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None

Description**TECHNICAL FIELD OF THE INVENTION**

5 **[0001]** The present invention relates, in general, to a radio communications system and method, namely a system and a method for implementing communications at Radio Frequency (RF) (including frequencies from a few KHz to hundreds of GHz) with increased transmission capacity.

[0002] In particular, the present invention concerns a radio communications system and method exploiting twisted signals in frequency domain for increasing transmission capacity.

10 **[0003]** The present invention can be advantageously exploited, in general, in all kinds of radio communications, and, in particular, in radio communications based:

- in general, on Orthogonal Frequency-Division Multiplexing (OFDM) and/or Orthogonal Frequency-Division Multiple Access (OFDMA); and,
- 15 • specifically, on Long Term Evolution (LTE) standard and/or Worldwide Interoperability for Microwave Access (WiMAX) standard.

BACKGROUND ART

20 **[0004]** In consideration of Orbital Angular Momentum (OAM) potentialities of increasing transmission capacity and since RF spectrum shortage problem is deeply felt in radio communications sector, recently a lot of experimental studies have been carried out on the use of OAM states, or modes, at RF (also known as radio vortices) in order to try to enhance RF spectrum reuse.

[0005] In this connection, reference may, for example, be made to:

- 25 • Mohammadi S. M. et al., "Orbital Angular Momentum in Radio - A System Study", IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION, IEEE SERVICE CENTER, PISCATAWAY, NJ, US, vol. 58, no. 2, 1 February 2010, pages 565-572, which shows that standard antennas arranged in circular arrays can be used to generate RF beams carrying OAM;
- 30 • Tamburini F. et al., "Encoding many channels in the same frequency through radio Vorticity: first experimental test", arXiv.org, 12 July 2011, Ithaca, NY, USA, which experimentally shows that it is possible to propagate and use the properties of twisted non-monochromatic incoherent radio waves to simultaneously transmit several radio channels on one and the same frequency by encoding them in different (and, thence, orthogonal) OAM states (even without using polarization or dense coding techniques);
- 35 • GB 2 410 130 A, which discloses a planar phased array antenna for transmitting and receiving OAM radio vortex modes, which antenna comprises a circular array of cavity backed axial mode spiral antenna elements whose phase is controlled such that the phase of each antenna element changes sequentially about the array; and
- WO 2012/084039 A1, which discloses a transmit antenna arrangement comprising N antenna elements arranged along a circumference with an angular separation of α degrees between neighboring antenna elements, the antenna arrangement comprising an OAM encoder arranged to receive N input signals for transmission, indexed from $M=-(N-1)/2$ up to $M=(N-1)/2$ for odd N and from $M=-(N-2)/2$ up to $N/2$ for even N; the OAM encoder connecting each input signal to each antenna element and giving each input signal M at each antenna element a phase shift of $M\alpha$ relative to the phase of the same input signal M at an adjacent antenna element; wherein two or more antenna elements are directional, have their directivity in the same direction, and have an antenna aperture higher than, or equal to, 5λ , where λ is the wavelength of the N input signals.

[0006] From a mathematical perspective, the transmission of an OAM mode (or state) at a single RF (i.e., by using a pure tone) implies that the electrical field on the radiating aperture can be represented as:

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$$F(\rho, \phi) = F(\rho)e^{jk\phi},$$

where ρ and ϕ are the cylindrical coordinates on the radiating aperture, j is the imaginary unit, and k is a positive or negative integer.

55 **[0007]** The radiated field can be represented in the far zone as:

$$E(\vartheta, \varphi) = \frac{1}{R} \iint_S F(\rho, \phi) e^{-j2\pi \frac{D}{\lambda} \sin(\vartheta) \cos(\varphi - \phi)} \rho d\rho d\phi ,$$

where ϑ and φ are the spherical coordinates in the far field, R denotes the radius of the sphere centered on the radiating aperture, S denotes the integration surface used at reception side, and λ denotes the wavelength used.

[0008] As is known, due to intrinsic characteristics of OAM, an OAM mode transmitted at a single RF (i.e., by using a pure tone) is affected by a phase singularity which creates a null at the bore-sight direction, thereby resulting that

$$E(0,0) = 0 .$$

[0009] In order for said phase singularity to be compensated, the integration surface S used at reception side should be sized so as to include the crown peak generated by the OAM mode.

[0010] In particular, the integration surface S used at reception side should be different for each OAM mode and, considering the sampling theorem applied to the radiating antenna, should have an area given by:

$$\Delta S = \Delta \Omega R^2 = 2 \left(\frac{\lambda}{D} R \right)^2 ,$$

where D denotes the diameter of the radiating antenna.

[0011] Therefore, the price to be paid with pure OAM modes transmitted by using pure tones (i.e., single radiofrequencies) is that the dimensions of the equivalent receiving antenna depend on the distance R from, and on the diameter D of, the transmitting antenna.

[0012] This solution is impractical for satellite communications, where the aperture efficiency and the size of the antennas are very critical issues. For example, in geostationary-satellite-based communications in Ka band, for a ground antenna having a diameter D of about 9 m, the diameter of the receiving ring on board the geostationary satellite should be of the order of 50 Km, thereby resulting impractical.

[0013] Thence, in view of the foregoing, the main criticality in using radio vorticity in practical systems is that the orthogonality between OAM modes depends on the size of antennas, on the distance between the transmitting and receiving antennas, and on the need for the receiving antenna to operate as an interferometer basis (as, for example, disclosed in the aforesaid papers "*Orbital Angular Momentum in Radio - A System Study*" and "*Encoding many channels in the same frequency through radio Vorticity: first experimental test*", in GB 2 410 130 A and in WO 2012/084039 A1). These constraints result in OAM-based radio communication systems which are inefficient and unusable for very long distances such as the ones involved in satellite communications.

[0014] Moreover, further criticalities in the use of radio vorticity for satellite communications are represented by the need of an extremely accurate mutual pointing of the transmitting and receiving antennas, and by the unfeasibility of the geometry for Earth-satellite configurations due to the criticality of the positioning of the receiving antennas (or the receiving antenna elements).

[0015] A solution to the aforesaid technical problems is provided in the International Application No. PCT/IB2012/056804 filed on 28th November 2012 in the name of EUTELSAT S.A. and concerning a multidimensional space modulation technique for transmitting and/or receiving radio vortices at frequencies ranging from a few KHz to hundreds of GHz. Specifically, the multidimensional space modulation technique according to the International Application PCT/IB2012/056804 allows to transmit and/or receive orthogonal RF OAM modes in one and the same direction (i.e., the bore-sight direction) and to overcome, at the same time, the aforesaid technical problems caused by OAM phase singularity at the bore-sight direction, thereby allowing the use of radio vortices also for long-distance radio communications, such as satellite communications.

[0016] In particular, the multidimensional space modulation according to the International Application PCT/IB2012/056804 is actually a phase modulation applied to signals to be transmitted at RF such that to result in orthogonal radio vortices along the bore-sight direction. Therefore, the modulation according to the International Application PCT/IB2012/056804 is called multidimensional space modulation because it allows orthogonal RF OAM modes to be transmitted and/or received in one and the same direction, namely the bore-sight direction, wherein each OAM mode represents a specific space channel along the bore-sight direction, which specific space channel is orthogonal to all the other space channels represented by the other OAM modes.

[0017] In order for the multidimensional space modulation according to the International Application

PCT/IB2012/056804 to be understood, attention is drawn, by way of example, to the fact that, as is known, a twisted RF signal having, or carrying, the OAM mode $m=+1$ is characterized by only one clockwise rotation of 360° of the Poynting vector around the propagation axis per period T and, thence, it can be generated by transmitting, for example by means of four ring-arranged transmitting antenna elements, RF signals associated with phases of 0° , 90° , 180° , and 270° clockwise distributed among said four ring-arranged transmitting antenna elements. Instead, the International Application PCT/IB2012/056804 proves that it is possible and convenient, in order to transmit at RF the OAM mode $m=+1$ and, at the same time, to solve the problem caused by OAM phase singularity at the bore-sight direction, to exploit only one antenna transmitting the four different phases 0° , 90° , 180° , and 270° at different times (or at different frequencies) with a time step of $T'=T/4$. This possibility increases the efficiency of the transmitting and receiving configuration, which can work regardless of the elementary antenna element spacing in an antenna array.

