i}^{prelim}(n) & \text{for} \quad n \in [0; L_{CP} + i_0] \\ 0 & \text{for} \quad n \in \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \\ h_{k_r,i}^{prelim}\left(n + \frac{L_{REF}}{2}\right) & \text{for} \quad n \in \left[\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1\right] \end{cases} \quad (25)$$

$$h_{2,k_r,i}(n) = \begin{cases} h_{k_r,i}^{prelim}\left(n + \frac{L_{REF}}{2}\right) & \text{for} \quad n \in [0; L_{CP} + i_0] \\ 0 & \text{for} \quad n \in \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \\ h_{k_r,i}^{prelim}(n) & \text{for} \quad n \in [\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1] \end{cases} \quad (26)$$

Where i_0 and i_1 are two positive or zero integers making it possible to retain a larger portion of the time IR of the channel than the portion corresponding to the cyclic prefix (to manage edge effects) and satisfying the following conditions:

$0 \leq i_0 \leq \frac{L_{REF}}{2} - L_{CP} - 1$ and $0 \leq i_1 \leq \frac{L_{REF}}{2} - L_{CP} - 2 - i_0$

- if $i_0 = \frac{L_{REF}}{2} - L_{CP} - 1$, this amounts to retaining all of the elements of the time IR (and i_1 is not used)
- if $i_1 = 0$, no sample belongs to the interval $[\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1]$.

At the end of step 4, the algorithm thus obtains, like in the first embodiment of the invention, the time IR associated with each transmit path k_e, with each receive path k_r and with each training sequence i, the only difference being that these time IRs are estimated on a length $L_{REF}/2$ instead of L_{REF}. Steps 5 to 7 remain unchanged (while replacing L_{REF} with $\frac{L_{REF}}{2}$ in equations (15) and (16) of steps 5 and 6 above) and step 8 is modified so as to take into account the fact that the noise samples are no longer located on one interval but on two intervals, after each of the two channel time IRs associated with each transmit path k_e that are present in the overall time IR $h_{k_r,i}^{prelim}$.

Step 6 of zero setting and of zero padding is illustrated in Figure 9 for $h_{1,k_r}^{ZP}(n)$ and $h_{2,k_r}^{ZP}(n)$.

Step 8: In order to estimate the total noise correlation matrix R, firstly, for each receive antenna k_r and on each REF sequence i, the noise samples $b_{k_r,i}(n)$ are initialised on the basis of the samples of the estimated channel time IR $h_{k_r,i}^{prelim}$:

$$b_{k_r,i}(n) = h_{k_r,i}^{prelim}(n) \quad (27)$$

for

$$n \in I_b = \left[L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1\right] \cup \left[\frac{L_{REF}}{2} + L_{CP} + i_0 + 1; L_{REF} - 1 - i_1\right]$$

with:

$$i = 1, \ldots, N_{REF}$$

$$k_r = 1, \ldots, K_r$$

Let $\boldsymbol{b}_i(n)$ denote the noise vector received on each sequence i

$$\boldsymbol{b}_i(n) = \left[b_{1,i}(n) \ldots b_{K_r,i}(n)\right]^T, \text{ for } n \in I_b \quad (28)$$

Next, the correlation matrix $\hat{R}$ is estimated by using the noise samples estimated on all of the training sequences:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H\right) \qquad (29)$$

The method according to the invention notably affords the advantage of equalising signals received according to a method that is inexpensive in terms of computing time.

PATENTTIVAATIMUKSET

1. Menetelmä signaalin vastaanottamiseksi ja tasaamiseksi interferenssiä ehkäisevästi, joka signaali lähetetään Alamouti-koodauksen mukaisesti lohkoissa monitiejärjestelmässä, joka käsittää vähintään yhden lähettimen, jossa on K_e = kaksi lähetystietä, ja vastaanottimen, jossa on K_r kappaletta vastaanottoteitä, jossa mainittu signaali käsittää vähintään yhden opetussekvenssin, joka muodostuu tunnetuista symboleista, ja vähintään yhden hyötydatasekvenssin, joka muodostuu tuntemattomista sekvensseistä, jotka on tarkoitus demoduloida, ja joka menetelmä käsittää ainakin seuraavat vaiheet:

- synkronoidaan ajan/taajuuden suhteen signaalit, jotka vastaanotetaan vastaanottimella,
- estimoidaan taajuusimpulssivasteet etenemiskanaville, jotka liittyvät kuhunkin K_e kappaleesta lähetysteitä,
- estimoidaan kokonaiskohinan korrelaatiomatriisi, $R(f)$,
- kullekin vastaanottotielle k_r, joka sisältyy joukkoon $[1 \ldots K_r]$, suoritetaan:
- diskreetti Fourier-muunnos DFT ensimmäisen tyyppiselle lohkolle hyötydataa $2i$, (63), $x'(2i)(f)=\mathrm{TFD}(x(2i)(n))$, jossa $x(2i)(n)$ on signaalivektori, joka vastaanotetaan lohkolla $2i$, jonka pituus on L_{DATA}, jossa $f=0,\ldots,L_{DATA}-1$ ja $n=0,\ldots,L_{DATA}-1$ ja
- konjugaatio-operaatio, jota seuraa permutaatio syklisellä permutaatiomatriisilla P ja ajankäännöllä, jota käytetään Alamouti-koodaukseen lohkoissa, joka suoritetaan lähetyksessä (64), ja diskreetti Fourier-muunnos DFT –operaatio (66) toiselle lohkolle hyötydataa $2i+1$, $x'(2i+1)(f)=\mathrm{TFD}(Px^*(2i+1)(n))$, jossa $x(2i+1)(n)$ on signaalivektori, joka

