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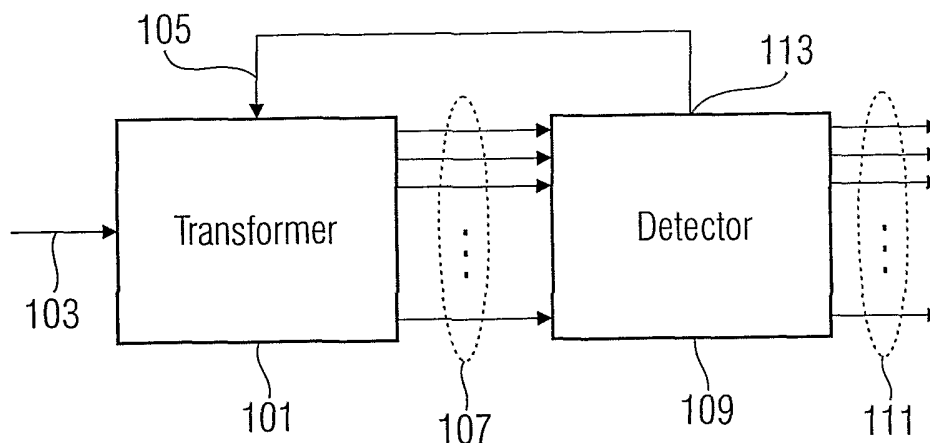
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(54) Title: APPARATUS AND METHOD FOR REDUCING A PHASE DRIFT



(57) Abstract: An apparatus for reducing a phase drift in a spectrum of a time-domain signal comprises a transformer (101) for time-frequency transforming the time-domain signal into a transformed signal in the frequency domain, the transformed signal representing a spectrum of the time domain signal, a detector (109) for detecting the phase drift, the detector (109) being further configured for generating a control signal indicating the phase drift, wherein the transformer (101) is further configured, in response to the control signal, for pre-processing the time-domain signal before time-frequency transforming or for post-processing the transformed signal after time-frequency transforming the time-domain signal in order to introduce a correction phase drift to the transformed signal for phase drift reduction, the correction phase drift at least partly compensating the phase drift.



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Apparatus and method for reducing a phase drift

The present invention is in the field of telecommunications and, in particular, in the field of digital signal processing.
5 ing.

Multi-carrier modulation, in particular, orthogonal frequency division multiplexing (OFDM) has been successfully applied for transmission in a wide variety of digital communication systems. In an OFDM transmitter, an OFDM symbol
10 to be transmitted, the OFDM symbol comprising a number of complex values resulting from mapping a group of data values onto signal space constellation points being associated with a chosen mapping scheme, e.g. QAM (QAM = Quadrature
15 Amplitude Modulation), is transformed into so-called time domain transmit signal using e.g. an inverse Fourier Transform to obtain a transmit signal to be transmitted. In order to eliminate inter-symbol interferences (ISI) and to preserve orthogonality of a received signal at a receiver,
20 cyclic prefix is inserted into a guard interval (GI) of the transmit signal, the guard interval being preferably longer than a maximum delay of the communication channel through which the transmit signal is to be transmitted.

Fig. 13 shows, by the way of example, a block diagram of a
25 conventional OFDM receiver. Firstly, the guard interval is removed from a signal received by an antenna. After a serial to parallel conversion (S/P), a Fourier Transform, for example a Fast Fourier Transform (FFT) is performed in order
30 to obtain a received version of the multi-carrier signal in frequency domain.

When transmitting an OFDM modulated signal over a multipath fading channel, the received signal will have unknown
35 amplitude and phase variations. After OFDM demodulation the channel response is described by the channel transfer function (CTF) for sub-carrier i , $H_i = H(f = i/T)$, where $1/T$ denotes the sub-carrier spacing. A snapshot of the magnitude

and phase of the observed CTF is shown in Fig. 24 illustrating that the phase of the CTF, $\phi_i = \arg(H_i)$, experiences a phase drift, defined by

$$E[\Delta\phi_i] \stackrel{\Delta}{=} \arg\{E[H_i H_{i-1}^*]\} < 0 \quad (1)$$

where $\Delta\phi_i = \arg\{H_i H_{i-1}^*\}$ accounts for the phase between sub-carriers $i-1$ and i . The phase drift defines the average change of phase between two adjacent sub-carriers. For many applications this phase drift should be as small as possible.

Phase drift in frequency domain results e.g. from frame synchronization errors, i.e. from timing errors introducing a phase shift in spectral domain, the phase shift increasing e.g. linearly over frequency.

It should be noted that non-perfect timing synchronization will also result in a change of the phase drift $E[\Delta\phi_i]$. In fact, if the timing offset, T_{off} , does not exceed the guard interval length, T_{GI} , minus the maximum delay of the channel τ_{max} , so $T_{off} \leq T_{GI} - \tau_{max}$, there will be no loss in orthogonality of the OFDM signal. However, $T_{off} \neq 0$ will result in a different phase drift $E[\Delta\phi_i]$, as is described in M. Hsieh and C. Wei, "Channel Estimation for OFDM Systems Based on Comb-Type Pilot Arrangement in Frequency Selective Fading Channels," IEEE Transactions on Consumer Electronics, vol. 44, pp. 217-225, Feb. 1998.

The phase drift may degrade the performance if space frequency codes or differential modulation is used. Furthermore, the phase drift also degrades the performance of polynomial interpolation algorithms, e.g. linear or spline interpolation.

Fig. 14 shows a phase and a magnitude of a snapshot of a channel transfer function (CTF). Fig. 15 shows a corre-

sponding time domain snapshot of a channel impulse response (CIR).

The non-zero phase drift $E[\Delta\phi_i]$ is typical for any OFDM
5 system. It is due to the structure of the channel impulse response (CIR), which is the inverse Fourier Transform of the CTF. The CIR which generates the CTF in Fig. 14 is shown in Fig. 15. As a general property, the CIR is only non-zero within the range $[0, \tau_{\max}]$, where $\tau_{\max}$ denotes the
10 maximum delay of the channel, or $[T_{\text{off}}, \tau_{\max} + T_{\text{off}}]$, T_{off} being a frame offset.

Since the CIR is related to the CTF by a Fourier Transform it may be viewed as the spectrum of the CTF.

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A non-zero phase drift can degrade a performance of an OFDM system, for example in a case of a differential phase coding. This is due to the fact that a phase drift introduces an additional phase term leading to phase errors when performing a differential phase demodulation, so that information detection errors occur.

Furthermore, the phase drift can introduce channel estimation errors when estimating the communication channel, for example when estimating the channel transfer function, in order to e.g. equalize the received multi-carrier signal in frequency domain.

In order to estimate e.g. the channel transfer function, an OFDM transmitter may introduce so-called pilot symbols being known in the receiver for channel estimation. Usually, the pilot symbols are used for modulating sub-carriers of a OFDM signal to be transmitted. At a receiver, the pilot symbols are removed from the modulated sub-carriers by the means of demodulation, e.g. by dividing the modulated sub-carriers by corresponding pilot symbols in order to obtain sub-carrier values comprising information with respect to the channel transfer function.

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In Fig. 16, a block diagram of pilot symbol based channel estimation for OFDM is depicted. After transforming the received signal to frequency domain by the means of the Fast Fourier Transform (FFT) a transformed signal is obtained having values associated with sub-carriers, wherein only certain sub-carriers are modulated by pilot symbols for channel estimation. In order to extract the modulated sub-carriers, a de-multiplexer can be used in order to de-multiplex (DMUX) the modulated sub-carriers, which are, subsequently, provided to a channel estimator being configured for channel estimation. The channel estimator may perform the pilot demodulation mentioned above in order to extract the sub-carrier values, wherein, after having performed the de-modulation, the sub-carrier values are estimates of the channel transfer function at frequency points associated with sub-carriers modulated by the pilot symbols. In order to obtain a channel estimate for all sub-carriers, the channel estimator may, for example, be configured for performing an interpolation in order to provide channel estimates for all sub-carriers by means of interpolating between two subsequent channel estimates obtained from the modulated sub-carriers being spaced apart by a number of sub-carriers. In order to detect information contained by the transformed signal bypassed by the de-multiplexer, a detection unit (DT) is used. For example, the detection unit is configured for receiving the channel estimates in order to equalize the transformed signal. However, if a phase drift occurs, then the channel estimates are erroneous due to an additional phase term. This leads to a performance degradation while detecting the information comprised by the transform signal after having equalized the transformed signal using the erroneous channel estimates.

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In order to reduce the phase drift, a phase compensation can be performed, as is described e.g. in M. Hsieh and C. Wei, "Channel Estimation for OFDM Systems Based on Comb-

Type Pilot Arrangement in Frequency Selective Fading Channels," IEEE Transactions on Consumer Electronics, vol. 44, pp. 217-225, Feb. 1998, for the case of channel estimation. In particular, after a DFT and after a subsequent pilot
5 signal extraction, a least squares estimate of pilot signals is performed in order to provide an estimated channel transfer function. In a next step, a change in phase caused by frame error is determined from subsequent values of the channel transfer function. In a next step, the estimated
10 change in phase is removed from the estimated channel transfer function and the resulting channel transfer function is provided to a minimum mean squared error estimator for further channel estimation. After the MMSE channel estimation, the channel transfer functions of data carriers
15 are interpolated using linear or higher order interpolations. In a next step, a phase post-compensation is performed where the previously removed phase change is restored in order to provide an interpolated channel transfer function comprising the change in phase.

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For differentially modulated signals, S. M. Weinfurtnner, "Frequency-Domain Frame Synchronization for Optimum Frequency-Differential De-modulation of OFDM," in Proc. IEEE Global Telecommunications Conference (GLOBECOM '99), Rio de
25 Janeiro, Brazil, pages 857-862, 1999 discloses applying a phase correction coefficient in the differential demodulation process in order to perform a phase rotation on sub-carrier amplitudes in frequency domain for compensating frame synchronization errors.

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It is the object of the present invention to provide a concept for an efficient phase drift reduction.

This object is obtained by an apparatus for reducing a
35 phase drift according to claim 1, or by a multi-carrier receiver according to claim 19, or by a method for reducing a phase drift according to claim 27, or by a method for re-

ceiving multi-carrier signals according to claim 28, or by a computer program according to claim 29.

The present invention is based on the finding that the
5 phase drift can efficiently be reduced when a feed-back
loop phase drift reduction structure is used. More specifi-
cally, the phase drift can efficiently be reduced when a
detector for detecting the phase drift in a signal is ar-
ranged after a phase drift compensating entity in order to
10 detect the phase drift or a remaining phase drift in the
signal provided by the phase drift compensating entity and
in order to control the phase drift compensating entity in
dependence on a phase drift detected in the signal via a
feed-back loop so that e.g. in each compensation step a
15 further phase drift reduction can be achieved.

In accordance with the present invention, the functionality
of the phase drift compensating entity is comprised by a
transformer for time-frequency transforming a time domain
20 signal into a transformed signal in frequency domain. For
example, the time domain signal may be a received version
of a multi-carrier signal, the received version suffering
from e.g. frame synchronization error. Therefore, the
transformed signal representing the spectrum of the time
25 domain signal suffers from the phase drift influencing a
phase of each sub-carrier.

The detector may be configured for directly detecting the
phase drift in frequency domain in the transformed signal
30 provided by the transformer.

Alternatively, the detector may be configured for detecting
the phase drift indirectly in time domain from a pre-
processed time-domain signal. For example, the detector may
35 be configured for detecting a delay resulting from a frame
synchronisation error, and for calculating the phase drift
resulting from the delay sing e.g. a Fourier transform.