[0018] From a conceptual perspective, according to the International Application PCT/IB2012/056804, in order to manage OAM rotation, namely in order to control the speed of rotation of an RF OAM mode about the bore-sight direction, a supplementary phase modulation is introduced, which leaves only a residue of the OAM twist and keeps the OAM signature in a limited bandwidth. This residual rotation achieved by means of the supplementary phase modulation allows a signal having a proper bandwidth to be orthogonal to another signal having a different rotation (multiple of the minimum one). Therefore, an RF twisted wave can be transmitted by means of a modulated waveform and can be received by an antenna operating in the complex conjugate mode. The received signal is equal to the transmitted one, apart from standard attenuation and transmission and reception gains in a time period T_{mod} . The bandwidth increase does not prevent the transmission of plane waves (i.e., the OAM mode $m=0$), but limits the number of OAM modes at different central frequencies in the available bandwidth. The multidimensional space modulation according to PCT/IB2012/056804 allows to use a standard antenna in place of a phased array antenna, since the used signals are native orthogonal.

[0019] It is important to underline the fact that the generation of RF OAM modes by means of the multidimensional space modulation according to PCT/IB2012/056804 allows to drastically simplify the antenna design. In fact, the antenna does not need to take memory at the period of the carrier frequency of the phase between elements $f_0=1/T_0$. This duty is performed by the sampling frequency of the twisted waves, which is at least 3 times the signal bandwidth; therefore the phase shift assigned to the sampling is already orthogonal in time; it follows that the antenna can be a standard one without the need of using a phased array configuration on either the antenna aperture, or, in case of a reflector antenna, the focal plane. Therefore, the multidimensional space modulation according to PCT/IB2012/056804 can be exploited in satellite communications by using already existing satellite and ground antennas.

[0020] In order for the multidimensional space modulation according to PCT/IB2012/056804 to be better understood, reference is made to Figure 1, which shows a functional block diagram of a transmitting system (denoted as whole by 1), which is disclosed in PCT/IB2012/056804 and which exploits the aforesaid multidimensional space modulation for transmitting radio vortices at frequencies ranging from a few KHz to hundreds of GHz.

[0021] In particular, the transmitting system 1 comprises:

- a signal generation section 10 designed to generate
 - a first digital signal $s_0(t)$ carrying an information stream, having a given sampling period T_0 and occupying a given frequency bandwidth W centered on a predefined frequency f_0 , and
 - up to $2N$ second digital signals $s_m(t)$, with $-N \leq m \leq +N$ and $N \geq 1$ (for the sake of illustration simplicity in Figure 1 only signals $s_{+1}(t)$, $s_{-1}(t)$, $s_{+N}(t)$ and $s_{-N}(t)$ are shown), each carrying a respective information stream, having a respective sampling period $T_m=4|m|T_0$ (or $T_m=3|m|T_0$) and occupying a respective frequency bandwidth $W/4|m|$ (or $W/3|m|$) centered on said predefined frequency f_0 (which can, conveniently, be an Intermediate Frequency (IF) thereby resulting that the first and second digital signals are IF digital signals);
- a device 100 for generating OAM modes, which is coupled with said signal generation section 10 to receive the first and second digital signals generated by the latter, and which is designed to
 - apply, to each second digital signal $s_m(t)$ received from the signal generation section 10, a respective space modulation associated with a respective OAM mode m so as to generate a corresponding modulated digital signal carrying said respective OAM mode m , having the given sampling period T_0 , and occupying the given frequency bandwidth W , and
 - provide an output digital signal $s_{\text{out}}(t)$ based on the modulated digital signals and on the first digital signal $s_0(t)$ received from the signal generation section 10; and
- an RF transmission section 1000, which is coupled with the device 100 to receive therefrom the output digital signal $s_{\text{out}}(t)$, and which is designed to transmit at predefined radio frequencies the output digital signal $s_{\text{out}}(t)$ by means

of a single antenna (which is not shown in Figure 1 for the sake of illustration simplicity and which can be also a reflector antenna with a single feed) or an antenna array (which is not shown in Figure 1 for the sake of illustration simplicity and which can be also a multi-feed reflector antenna), thereby transmitting an overall RF signal carrying

- said first digital signal $s_0(t)$ by means of a plane wave, and
- said second digital signals $s_m(t)$, each by means of a corresponding radio vortex having the respective OAM mode m .

[0022] The aforesaid predefined radio frequencies can conveniently range from a few KHz to hundreds of GHz depending on the specific application for which the overall transmitting system 1 is designed.

[0023] Conveniently, the signal generation section 10 can be a signal generation section of a transmitting system for satellite communications (such as a transmitting system of a feeder link Earth station, of a satellite, or of a ground apparatus for satellite communications), or of a device for wireless communications, such as LTE-based communications.

[0024] Accordingly, the RF transmission section 1000 can conveniently be an RF transmission section of a transmitting system for satellite communications (such as a transmitting system of a feeder link Earth station, of a satellite, or of a ground apparatus for satellite communications), or of a device for wireless communications, such as LTE-based communications.

[0025] Additionally, Figure 2 shows in greater detail the device 100 for generating OAM modes, which device 100 comprises $2N$ OAM mode generation modules. In particular, Figure 2 shows, for the sake of illustration simplicity, only:

- an OAM mode generation module 110 for generating OAM mode $m=+1$;
- an OAM mode generation module 120 for generating OAM mode $m=-1$;
- an OAM mode generation module 130 for generating OAM mode $m=+N$; and
- an OAM mode generation module 140 for generating OAM mode $m=-N$.

[0026] In detail, a generic OAM mode generation module for generating OAM mode m is operable to apply to a respective second digital signal $s_m(t)$ received from the signal generation section 10 a respective space modulation associated with said OAM mode m so as to generate a corresponding space-modulated digital signal $sms_m(t)$ carrying said OAM mode m , having the given sampling period T_0 , and occupying the whole given frequency bandwidth W centered on said predefined frequency f_0 .

[0027] More in detail, the generic OAM mode generation module for generating the OAM mode m is operable to:

- receive a synchronization signal $synch_m$ (not shown in Figure 2 for the sake of illustration clarity) indicating the given sampling period T_0 and, conveniently, also the sampling period T_m of the respective second digital signal $s_m(t)$ received from the signal generation section 10; and
- apply the respective space modulation to said respective digital signal $s_m(t)$ by
 - digitally interpolating said respective second digital signal $s_m(t)$ on the basis of the received synchronization signal $synch_m$ so as to generate a corresponding digitally-interpolated signal having the given sampling period T_0 ;
 - applying to the digitally-interpolated signal a respective digital phase modulation associated with said OAM mode m such that to generate a corresponding phase-modulated signal carrying said OAM mode m with a predefined OAM mode rotation speed; and
 - digitally filtering the phase-modulated signal thereby obtaining a filtered signal which represents the aforesaid space-modulated digital signal $sms_m(t)$.

[0028] For example, the OAM mode generation module 110 is conveniently configured to:

- receive, from the signal generation section 10, the second digital signal $s_{+1}(t)$ and a synchronization signal $synch_{+1}$ indicating the given sampling period T_0 and, conveniently, also the sampling period $T_{+1}=4T_0$ (or $T_{+1}=3T_0$) of the second digital signal $s_{+1}(t)$;
- digitally interpolate the second digital signal $s_{+1}(t)$ by outputting, for each digital sample of said second digital signal $s_{+1}(t)$, four corresponding digital samples with time step (i.e., time distance) T_0 , thereby generating a corresponding digitally-interpolated signal having the given sampling period T_0 ;
- apply to each set of four digital samples obtained by means of the digital interpolation digital phase shifts related to the OAM mode $+1$ with the predefined OAM mode rotation speed (namely, digital phase shifts related to phase values 0 , $\pi/2$, π and $3\pi/2$) so as to generate a corresponding set of four phase-shifted digital samples, which corresponding set of four phase-shifted digital samples carries said OAM mode $+1$ with the predefined OAM mode rotation speed;

- digitally filter each set of four phase-shifted digital samples obtained by means of the digital phase shifting so as to output a corresponding set of four filtered digital samples; and
- combine the sets of four filtered digital samples obtained by means of the digital filtering into a single filtered signal which represents the space-modulated digital signal $sms_{+1}(t)$.

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[0029] Accordingly, the OAM mode generation module 120 is conveniently configured to:

- receive, from the signal generation section 10, the second digital signal $s_{-1}(t)$ and a synchronization signal $synch_{-1}$ indicating the given sampling period T_0 and, conveniently, also the sampling period $T_{-1}=4T_0$ (or $T_{-1}=3T_0$) of the second digital signal $s_{-1}(t)$;
- digitally interpolate the second digital signal $s_{-1}(t)$ by outputting, for each digital sample of said second digital signal $s_{-1}(t)$, four corresponding digital samples with time step (i.e., time distance) T_0 , thereby generating a corresponding digitally-interpolated signal having the given sampling period T_0 ;
- apply to each set of four digital samples obtained by means of the digital interpolation digital phase shifts related to the OAM mode -1 with the predefined OAM mode rotation speed (namely, digital phase shifts related to phase values $0, 3\pi/2, \pi$ and $\pi/2$) so as to generate a corresponding set of four phase-shifted digital samples, which corresponding set of four phase-shifted digital samples carries said OAM mode -1 with the predefined OAM mode rotation speed;
- digitally filter each set of four phase-shifted digital samples obtained by means of the digital phase shifting so as to output a corresponding set of four filtered digital samples; and
- combine the sets of four filtered digital samples obtained by means of the digital filtering into a single filtered signal which represents the space-modulated digital signal $sms_{-1}(t)$.