vastaanotetaan lohkolla $2i$+1, jonka pituus on L_{DATA}, jossa f=0,…,L_{DATA}−1 ja n=0,…,L_{DATA}−1, jossa * on konjugaatio-operaattori ja matriisi P on syklinen permutaatiomatriisi, ajankäännöllä, määriteltynä seuraavalla suhteella lohkolle hyötysymboleita $a(n)$, jonka pituus on L_{DATA}:

$$P\,[a(0),\dots,a(L_{DATA}-1)] = [a(n),\dots,a(0),a(L_{DATA}-1),\dots,a(n+1)])$$

jossa mahdollisia matriiseja P on L_{DATA} kappaletta, kunkin ollessa määritelty ensimmäisellä alkiolla $a(n)$ matriisilla P tehdyn permutaation ulostulolohkosta, jossa lähetyksessä ja vastaanotossa käytetään samaa permutaatiomatriisia;

- lasketaan painotusvektorit $w_{ke}(f)$, jotka liittyvät kuhunkin lähetystiehen (k_e=1,…,Ke=2), (601, 602):

$$\boldsymbol{w}_{\boldsymbol{k}_e}(f) = \frac{R(f)^{-1}\boldsymbol{h}_{\boldsymbol{k}_e}(f)}{1+\sum_{e=1}^{Ke}\boldsymbol{h}_e(f)^H R(f)^{-1}\boldsymbol{h}_e(f)}$$

jossa $R(f)^{-1}$ on kokonaiskohinan korrelaatiomatriisin käänteisarvo, $h_{ke}(f)$ on vektori, joka sisältää etenemiskanavan taajuusimpulssivasteet K_r kappaleella vastaanottoteitä lähetystielle k_e, f=0,…,L_{DATA}−1 taajuuden indeksi,

- suoritetaan suodatus kullekin taajuudelle kahdesta lohkosta hyötydataa käyttäen painotuksia $w_{ke}(f)$, kun k_e=1 ja k_e=2

lohkolle $2i$, jossa:

$y(2i)(f) = w_1^H(f)x'(2i)(f) + w_2^T(f)x'(2i+1)(f)$, kun f=0,…,L_{DATA}−1 (603), w_1^H painotusvektorin konjugaattitranspoosi, kun k_e=1, w_2^T painotusvektorin transpoosi, kun k_e=2,

lohkolle $2i$+1, jossa:

$y(2i+1)(f) = w_2^H(f)x'(2i)(f) - w_1^T(f)x'(2i+1)(f)$, kun f=0,…,L_{DATA}−1 (604), jotta saadaan estimoitua vastaavasti symbolit, jotka vastaavat lohkoa $2i$ ja lohkoa $2i$+1,

- suoritetaan käänteinen Fourier-muunnos DFT^{-1} –operaatio signaaleille $y(2i)(f)$ ja $y(2i+1)(f)$ (65, 68), jotta saadaan estimaatti symboleista, jotka lähetetään vastaavasti lohkoilla $2i$ ja $2i$+1, ja dekoodataan (69) lähetetyt hyötysymbolit.

2. Patenttivaatimuksen 1 mukainen menetelmä, **tunnettu siitä, että** käytetään kahta peräkkäistä datalohkoa $2i$ ja $2i$+1 koodaamiseen lähetyksessä ja symbolien estimoimiseen vastaanotossa.

3. Jonkin patenttivaatimuksista 1 tai 2 mukainen menetelmä, **tunnettu siitä, että** se käsittää edelleen vaiheen, jossa normalisoidaan symbolit, jotka on estimoitu ulostulossa käänteisellä Fourier-muunnoksella TFD^{-1}, kertoimella η, joka määritellään käyttäen sen estimaattia:

$$\hat{\eta} = \frac{1}{L_{DATA}} \sum_{f=0}^{L_{DATA}-1} \frac{\boldsymbol{h}_1(f)^H R(f)^{-1} \boldsymbol{h}_1(f) + \boldsymbol{h}_2(f)^H R(f)^{-1} \boldsymbol{h}_2(f)}{1 + \boldsymbol{h}_1(f)^H R(f)^{-1} \boldsymbol{h}_1(f) + \boldsymbol{h}_2(f)^H R(f)^{-1} \boldsymbol{h}_2(f)}.$$

4. Patenttivaatimuksen 3 mukainen menetelmä, **tunnettu siitä, että** vaiheessa, jossa dekoodataan (69) symbolit, otetaan huomioon estimoidun signaali-kohinasuhteen

$$\widehat{SNR} = \frac{\hat{\eta}}{1-\hat{\eta}}$$

arvo, joka saadaan kullekin kahden symbolien lohkon ryhmälle dekoodauksessa käytettyjen uskottavuussuhteiden LLR laskemiseksi.