The transformed signal is provided to a detector connected to an output of the transformer for detecting the phase drift in the transformed signal, wherein the detection of the phase drift is preferably performed in frequency domain. In accordance with the present invention, the detector is configured for generating a control signal indicating the phase drift upon detecting the phase drift. The control signal is then fed-back to the transformer which is configured, in response to the control signal, for pre-processing the time-domain signal in time-domain before time-frequency transforming in order to introduce a correction phase drift to the transformed signal in frequency domain, or for post-processing the transformed signal after time-frequency transforming the time domain signal in order to directly introduce the correction phase drift to the transformed signal, the correction phase drift, in both cases, at least partly compensating the phase drift so that the phase drift is reduced.

In accordance with the present invention, the transformed signal obtained from time-frequency transforming the time domain signal, e.g. obtained from applying a Fourier Transform to the time-domain signal, may be post-processed in frequency domain by e.g. manipulating the phase of the transformed signal so that a correction phase drift is introduced in frequency domain to the transformed signal, i.e. so that a correction phase is introduced to the phase of the transformed signal in order to reduce the phase drift.

In addition, the inventive transformer has the ability of pre-processing the time domain signal in the time domain in order to indirectly introduce the correction phase drift to the transformed signal resulting from transforming the time domain signal into frequency domain after pre-processing. In this case, the time delay-phase shift correspondence of a time-frequency transform, for example of a Fourier Transform, is exploited since e.g. a time delay always intro-

duces an additional phase shift in a spectrum of a time delayed signal.

As mentioned above, the detector is configured for detecting the phase drift in the transformed signal, the transformed signal resulting from transforming the time domain signal into frequency domain signal after pre-processing the time domain signal or resulting from transforming the time domain signal into frequency domain. Therefore, the phase drift can also adaptively be reduced, wherein, in a plurality of reduction steps, a phase drift remaining after a previous compensation step may be detected and reduced.

For example, in a first step, the detector receives the transformed signal comprising the uncompensated phase drift. Then, the detector determines, on the basis of the unprocessed transformed signal, the phase drift comprised by the transformed signal and generates a control signal, which is provided to the transformer for controlling the pre-processing or post-processing or both. In response to the control signal comprising e.g. information on the detected phase drift, the transformer is configured for pre-processing or post-processing or pre- and/or post-processing the transformed signal in order to reduce the phase drift. In a next step, the detector receives the processed transformed signal comprising reduced phase drift. In a next step, the detector detects the remaining phase drift comprised by the (processed) transformed signal provided by the transformer, generates a further control signal indicating the remaining phase drift and feeds the control signal back to the transformer for further reduction of the remaining phase drift and so forth.

In addition, the phase drift can effectively be reduced when the transformer applies all inventive processing functionalities for phase drift reduction. For example, the transformer may be configured for coarsely reducing the phase drift by the means of pre-processing the time domain

signal before time frequency transforming, for transforming
the time domain signal after pre-processing and for post-
processing the resulting transformed signal in frequency
domain for obtaining a fine phase drift reduction, or vice
5 versa.

Further embodiments of the present invention will be de-
scribed with respect to the following figures, in which:

10 Fig. 1 shows a block diagram of an apparatus for reduc-
ing a phase drift in accordance with an embodi-
ment of the present invention.

Fig. 2a shows the inventive post-processing approach;

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Fig. 2b shows the inventive pre-processing approach;

Fig. 3a shows a phase of a channel transfer function;

20 Fig. 3b shows a magnitude of a channel transfer function;

Fig. 4 shows a corresponding spectrum of the channel
transfer function of Figs. 3a and 3b;

25 Fig. 5 shows a block diagram of an apparatus for reduc-
ing a phase drift in accordance with a further
embodiment of the present invention;

Fig. 6 shows a block diagram of an apparatus for reduc-
30 ing a phase drift in accordance with a further
embodiment of the present invention;

Fig. 7 shows OFDM system parameters;

35 Fig. 8 shows a power delay profile of a channel;

Fig. 9 demonstrates a performance of the inventive ap-
proach;

Fig. 10 shows an effective channel impulse response resulting when cyclically shifting a received signal in time domain;

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Fig. 11 demonstrates the performance of the inventive approach;

Fig. 12 shows a block diagram of the inventive apparatus for reducing a phase shift in accordance with a further embodiment of the present invention;

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Fig. 13 shows a block diagram of an OFDM receiver;

Fig. 14 shows a phase and a magnitude of a channel transform functions;

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Fig. 15 shows a corresponding time domain channel impulse response; and

20

Fig. 16 shows a channel estimation approach.

Fig. 1 shows a block diagram of the inventive apparatus for reducing a phase drift in accordance with an embodiment of the present invention.

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The apparatus shown in Fig. 1 comprises a transformer 101, the transformer 101 having an input 103, a control input 105 and a plurality of outputs 107 connected to a plurality of inputs of a detector 109. The detector 109 comprises a plurality of outputs 111 and a control output 113, the control output 113 being coupled to the control input 105 of the transformer 101.

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The transformer 101 is configured for receiving a time-domain signal via the input 103. If, for example, during frame synchronization being performed in order to obtain the time-domain signal an error occurs, then a spectrum of

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the time-domain signal comprises a phase drift superimposing the phase of the spectrum of the time-domain signal. For example, the phase drift may be described as a linearly increasing phase ramp superimposing the phase of the spectrum of the time domain signal. In this case, the phase change between two subsequent spectral values is constant.

In order to reduce the phase drift affecting the spectrum of the time-domain signal, which can be a received version of a transmit signal, the transformer is configured for transforming the time-domain signal into a transformed signal in frequency domain, the transformed signal representing a spectrum of the time-domain signal.

The transformed signal in frequency domain comprises a number of spectral coefficients which are provided via the plurality of outputs 107 to the detector 109 for detecting the phase drift in the transformed signal. The detector 109 is configured for detecting the phase drift in the transformed signal, for generating a control signal indicating the phase drift and to providing the control signal via the control output 113 to the transformer 101. In response to the control signal, the transformer 101 is configured for pre-processing the time domain signal before time-frequency transforming or for post-processing the transformed signal after time-frequency transforming the time domain signal in order to introduce a correction phase drift to the transformed signal, the correction phase drift at least partly compensating the phase drift for phase drift reduction.

In order to pre-process the time-domain signal for achieving the effect of introducing the correction phase drift to the transformed signal after time-frequency transforming the (pre-processed) time domain signal, the transformer may comprise a delay element for delaying the time-domain signal so that, upon exploiting the time delay/frequency shift correspondence the correction phase shift is introduced.

In accordance with a preferred embodiment of the present invention, the transformer comprises, as a delay element, a cyclic shift element for pre-processing the time-domain signal, wherein the cyclic shift element is configured for
5 receiving the control signal from the detector 109 and, in response to the control signal, for cyclically shifting the time domain signal in order to introduce the correction phase drift to the transformed signal.

10 For example, the cyclic shift element is configured for cyclically shifting the time-domain signal by a number of values, wherein the number of values depends on the phase drift detected by the detector. For example, the detector 109 is configured for determining the number of values the
15 time-domain signal is to be shifted by using the detected phase drift and exploiting the cyclic shift/frequency shift property of a time-frequency transform, for example of a Fourier Transform. The control signal provided to the transformer may in this case comprise the information on
20 the number of values the time domain signal is to be shifted by.

In accordance with a further embodiment of the present invention, the detector is configured for detecting the phase
25 shift from the transformed signal and for providing information on the detected phase drift to the transformer. The transformer, in response to the received information on the phase drift, may be configured for determining the number of values the time-domain signal is to be shifted by in exactly the same way as described above in connection with
30 the detector 109.

Preferably, the transformer comprises a Fourier transformer coupled to an output of the delay elements, e.g. to an out-
35 put of the cyclic shift element, for transforming the time-domain signal provided by the cyclic shift element into the transformed signal.

In this case, the plurality of outputs 107 comprised by the transformer 101 corresponds to a plurality of outputs of the Fourier transformer, the Fourier transformer outputting the transformed signal comprising a number of spectral values, wherein a number of the outputs 107 corresponds to the number of spectral values. In this case, the plurality of inputs of the detector is coupled to the plurality of outputs of the Fourier transformer. For example, the plurality of outputs of the Fourier transformer is directly connected with the plurality of inputs of the detector 109. In order to control the pre-processing of the time-domain signal, the control output 113 of the detector 109 may be connected to a control input comprised by the cyclic shift element, so that the detector 109 directly controls the pre-processing of the time-domain signal.

As has been described above, either the detector 109 or the transformer 101 may be configured for determining the number of values the time-domain signal is to be shifted by. For example, the number of values is determined based on an expectation value for a phase drift, which expectation value is calculated in frequency domain by, for example, averaging over a phase change between two subsequent sub-carriers. In order to obtain the number of values the time domain signal is to be shifted by, the expectation value for the phase drift may be multiplied by a number of values comprised by the transformed signal and divided by a factor of 2π , wherein the number of values of the transformed signal corresponds for example to a number of sub-carriers associated with a multi-carrier transmission scheme.

For example, the cyclic shift element is configured for performing a left shift or a right shift in dependence on a sign of the phase drift. For example, the cyclic shift element is configured for performing a left shift in order to introduce a positive correction phase drift to the transformed signal for compensating a negative phase drift, or for performing a right shift in order to introduce a nega-

tive correction phase drift to the transformed signal for compensating a positive phase drift.

5 In order to cyclically shift the time-domain input signal, the cyclic shift element may comprise a shift register, the shift register having an input for receiving the time-domain signal and an output coupled to the input, so that a cyclic shift can be performed.

10 As has been mentioned above, the transformer may be configured for post-processing the transformed signal in order to introduce the correction phase drift directly in the frequency domain for compensating the phase drift. For example, the transformer comprises a phase compensator for
15 post-processing the transformed signal in frequency domain, the phase compensator being configured for changing a phase of the transformed signal in frequency domain in order to introduce the correction phase drift. For example, the phase compensator is configured for multiplying each transformed
20 signal value by a complex value for phase shifting the transformed signal values in frequency domain.

In accordance with an embodiment of the present invention, the transformer 101 comprises either the cyclic shift element
25 for pre-processing the time domain signal or the phase compensator for post-processing the transformed signal. In the latter case, the transformer may comprise a Fourier transformer for time-frequency transforming the time domain signal, wherein the phase compensator is connected to an
30 output or to a plurality of outputs of the Fourier transformer, wherein the detector 109 is connected to an output or to a plurality of outputs of the phase compensator, so that the detector 109 only receives the transformed signal provided by the phase compensator, wherein the phase com-
35 pensator is configured for receiving the control signal generated by the detector 109 upon detecting a phase drift in the transformed signal. In order to receive the control signal, the phase compensator may comprise a control input

for receiving the control signal from the detector 109. For example, the control input of the phase compensator is directly connected to the control input 105 of the transformer 101.

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The compensator may be configured for adding the correction phase drift to a phase of the transformed signal for compensating a negative phase drift, or for subtracting the correction phase drift from the phase of the transformed signal for compensating a positive phase drift.

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In accordance with an embodiment of the present invention, the correction phase drift to be introduced by the phase compensator may be determined by the detector 109 upon detecting the phase drift. Furthermore, the correction phase drift may be determined by the transformer 109 in response to information on the phase drift provided by the detector 109. For example, the correction phase drift to be introduced by the phase compensator corresponds to the phase drift detected by the detector 109, except for the sign. In accordance with a further aspect of the present invention, the transformer may comprise both: the cyclic shift element for pre-processing the time domain signal and the phase compensator for post-processing the transformed signal for two-stage reduction of the phase drift. For example, the cyclic shift element cyclically shifts the time domain signal in order to coarsely reduce the phase drift. The time domain signal provided by the cyclic shift element is then transformed into frequency domain by e.g. a Fourier transformer coupled to the cyclic shift element in order to obtain the transformed signal. The phase compensator is coupled to an output or to a plurality of outputs of a Fourier transformer for post-processing the transformed signal, wherein an output or a plurality of outputs of the phase compensator is coupled to the detector 109 for detecting the phase drift. Preferably, the detector 109 simultaneously controls the cyclic shift element and the phase compensator for phase drift reduction.