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[0030] The OAM mode generation modules for generating higher-order OAM modes (i.e., with $|m| > 1$) operate, mutatis mutandis, conceptually in the same way as the OAM mode generation modules 110 and 120.

[0031] Moreover, again with reference to Figure 2, the device 100 further comprises:

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- a combining module 150 operable to combine the first digital signal $s_0(t)$ received from the signal generation section 10 and all the space-modulated digital signals $sms_m(t)$ generated by the OAM mode generation modules into a corresponding combined digital signal $s_c(t)$; and
- a transmission filtering module 160, which is operable to digitally filter the combined digital signal $s_c(t)$ by means of a predefined transmission filter such that to adjust the signal bandwidth to the bandwidth of transmission radio channel (i.e., the specific radio channel used in transmission) so as to reduce Inter-Symbol Interference (ISI), thereby obtaining a corresponding output digital signal $s_{out}(t)$; wherein the transmission filtering module 160 is coupled with the RF transmission section 1000 to provide the latter with the output digital signal $s_{out}(t)$.

[0032] For example, in case of (free-space) satellite communications on a radio channel having the given frequency bandwidth W , the transmission filter can be a predefined root raised cosine filter adapted to said given frequency bandwidth W .

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[0033] As far as reception is concerned, reference is made to Figure 3, which shows a functional block diagram of a receiving system (denoted as whole by 2), which is disclosed in PCT/IB2012/056804 and which exploits the aforesaid multidimensional space modulation for receiving radio vortices at frequencies ranging from a few KHz to hundreds of GHz.

[0034] In particular, the receiving system 2 comprises:

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- an RF reception section 2000, which is designed to receive signals at predefined radio frequencies by means of a single antenna (which is not shown in Figure 3 for the sake of illustration simplicity and which can be also a reflector antenna with a single feed) or an antenna array (which is not shown in Figure 3 for the sake of illustration simplicity and which can be also a multi-feed reflector antenna), and which is designed to obtain an incoming digital signal $u_{in}(t)$ on the basis of the received signals;
- a device 200 for demodulating OAM modes, which is coupled with said RF reception section 2000 to receive the incoming digital signal $u_{in}(t)$ therefrom, and which is designed to process said incoming digital signal $u_{in}(t)$ so as to output useful signals (in Figure 3 useful signals $u_0(t), u_{+1}(t), u_{-1}(t), u_{+N}(t)$ and $u_{-N}(t)$ outputted by the device 200 are shown); and
- a signal processing section 20, which is coupled with said device 200 to receive the useful signals outputted by the latter and which is designed to process said useful signals.

[0035] The aforesaid predefined radio frequencies can conveniently range from a few KHz to hundreds of GHz depending on the specific application for which the overall receiving system 2 is designed.

[0036] Conveniently, the RF reception section 2000 can be an RF reception section of a receiving system for satellite communications (such as a receiving system of a feeder link Earth station, of a satellite, or of a ground apparatus for satellite communications), of a device for wireless communications (such as LTE-based communications), of a radar system, of a Synthetic Aperture Radar (SAR) system, or of a radio astronomy receiving system.

[0037] Accordingly, the signal processing section 20 can conveniently be a signal processing section of a receiving system for satellite communications (such as a receiving system of a feeder link Earth station, of a satellite, or of a ground apparatus for satellite communications), of a device for wireless communications (such as LTE-based communications), of a radar system, of a SAR system, or of a radio astronomy receiving system.

[0038] Additionally, Figure 4 shows in greater detail the device 200 for demodulating OAM modes. In particular, as shown in Figure 4, the device 200 comprises a reception filtering module 210, which is operable to digitally filter the incoming digital signal $u_{in}(t)$ by means of a predefined reception filter such that to equalize the incoming digital signal $u_{in}(t)$ with respect to reception radio channel (i.e., the specific radio channel used in reception) and, conveniently, also with respect to transmission filter (i.e., the specific filter used in transmission), thereby obtaining a corresponding filtered incoming digital signal $u_f(t)$.

[0039] For example, in case of (free-space) satellite communications on a radio channel having the given frequency bandwidth W , wherein the transmission filter is a predefined root raised cosine filter adapted to said given frequency bandwidth W , the reception filter can be the complex conjugate of said predefined root raised cosine filter so as to reduce ISI.

[0040] Additionally, again with reference to Figure 4, the device 200 further comprises a digital oversampling module 220 operable to digitally oversample the filtered incoming digital signal $u_f(t)$ on the basis of a predefined oversampling period T_{over} , thereby outputting a corresponding set of digital samples.

[0041] For example, in case the receiving system 2 is configured to receive the RF signals transmitted by the transmission system 1, the predefined oversampling period T_{over} can conveniently be equal to T_0/Q , wherein T_0 is the given sampling period previously introduced in connection with the transmission system 1, and Q denotes an integer higher than one.

[0042] Furthermore, again with reference to Figure 4, the device 200 comprises also a processing module 230 configured to:

- provide a linear system of M equations (where M denotes an integer higher than one) relating
 - the set of digital samples outputted by the digital oversampling module 220
 - to X unknown digital values (where X denotes an integer higher than one and lower than M) of useful signals associated, each, with a respective predefined OAM mode m with a predefined OAM mode rotation speed;
 - wherein said linear system of M equations relates the set of digital samples outputted by the digital oversampling module 220 to the X unknown digital values through
 - first predefined parameters related to the predefined OAM modes with the predefined OAM mode rotation speed, and
 - second predefined parameters related to the predefined reception filter, to the reception radio channel and, conveniently, also to the transmission filter;
- compute the X digital values by solving the linear system of M equations; and
- digitally generate and output the useful signals (for example the useful signals $u_0(t)$, $u_{+1}(t)$, $u_{-1}(t)$, $u_{+N}(t)$ and $u_{-N}(t)$ shown in Figure 4) on the basis of the corresponding digital values computed.

[0043] In this connection, it is important to underline the fact that, in order to extract the useful signals (i.e., in order to solve the linear system of M equations thereby computing the X digital values, and, thence, to generate and output the useful signals), the processing module 230 is conveniently configured to operate as a generalized matched filter which exploits one or more mathematical processing techniques, such as the pseudo-inverse technique.

[0044] Moreover, it is also important to underline the fact that the oversampling operation performed by the digital oversampling module 220 allows to increase redundancy of the linear system of M equations (i.e., it allows to obtain a number M of independent equations higher and higher than the number X of the unknown digital values), thereby allowing to find more robust solutions to said linear system of M equations.

[0045] Furthermore, the better the characterization of the OAM modes and of the radio channel in the linear system of M equations, the more robust the resolution of said linear system of M equations. Specifically, an increase of the number of first and second predefined parameters used in the linear system of M equations allows to increase redundancy of said linear system of M equations (i.e., it allows to obtain a number M of independent equations higher and higher than the number X of the unknown digital values), thereby allowing to optimize the resolution of, i.e., to find optimum

solutions to, said linear system of M equations in terms of energy per bit to noise power spectral density ratio E_b/N_0 .

[0046] In case the receiving system 2 is configured to receive the RF signals transmitted by the transmission system 1, the first predefined parameters are related to the sampling periods T_0 and T_m previously introduced in connection with the device 100, and to the digital phase shifts applied by the OAM mode generation modules of the device 100 to the digital samples of the digitally-interpolated signals.

[0047] Moreover, again in case the receiving system 2 is configured to receive the RF signals transmitted by the transmission system 1, the useful signals generated and outputted by the processing module 230 (such as the signals $u_0(t)$, $u_{+1}(t)$, $u_{-1}(t)$, $u_{+N}(t)$ and $u_{-N}(t)$ shown in Figure 4) are the digital signals transmitted by said transmission system 1 by means of the plane wave and the several radio vortices (namely the signals $s_0(t)$, $s_{+1}(t)$, $s_{-1}(t)$, $s_{+N}(t)$ and $s_{-N}(t)$ shown in Figures 1 and 2).

[0048] Preferably, the device 100 for generating OAM modes and the device 200 for demodulating OAM modes are implemented by means of Field-Programmable Gate Array (FPGA), Application-Specific Integrated Circuit (ASIC), and Software Defined Radio (SDR) technologies.