5. Jonkin patenttivaatimuksista 1–4 mukainen menetelmä, **t u n n e t t u siitä, että** käytetään opetussekvensseihin riippumattomia sekvenssejä lähetysteillä.

6. Patenttivaatimuksen 5 mukainen menetelmä, **t u n n e t t u siitä, että** estimoidaan kanavan taajuusimpulssivaste, joka liittyy kuhunkin lähetysantenniin k_e ja kuhunkin vastaanottoantenniin k_r, jossa f=0,…,L_{DATA}-1, suorittamalla seuraavat vaiheet:

- suoritetaan synkronoinnin jälkeen vastaanotetun signaalin diskreetti Fourier-muunnos DFT indeksin i kullakin referenssisekvenssillä REF, i=1,…,N_{REF}, jossa N_{REF} on referenssisekvenssien lukumäärä yhdellä tasolla (N_{REF}≥1), kullekin vastaanottoantennille k_r: $y_{k_{r},i}(f) = (DFT)[x_{kr,i}(n)]$, jossa $x_{kr,i}$ on vastaanotettu signaali vastaanottotiellä k_r ja sekvenssillä REF i, jossa $n_0 \leq n < n_0 + L_{REF} - 1$, jossa n_0 on synkronointiasema, n kappaleen indeksejä vastatessa ajallisia näytteitä, jotka on näytteistetty symbolinopeudella, jossa $0 \leq f < L_{REF} - 1$, jossa L_{REF} on sekvenssin REF symbolien lukumäärä
- lasketaan alustavasti lähetysantennilla k_e lähetetyn referenssisekvenssin diskreetti Fourier-muunnos DFT indeksin i lohkolle REF, jossa:

$d_{k_{e},i}(f) = DFT\left[d_{k_{e},i}(n)\right]$ kun $0 \leq f < L_{REF} - 1$,
jossa $d_{k_{e},i}(f)$ on lähetysantennilla k_e lähetetyn referenssisekvenssin DFT indeksin i sekvenssille REF:
- suoritetaan kanavan taajuusimpulssivasteen alustava estimointi:

$$h_{k_e,k_r,i}^{alustava}(f) = \frac{y_{k_e,i}(f)}{d_{k_e,i}(f)}, \text{ kun } f=0,\dots,L_{REF}-1$$

- suoritetaan kanavan ajallisen impulssivasteen alustava estimointi perustuen käänteisen diskreetin Fourier-muunnoksen, DFT^{-1}, operaatioon indeksin i kullekin lohkolle REF, kullakin lähetysantennilla k_e ja kullakin vastaanottoantennilla k_r:

$$h_{k_e,k_r,i}^{alustava}(n) = DFT^{-1}\left[h_{k_e,k_r,i}^{alustava}(f)\right], \text{ kun } n, f=0,\dots,L_{REF}-1,$$

- vähennetään alustavan ajallisen impulssivasteen kohina ja saadaan ajalliset impulssivasteet, jotka liittyvät kuhunkin lähetystiehen, seuraavissa kahdessa vaiheessa:

- supistetaan nollaan kanavan ajalliset impulssivasteet $h_{k_e,k_r,i}$ alueella, jossa vasteet eivät sisällä hyödyllisten moniteiden vaikutusta, poislukien syklisen etuliitteen CP koko:

$$h_{k_e,k_r,i}(n) = \begin{cases} h_{k_e,k_r,i}^{alustava}(n) \; kun \; n \in [0;\, L_{CP}+i_0] \cup [L_{REF}-i_1;\, L_{REF}-1] \\ 0 \; kun \; n \in [L_{CP}+i_0+1; L_{REF}-1-i_1]] \end{cases}$$

jossa i_0 ja i_1 ovat kaksi positiivista kokonaislukua tai nolla, L_{CP} on syklisen etuliitteen CP symbolien lukumäärä, jota käytetään kullakin referenssisekvenssillä REF ja kullakin hyötydatasekvenssillä DATA, ja sekvenssin REF kokonaiskoko on yhtä kuin $L_{REF}+L_{CP}$

- kullekin lähetystielle k_e ja kullekin vastaanottotielle k_r keskiarvoistetaan saadut kanavan ajalliset impulssivasteet kullakin sekvenssillä REF tason N_{REF} kappaleella sekvenssejä (kun $N_{REF}>1$):

$$h_{k_e,k_r}(n) = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} h_{k_e,k_r,i}(n),\; kun \; n \in [0; L_{REF}-1]$$

- määritetään käytettävä kanavan taajuusimpulssivaste datalohkojen demoduloimiseksi seuraavissa kahdessa vaiheessa:

- sovelletaan ”zero-padding”-operaatiota keskiarvoistettuihin ajallisiin impulssivasteisiin, niin että saadaan interpoloitua impulssivaste pituudesta L_{REF} pituuteen L_{DATA} (kun $L_{DATA}>L_{REF}$):

$$\begin{cases} h^{ZP}_{k_{e},k_{r}}(n) = h_{k_{e},k_{r}}(n) kun\, n = [0;\, L_{CP} + i_0] \\ h^{ZP}_{k_{e},k_{r}}(n) = 0\; kun\, n = [L_{CP} + i_0 + 1;\, L_{DATA} - 1 - i_1] \\ h^{ZP}_{k_{e},k_{r}}(n) = h_{k_{e},k_{r}}(n - L_{DATA} + L_{REF}) kun\, n = [L_{DATA} - i_1;\, L_{DATA} - 1] \end{cases}$$

jossa ZP: zero-padding tai nollien lisääminen (ZP)

- estimoidaan kanavan impulssivasteet pituudella L_{DATA}, merkitty $h_{k_{e},k_{r}}(f)$, käyttäen Fourier-muunnosta kanavan ajallisiin impulssivasteisiin lisättynä nollilla $h_{k_{e},k_{r}}(f) = DFT\left[h^{ZP}_{k_{e},k_{r}}(n)\right]$, kun n, f=0,…,L_{DATA}-1.

7. Patenttivaatimuksen 6 mukainen menetelmä, **tunnettu siitä, että** estimoidaan kokonaiskohinan korrelaatiomatriisi $R(f)=\hat{R}$ käyttämällä estimoituja kohinanäytteitä joukolle opetussekvenssejä:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H \right)$$

jossa vastaanotetun kohinan vektori kullekin näytteelle n kustakin opetussekvenssistä i=1,…,N_{REF}, estimoidaan näytteille kanavan alustavasta ajallisesta impulssivasteesta, joka liittyy yhteen kahdesta lähetysantennista, joka vastaa kohinanäytteitä, kun $n \in I_b = [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1]$

$$b_i(n) = [b_{1,i}(n) \dots b_{K_{r},i}(n)]^T, kun\ n \in l_b = [L_{CP} + i_0 + 1; L_{REF} - 1 - i_1]$$

jossa $b_{k_{r},1}(n) = h^{alustava}_{1,k_{r},1}(n)$.

8. Jonkin patenttivaatimuksista 1 - 4 mukainen menetelmä, **tunnettu siitä, että** käytetään Cazac-sekvenssejä opetussekvensseille, jossa toinen lähetystie lähettää saman sekvenssin kuin ensimmäinen tie, sen jälkeen kun on sovellettu syklistä siirtoa $L_c=L_{REF/2}$.

9. Patenttivaatimuksen 8 mukainen menetelmä, **tunnettu siitä, että** estimoidaan kanavan taajuusimpulssivaste, joka liittyy kuhunkin lähetysantenniin k_e ja kuhunkin vastaanottoantenniin k_r, kun f=0,…,L_{DATA}−1, suorittamalla seuraavat vaiheet:

- suoritetaan synkronoinnin jälkeen vastaanotetun signaalin diskreetti Fourier-muunnos DFT kullekin referenssisekvenssille REF, joka liittyy datalohkoon:

$y_{kr,i}(f)=DFT[x_{kr,i}(n)]$, kun $n_0 \leq n < n_0+L_{REF}-1$
jossa n_0 on synkronointiasema, n kappaleen indeksejä vastatessa ajallisia näytteitä, jotka on näytteistetty symbolinopeudella, ja jossa $0 \leq f < L_{REF}-1$,
- lasketaan alustavasti ensimmäisellä lähetysantennilla lähetetyn referenssisekvenssin diskreetti Fourier-muunnos indeksin i lohkolle:

$d_i(f)=DFT[d_i(n)]$, kun $0 \leq f < L_{REF}-1$:
- suoritetaan kanavan taajuusimpulssivasteen alustava estimointi:

$h^{alustava}_{k_{r},i}(f) = \frac{y_{k_{r},i}(f)}{d_i(f)}$, kun f=0,…,L_{REF}−1,

- suoritetaan kanavan ajallisen impulssivasteen alustava estimointi käyttäen käänteisen Fourier-muunnoksen, DFT^{-1}, operaatiota:

$h_{k_{r,i}}^{alustava}(n) = DFT^{-1}[h_{k_{r,i}}^{alustava}(f)]$, kun $n, f=0,\ldots,L_{REF}-1$