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For example, the cyclic shift element is configured for receiving the control signal from the detector, wherein the control signal indicates a number of values the time domain
5 signal is to be shifted by in order to introduce a correction phase drift to the transformed signal for coarsely compensating the phase drift. In response to the control signal, the cyclic shift element may be configured for cyclically shifting the time domain signal by the number of
10 values in order to coarsely compensate the phase drift.

Accordingly, the phase compensator may be configured for receiving a further control signal from the detector 109, the further control signal indicating a further correction
15 phase drift to be introduced to a phase of the transformed signal for a fine compensation of the phase drift. In response to the further control signal, the phase compensator may be configured for introducing the further correction phase drift to the phase of the transformed signal.

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It is to be noted that the control signal and the further control signal may be comprised by a common control signal generated by the detector 109 upon detecting phase drifts affecting the transformed signal.

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As has been mentioned above, the number of values the time domain signal is to be shifted by or the correction phase drift to be added to a phase of the transformed signal may be determined by the detector 109 upon detecting the phase
30 drift or by the transformer 101 upon receiving a control signal indicating a detected phase drift.

In order to detect the phase drift, the detector 109 may be configured for determining an average change in phase between two subsequent values of the transformed signal,
35 wherein the average change in phase represents the phase drift.

For example, the transformed signal may comprise all spectral values resulting from transforming the time domain signal into frequency domain. In accordance with a further aspect of the present invention, the transformed signal may be composed of a set of spectral values associated with sub-carriers being, in a transmitter, modulated by pilot symbols for channel estimation purposes. For example, two subsequent sub-carriers being modulated by pilot symbols may be spaced apart by a number of sub-carriers being used e.g. for information transmission. In this case, the detector is operative for detecting the phase drift from the spectral values associated with sub-carriers being modulated by the pilot symbols. In order to extract the modulated sub-carriers, the transformer may further comprise a selector, the selector being configured for selecting the sub-carriers modulated by the pilot symbols from spectral values obtained from time-frequency transforming the time-domain signal. In this case, the set of spectral values representing the sub-carriers modulated by the pilot symbols constitutes the transformed signal. In order to demodulate the modulated sub-carriers, the transformer may further be configured for demodulating the set of spectral values constituting the transformed signal using pilot symbols wherein, by way of example only, each spectral value may be divided by an associated pilot symbol. Alternatively, each spectral value may be multiplied by a conjugate complex version of the associated pilot symbol.

In order to determine the phase drift, the detector 109 may be configured for determining an average value of a product between two subsequent values of the transformed signal, wherein one of these values may complex conjugated. For example, the detector 109 may be configured for determining an average value of a product between a value associated with a certain sub-carrier and a complex conjugate version of a value associated with a sub-carrier preceding the sub-carrier.

In order to determine the phase drift, the detector 109 may be configured for determining a phase of the average value, the phase of the average value representing the phase drift. In the case that the transformed signal comprises a set of values associated with sub-carriers being spaced apart, the phase of the average value is preferably defined by a spacing, the spacing indicating a number of sub-carriers separating two subsequent sub-carriers used for pilot symbol transmission.

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In accordance with a further aspect of the present invention, the detector may be configured for determining an average phase change between two values of the transformed signal, the two values being spaced apart by a number of values, and for dividing the average phase change by the number of values in order to detect the phase shift for each value of the transformed signal. This case may, for example, correspond to the above-mentioned case when only certain sub-carriers are used for pilot symbol transmission, and, unlike in the above embodiment, the transformer does not comprise any selector. However, the phase drift may be also derived from spectral values of the transformed signal, which are not associated with sub-carriers being used for pilot symbol transmission. In any case, a complexity reduction can be introduced since only a certain subset of values of the transformed signal will be used for detecting the phase drift.

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The present invention further provides a multi-carrier receiver comprising the apparatus for reducing a phase drift as has been described above.

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Preferably, the apparatus for reducing the phase drift is configured for reducing the phase drift in a receivable multi-carrier signal in time domain and for providing a transformed signal representing a receivable multi-carrier signal in frequency domain, the transformed signal having reduced phase drift. In other words, the inventive appara-

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tus for reducing the phase drift is utilized for transforming the receivable multi-carrier signal in time domain into frequency domain in accordance with a multi-carrier receiving scheme, for example OFDM, and simultaneously, for reducing a phase drift.

In addition, the multi-carrier receiver may comprise means for extracting information comprised by the transformed signal.

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In accordance with a further aspect of the present invention, the multi-carrier receiver may comprise a synchronizer for performing a frame synchronization in time domain, wherein the phase drift results from a frame synchronization error due to an erroneous sample time instant being considered the time instant determining a beginning of a frame in time domain.

In order to reduce the frame synchronization error, the detector 109 comprised by the apparatus for reducing the phase drift may be configured for generating a frame synchronization control signal indicating the phase drift, and providing the frame synchronization control signal to the synchronizer. The synchronizer may be configured, in response to the frame synchronization control signal, for adjusting a sample time instant in order to reduce the frame synchronization error. Therefore, the detector may also control the operation of the synchronizer so that frame synchronization errors are reduced in dependence on a detected phase drift, wherein the synchronizer is arranged before the inventive transformer for performing a frame synchronization in time domain.

In order to reduce the remaining phase drift in frequency domain, the functionality of the apparatus for reducing the phase drift can be exploited.

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In accordance with a further aspect of the present invention, the means for extracting information may comprise a differential de-modulator for differentially de-modulating the transformed signal in order to extract information from a phase change between two subsequent values of the transformed signal.

In accordance with a further aspect of the present invention, the means for extracting information may comprise a low-pass filter for filtering the transformed signal in frequency domain. Preferably, the low-pass filter comprises real valued coefficients representing a real part of a filter response, wherein filter coefficients representing the imaginary part of the filter response are set to zero. In this case, the transformer is further configured for pre-processing the receivable multi-carrier signal in time domain or for post-processing the transformed signal in order to introduce a phase shift to the transformed signal for shifting a spectrum of the transformed signal towards the passband of the low-pass filter.

Therefore, frequency domain filtering using real valued filters having symmetrical two-sided filter transfer function can be performed which further reduces a receiver's complexity.

In accordance with a further aspect of the present invention, the low-pass filter comprised by the means for extracting information may be configured for channel estimation in order to extract channel information from, for example, the receivable multi-carrier signal in time domain is a receivable version of a transmit signal being transmitted through a communication channel. For example, the low-pass filter is a Wiener filter for channel estimation.

In accordance with a further aspect of the present invention, the low-pass filter is an interpolation filter for interpolating between estimated values of the channel

transfer function, when the estimated values of the channel transfer function in frequency domain are associated with sub-carriers which are spaced apart by a number of sub-carriers. In this case, the interpolation process is performed in order to obtain estimates of intermediate values of the channel transfer function.

In accordance with a further aspect of the present invention, the means for extracting information comprises a space frequency block decoder for space-frequency block decoding the transformed signal in frequency domain when, in a receiver, space-frequency block coding is employed.

In the following, further embodiments of the present invention will be described with respect to Figs. 2 to 12.

Fig. 2a shows an OFDM receiver with phase compensation after Fourier Transform, wherein the receiver comprises an antenna 201 coupled to a means 203 for removing guard interval, the means 103 for removing the guard interval being coupled to a serial to parallel converter 205 (S/P). The serial to parallel converter 205 has a plurality of outputs coupled to a plurality of inputs of a Fourier transformer 207, the Fourier transformer 207 being configured for performing a Fast Fourier Transform (FFT). The Fourier transformer 207 has a plurality of outputs coupled to a phase compensator 209, the phase compensator 209 having a plurality of inputs being coupled to a detector not shown in Fig. 2a. It is to be noted that the Fourier transformer 207 and the phase compensator 209 are comprised by the inventive transformer mentioned above.

Fig. 2b shows a block diagram of an OFDM receiver having cyclic shift before performing the Fourier Transform by the Fourier transformer 207. The cyclic shift is performed by a cyclic shift element 301 coupled between the means 203 for removing the guard interval and the serial to parallel converter 205. In this embodiment, the cyclic shift element

301, the serial to parallel converter 205 and the Fourier transformer 207 constitute the inventive transformer, wherein the plurality of outputs of the transformer 207 is coupled to the detector, which is not shown in Fig. 2b. Furthermore, it is not shown in Figs. 2a and 2b that the detector may be configured for controlling the phase compensator 209 and the cyclic shift element 301.

As is depicted in Figs. 2a and 2b, the solutions shown therein are applicable to a wide range of OFDM receivers due to the OFDM standard conform structure. Moreover, if the phase compensation is applied according to the embodiment of Fig. 2a, there is no need to compensate the induced phase shift after channel estimation and interpolation, contrary to the prior art approaches mentioned above.

In particular, the negative effects of a non-zero phase drift from (1) can be compensated by a phase compensation unit after the FFT at the OFDM receiver, as is shown in Figs. 2a and 2b.

Another possibility to solve the problem related to the one-sided spectrum of the frequency response is to match the filter response to the observed characteristics of the OFDM signal. This is indicated in Fig. 5, where the pass-band of the channel estimation filter is chosen within the range $[0, \tau_{max}]$. However, this will always result in complex valued filter coefficients.

In order to obtain a channel estimate with a filter having real valued coefficients, the inventive approach described above can be applied, wherein cyclic shift at the receiver or phase compensation at the receiver can be applied. Cyclic shifts in the receiver are known from M. I. Rahman, K. Witrisal, D. Prasad, O. Olsen, and R. Prasad, "Performance Comparison between MRC Receiver Diversity and Cyclic Delay Diversity in OFDM WLAN Systems," in Proc. Int. Symposium on Wireless Personal Multimedia Communications (WPMC'03), Yo-

kosuka, Japan, Oct. 2003 and from A. Dammann and S. Kaiser, "Standard Conformable Antenna Diversity Techniques for OFDM and its Application to the DVB-T System," in Proc. IEEE Global Telecommunications Conference (GLOBECOM 2001), San Antonio, TX, USA, pages 3100-3105, Nov. 2001, wherein cyclic shift is applied on the receiver side in order to exploit spatial diversity. In accordance with the present invention, however, cyclic delays are not inserted in order to exploit spatial diversity, but in order to reduce phase drift or to obtain filter having real valued filter coefficients.

We propose to cyclically shift the received signal before the FFT, as shown in Fig. 2b. Cyclically shifting the received signal by $-\delta_{cyc}$ samples before the FFT results in the following phase shift after the FFT:

$$\theta_{cyc} = 2\pi\delta_{cyc}/N_{FFT} \quad (2)$$

Fig. 3a shows a phase of a channel transfer function (CTF) after cyclically shifting a time domain signal before FFT. Fig. 3b shows a corresponding magnitude of the channel transfer function.

The effect of the cyclic shift to the received signal is $Y'_i = Y_i e^{j\theta_{cyc}i}$. If δ_{cyc} is chosen according to (2), the effect to the received signal is identical, regardless whether a phase compensation after the FFT (Fig. 2a) or a cyclic shift before the FFT (Fig. 2b) is chosen.

30

In contrast to the cyclic shift operation as proposed in Fig. 2b, a phase shift according to Fig. 2a is more computationally complex. While a cyclic shift can be performed very efficiently using a shift register of $-\delta_{cyc}$ samples, a phase compensation of θ_{cyc} degree per sub-carrier requires one multiplication by $e^{j\theta_{cyc}i}$ per sub-carrier. On the other hand, given the overall complexity of an OFDM receiver, one

additional multiplication per sub-carrier may not be that significant.