[0049] Finally, according to a further aspect of to the International Application PCT/IB2012/056804, an overall radio communication system including both the transmission system 1 and the receiving system 2 is preferably designed to:

- monitor interference experienced by the radio vortices transmitted; and,
- if the interference experienced by a radio vortex carrying a given digital signal $s_m(t)$ by means of a given OAM mode m meets a given interference-related condition (for example, if it exceeds a given interference level),
 - start using an OAM mode m^* different from the given OAM mode m for transmitting the information stream previously carried by said given digital signal $s_m(t)$ by means of said given OAM mode m , and
 - stop using said given OAM mode m .

[0050] In case said further aspect of PCT/IB2012/056804 is used for satellite communications, it is possible to mitigate jammer, since said further aspect of PCT/IB2012/056804 allows to reject a jammed OAM mode. Moreover, said further aspect of PCT/IB2012/056804 can be used also in combination with other anti-jamming capabilities of the receiving system.

[0051] Additionally, it is worth noting that US 2013/235885 A1 relates to channel bonding with Orbital Angular Momentum (OAM). In particular, US 2013/235885 A1 discloses a system comprising:

- a distributor;
 - an input interface; and
 - output interfaces to corresponding communication channels that form a bonded channel group;
- wherein:
- the distributor is in communication with the input interface and the output interfaces;
 - the distributor is configured to receive source data from the input interface and distribute the source data across the communication channels; and
 - the communication channels comprise orbital OAM channels.

[0052] US 2013/235885 A1 further discloses a method comprising:

- receiving a packet stream;
- dividing the packet stream into communication units;
- identifying a bonded channel group of communication channels among a set of available communication channels; and
- distributing the communication units across the bonded channel group, where the communication channels of the bonded channel group have a different orbital state from one another.

[0053] More specifically, the method according to US 2013/235885 A1 uses orbital multiplexing on top of other multiplexing techniques, such as Orthogonal Frequency-Division Multiplexing (OFDM). According to US 2013/235885 A1, the orbital multiplexing codes the information across the orbit and spectral domains by using multiple transmit and receive "orbital" devices.

[0054] Furthermore, it is worth noting also that the paper by Masahichi Kishi entitled "The application of a short-time DFT to a Hilbert transformer", Electronics and Communications in Japan (Part I - Communications), Wiley, Hoboken, NJ, US, vol.75, no. 5, 1 May 1992, pages 13-22, discloses a Hilbert transformer implemented using the instantaneous spectrum concept and then applied in communications systems, wherein said Hilbert transformer is based on the concept

that the Hilbert transform is fully realized in the frequency domain by exchanging the complex real part with the complex imaginary part of the instantaneous spectrum, which is then analyzed using a short-time DFT.

[0055] Finally, the paper by Jun Jing et al. entitled "Feature Extraction of Pulse Signal Based on Hilbert-Huang Transformation and Singular Value Decomposition", Proceedings of ICBBE 2007, IEEE, Piscataway, NJ, USA, 1 July 2007, pages 1007 - 1010, discloses a pulse diagnosis approach based on Hilbert-Huang transformation method and Singular Value Decomposition (SVD) technique, wherein the Empirical Mode Decomposition (EMD) method is used to decompose the signal into a number of Intrinsic Mode Function (IMF) components, then the Hilbert transformation is applied to create analytic signal and obtain instantaneous frequency and instantaneous amplitude, from which the initial feature vector matrices are formed. By applying the SVD technique to the initial feature vector matrices, the singular values are obtained, which are regarded as the state feature vectors of the human pulse signals.

OBJECT AND SUMMARY OF THE INVENTION

[0056] The Applicant has carried out an in-depth study in order to develop a practical, efficient mode for carrying out the multidimensional space modulation disclosed in the International Application PCT/IB2012/056804, and this in-depth study has led the Applicant to develop a new, inventive system and method for transmitting and receiving signals at Radio Frequency (RF) (including frequencies from a few KHz to hundreds of GHz) with increased transmission capacity.

[0057] Therefore, an object of the present invention is that of providing a system and a method for transmitting and receiving RF signals with increased transmission capacity.

[0058] This and other objects are achieved by the present invention in so far as it relates to a method and a system for radio communications, as defined in the appended claims.

BRIEF DESCRIPTION OF THE DRAWINGS

[0059] For a better understanding of the present invention, preferred embodiments, which are intended purely by way of non-limiting example, will now be described with reference to the attached drawings (all not to scale), wherein:

- Figure 1 schematically illustrates a transmitting system for transmitting radio vortices according to the background art;
- Figure 2 schematically illustrates a device for generating OAM modes, which device is exploited by the transmitting system shown in Figure 1;
- Figure 3 schematically illustrates a receiving system for receiving radio vortices according to the background art;
- Figure 4 schematically illustrates a device for demodulating OAM modes, which device is exploited by the receiving system shown in Figure 2;
- Figure 5 schematically illustrates time behavior of a complex frequency function constituted by a real cosinusoidal function and an imaginary sinusoidal function;
- Figures 6 and 7 schematically illustrate complex frequency samples of a signal limited in time and of its corresponding analytical signal, respectively (assuming that the signal can be approximated as a band-limited signal too);
- Figures 8 and 9 schematically illustrate time inverse Fourier transform of a complex frequency function and time inverse Fourier transform of its frequency Hilbert transform, respectively;
- Figure 10 schematically illustrates a comparison between a process for generating twisted waves in time domain according to International application PCT/FR2013/052636, and a process for generating twisted waves in frequency domain according to the present invention;
- Figures 11 and 12 schematically illustrate a complex frequency function of an analytical signal and its Hilbert transform in frequency domain, respectively;
- Figure 13 schematically illustrates frequency Hilbert transform and approximations of the latter obtained by using, respectively, one, two, and five orthogonal exponential modes;
- Figures 14, 15 and 16 schematically illustrate frequency behavior of three different orthogonal exponential modes;
- Figure 17 schematically illustrates powers associated with orthogonal harmonic modes used to develop frequency Hilbert transform;
- Figures 18 and 19 schematically illustrate frequency twist complex functions related to two orthogonal harmonic modes;
- Figure 20 schematically illustrates an OFDM super frame with additional Twisted frame Frequency Units (TFUs);
- Figure 21 schematically illustrates an approximation of the impulse response of the frequency Hilbert transform using three complex harmonic modes;
- Figure 22 schematically represents a radio communications method according to the present invention as a generalization of the traditional OFDM (or OFDMA) technique;
- Figures 23 and 24 schematically illustrate time behavior of a time twisted mode +1 complex signal and of a frequency twisted mode +1 complex signal, respectively;

- Figures 25 and 26 schematically illustrate symbol and clock time relationship according to an aspect of the present invention;
- Figure 27 schematically illustrates a traditional scheme of cyclic prefix for OFDM-OFDMA;
- Figure 28 schematically illustrates an example of two delay spread components with cyclic prefix;
- 5 • Figure 29 schematically illustrates time frame duration increase due to cyclic prefix;
- Figures 30 and 31 schematically illustrate time behavior of main mode and twisted modes, respectively, when a cyclic prefix is used;
- Figure 32 schematically illustrates noise bandwidths for time twisted waves;
- Figure 33 schematically illustrates noise impact in the case of frequency twisted waves;
- 10 • Figure 34 schematically illustrates spectral efficiency as a function of the energy per symbol over noise density for frequency twisted waves;
- Figure 35 schematically illustrates a frequency structure of frequency twisted waves according to an illustrative embodiment of the present invention;
- Figure 36 schematically illustrates a transmitting system according to an illustrative embodiment of the present invention;
- 15 • Figure 37 schematically illustrates a receiving system according to an illustrative embodiment of the present invention;
- Figure 38 schematically illustrates an example of square matrix resulting from the multiplication of the transpose of a transmission matrix according to an aspect of the present invention by said transmission matrix;
- Figure 39 schematically illustrates a multilayer architecture wherein a frame structure of frequency twisted waves according to an aspect of the present invention is embedded in a traditional OFDM architecture;
- 20 • Figure 40 schematically illustrates computational complexity of the present invention and frequency reuse according to the present invention as a function of the number of frequency twisted modes used; and
- Figure 41 schematically illustrates flexibility in using OFDM modularity, complex equation number and implementation criticality of the present invention as a function of the number of frequency twisted modes used.

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DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS OF THE INVENTION

[0060] The following discussion is presented to enable a person skilled in the art to make and use the invention. Various modifications to the embodiments will be readily apparent to those skilled in the art, without departing from the scope of the present invention as claimed. Thus, the present invention is not intended to be limited to the embodiments shown and described, but is to be accorded the widest scope consistent with the principles and features disclosed herein and defined in the appended claims.

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[0061] The present invention relates to a practical, efficient mode, in general, for increasing transmission capacity and, in particular, for increasing RF spectrum reuse. In this connection, International Application No. PCT/FR2013/052636 (whose content is herewith enclosed by reference) filed on 5th November 2013 in the name of EUTELSAT S.A. discloses the feasibility of increasing transmission capacity at Radio Frequency (RF) (including frequencies from a few KHz to hundreds of GHz) by exploiting a proper approximation in time domain of the Hilbert transform of digital analytical signals, wherein said approximation of the Hilbert transform is implemented by using twisted waves, specifically orthogonal harmonic modes.