- erotetaan kanavan ajalliset impulssivasteet, jotka vastaavat kutakin lähetystietä ajallisessa impulssivasteessa $h_{k_{r,i}}^{alustava}(n)$, ja supistetaan nollaan kanavan ajalliset impulssivasteet alueella, jossa vasteet eivät sisällä hyödyllisten moniteiden vaikutusta, poislukien CP:n koko:

$$h_{1,k_{r,i}}(n) = \begin{cases} h_{k_{r,i}}^{alustava}(n), kun\ n \in [0; L_{CP} + i_0] \\ 0, kun\ n \in [L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1] \\ h_{k_{r,i}}^{alustava}\left(n + \frac{L_{REF}}{2}\right), kun\ n \in [\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1] \end{cases}$$

$$h_{2,k_{r,i}}(n) = \begin{cases} h_{k_{r,i}}^{alustava}\left(n + \frac{L_{REF}}{2}\right), kun\ n \in [0; L_{CP} + i_0] \\ 0, kun\ n \in [L_{CP} + i_0 + 1; \frac{L_{REF}}{2} - 1 - i_1] \\ h_{k_{r,i}}^{alustava}(n), kun\ n \in [\frac{L_{REF}}{2} - i_1; \frac{L_{REF}}{2} - 1] \end{cases}$$

- kullekin lähetystielle k_e ja kullekin vastaanottotielle k_r, keskiarvoistetaan kanavan ajalliset impulssivasteet, jotka on saatu kullakin sekvenssillä REF tason N_{REF} kappaleella sekvenssejä (kun $N_{REF}>1$):

$$h_{k_e,k_r}(n) = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} h_{k_{e} k_{r,i}}(n),\ kun\ n \in [0; L_{REF}/2 - 1]$$

- määritetään kanavan taajuusimpulssivaste, jota käytetään datalohkojen demoduloimiseen, kahdessa seuraavassa vaiheessa:

- sovelletaan ”zero-padding”-operaatiota keskiarvoistettuihin ajallisiin impulssivasteisiin, niin että saadaan interpoloitua impulssivaste pituudesta $L_{REF}/2$ pituuteen L_{DATA} (kun $L_{DATA}>L_{REF}/2$):

$$\begin{cases} h^{ZP}_{k_e,k_r}(n) = h_{k_e,k_r}(n), kun\, n = [0;\, L_{CP}+i_0] \\ h^{ZP}_{k_e,k_r}(n) = 0, kun\, n = [L_{CP}+i_0+1;\, L_{DATA}-1-i_1] \\ h^{ZP}_{k_e,k_r}(n) = h_{k_e,k_r}(n - L_{DATA} + L_{REF}/2), kun\, n = [L_{DATA}-i_1;\, L_{DATA}-1] \end{cases}$$

jossa ZP: zero-padding tai nollien lisääminen (ZP),

jossa i_0 ja i_1 ovat kaksi positiivista kokonaislukua tai nolla

- estimoidaan kanavan impulssivasteet pituudella L_{DATA}, merkitty $h_{k_e,k_r}(f)$, käyttäen Fourier-muunnosta kanavan ajallisiin impulssivasteisiin lisättynä nollilla

$h_{k_e,k_r}(f) = DFT\left[h^{ZP}_{k_e,k_r}(n)\right]$, kun $n, f=0,\dots,L_{DATA}-1$.

10. Patenttivaatimuksen 9 mukainen menetelmä, **t u n n e t t u siitä, että** estimoidaan kokonaiskohinan korrelaatiomatriisi $R(f)=\hat{R}$ käyttämällä estimoituja kohinanäytteitä joukolle opetussekvenssejä:

$$\hat{R} = \frac{1}{N_{REF}} \sum_{i=1}^{N_{REF}} \left(\sum_{n \in I_b} \boldsymbol{b}_i(n) \boldsymbol{b}_i(n)^H \right)$$

jossa vastaanotetun kohinan vektori kullekin näytteelle n kustakin opetussekvenssistä $i=1,\dots,N_{REF}$, estimoidaan näytteille kanavan alustavasta ajallisesta impulssivasteesta, joka vastaa kohinanäytteitä, kun

$$n \in I_b = \left[L_{CP}+i_0+1; \frac{L_{REF}}{2} - 1 - i_1\right] \cup \left[\frac{L_{REF}}{2} + L_{CP} + i_0 + 1; L_{REF} - 1 - i_1\right]$$

$$\boldsymbol{b}_i(n) = [b_{1,i}(n) \dots b_{K,i}(n)]^T,$$

jossa

$$b_{k_{r,i}}(n) = h_{k_{r,i}}^{alustava}(n).$$

11. Vastaanotto- ja tasausjärjestelmä signaalin interferenssin ehkäisemiseksi, joka signaali lähetetään Alamouti-koodauksen mukaisesti lohkoissa monitiejärjestelmässä, joka käsittää vähintään yhden lähettimen, jossa on K_e = *kaksi* vastaanottotietä, ja vastaanottimen, jossa on K_r kappaletta vastaanottoteitä, jossa mainittu signaali käsittää vähintään yhden opetussekvenssin, joka muodostuu tunnetuista symboleista, ja vähintään yhden hyötydatasekvenssin, joka muodostuu tuntemattomista sekvensseistä, jotka on tarkoitus demoduloida, **tunnettu siitä, että** vastaanotin (52) käsittää prosessorin (523), joka on kytketty vastaanottoteihin ja konfiguroitu suorittamaan jonkin patenttivaatimuksista 1 – 10 mukaisen menetelmän mukaiset vaiheet.