Note, the cyclic shift only changes the phase of the CTF,
 5 the magnitude remains unaffected. This can be checked by
 comparing the CTF of an OFDM signal without and with cyclic
 shift shown in Figs. 14, 3a and 3b. Since the effects of
 the frequency selective channel are compensated by the
 channel estimator anyway, no other operations are neces-
 10 sary.

A snapshot of the magnitude and phase of the CTF and the
 corresponding CIR after cyclically shifting the received
 signal is shown in Figs. 3a, 3b and Fig. 4, respectively.
 15 While there are still strong variations in amplitude and
 phase due to frequency selective fading, the phase drift
 $E[\Delta\phi_i]$ is compensated. The effective CIR of Fig. 4 is
 shifted towards negative delays. Instead of a one-sided
 spectrum, the received signal now has a two-sided spectrum.

20 The cyclic shift at the receiver side can also be applied
 to channel estimation based on the discrete cosine trans-
 form (DCT). Since the DCT operates on a two-sided spectrum,
 the cyclic shift before OFDM demodulation may be very bene-
 25 ficial.

For pilot-symbol aided channel estimation (PACE) known sym-
 bols (pilots) are inserted, with an equidistant spacing of
 D_f sub-carriers. In order to reconstruct the signal, fil-
 30 tering and interpolation between the pilots is necessary.
 For channel estimation the received signal After OFDM de-
 modulation at the pilot positions, $\tilde{H}_{\tilde{i}} = H_{iD_f} + N_{iD_f}$, with
 $\tilde{i} = \lfloor i/D_f \rfloor$, are used. An FIR filter with dimension M_f can
 be written in the general form

35

$$\hat{H}_i = \hat{H}_{D_f\tilde{i} + \Delta i} = \sum_{m=0}^{M_f-1} W_m \cdot \tilde{H}_{\tilde{i}-m}, \quad i = D_f\tilde{i} + \Delta i \quad (3)$$

If the symmetries of the spectrum of W_m are such that the real and imaginary parts are an even and odd function, respectively, the coefficients of FIR interpolation and/or smoothing filters will be real valued. In general a real
 5 valued filter will only have half the computational cost of a complex valued filter.

In P. Höher et al, „Pilot-Symbol-Aided Channel Estimation in Time and Frequency“, in Proc. Communication Theory Mini-
 10 Conference (CTMC) within IEEE Global Telecommunications Conference (Globecom 97), Phoenix, AZ, USA, pp.90-96, 1997, PACE by Wiener filtering was proposed.

In accordance with a further aspect of the present invention, if a Wiener interpolation filter (WIF) with model mismatch is chosen, the filter $W = [W_1, \dots, W_{M_f}]^T$ is designed such that it covers a great variety of power delay profiles. Accordingly, a rectangular shaped power delay profile with maximum delay T_w fulfils this requirement. The
 15 Fourier Transform of a uniform power delay profile, which is non-zero within the range $[0, T_w]$, yields the frequency correlation
 20

$$R_{HH}[\Delta i] = \frac{T \sin(\pi T_w \Delta i / T)}{\pi T_w \Delta i} \cdot e^{-j\pi T_w \Delta i / T} \quad (4)$$

25

Due to the complex phase term in $R_{HH}[\Delta i]$ the corresponding Wiener filter will also have complex coefficients.

However, a complexity of a filtering operation or of a
 30 channel estimation operation when using filter having complex valued coefficients is insignificantly increased when compared with filtering or channel estimation using filters having real valued coefficients only.

35 Suppose the CIR without cyclic shift is within the range $[0, \tau_{\max}]$ then a cyclic shift of $-\delta_{\text{cyc}}$, denoting a left shift, will result in an effective CIR which is non-zero within

the range $[-\delta_{cyc}, \tau_{max} - \delta_{cyc}]$. If the cyclic shift is chosen such that

$$\delta_{cyc} = \frac{T_w - \Delta_w}{2} \quad (5)$$

5

the effective CIR will now be non-zero only within the passband of the channel estimation filter. Given a filter with low-pass characteristics, having a symmetric two-sided passband within the range $[-T_w/2, T_w/2]$, the effective CIR of the cyclically shifted signal will pass through the filter undistorted, as illustrated in Fig. 4. If the maximum delay of the channel τ_{max} can be estimated, T_w could be chosen according to $T_w = \tau_{max} + \Delta_w$, where Δ_w accounts for a roll-off delay, which may be inserted to non-perfect slope of the filter response. On the other hand, if τ_{max} is not known it can be upper bounded by T_{GI} , so $T_w = T_{GI}$.

10

Now we can use a real valued frequency correlation function to generate the filter coefficients

20

$$R_{HH}^{[cyc]}[\Delta i] = \frac{T \sin(\pi T_w \Delta i / T)}{\pi T_w \Delta i} \quad (6)$$

Therefore, the WIF matched to the uniform power delay profile $[-T_w/2, T_w/2]$ is also real valued. This means that the computational cost is cut by half.

25

Alternatively to a mismatched WIF any FIR low pass interpolation filter benefits from the proposed cyclic shift. As long as the filter is matched to a passband of $[-T_w/2, T_w/2]$ the filter will be real valued. For such a filter, the performance will be optimum if the signal to be filtered passes through the filter unchanged.

30

35

Fig. 5 shows an OFDM receiver with phase compensation after the FFT in accordance with a further embodiment of the present invention.

Unlike the embodiment of Fig. 2a, the OFDM receiver shown in Fig. 5 comprises a phase compensator 501 having a plurality of outputs coupled to a detector 503, the detector 503 having a control output 505 coupled back to a control input of the phase compensator for providing information on the detected phase drift Θ_{cyc} .

Fig. 6 shows an OFDM receiver with cyclic shift before the FFT. Unlike the embodiment shown in Fig. 2b, the OFDM receiver shown in Fig. 6 comprises a cyclic shift element 601 coupled between the means 203 for removing guard interval and the serial to parallel converter 205, and a detector 603 coupled to the plurality of outputs of the Fourier transformer 207, wherein the detector 603 comprises a control output 605 coupled back to a control input of the cyclic shift element 601 for providing information on the number of values the time domain signal provided by the means 203 for removing the guard interval is to be shifted by.

Dependent on the application the cyclic shift may be chosen to compensate the phase drift of (1), that is

$$\delta_{cyc} = E[\Delta\phi_i] \cdot \frac{N_{FFT}}{2\pi} \quad (7)$$

OFDM systems with differential modulation or space-frequency coded OFDM systems will have an improved performance if $E[\Delta\phi_i]$ is estimated sufficiently well.

In order to integrate this phase drift compensation in the proposed receiver structure, an adaptive implementation appears attractive, as shown in Fig. 5. For the proposed receiver performing a cyclic shift before the FFT, this solution can be implemented as shown in Fig. 6. Initially, the cyclic shift, δ_{cyc} may be set to a default value within the range $[0, T_{GI}/2]$. Then, δ_{cyc} can be estimated and fed back to

the cyclic shifting unit. Note, the phase drift is expected to change only on a long term basis, so no frequent updates are necessary.

- 5 In the following, a performance of the inventive approach for space-frequency block codes (SFBC) will be described.

In order to generate BER curves, an OFDM system using a space-frequency block code (SFBC) with $N_T = 2$ transmit antennas has been implemented. An OFDM system with $N_T = 2$ transmit and $N_R = 1$ receive antennas is used. The system parameters of the OFDM system and of the channel model are shown in Fig. 7. The channel is modelled by a tap delay line model with $Q_0 = 12$ taps, a tap spacing of
10 $\Delta\tau = 16 \cdot T_{sp1}$, with an exponential decaying power delay profile, illustrated in Fig. 8. The total transmit power of the system is fixed, such that the total transmit power of a N_T antenna system is equivalent to a single antenna system. No outer channel coding has been employed.

20

It is seen that for the considered system the BER floor can be somewhat reduced by optimizing the cyclic shift δ_{cyc} . For the considered parameters the optimum cyclic shift is about $44T_{sp1}$, which result in the best performance. It can also be
25 seen that the accuracy of δ_{cyc} does not need to be high.

In accordance with the present invention, a performance of a polynomial interpolator can significantly be optimized. For PACE interpolation in frequency and time direction is
30 necessary. While in time direction the Doppler power spectrum has in general a symmetric two-sided and real valued profile (at least approximately), the power delay profile is real valued but one-sided. In the following, the benefit of inserting a cyclic shift, δ_{cyc} will be described for a
35 linear interpolator. The results are applicable for higher order polynomial interpolators as well.

Fig. 9 shows BER vs E_b/N_0 for $N_T = 2$, and various cyclic shifts δ_{cyc} .

In accordance with the present invention, a performance of
 5 a polynomial interpolator can significantly be optimised. For linear interpolation, two successive pilot sub-carriers are used to determine the channel response for sub-carrier located in between these two pilots. For sub-carrier i , the channel estimate is given by

10

$$\hat{H}_i = \hat{H}_{D_f \tilde{i} + \Delta i} = \left(1 - \frac{\Delta_i}{D_f}\right) \tilde{H}_{\tilde{i}} + \frac{\Delta_i}{D_f} \tilde{H}_{\tilde{i}+1}, \quad i = D_f \tilde{i} + \Delta i$$

15 It is instructive to describe linear interpolation by a FIR filter from (3). A linear interpolator can be described by the filter response

20

$$W_m^{[lin]} = \begin{cases} 1 - |m|/D_f, & |m| \leq D_f \\ 0, & \text{elsewhere} \end{cases} \quad (8)$$

Inserting $W_m^{[lin]}$ into FIR filter equation from (3) will accomplish linear interpolation. The distortion imposed by
 25 the linear interpolator can be assessed by considering the spectrum of (8), i.e. the inverse Fourier Transform of $W_m^{[lin]}$, given by

30

$$w^{[lin]}(\tau) = \frac{1}{D_f} \left(\frac{\sin(2\pi\tau D_f/2)}{\sin(2\pi\tau/2)} \right)^2 \quad (9)$$

The spectrum of the linear interpolator is plotted in Fig. 10. First of all, a higher over-sampling rate compared
 35 to an ideal low-pass filter will be required. Second, for the best performance the spectrum of the signal to be interpolated should be concentrated around $\tau = 0$. Since, the CIR is zero only within the range $[0, \tau_{max}]$, a cyclic shift of

$\delta_{cyc} = E[\Delta\varphi_i] \frac{N_{FFT}}{2\pi}$, according to (7), will improve the interpolation performance.

Fig. 10 shows a time domain channel impulse response snapshot after cyclically shifting the signal before the FFT. In Fig. 10, also a frequency response of a linear interpolator is shown.

Fig. 11 shows MSE vs SNR for polynomial interpolation algorithms with a cyclic shift of $\delta_{cyc}=0$ (dashed lines) and $\delta_{cyc}=44$ (solid lines).

In the following, a performance of the inventive approach for some polynomial interpolation algorithms will be described. In Figure 13 the MSE is plotted against the SNR for various polynomial interpolation algorithms. It is seen that the linear interpolator with cyclic shift has a significantly lower error floor. The same is true for the spline interpolator, which gets close to the optimum Wiener filter if a cyclic shift is inserted.

In the following, further advantages associated with the inventive concept will be indicated.

25

Possible applications of the proposed cyclic shift at the OFDM receiver are e.g.:

- Providing real valued filter coefficients for PACE. This will cut the number of required multiplication per channel estimate by a factor of two, with identical performance.
- Improving the performance of polynomial interpolation algorithms. The error floor caused by the interpolation error of polynomial interpolation algorithms can be improved significantly.