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[0062] In particular, PCT/FR2013/052636 describes a radio communications system, which comprises a transmitter and a receiver, wherein the transmitter is configured to:

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- generate or receive digital symbols having a given symbol rate associated with a corresponding symbol period;
- generate, every S digital symbols generated/received, a respective multi-mode digital signal, which has a predefined time length shorter than S times the symbol period, which is sampled with a predefined sampling rate higher than the symbol rate, and which carries said S digital symbols by means of a plurality of orthogonal harmonic modes comprising
 - a main mode which is a real harmonic mode and carries P of said S digital symbols, and
 - one or more secondary modes carrying the other S-P digital symbols, each secondary mode being a complex harmonic mode time-shifted by half the symbol period with respect to the main mode (wherein S is an integer higher than three and P is an integer lower than S); and
- transmit a radio frequency signal carrying a sequence of the generated multi-mode digital signals.

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[0063] Moreover, the receiver of the radio communications system according to PCT/FR2013/052636 is configured to:

- receive the radio frequency signal transmitted by the transmitter;

- process the received radio frequency signal so as to obtain a corresponding incoming digital signal; and
- extract, from successive, non-overlapped portions of the incoming digital signal sampled with the predefined sampling rate, the S digital symbols respectively carried by each incoming digital signal portion by means of the orthogonal harmonic modes; wherein each of the successive, non-overlapped portions of the incoming digital signal has the predefined time length.

[0064] Preferably, the transmitter according to PCT/FR2013/052636 is configured to generate a multi-mode digital signal carrying S digital symbols by:

- allocating P of the S digital symbols to the main mode by providing, for each of said P digital symbols, a corresponding complex value which represents said digital symbol and is related to the main mode;
- allocating each of the other S-P digital symbols to a corresponding secondary mode by providing, for each of said S-P digital symbols, a corresponding complex value which represents said digital symbol and is related to the secondary mode to which said digital symbol is allocated;
- computing, by using a predefined transmission matrix, M multi-mode complex values related to M successive time instants which, within the predefined time length, are separated by half the symbol period, wherein M is an integer equal to or higher than S, and wherein the predefined transmission matrix relates
 - the S complex values representing the S digital symbols and related to the harmonic modes
 - to the M successive time instants
 - through complex coefficients each of which is related to a respective harmonic mode and to a respective time instant; and
- generating a multi-mode digital signal having the predefined time length and sampled with the predefined sampling rate on the basis of the M multi-mode complex values computed.

[0065] Again preferably, the receiver according to PCT/FR2013/052636 is configured to extract the S digital symbols carried by an incoming digital signal portion having the predefined time length and sampled with the predefined sampling rate by:

- extracting, from said incoming digital signal portion, M multi-mode complex values related to M successive time instants which are, within the predefined time length, separated by half the symbol period;
- computing, by using a reception matrix derived from the predefined transmission matrix, S complex values representing the S digital symbol carried by said incoming digital signal portion by means of the orthogonal harmonic modes, wherein said reception matrix relates
 - the M extracted multi-mode complex values related to the M successive time instants
 - to the S complex values to be computed
 - through complex coefficients each of which is related to a respective harmonic mode and to a respective time instant; and
- determining the S digital symbols represented by the S complex values computed.

[0066] Conveniently, the reception matrix used by the receiver according to PCT/FR2013/052636 is derived from the predefined transmission matrix through a generalized inversion technique.

[0067] More conveniently, according to PCT/FR2013/052636, the predefined transmission matrix is such that the matrix resulting from the multiplication of the transpose of said predefined transmission matrix and said predefined transmission matrix has a determinant different from zero, and the reception matrix is derived from the predefined transmission matrix through a pseudo-inverse technique.

[0068] More and more conveniently, according to PCT/FR2013/052636, the reception matrix is computed on the basis of the following formula:

$$[[GMF]] = ([[A]]^T [[A]])^{-1} [[A]]^T ,$$

where $[[GMF]]$ denotes the reception matrix, $[[A]]$ denotes the predefined transmission matrix, $[[A]]^T$ denotes the

transpose of the predefined transmission matrix, and $(\mathbb{[A]}^T \mathbb{[A]})^{-1}$ denotes the operation of inversion of the matrix resulting from the multiplication of the transpose of the predefined transmission matrix and the predefined transmission matrix.

[0069] Preferably, according to PCT/FR2013/052636, the main mode comprises, within the predefined time length, P samples with sampling period equal to the symbol period, the secondary modes comprise, within the predefined time length, P-1 samples with sampling period equal to the symbol period, each secondary mode is time-shifted by half the symbol period with respect to the main mode, and said M successive time instants, which, within the predefined time length, are separated by half the symbol period, are the sampling times of the main mode and of the secondary modes, thereby resulting that $M=2P-1$.

[0070] More preferably, according to PCT/FR2013/052636, the harmonic modes comprise 2N secondary complex harmonic modes each of which carries a respective Orbital Angular Momentum (OAM) mode and has a respective topological-charge-related index n comprised between -N and +N, wherein N is an integer higher than one; moreover, the main mode carries $P=2^{N+1}+1$ digital symbols and each secondary complex harmonic mode having topological-charge-related index n carries 2^{N-n+1} digital symbols, thereby resulting that $M=2^{N+2}+1$ and $S=2^{N+2}-1$.

[0071] Conveniently, according to PCT/FR2013/052636, the predefined sampling rate depends at least on the predefined time length of each multi-mode digital signal and of each of the successive, non-overlapped portions of the incoming digital signal.

[0072] More conveniently, according to PCT/FR2013/052636, the predefined time length is equal to P times the symbol period.

[0073] More and more conveniently, according to PCT/FR2013/052636, the predefined sampling rate is determined on the basis of the following formula:

$$CR = \frac{2P + u}{2P} \cdot \frac{1}{T_s},$$

where CR denotes said predefined sampling rate, T_s denotes the symbol period, and u denotes a digital-vestigial-component-related parameter whose value is an integer and depends at least on the predefined time length.

[0074] Preferably, the transmitter according to PCT/FR2013/052636 is configured to generate a multi-frame digital signal comprising successive, non-overlapped time frames each of which has the predefined time length and is occupied by a respective multi-mode digital signal; moreover, the multi-frame digital signal carries frame synchronization data related to frame synchronization of its time frames; accordingly, the radio frequency signal transmitted by the transmitter carries the multi-frame digital signal.

[0075] Additionally, the receiver according to PCT/FR2013/052636 is further configured to:

- extract the frame synchronization data from the incoming digital signal;
- detect, on the basis of the extracted frame synchronization data, successive, non-overlapped time frames of the incoming digital signal with the predefined time length; and,
- for each detected time frame of the incoming digital signal, extract, from the incoming digital signal portion within said time frame, the S digital symbols carried by said incoming digital signal portion by means of the orthogonal harmonic modes.

[0076] More preferably, according to PCT/FR2013/052636, the multi-frame digital signal comprises a preamble followed by F successive, non-overlapped time frames occupied, each, by a respective multi-mode digital signal, F being an integer higher than one; in particular, the preamble carries frame synchronization data related to frame synchronization of the F following time frames.

[0077] More and more preferably, according to PCT/FR2013/052636, the frame synchronization data indicate time frame beginning and/or the predefined time length of the time frames.

[0078] In order to increase, in general, transmission capacity at Radio Frequency (RF) (including frequencies from a few KHz to hundreds of GHz) and, in particular, RF spectrum reuse, the present invention, by exploiting duality between time and frequency, teaches to use a twisted-wave-based approximation of the Hilbert transform in frequency domain.

[0079] In particular, thanks to duality principle between time and frequency it is possible to exploit twisted wave functions also in frequency domain. The results are very interesting and promising with features which are, on the one hand, similar to time domain case, but, on the other hand, rather different, for practical applications, from time domain case.

[0080] In detail, "frequency twist" can be seen as a generalisation of the well-known OFDM approach, introducing an absolute novelty in the analysis and design of OFDM signals.

[0081] Theory underlying the present invention will be presented in the following.

[0082] As is known, a signal can be represented in time or frequency domain, time and frequency being conjugate variables.

5 [0083] Considering a time-limited signal within a time window T (as usual technique in the case of OFDM-OFDMA signals), in frequency domain said signal can be represented by a series of sinc functions:

$$10 \quad X(f) = \sum_{k=-\infty}^{+\infty} a_k e^{j\varphi_k} \frac{\sin\left[\pi T\left(f - \frac{k}{T}\right)\right]}{\pi T\left(f - \frac{k}{T}\right)} .$$

15 [0084] In the case the signal $X(f)$ can be approximated with a band-limited signal with bandwidth B , this implies that:

$$X(f - f_0) = X^+(f - f_0) + X^-(f - f_0) ,$$

where + and - denote positive and negative frequencies, respectively.