12. Patenttivaatimuksen 11 mukainen vastaanotto- ja tasausjärjestelmä, **tunnettu siitä, että** järjestelmä on 4G-tiedonsiirtojärjestelmä.

Fig. 1

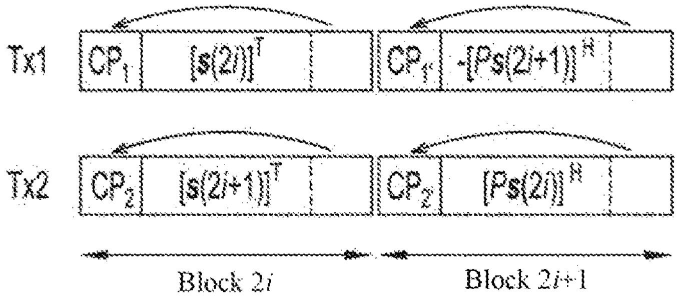


Fig. 2

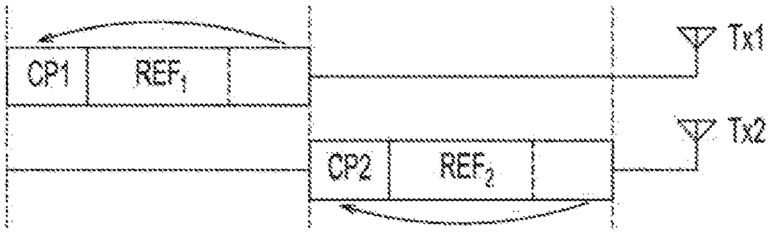


Fig. 3

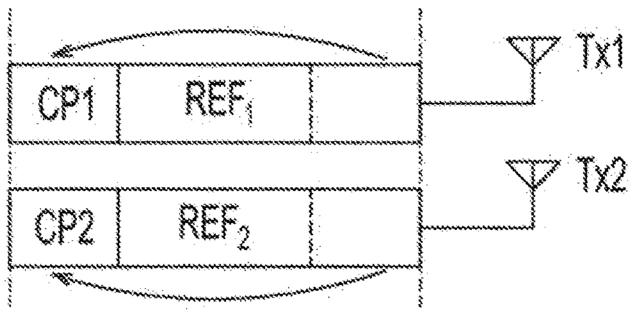


Fig. 4

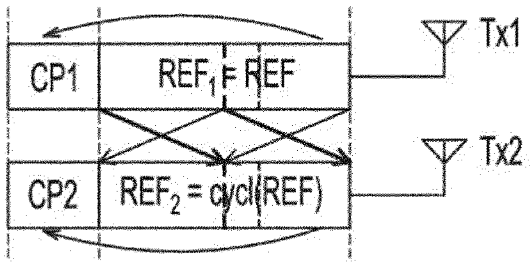


Fig. 5

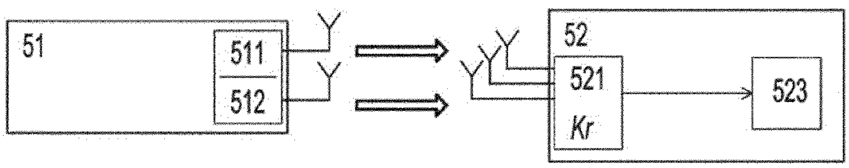


Fig. 6

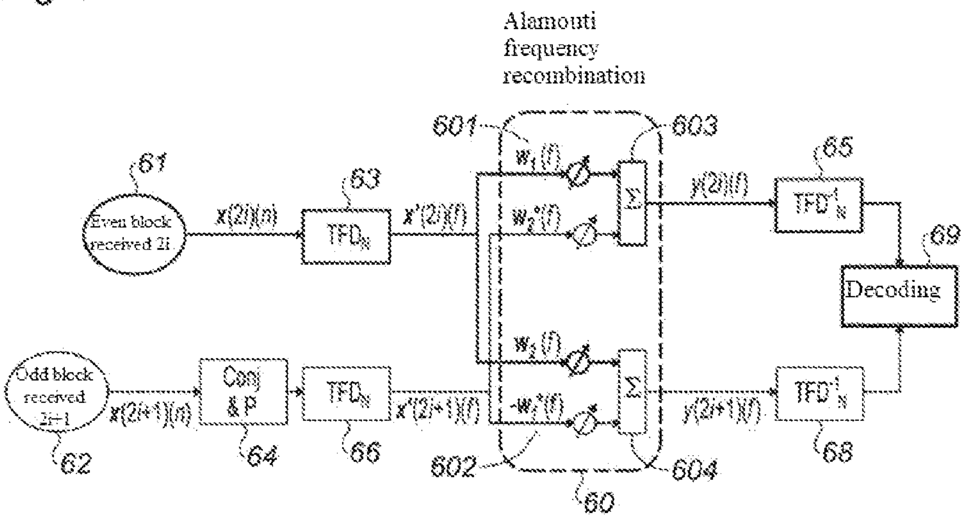

Alamouti
frequency
recombination
601
603
61
65
63
69
Even block
received 2i
x(2i)(n)
TFD_N
x'(2i)(f)
w_1(f)
w_2*(f)
Σ
y(2i)(f)
TFD^-1_N
Decoding
w_2(f)
-w_1*(f)
Odd block
received
2i+1
x(2i+1)(n)
Conj
& P
TFD_N
x'(2i+1)(f)
Σ
y(2i+1)(f)
TFD^-1_N
62
64
66
602
604
68
60