- Improving performance of differentially modulated signals and space frequency codes.

Compared to inserting a phase shift, which requires one
 5 complex multiplication per sub-carrier, a cyclic shift operation is significantly easier to implement. Furthermore, no post-processing after channel estimation and/or demodulation is required. For instance, the solution proposed in
 10 M. Hsieh and C. Wei, "Channel Estimation for OFDM Systems Based on Comb-Type Pilot Arrangement in Frequency Selective Fading Channels," IEEE Transactions on Consumer Electronics, vol. 44, pp. 217-225, Feb. 1998. requires an additional multiplication per sub-carrier after interpolation.

15 In the following, orthogonal frequency division multiplexing (OFDM) will be described.

For OFDM the signal stream is divided into N_c parallel sub-streams, typically for any multi-carrier modulation
 20 scheme. The i^{th} sub-stream, commonly termed sub-carrier, of the ℓ^{th} symbol block, named OFDM symbol, is denoted by $X_{\ell,i}$. An inverse DFT with N_{FFT} points is performed on each block, and subsequently the guard interval having N_{GI} samples is inserted to obtain $x_{\ell,n}$. After D/A conversion, the
 25 signal $x(t)$ is transmitted over a mobile radio channel with response $h(t, \tau)$. Assuming perfect synchronization, the received signal of the equivalent baseband system at sampling instants $t = [n + \ell N_{sym}]T_{sp1}$ is in the form

30

$$y_{\ell,n} \stackrel{\Delta}{=} y([n + \ell N_{sym}]T_{sp1}) = \int_{-\infty}^{\infty} h(t, \tau) \cdot x(t - \tau) d\tau + n(t) \Big|_{t=[n+\ell N_{sym}]T_{sp1}} \quad (10)$$

where $n(t)$ represents additive white Gaussian noise, and
 35 $N_{sym} = N_{FFT} + N_{GI}$ accounts for the number of samples per OFDM symbol. At the receiver, the guard interval is removed and the information is recovered by performing a DFT on the received block of signal samples, to obtain the output of the

OFDM demodulation $Y_{\ell,i}$. The received signal after OFDM demodulation is given by

$$Y_{\ell,i} = X_{\ell,i}H_{\ell,i} + N_{\ell,i} \quad (11)$$

5

where $X_{\ell,i}$ and $H_{\ell,i}$ denotes the transmitted information symbol and the channel transfer function (CTF) at sub-carrier i of the ℓ^{th} OFDM symbol, respectively. The term $N_{\ell,i}$ accounts for additive white Gaussian noise (AWGN) with zero mean and variance N_0 . It is assumed that the transmitted signal consists of L OFDM symbols, each having N_c sub-carriers.

We focus on channel estimation in frequency direction (sub-carrier index i). Thus, the index denoting the OFDM symbol, ℓ , will be dropped in the following.

We consider a time-variant frequency selective fading channel, modeled by a tapped delay line with Q_0 non-zero taps. It is generally assumed that the channel is time limited by the maximum delay $\tau_{\max}$. Then the channel impulse response (CIR) $h(t, \tau)$ is zero outside the range $[0, \tau_{\max}]$. It is commonly assumed that the channel impulse response (CIR) is approximately constant during one OFDM symbol, so the time dependency of the CIR within one OFDM symbol can be dropped, i.e. $h(t, \tau) \approx h(\tau)$.

The channel transfer function (CTF) of (11), is the Fourier Transform of the CIR $h^{(\mu)}(\tau)$. Sampling the result at frequency $f = i/T$, the CTF at sub-carrier i is defined by $H_i = H(f = i/T)$, where $H(f)$ accounts for the analog CTF, and $T_{\text{sym}} = (N_{\text{FFT}} + N_{\text{GI}})T_{\text{sp1}}$ and $T = N_{\text{FFT}}T_{\text{sp1}}$ represents the OFDM symbol duration with and without the guard interval, respectively.

35

If the guard interval is longer than the maximum delay of the channel, i.e. $T_{\text{GI}} \geq \tau_{\max}$, the orthogonality at the re-

ceiver after OFDM demodulation is maintained, and the received signal of (11) is obtained.

In the following, an OFDM receiver with a cyclic shift at the receiver side will be described.

In this section the effective channel model after the cyclic shift operation before the FFT. After the cyclic prefix is removed the received signal from (10) is cyclically delayed by $-\delta_{cyc}$ samples

$$\begin{aligned} Y_n^{[cyc]} &\stackrel{\Delta}{=} Y_{(n+\delta_{cyc}) \bmod N_{FFT}} = \int_{-\infty}^{\infty} h(t - \tau) \cdot x(\tau) d\tau + n(t) \Big|_{t=[n+\delta_{cyc}]T_{sp1}} \\ &= \int_{-\infty}^{\infty} h(t + \delta_{cyc}T_{sp1} - \tau) \cdot x(\tau) d\tau + n(t + \delta_{cyc}T_{sp1}) \Big|_{t=nT_{sp1}} \end{aligned} \quad (12)$$

It is instructive to define an effective CIR, which describes the equivalent CIR of the cyclically shifted signal

$$h^{[cyc]}(\tau) = h(\tau + \delta_{cyc}T_{sp1}) \quad (13)$$

While the maximum delay of the channel τ_{max} is not changed, the effective CIR, $h^{[cyc]}(\tau)$, is now non-zero within the range $[-\delta_{cyc}, \tau_{max} - \delta_{cyc}]$.

The FFT translates a cyclic delay into phase shifts. The effective CTF of the cyclic receiver is described by

$$H_i^{[cyc]} = H_i e^{j2\pi i \delta_{cyc} / N_{FFT}} \quad (14)$$

Now the cyclically shifted received signal after OFDM demodulation can be expressed as

$$Y_i = X_i H_i^{[cyc]} + N_i = X_i H_i e^{j2\pi i \delta_{cyc} / N_{FFT}} + N_i \quad (15)$$

In the following, the principles of pilot-symbol aided channel estimation for OFDM will be described

For pilot-symbol aided channel estimation (PACE), known symbols (pilots) are multiplexed into the data stream, which are used as side information to estimate the channel. PACE was first introduced for single carrier systems and
 5 required a flat-fading channel. To describe pilot symbol-assisted channel estimation it is useful to define a subset of the received signal sequence containing only the pilots, $\{\tilde{X}_i^{(\mu)}\} = \{X_i^{(\mu)}\}$, with $i = \tilde{i}D_f$. So, the pilot sequence is transmitted at a D_f times lower rate $\tilde{i} = \lfloor i/D_f \rfloor$ in frequency di-
 10 rection. (As a general convention, variables describing pilot symbols will be marked with a $\sim$ in the following.) It is assumed that the pilots $\tilde{X}_i$ are chosen from a PSK constellation, so $|\tilde{X}_i| = 1$.

15 After OFDM demodulation, the received signal Y_i of (11) is obtained. For channel estimation the received signal at the pilot positions are de-multiplexed from the data stream, to obtain the received pilot sequence

$$20 \quad \tilde{Y}_i \stackrel{\Delta}{=} \tilde{X}_i \tilde{H}_i + \tilde{N}_i = X_i H_i + N_i, \quad \text{with } \{i\} \in G \quad (16)$$

where G is the subset of the OFDM frame containing the pilots.

25 In the following, OFDM channel estimation by FIR filtering will be described.

The first step in the channel estimation process is to remove the modulation of the pilot symbols, which provides an
 30 initial estimate of the CTF at pilot positions

$$\tilde{H}_i = \tilde{X}_i^* \tilde{Y}_i = \tilde{H}_i + \tilde{X}_i^* \tilde{N}_i \quad (17)$$

with $\tilde{X}_i \tilde{X}_i^* = 1$.

35

The channel estimator uses the demodulated pilots $\tilde{H}_i$ from (17) to yield the channel estimate

$$\hat{H}_i = \sum_{m=0}^{M_f-1} W_m \cdot \tilde{H}_{i-m} \quad (18)$$

where M_f denotes the filter order, i.e. the number of coefficients of the FIR filter W_m .

5

The FIR filter $W = [W_0, \dots, W_{M_f-1}]^T$ may be implemented as e.g. low-pass interpolation filters, polynomial interpolators, or Wiener interpolation filters. The Wiener interpolation filter minimizes the mean squared error (MSE) between the
 10 desired response H_i and the observation, i.e. the received pilot symbols. This means that knowledge about the channel statistics is required. In contrast, low-pass interpolation filters and polynomial interpolators do not assume any knowledge of the channel statistics.

15

For multi-carrier systems the observed channel is typically correlated in two dimensions, frequency and time. Moreover, the extension to PACE in two dimensions is possible.

20 In the following, Wiener filtering will be described.

The Wiener interpolation filter (WIF) is implemented by a FIR filter with M_f taps, according to (18). The WIF, $W'[\Delta i]$, is obtained by solving the Wiener-Hopf equation

25

$$W' = R'_{HH}[\Delta i] R'_{HH}^{-1} \quad (19)$$

In order to generate WIF, knowledge of the auto-correlation matrix and the cross-correlation vector are required

30

$$\begin{aligned} R'_{HH} &= E\{\tilde{H}\tilde{H}^H\} = R'_{HH} + N_0 I \quad \in C^{M_f \times M_f} \\ R'_{HH}[\Delta i] &= E\{H_i \tilde{H}^H\} = R'_{HH}[\Delta i] \quad \in C^{1 \times M_f} \end{aligned} \quad (20)$$

35 The entry of the m^{th} row and n^{th} column of the auto-correlation matrix of the CTF at pilot positions is given by

$$\{R'_{\tilde{H}\tilde{H}}\}_{m,n} = E[\tilde{H}_{i-n}\tilde{H}_{i-m}^*] = E[H_{(i-n)D_f}H_{(i-m)D_f}^*] = R'_{HH}[(m-n)D_f] \quad (21)$$

Provided the channel can be described by a tap delay line, the frequency correlation, $R'_{HH}[(m-n)D_f]$, between sub-carriers spaced $\Delta f = \Delta i/T$ Hz apart becomes

$$R'_{HH}[\Delta i] \stackrel{\Delta}{=} R'_{HH}(\Delta i/T) = \sum_{q=1}^{Q_0} \sigma_q^2 e^{-j2\pi\tau_q \Delta i/T} \quad (22)$$

The m^{th} entry of the cross-correlation vector $R'_{\tilde{H}\tilde{H}}[\Delta i] = E[H_i\tilde{H}^H]$ can be expressed as

$$\begin{aligned} \{R'_{\tilde{H}\tilde{H}}[\Delta i]\}_m &= E[H_i\tilde{H}_{i-m}^*] = E[H_iH_{(i+m)D_f}^*] \\ &= R'_{HH}[\Delta i - mD_f] \quad \Delta i = i - D_f\tilde{i} \end{aligned} \quad (23)$$

In the following, the inventive mismatched estimator will be described

20

For the WIF the auto and cross-correlation functions need to be estimated at the receiver. It may be prohibitive to estimate the filter coefficients during operation in real time. Alternatively, a robust estimator with a model mismatch may be chosen. The filter W is designed such that it covers a great variety of power delay profiles. For example, a rectangular shaped power delay profile with maximum delay T_w fulfils this requirement. This assumption provides the frequency correlation function of the mismatched estimator, $R_{HH}^{(cyc)}[\Delta i]$, from (6). The mismatched estimator is determined by substituting $R_{HH}^{(cyc)}[\Delta i]$ from (6) into (21) and (23). Then the Wiener-Hopf equation (19) needs to be determined only once. By using a mismatched estimator the filter coefficients can be pre-computed and stored.