20 [0085] Taking into consideration only the positive frequencies, it is possible to write:

$$X^+(f - f_0) = X(f - f_0) \quad \text{for } f \geq 0 ,$$

25 and

$$X^+(f - f_0) = 0 \quad \text{for } f < 0 ,$$

30 and also

$$35 \quad X^+(f) = \sum_{k=-N}^{+N} a_k e^{j\varphi_k} \frac{\sin\left[\pi T\left(f - f_0 - \frac{k}{T}\right)\right]}{\pi T\left(f - f_0 - \frac{k}{T}\right)} ,$$

where $N = TB$.

40 [0086] Thence, each sample is constituted by a real part given by $a_k \cos \varphi_k$, and an imaginary part given by $a_k \sin \varphi_k$. The time representation of such a sample is given by one cosinusoidal function in the time window T , having an amplitude of $a_k \cos \varphi_k$, and one sinusoidal function having an amplitude of $a_k \sin \varphi_k$, as shown in Figure 5 where time behavior of said complex frequency sample is illustrated.

45 [0087] The frequency pattern is given by two couples of sinc functions, namely one for the real part and one for the imaginary part, as shown in Figures 6 and 7 which illustrate complex frequency samples of a signal limited in time and of its corresponding analytical signal, respectively (assuming that the signal can be approximated as a band-limited signal too).

50 [0088] As far as analytical signals are concerned, the traditional Hilbert transform is applied in time, assuming that the total bandwidth of the signal can be considered limited and that the baseband signal has been shifted to a proper frequency such that to allow the full bandwidth to be on the positive frequency semiaxis (and, of course, replicated on the negative one). On the positive frequency semiaxis, with respect to the central frequency sample for $k=0$, frequency samples are complex and there results that $a_k e^{j\varphi_k} \neq a_k e^{j\varphi-k}$.

55 [0089] Taking into consideration a complex frequency sample related to a limited time window, it is possible to apply a second Hilbert transform to the function $X^+(f - f_0)$ in frequency domain (on the assumption that the signal is a limited time duration signal):

$$X_H^+(f - f_0) = \int_{-\infty}^{+\infty} \frac{X(\varphi)}{\pi(f - f_0 - \varphi)} d\varphi ,$$

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where the integral can be understood as the main Cauchy value.

[0090] Thence, the time function results to be:

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$$x_H(t) = x(t)e^{j2\pi f_0 t} (u_0(t) - u_0(-t)) ,$$

where $u_0(t)$ and $u_0(-t)$ are the step functions for $t > 0$ and $t < 0$, respectively.

[0091] On the assumption that

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$$X(f) \Big|_{f_k = \frac{k}{T}} = \frac{\sin[\pi(fT - k)]}{\pi(fT - k)} ,$$

then the time function is given by:

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$$x(t) = e^{j2\pi k \frac{t}{T}} , \quad \text{for } |t| < \frac{T}{2} ,$$

25

and

$$x(t) = 0 , \quad \text{for } |t| \geq \frac{T}{2} .$$

30

[0092] Thence, the time transform of the frequency Hilbert transform results to be:

35

$$x(t)(u_0(t) - u_0(-t)) = e^{j2\pi k \frac{t}{T}} \quad \text{for } 0 < t < \frac{T}{2} ,$$

$$x(t)(u_0(t) - u_0(-t)) = -e^{j2\pi k \frac{t}{T}} \quad \text{for } -\frac{T}{2} < t < 0 ,$$

40

and

$$x(t)(u_0(t) - u_0(-t)) = 0 \quad \text{for } |t| \geq \frac{T}{2} .$$

45

[0093] In this respect, Figures 8 and 9 show time inverse Fourier transform of a complex frequency function and time inverse Fourier transform of its frequency Hilbert transform, respectively.

[0094] The analysis of properties of this signal family is thence based on a sequential application of one time Hilbert transform to get the analytical signal and one frequency Hilbert transform to get the twisted wave signals, which are orthogonal to the original samples.

[0095] The process just described is similar but somewhat substantially different from time twist case. In fact, as described in PCT/FR2013/052636, in time twist the Hilbert transform is applied twice in time: the first time Hilbert transform is used to get the analytical signal, and the second time Hilbert transform is used to create the family of twisted waves orthogonal to the original samples.

[0096] In this respect, Figure 10 schematically shows a comparison of the processes for generating twisted waves in time domain according to PCT/FR2013/052636 and in frequency domain according to the present invention.

[0097] In particular, as shown in Figure 10, both the process according to PCT/FR2013/052636 (denoted as a whole

by 300) and the process according to the present invention (denoted as a whole by 400) are applied to a limited-band signal $x(t)$ and include:

- applying to the limited-band signal $x(t)$ a frequency shift (block 301 and 401, respectively); and
- performing a time Hilbert transform of the frequency-shifted signal to get the analytical signal (block 302 and 402, respectively).

[0098] Instead, the two processes are differentiated by the fact that:

- the process 300 according to PCT/FR2013/052636 exploits an approximation in time domain of the Hilbert transform of the analytical signal to create time twisted waves (block 303); and
- the process 400 according to the present invention exploits an approximation in frequency domain of the Hilbert transform of the analytical signal to create frequency twisted waves (block 403).

[0099] In detail, as far as the process 400 according to the present invention is concerned, the Hilbert transform in frequency domain can be seen as an inverse Fourier transform of the analytical signal previously described. In this respect, Figures 11 and 12 show complex frequency function of the analytical signal and its Hilbert transform in frequency domain, respectively.

[0100] The frequency Hilbert transform increases the bandwidth necessary to represent the signal, due to the presence of a discontinuity in the time function at the origin (i.e., taking into consideration the meaning of the main Cauchy value, the position of the symmetry/asymmetry axis of the integration). This aspect is similar to the situation of the Hilbert transform in time domain, and can be handled by considering a development into a series of orthogonal modes. In this respect, Figure 13 shows frequency Hilbert transform and approximations of the latter obtained by using, respectively, one, two, and five orthogonal exponential modes.

[0101] Each mode higher than mode 0 is represented by a couple of odd pulses in frequency domain, centered with respect to the frequency f_0 . In this respect, Figures 14, 15 and 16 show frequency behavior of the mode 0, 1 and 2, respectively. The modes are mutually orthogonal, but the orthogonality between each of them and the main mode frequency pulses is given by the symmetry property of the full bandwidth and this could be defined as a synoptic orthogonality.

[0102] Similarly to the time domain case, the above property is similar to interferometry, which is a property depending on the space geometry and not directly on the signal.

[0103] Therefore, frequency domain can be assimilated to a sort of space (specifically, a "frequency space"), similarly to the situation of the time twist where the time is considered a space (specifically, a "time space"), with additional degrees of freedom.

[0104] It is important to note a basic difference between time and the frequency pulses: time pulses are real, while frequency pulses are in general complex.

[0105] Therefore, the frequency twist shows a more robust capability to carry an additional information channel. In fact, while for time twist it is necessary to increase the nominal Nyquist bandwidth (approximately of 33%), the frequency twist can work without this limitation.

[0106] The frequency Hilbert transform allows, theoretically, to maintain all the information content of the original signal. Therefore, also the orthogonal harmonic mode development up to infinity of the frequency Hilbert transform allows, theoretically, to maintain all the information content of the original signal. Each mode contributes to the information content proportionally to the respective power of the mode (assuming that the overall power of the signal is equal to one). In this respect, Figure 17 schematically shows the respective power (i.e., information content) associated with each mode up to the eleventh mode. From Figure 17 it can be noted that with the first two or three modes it is possible to maintain about 90% of the information content of the original signal with a potential frequency reuse of 1.9. The use of additional higher-order modes results in an increase in implementation complexity rather than in an effective improvement in the performances.

[0107] Generation of frequency twist, in analogy with time twist generation, is organized by associating the complex symbol value $a_k e^{j\omega k}$ to a set of frequency pulses, properly shifted and phase-rotated.

[0108] For the sake of simplicity, it is considered to operate in an OFDM signal structure, where the main signal is represented by the Inverse Fast Fourier Transform (IFFT) of the symbol time flow.

[0109] In addition to this frequency symbol set, it is added, for each mode, a set of frequency samples.