Fig. 7

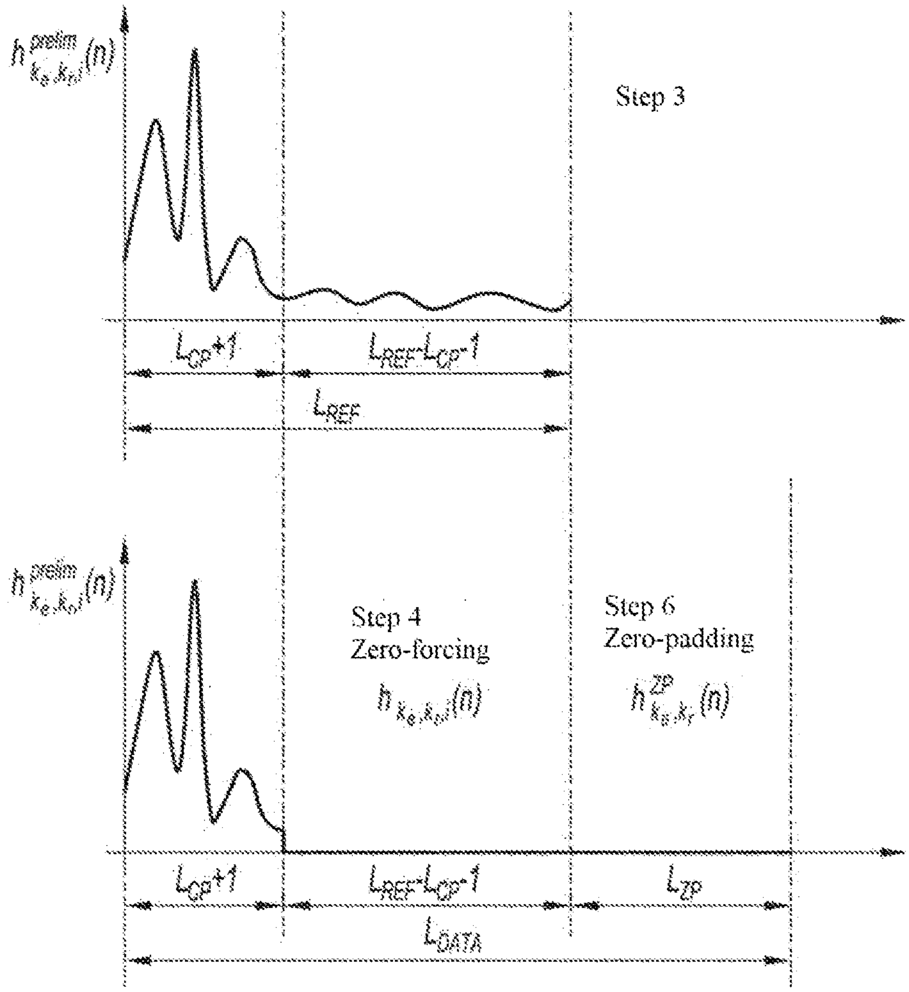

$h^{prelim}_{k_e,k_{rl}}(n)$
Step 3
$L_{CP}+1$
$L_{REF}-L_{CP}-1$
L_{REF}
$h^{prelim}_{k_e,k_{rl}}(n)$
Step 4
Zero-forcing
$h_{k_e,k_{rl}}(n)$
Step 6
Zero-padding
$h^{ZP}_{k_e,k_r}(n)$
$L_{CP}+1$
$L_{REF}-L_{CP}-1$
L_{ZP}
L_{DATA}