35

It is important to note that the parameters of the robust estimator should always be equal or larger than the worst case channel conditions, i.e. largest propagation delays

and maximum expected velocity of the mobile user. Furthermore, the average SNR at the filter input, γ_w , which is used to generate the filter coefficients, should be equal or larger than actual average SNR, so $\gamma_w \geq \gamma_c$. In order to
 5 determine the channel estimator only T_w , F_w , and γ_w are required. If the maximum delay of the channel τ_{max} is not known it can be upper bounded by the guard interval duration T_{GI} . Since the filter should also satisfy the sampling theorem, the filter pass-bands can be chosen within the
 10 range

$$\tau_{max} \leq T_w \leq T_{GI} \leq T/D_f \quad (24)$$

In the following, estimating the phase drift will be described
 15

The phase drift, $E[\Delta\phi_i]$, specifies the average change in phase between two adjacent sub-carriers, as defined in (1). One possibility to estimate the phase drift is

$$\Delta\hat{\phi}_i = \arg\left\{\frac{1}{N-1} \sum_{i=2}^N H_i H_{i-1}^*\right\} \approx E[\Delta\phi_i] \quad (25)$$

which provides a sufficiently accurate estimate. More sophisticated algorithms to estimate $E[\Delta\phi_i]$ are described in
 25 S. Kay, "A Fast and Accurate Single Frequency Estimator", IEEE Transactions in Acoustics, Speech and Signal Processing", vol. 37, pp. 1987-1990, December 1989.

Obviously, the CTF H_i is not available at the receiver.
 30 Instead a noisy estimate of H_i can easily be generated by $H_i \approx Y_i/\hat{X}_i$, where $\hat{X}_i$ is a hard decision of X_i . The accuracy of (25) will degrade due to noise and decision feedback effects. Fortunately, $\Delta\hat{\phi}_i$ does not need to be particularly accurate.

35 If pilot symbols are transmitted, e.g. for channel estimation, the phase drift may be estimated by

$$\Delta\hat{\phi}_i = \frac{1}{D_f} \cdot \arg\left\{\frac{1}{N-1} \sum_{i=2}^N \tilde{H}_i \tilde{H}_{i-1}^*\right\} \approx E[\Delta\phi_i] \quad (26)$$

5 where $\tilde{H}_i$ was defined in (17). The division through D_f becomes necessary, since the pilots are spaced D_f sub-carriers apart, i.e. D_f times the phase drift is estimated.

10 Fig. 12 shows a further embodiment of an OFDM receiver in accordance with the present invention.

Unlike the embodiment of Fig. 2a, the OFDM receiver shown in Fig. 12 comprises a de-multiplexer 1201 for de-
 15 multiplexing sub-carriers being modulated by pilot symbols for channel estimation. Optionally, the de-multiplexer 1201 (DMUX pilots) is configured for demodulating the sub-carriers being modulated by the pilot symbols. The de-multiplexer has an output coupled to the inventive apparatus 1203 for reducing the phase drift. The apparatus 1203
 20 for reducing the phase drift is coupled to a channel estimator 1205 being configured for estimating the channel transfer function in frequency domain. The channel estimator is coupled to means 1207 being configured for introducing back the compensated phase shift so that all channel
 25 influences may be taken into account. The means 1207 for phase post-compensation is coupled to a means 1209 for extracting information comprised by a transformed signal provided by the FFT 207. In other words, the means 1209 for
 30 extracting information is configured for determining an information amount comprised by a signal provided by the de-multiplexer 1201.

In accordance with a further aspect of the present invention,
 35 the apparatus 1203 for reducing the phase drift (phase drift compensator) may be configured for providing a signal containing information on the phase drift to the means 1207 for phase post-compensation, so that in response

to the signal provided by the apparatus 1203, phase post-compensation can be performed.

Moreover, the present invention provides concepts for filtering and interpolation for OFDM. Generally, the frequency response of the received signal after OFDM demodulation has a one-sided spectrum. By cyclically shifting the received signal before the FFT, the one-sided spectrum can be transformed into a symmetric two-sided spectrum. In general, the performance of standard interpolation algorithms such as linear or spline interpolation can be improved if the cyclic shift is appropriately chosen. Furthermore, the performance of space frequency codes and differential modulation can also be improved. If the symmetries of the two-sided spectrum are such that the real and imaginary parts are an even and odd function, respectively, the coefficients of FIR interpolation and/or smoothing filters become real valued, which cuts the computational cost by half.

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

CLAIMS

1. Apparatus for reducing a phase drift in a spectrum of
a time-domain signal, the apparatus comprising:
5 a transformer (101) for time-frequency transforming
the time-domain signal into a transformed signal in
the frequency domain, the transformed signal repre-
senting a spectrum of the time domain signal;
10 a detector (109) for detecting the phase drift, the
detector (109) being further configured for generating
a control signal indicating the phase drift;
15 wherein

the transformer (101) is further configured, in re-
sponse to the control signal, for pre-processing the
time-domain signal before time-frequency transforming
20 or for post-processing the transformed signal after
time-frequency transforming the time-domain signal in
order to introduce a correction phase drift to the
transformed signal for phase drift reduction, the cor-
rection phase drift at least partly compensating the
25 phase drift.
2. Apparatus according to claim 1, wherein the detector
(109) is configured for detecting the phase drift in
the transformed signal provided by the transformer
30 (101) or wherein the detector is configured for de-
tecting the phase drift from the time-domain signal
after pre-processing.
3. Apparatus according to claim 1 or 2, the transformer
35 (101) comprising a cyclic shift element (301) for pre-
processing the time-domain signal, the cyclic shift
element (301) being configured for receiving the con-
trol signal and, in response to the control signal,

for cyclically shifting the time-domain signal in order to introduce the correction phase drift to the transformed signal.

- 5 4. Apparatus according to claim 1 to 3, the detector (107) being coupled to an output of the transformer (101) for detecting the phase drift in the transformed signal.
- 10 5. Apparatus according to claim 3 or 4, the transformer (101) comprising a Fourier transformer coupled to an output of the cyclic shift element for transforming the time-domain signal provided by the cyclic shift element into the transformed signal.
- 15 6. Apparatus according to claim 5, wherein the detector is coupled to an output of the transformer (101) for detecting the phase drift in the transformed signal.
- 20 7. Apparatus according to claims 3 to 6, the detector (109) comprising a control output, the control output being connected to a control input of the cyclic shift element.
- 25 8. Apparatus according to claims 3 to 7, the cyclic shift element being configured for cyclically shifting the time-domain signal by a number of values, the detector (109) being configured for determining the number of values based on the following equation:

30

$$\delta_{cyc} = E[\Delta\phi_i] \cdot \frac{N_{FFT}}{2\pi}$$

35

wherein $E[\Delta\phi_i]$ denotes an expectation value for a phase drift $\Delta\phi_i$ detected for an i^{th} carrier of N_{FFT} sub-carriers.

9. Apparatus according to claims 3 to 8, the cyclic shift element being configured for performing a left shift

- in order to introduce a possible correction phase drift to the transformed signal for compensating a negative phase drift, or for performing a right shift in order to introduce a negative correction phase drift to the transformed signal for compensating a positive phase drift.
- 5
10. Apparatus according to claims 3 to 9, the cyclic shift element comprising a shift register, the shift register having an input and an output coupled to the input.
- 10
11. Apparatus according to claims 1 to 10, the transformer (101) comprising a phase compensator for post-processing the transformed signal in frequency domain, the phase compensator being configured for changing a phase of the transformed signal in frequency domain in order to introduce the correction phase drift.
- 15
12. Apparatus according to claim 11, the transformer (101) comprising a Fourier transformer for time-frequency transforming the time-domain signal, the phase compensator being connected to an output of the Fourier transformer, the detector (109) being connected to an output of the phase compensator, the phase compensator being configured for receiving the control signal from the detector (109).
- 20
- 25
13. Apparatus according to claim 12, the phase compensator comprising a control input for receiving the control signal from the detector (109).
- 30
14. Apparatus according to claims 11 to 13, the phase compensator being configured for adding the correction phase drift to a phase of the transformed signal for compensating a negative phase drift, or for subtracting the correction phase drift from the phase of the
- 35

transformed signal for compensating a positive phase drift.

15. Apparatus according to claims 1 to 14, the detector
5 (109) being configured for determining an average change in phase between two subsequent values of the transformed signal, the average change in phase representing the phase drift.
- 10 16. Apparatus according to claims 1 to 15, the detector being configured for determining an average value of a product between two subsequent values of the transformed signal, and for determining a phase of the average value, the phase of the average value representing
15 the phase drift.
17. Apparatus according to claims 1 to 16, the detector
(109) being configured for determining an average phase change between two values of the transformed
20 signal, the two values being spaced apart by a number of values, and for dividing the average phase change by the number of values in order to detect the phase drift for each value of the transformed signal.
- 25 18. Apparatus according to claims 1 to 17, the transformer comprising a cyclic shift element for pre-processing the time-domain signal, a Fourier transformer coupled to the cyclic shift element for time-frequency transforming the time-domain signal being cyclically
30 shifted by the cyclic shift element, a phase compensator coupled to an output of the Fourier transformer for post-processing a transformed signal provided by the Fourier transformer, wherein an output of the phase compensator is coupled to the detector (109),
35 the cyclic shift element being configured for receiving a control signal from the detector (109), the control signal indicating a number of values the time-domain signal is to be shifted by in order to intro-

duce a correction phase drift to the transformed signal for coarsely compensating the phase drift, and for cyclically shifting the time-domain signal by the number of values, the phase compensator being configured for receiving a further control signal indicating a further correction phase drift to be introduced to a phase of the transformed signal for fine compensation of the phase drift, and for introducing the further correction phase drift to the phase of the transformed signal.

19. Multi-carrier receiver, comprising:

the apparatus according to claims 1 to 16 for reducing a phase drift in a receivable multi-carrier signal in time-domain, the apparatus being configured for providing a transformed signal representing a receivable multi-carrier signal in frequency domain, the transformed signal having reduced phase drift; and

means for extracting information comprised by the transformed signal.

20. Multi-carrier receiver according to claim 19, the multi-carrier receiver further comprising a synchronizer for performing a frame synchronization, wherein the phase drift results from a frame synchronization error due to an erroneous sample time instant, wherein the detector (109) is configured for generating a frame synchronization control signal indicating the phase drift, and wherein the synchronizer is configured, in response to the frame synchronization control signal, for adjusting a sample time instant in order to reduce the frame synchronization error.

21. Multi-carrier receiver according to claim 19 or 20, the means for extracting information comprising a differential de-modulator for differentially de-

modulating the transformed signal in order to extract information from a phase change between two subsequent values of the transformed signal.

- 5 22. Multi-carrier receiver according to claims 19 to 20, wherein the means for extracting information comprises a low-pass filter for filtering the transformed signal in frequency domain, the low-pass filter comprising
10 real valued coefficients representing a real part of a filter response, wherein filter coefficients representing the imaginary part of the filter response are set to zero, wherein the transformer is further configured for pre-processing the receivable multi-carrier signal in time domain or for post-processing
15 the transformed signal in frequency domain in order to introduce a phase shift to the transformed signal for shifting a spectrum of the transformed signal towards a passband of the low-pass filter.
- 20 23. Multi-carrier receiver according to claim 22, wherein the receivable multi-carrier signal in time domain is a receivable version of a transmit signal being transmitted through a communication channel, wherein the low-pass filter is configured for channel estimation.
- 25 24. Multi-carrier receiver according to claim 23, wherein the low-pass filter is a Wiener filter for channel estimation.
- 30 25. Multi-carrier receiver according to claim 23 or 24, wherein the low-pass filter is an interpolation filter.
- 35 26. Multi-carrier receiver according to claims 19 to 25, wherein the means for extracting information comprises a space-frequency block decoder.