[0110] Modes ± 1 are generated repeating the same symbol at 4 different frequencies $\frac{1}{T} \left(\frac{1}{2} + k \right)$, changing each

time their phases according to $e^{\pm jk\frac{\pi}{2}}$, with $k=0,1,2,3$. This means that the associated IFFT is the sum of 4 decimated

5 IFFT, having only 1 row for each sample and each one is weighted by $0.5e^{\pm jk\frac{\pi}{2}}$.

[0111] Modes ± 2 are generated repeating the same symbol at 8 different frequencies $\frac{1}{T}\left(\frac{3}{4} + k\right)$, changing each

10 time their phases according to $e^{\pm jk\frac{\pi}{4}}$, with $k=0,1,\dots,7$. This means that the associated IFFT is the sum of 8 decimated

15 IFFT, having only 1 row for each sample and each one is weighted by $\frac{1}{\sqrt{8}}e^{\pm jk\frac{\pi}{4}}$.

[0112] In general, modes $\pm N$ are generated repeating the same symbol at 2^{N+1} different frequencies

20 $\frac{1}{T}\left(\frac{2^N - 1}{2^N} + k\right)$, changing each time their phases according to $e^{\pm jk\frac{\pi}{2^N}}$, with $k = 0, 1, \dots, 2^{N+1}-1$. This means that the associated IFFT is the sum of $4N$ decimated IFFT, having only 1 row for each sample and each one is weighted as

25
$$\frac{1}{2^{\frac{N+1}{2}}} e^{\pm jk\frac{\pi}{2^N}}$$

[0113] In practical terms, phases can be simplified (in terms of implementation) assuming the same value each $\pi/2$, in this way the rotation can be represented by a smaller number of bits.

30 [0114] In this respect, Figures 18 and 19 show frequency twist complex functions for mode 1 and 2, respectively. From Figures 18 and 19 it can be noted that the bandwidth increases each time that the mode grows. This implies a sort of increased rigidity of the traditional OFDM structure.

[0115] Then, let us take into consideration an OFDM signal architecture, which can be considered a sequence of frequency pulses having the shape of a sinc. In the same frequency band frequency twisted waves are added and these additional elements in the following will be called Twisted Frequency frame Units (TFUs). In this respect, Figure 20 shows an OFDM super frame with additional TFUs.

[0116] The structure of a TFU is given by the superposition of the OFDM structure and of the structure of the twisted modes previously defined.

[0117] The minimum length of a TFU bandwidth, where modes up to $\pm N$ are used, is given by:

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$$\frac{2^{N+1} + 1}{T},$$

45 where T is the time interval duration which is the inverse of the frequency pulse symbol bandwidth B_S (i.e., $T = 1/B_S$).

[0118] In this respect, Figure 21 show an approximation of the impulse response of the frequency Hilbert transform, which approximation uses three modes.

[0119] The mode structure in the TFU frame takes into account the length of each mode; therefore, using up to mode $\pm N$, the number M_{MFS} of the frequency samples of the main mode is:

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$$M_{MFS} = 2^{N+1} + 1.$$

[0120] It is worth recalling that, assuming $B_S = 1/T$, the frequency samples of generic frequency twisted mode $\pm N$ are at frequencies

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$$\frac{1}{T} \left(\frac{2^N - 1}{2^N} + k \right).$$

5
[0121] Moreover, the number of complex symbol values (or, at reception side, of complex unknowns) of the main mode $n = 0$ is $M_{MFS} = 2^{N+1} + 1$, the number of complex symbol values (or, at reception side, of complex unknowns) of the modes $+1$ and -1 is 2^N , the number of complex symbol values (or, at reception side, of complex unknowns) of the modes $+2$ and -2 is 2^{N-1} , the number of complex symbol values (or, at reception side, of complex unknowns) of the modes $+i$ and $-i$ is 2^{N-i+1} , and the number of complex symbol values (or, at reception side, of complex unknowns) of the modes $+N$ and $-N$ is $2^{N-N+1} = 2$.

10
[0122] Therefore, the overall number S_{TOT} of complex symbol values (or of complex unknowns) is given by:

$$15 \quad S_{TOT} = \left(2^{N+1} + 1 \right) + \sum_{i=1}^N 2^{N-i+1},$$

wherein the first addend represents the number M_{MFS} of symbols (or, at reception side, of complex unknowns) of the main mode $n=0$, while the second addend (i.e., the summation) represents the number $S_{TOT} - M_{MFS}$ of symbols (or, at reception side, of complex unknowns) of all the other modes with $n \neq 0$.

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[0123] The foregoing mathematical formula can be rewritten as:

$$25 \quad S_{TOT} = 1 + \sum_{i=0}^N 2^{N-i+1} = 1 + 2^{N+1} \cdot \sum_{i=0}^N \left(\frac{1}{2} \right)^i.$$

[0124] Thence, since it is known that

$$30 \quad \sum_{i=0}^N x^i = \frac{x^{N+1} - 1}{x - 1} \quad \text{if } x \neq 1,$$

35 then it results that:

$$S_{TOT} = 2^{N+2} - 1.$$

40
[0125] The overlapping of frequency pulses associated with different symbols creates a special form of orthogonality, which depends on the structure of the TFUs. In this sense the TFUs represent a "frequency space" and the different signals are orthogonal in this space according to symmetry and antisymmetry features of the signal structure. This property can be seen as equivalent to the interferometry in the traditional geometrical space.

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[0126] Anyway, even if the present invention deals with "frequency space", the procedure for determining the transmitted signals is performed in the time domain and not in the frequency domain.

[0127] In particular, as shown in Figure 22 which schematically represents the radio communications method according to the present invention (denoted as a whole by 600) as a generalization of the traditional OFDM (or OFDMA) technique (denoted as a whole by 500), the radio communications method 600 according to the present invention can be considered similar to the traditional OFDM (or OFDMA) technique 500, which, as is broadly known, comprises:

- 50
- at the transmission side, the conversion of a symbol serial time stream into a parallel independent frequency stream via an Inverse Fast Fourier Transform (IFFT) (block 501); and,
 - at the reception side, the back-transformation into the original symbol serial time stream via a Fast Fourier Transform (FFT) (block 502).

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[0128] Similarly, the radio communications method 600 according to the present invention exploits:

- at the transmission side, a "Generalized Inverse Fast Fourier Transform" (GIFFT) (block 601) which includes the

implementation of the previously described frequency Hilbert transform approximation based on frequency twisted modes; and,

- at the reception side, a "Generalized Fast Fourier Transform" (GFFT) (block 602) which includes the extraction of the symbols carried by the frequency twisted modes.

5 **[0129]** Let us now consider the structure of the twisted signals in time domain and in frequency domain (on the assumption that for both the domains the first mode ± 1 is used):

- a time twisted mode ± 1 signal can be expressed as a)

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$$x_{T\pm 1}(t) = a_k e^{j\varphi_k} \left\{ \frac{1}{2} \left[\text{rect}\left(t - \frac{T}{2}\right) - \text{rect}\left(t - \frac{3T}{2}\right) \right] \pm \frac{j}{2} \left[\text{rect}(t - T) - \text{rect}\left(t - \frac{5T}{2}\right) \right] \right\};$$

15

and

- a frequency twisted mode ± 1 signal can be expressed as

20

$$X_{F\pm 1}(f) = a_k e^{j\varphi_k} \left\{ \frac{1}{2} \left[\text{rect}\left(f - \frac{B}{2}\right) - \text{rect}\left(f - \frac{3B}{2}\right) \right] \pm \frac{j}{2} \left[\text{rect}(f - B) - \text{rect}\left(f - \frac{5B}{2}\right) \right] \right\}.$$

25 **[0130]** The frequency twisted mode ± 1 signal is analyzed in time domain thereby resulting that: b)

25

$$x_{F\pm 1}(t) = \frac{a_k e^{j\varphi_k}}{2} \left\{ \left[e^{-j2\pi\left(f - \frac{B}{2}\right)t} - e^{-j2\pi\left(f - \frac{3B}{2}\right)t} \right] \pm j \left[e^{-j2\pi(f - B)t} - e^{-j2\pi\left(f - \frac{5B}{2}\right)t} \right] \right\}$$

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for $|t| < \frac{T}{2}$.

35 **[0131]** From a comparison of the signals a) and b) it is evident that the frequency twist is more robust in keeping the independence of the signal equation system. These feature is evident also from the time behavior of the twisted signals, as shown in Figures 23 and 24 which illustrate the time behavior of a time twisted mode +1 signal and of a frequency twisted mode +1 signal, respectively.

40 **[0132]** From an ideal point of view the frequency Hilbert transform is applicable to a time-limited signal. Therefore, in order for the frequency Hilbert transform to be applicable to a continuous time symbol flow, it is necessary to apply said transform to successive time windows of said continuous time symbol flow and to identify the beginning and the end of each time window. This implies that the time window length is increased of a proper portion so as to render each time window detectable. This feature is similar to the bandwidth increase necessary in the case of time twisted waves.