Fig. 8

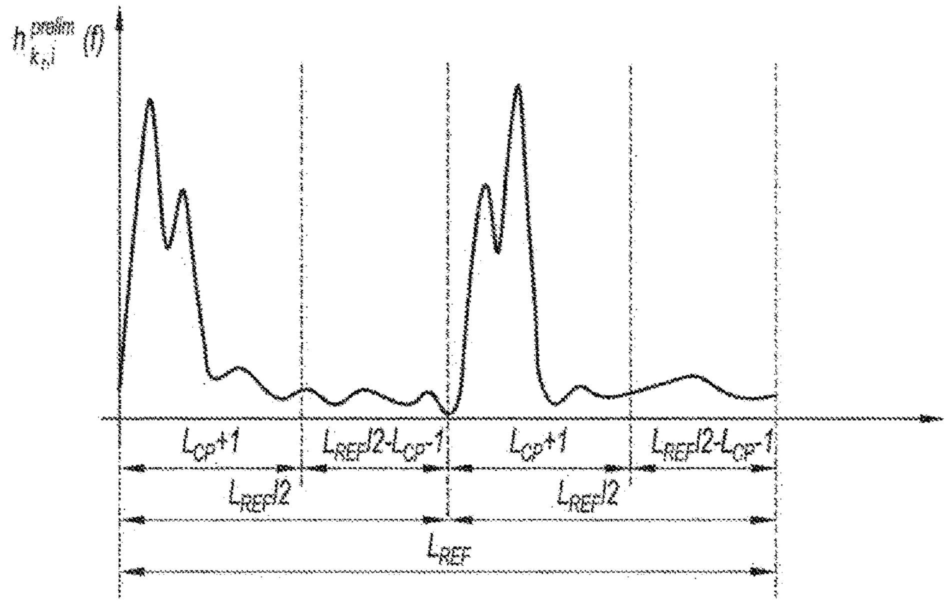
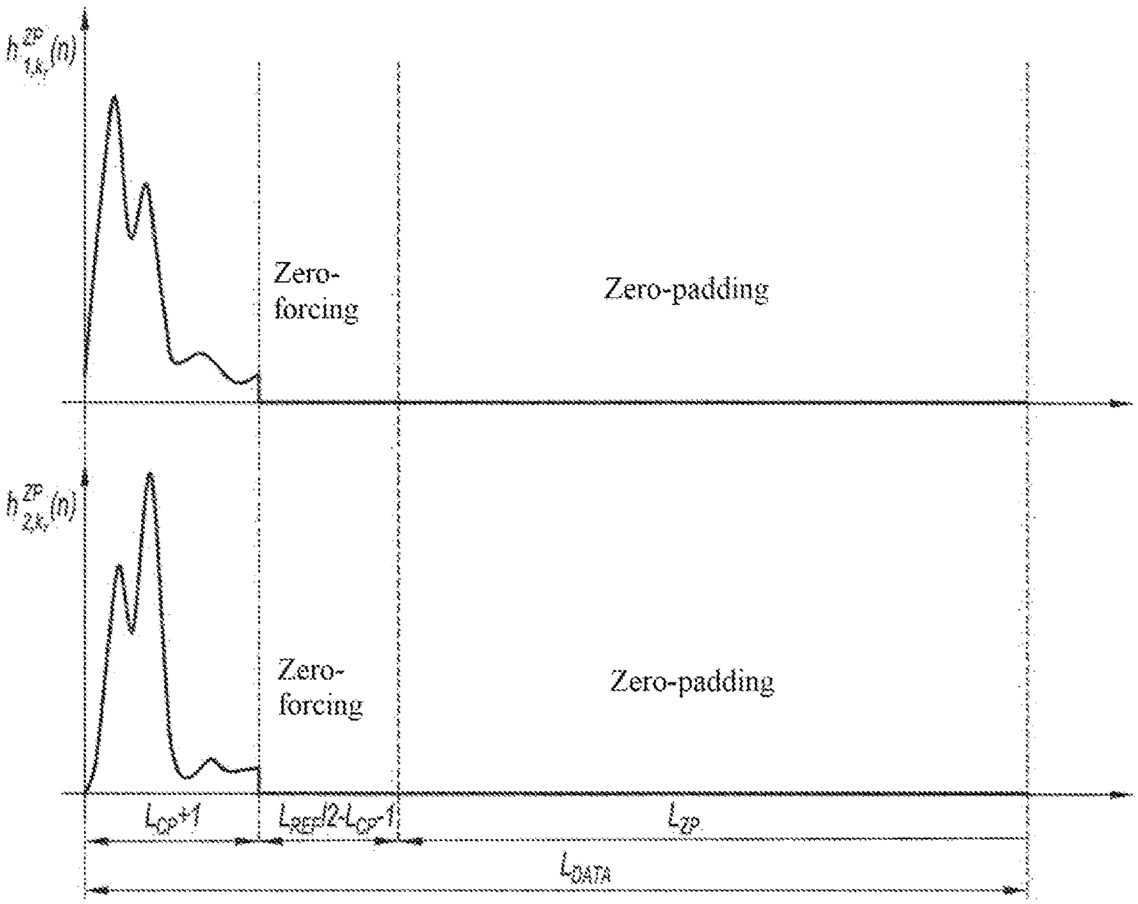

$h_{k_0,i}^{prelim}(l)$
$L_{CP}+1$
$L_{REF}/2-L_{CP}-1$
$L_{CP}+1$
$L_{REF}/2-L_{CP}-1$
$L_{REF}/2$
$L_{REF}/2$
L_{REF}

Fig. 9


$h^{ZP}_{1,k_r}(n)$
Zero-
forcing
Zero-padding
$h^{ZP}_{2,k_r}(n)$
Zero-
forcing
Zero-padding
$L_{CP}+1$
$L_{REF}/2-L_{CP}-1$
L_{ZP}
L_{DATA}