27. Method for reducing a phase drift in a spectrum of a time-domain signal, the method comprising:

5 transforming the time-domain signal into a transformed signal in frequency domain, the transformed signal representing a spectrum;

detecting the phase drift;

10 generating a control signal indicating the phase drift;

15 in response to the control signal, pre-processing the time-domain signal before time-frequency transforming or post-processing the transformed signal after time-frequency transforming the time-domain signal in order to introduce a correction phase drift to the transformed signal, the correction phase drift at least partly compensating the phase drift for phase shift reduction.

20

28. Method for receiving multi-carrier signals, comprising:

25 the method according to claim 27 for reducing a phase drift in a receivable multi-carrier signal in time domain and for providing a transformed signal representing a receivable multi-carrier signal in frequency domain, the transformed signal having reduced phase drift; and

30

extracting information comprised by the transformed signal.

35 29. Computer program having a program code for performing the method according to claim 27 or the method according to claim 28, when the program runs on a computer.

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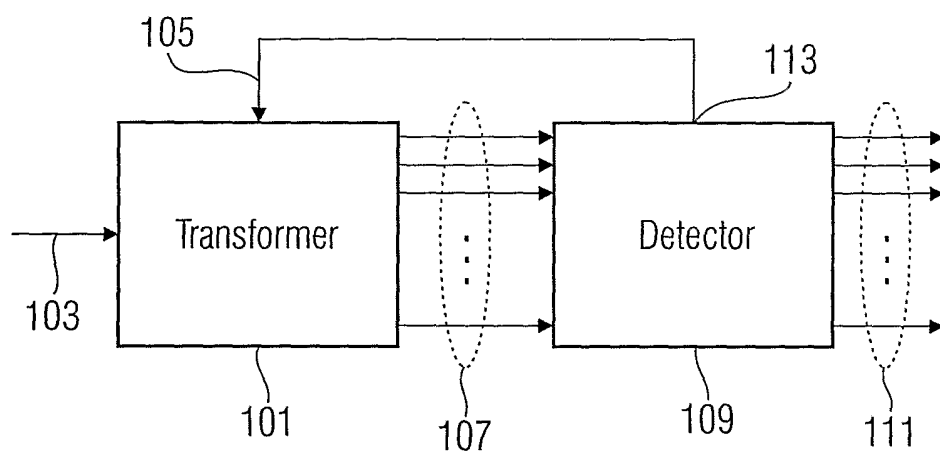


FIG. 1

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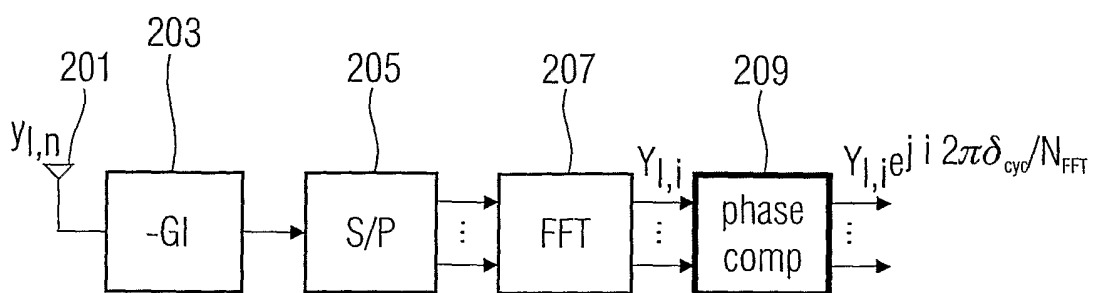


FIG. 2a

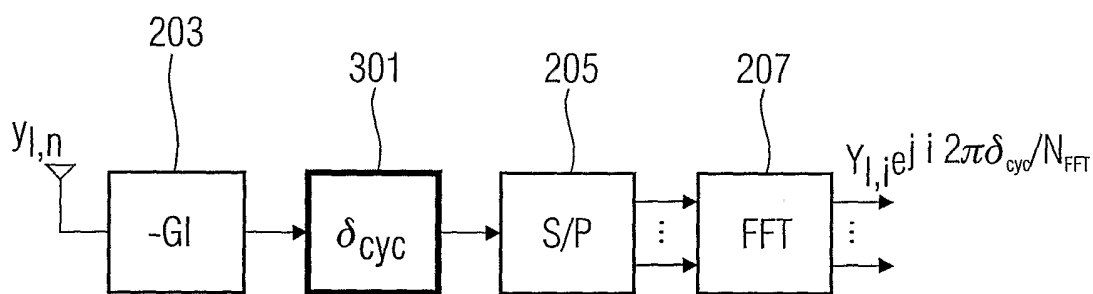


FIG. 2b

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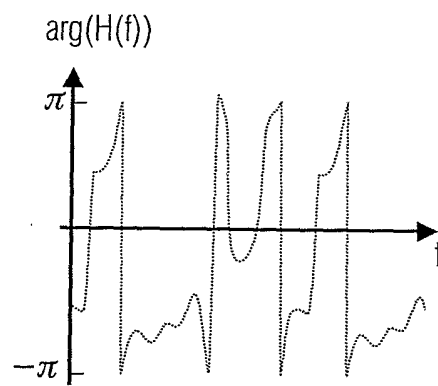


FIG. 3a

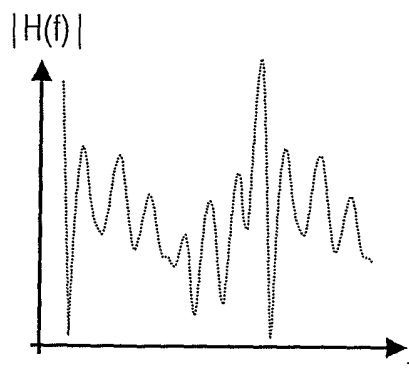


FIG. 3b

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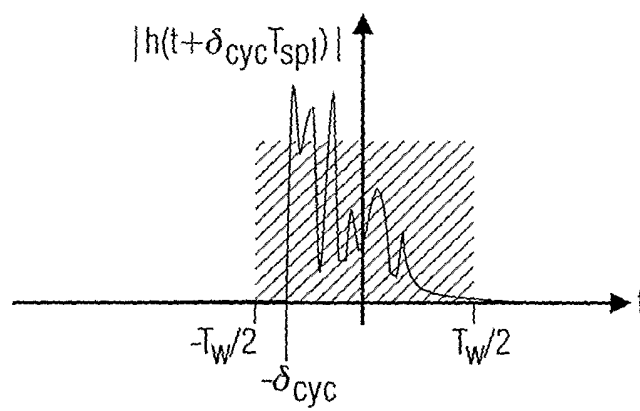


FIG. 4

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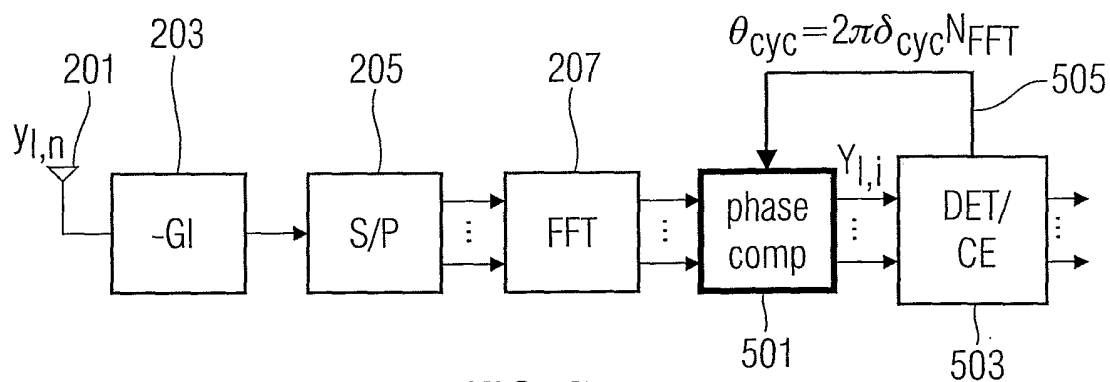


FIG. 5

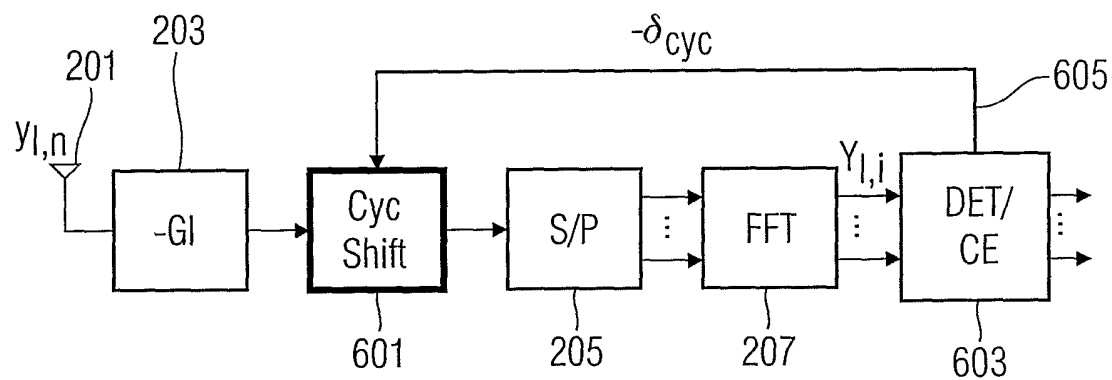


FIG. 6

Modulation	QPSK	
# Tx antennas	N_T	4
Bandwith	B	101.5 Mhz
# subcarriers	N_C	769
FFT length	N_{FFT}	1024
Guard interval (GI) length	N_{GI}	226
Sample duration	T_{spl}	7.4 ns
Pilot spacing freq.	D_f	2
Pilot spacing time	D_t	8

Parameters for mismatched channel estimator

Max. Channel delay	τ_{max_ω}	$N_{GI}T_{spl}$
Max. Doppler freq.	ν_{max_ω}	$0.013 \cdot T_{sym}$
Max. E_b/N_0	γ_ω	30 dB

FIG. 7

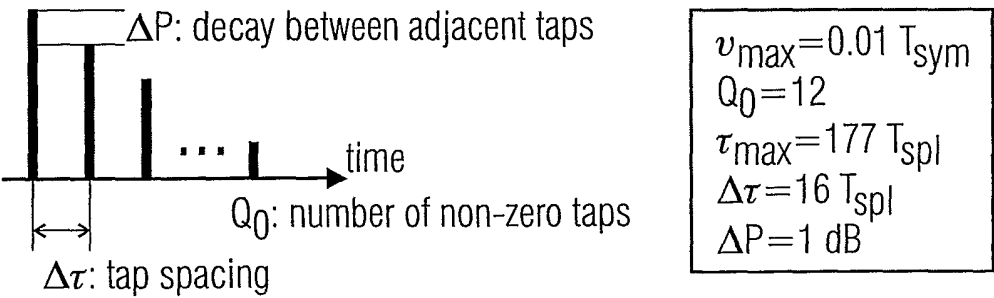


FIG. 8

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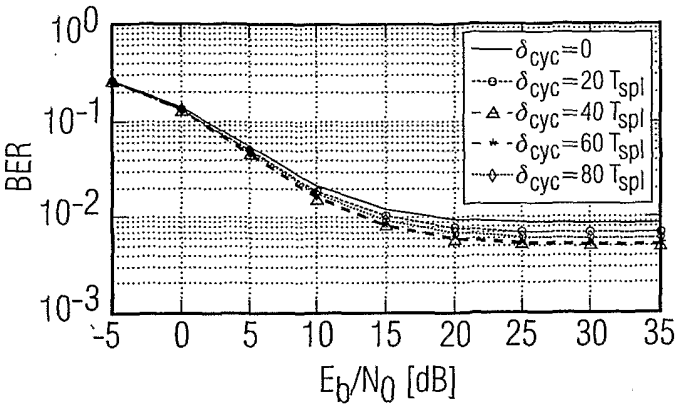


FIG. 9

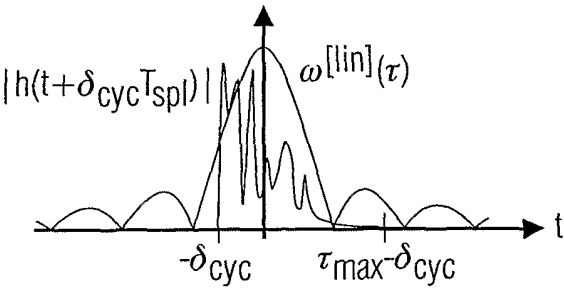


FIG. 10

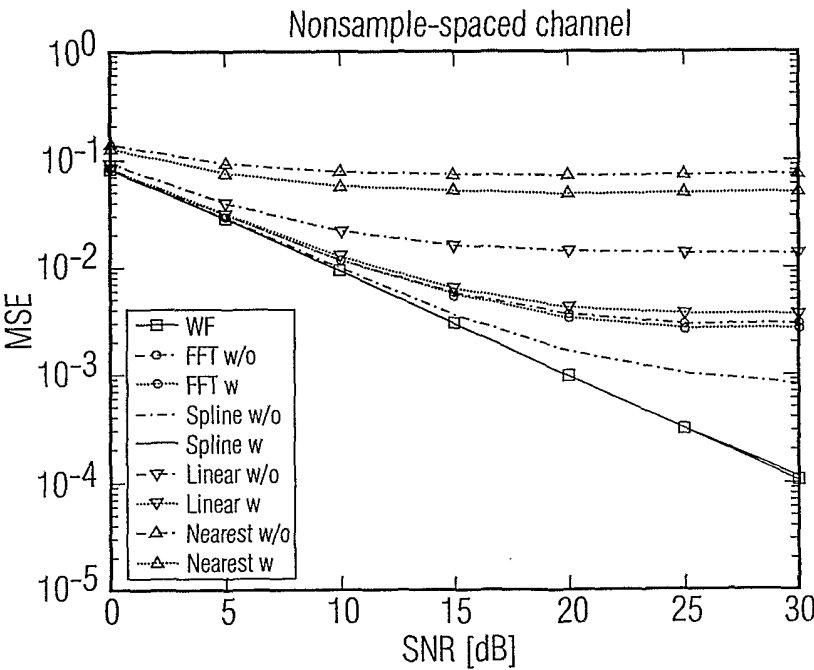


FIG. 11

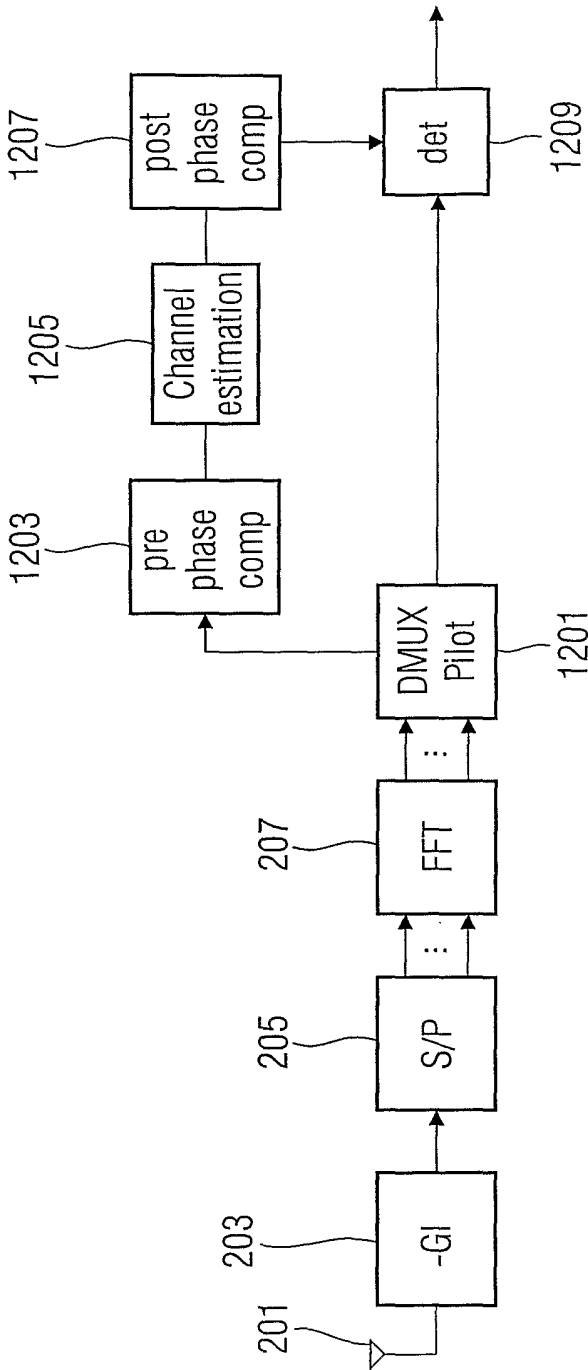


FIG. 12

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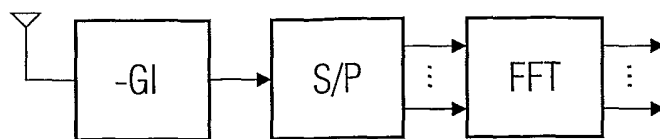


FIG. 13

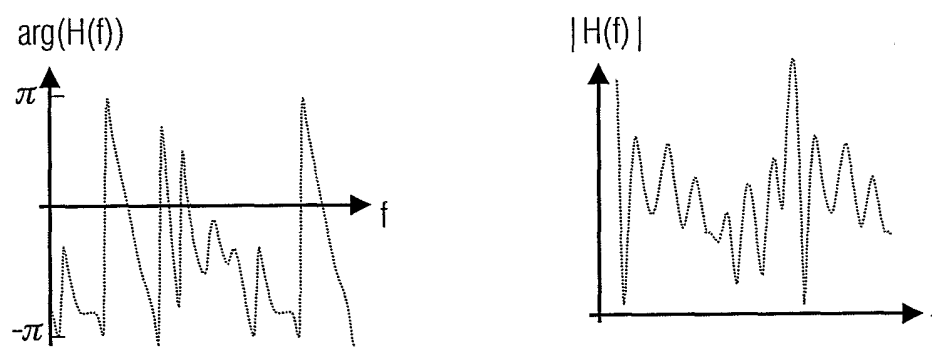


FIG. 14

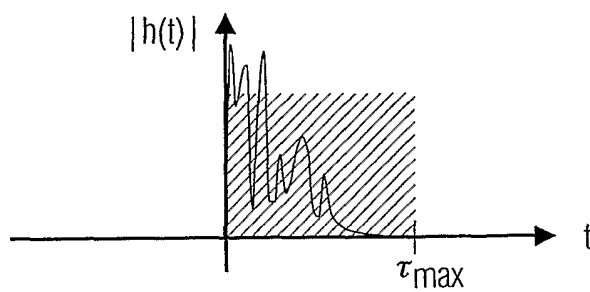


FIG. 15

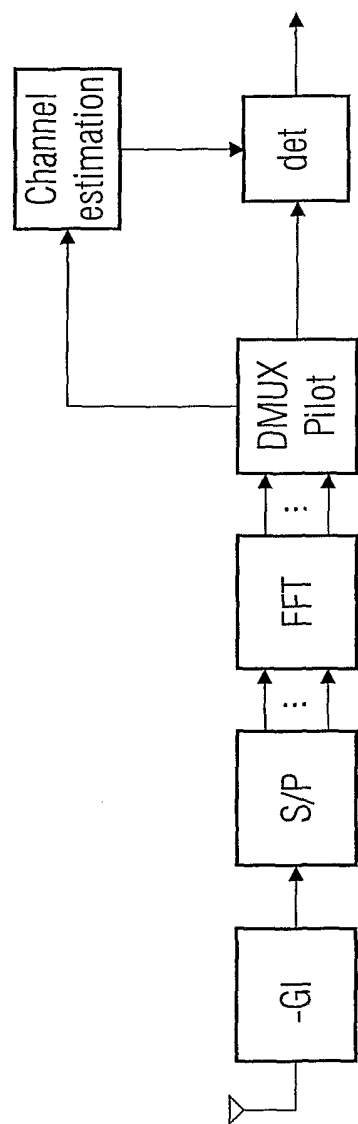


FIG. 16

INTERNATIONAL SEARCH REPORT

International Application No

PCT/EP2004/009373

A. CLASSIFICATION OF SUBJECT MATTER

IPC 7 H04L27/26

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

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC 7 H04L

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

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

EPO-Internal, COMPENDEX, 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 2004/008802 A1 (GALPERIN NOAM ET AL) 15 January 2004 (2004-01-15) page 1, right-hand column, paragraph 8 - page 2, right-hand column, paragraph 19 page 3, right-hand column, paragraph 59 - page 4, left-hand column, paragraph 62 page 4, left-hand column, paragraph 65 page 5, left-hand column, paragraph 71 - paragraph 73 claims 65,69-72; figure 2 ----- -/--	1,2,19, 27-29



Further documents are listed in the continuation of box C.



Patent family members are listed in annex.

° Special categories of cited documents:

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E earlier document but published on or after the international filing date

L document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

O document referring to an oral disclosure, use, exhibition or other means

P document published prior to the international filing date but later than the priority date claimed

T later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

X document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

Y document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

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Date of the actual completion of the international search

22 April 2005

Date of mailing of the international search report

29/04/2005

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Schiffer, A

INTERNATIONAL SEARCH REPORT

International Application No

PCT/EP2004/009373

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	SPETH M ET AL: "Optimum receiver design for OFDM-based broadband transmission.II. A case study" IEEE TRANSACTIONS ON COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 49, no. 4, April 2001 (2001-04), pages 571-578, XP002255287 ISSN: 0090-6778 page 572, right-hand column, paragraph III.C - page 577, left-hand column figure 2	1,2,19, 27-29
X	US 2003/053564 A1 (KIM YOUNGGYUN ET AL) 20 March 2003 (2003-03-20) page 1, left-hand column, paragraph 7 - right-hand column, paragraph 9 page 3, left-hand column, paragraph 34 - right-hand column, paragraph 41 page 4, left-hand column, paragraph 45 - paragraph 48 page 6, left-hand column, paragraph 77 - right-hand column, paragraph 82 claims 1,6-8; figure 7	1,2,19, 27-29
X	EP 0 795 985 A (NOKIA TECHNOLOGY GMBH) 17 September 1997 (1997-09-17) page 2, line 29 - line 34 page 3, line 3 - line 7 page 3, line 35 - line 53 claims 1-4	1,2,19, 27-29
X	US 2004/120412 A1 (BANERJEA RAJA) 24 June 2004 (2004-06-24) page 2, right-hand column, paragraph 18 page 3, left-hand column, paragraph 25 page 3, right-hand column, paragraph 34 - page 4, right-hand column, paragraph 45 page 5, left-hand column, paragraph 51 - right-hand column, paragraph 58 claims 1,3	1,2,19, 27-29
A	US 6 618 452 B1 (HUBER JOHANNES ET AL) 9 September 2003 (2003-09-09) column 10, line 5 - line 38 column 15, line 53 - column 16, line 41 column 40, line 8 - line 27 column 43, line 57 - column 44, line 30	1-29

INTERNATIONAL SEARCH REPORT

International Application No

PCT/EP2004/009373

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