45 **[0133]** Therefore, the frequency rotation requires a time interval slightly larger than the minimum one required by the sampling theorem. This condition is equivalent to consider a symbol duration T_{sym} longer than the system clock duration T_{clk} , as schematically illustrated in Figure 25.

[0134] The above condition implies that, for instance, every 18 frequency bands an additional one is necessary and

50 that, as a consequence, the bandwidth efficiency is given by $\frac{18}{19} \cong 0.95$. In this respect, Figure 26 show symbol and clock period relationship in the case of 18 signal samples and 19 filter samples per frame.

[0135] An interesting aspect of this condition applied to the frequency twisted waves is that it can be interpreted as equivalent to the well-known cyclic prefix already used with the OFDM technique.

55 **[0136]** For multiple path transmission the delay spread is generated by the set of different paths between the transmitter and receiver when those paths have different delays.

[0137] As an example, a signal following a direct line-of-sight path would arrive before a different version of the same signal which is reflected by a distant building.

[0138] Time domain receivers typically synchronize with each delay spread component and adjust their individual

timings before combining the received signals.

[0139] When using a rake receiver, each finger belonging to the rake receiver synchronizes itself with a specific delay spread component. The number of delay spread components which can be combined is, thence, limited to the number of rake fingers. Any delay spread component which is not combined appears as interference.

[0140] LTE receivers do not need to synchronize themselves with individual delay spread components, i.e., it is not necessary to adjust the timing of delay spread components, nor it is necessary to do any combining of delay spread components. An LTE receiver can operate directly on the aggregate received signal without considering delay spread components.

[0141] The cyclic prefix represents a guard period at the start of each OFDMA symbol which provides protection against multi-path delay spread. The cyclic prefix also represents an overhead which should be minimized.

[0142] The duration of the cyclic prefix should be greater than the duration of the multi-path delay spread.

[0143] LTE specifies both normal and extended cyclic prefix lengths. The normal cyclic prefix is intended to be sufficient for the majority of scenarios, while the extended cyclic prefix is intended for scenarios with particularly high delay spread. Durations for the normal and extended cyclic prefix vary from 7% in the standard case up to 25% in the extended case. The cyclic prefix is generated by copying the end of the main body of the OFDMA symbol at the beginning, as shown in Figure 27 which illustrates the traditional scheme of cyclic prefix for OFDM-OFDMA.

[0144] The signal is always continuous at the interface between the cyclic prefix and the main body of the symbol. This results from the main body of the symbol always including an integer number of subcarrier cycles.

[0145] Figure 28 shows an example of 2 delay spread components. The second delay spread component is received later than the first delay spread component. An FFT processing window is defined at the receiver:

- the processing window captures the main body of the OFDMA symbol belonging to the first delay spread component; the cyclic prefix belonging to the first delay spread component is discarded;
- the processing window captures part of the cyclic prefix and the majority of the main body of the OFDMA symbol belonging to the second delay spread component; sections of the cyclic prefix and main body of the OFDMA symbol which fall outside the processing window are discarded; and,
- in the extreme case, where the delay spread is equal to the duration of the cyclic prefix, the FFT processing window fully captures the cyclic prefix belonging to the delay spread component and discards a section of the main body of the OFDMA symbol which has a duration equal to the cyclic prefix.

[0146] The time domain representation of each delay spread component within the processing window is different, however, the frequency domain representation of each delay spread component within the processing window is identical.

[0147] Let us now come back to the description of the present invention and let us assume that modes up to $N=\pm 2$ are used, then the band occupied by the TFU configuration is given by $(2^{2+1}+1)=9$ frequency slots. To this TFU corresponds a Twisted Time frame Unit (TTU), which is increased to avoid time duration ambiguities. If one half slot is considered, the TTU increases of $1/(2B_{sym})$ and the total length of the TTU is $9.5/B_{sym}$. In this respect, Figure 29 shows time frame duration increase due to cyclic prefix (CP) .

[0148] This increase is much lower than the one required by OFDMA. This implies that in practical system there is no additional loss for including frequency twisted waves in the OFDM (or OFDMA) super frame.

[0149] The increase of time interval duration creates automatically a replica of the signal at the beginning of the time interval itself, without any change in the occupied frequency bandwidth.

[0150] This approach is, thence, much more interesting for the understanding of the physical meaning of the cyclic prefix, than the ordinary explanation about its use.

[0151] Considering sampling in frequency domain at a symbol rate slightly smaller than the clock rate, the signal in time domain, on the assumption that only the main mode is used, has the time behavior shown in Figure 30, where the sinusoid are not exactly a multiple of the symbol period: i.e., what the cyclic prefix is performing.

[0152] Adding the FTUs, the twisted mode signals present the same behavior of the main mode signal, as shown in Figure 31.

[0153] Increasing the number of TTUs, the number of sinusoidal signal increases, but the ratio between T_{cl} and T_{sym} remains unchanged.

[0154] The OFDM-twisted frequency has two hierarchical levels:

- the former is related to the TTUs structure; and
- the latter is related to the assembly of the TFUs constituting the OFDM structure.

[0155] Both the levels have the same time duration and the difference is given by the component frequency blocks:

- each TFU corresponds to a number of samples defined by the twisted frequency structure, which introduces additional

frequency samples located between the main frequency samples;

- the super frame structure is a set of TTUs, centered at the proper frequency, and a set of traditional OFDM frequency samples, if wished;
- the standard frequency samples can simplify the process of synchronization and phasing.

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[0156] In order to consider the impact of thermal noise on the twisted waves, it is important to consider what happens on the time twisted waves, because there is a very interesting difference between the two families of twisted waves, which can have important applications in telecommunications, especially in the case of mobile LTE ones.

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[0157] The noise level for the time twisted waves can be represented as divided into two parts:

- a first part related to the symbol rate bandwidth; and
- a second part related to the difference between the symbol rate and the clock bandwidth.

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[0158] In this respect, Figure 32 schematically illustrates noise bandwidths for time twisted waves: one defined according to minimum Nyquist bandwidth, the other related to the bandwidth increase for solving ambiguity aspects.

[0159] The above noise structure can be written as:

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$$n_t(t)e^{j2\pi f_0 t} = n_{\text{int } B}(t)e^{j2\pi f_0 t} + \frac{1}{\sqrt{2}} n_{\text{ext}}(t) \left[e^{j2\pi \left(f_0 + \frac{B+\Delta B/2}{2} \right) t} + e^{j2\pi \left(f_0 - \frac{B+\Delta B/2}{2} \right) t} \right],$$

where $n_{\text{int } B}(t)$ denotes the noise part related to the symbol rate bandwidth, and $n_{\text{ext}}(t) = n_{\text{ext}\Delta B}(t)e^{j\varphi_{\Delta B}(t)}$ denotes the additional noise part due to the need of avoiding ambiguities on the symbol rate phasing.

25
 $n_{\text{int } B}(t)$, when sampled at the symbol rate, is an even function (no information on the odd sampling).

$$\sqrt{2} n_{\text{ext}\Delta B}(t) e^{j2\pi f_0 t} \cos \left[2\pi \left(\frac{B + \Delta B/2}{2} \right) t + \varphi_{\Delta B}(t) \right]$$

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[0160] Moreover, causes an additional contribution on the even and odd components.

[0161] With reference to Figure 32, the noise in the Nyquist band can be represented by a sinc time pulse, which is an even function, while the two sidebands can be represented by an even and an odd component.

[0162] In terms of relationship between the noise components, it can be written:

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$$\frac{\text{noise}_{\text{even}}}{\text{noise}_{\text{odd}}} \cong \frac{B + \Delta B}{\Delta B} = 1 + \frac{B}{\Delta B} .$$

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[0163] Considering the above for a simplified rect filter, the odd modes can be presented as:

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$$\text{POWER MODES}_N = 2 \sum_{k=1}^N \left(\frac{2}{k\pi} \right)^2, \quad \text{for the first modes } \pm 1 \text{ there results } \frac{8}{\pi^2} \cong 0.81 ;$$

45

$$\left(\frac{C}{N} \right)_{\text{odd}} = \frac{0.81 \Delta B}{B + \Delta B} \sigma_t, \quad \text{for a bandwidth increase of 1 over 18 there results}$$

$$\left(\frac{C}{N} \right)_{\text{odd}} = -11.8 + \sigma_{\text{idB}} \text{dB} = -11.8 + \left(\frac{C}{N} \right)_{\text{even}} \text{dB}$$

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- the 11.8 dB term can be considered as the minimum gain against unwished interferences.

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[0164] Instead, as far as the case of frequency twisted waves is concerned, the noise spectrum occupies the bandwidth W and there is not any possibility of separating, in the time domain, its even and odd components for a single frequency pulse, as shown in Figure 33 (which schematically illustrates noise impact in the case of frequency twisted waves).

[0165] In fact, in the time domain the noise signal samples are not associated with the main signal samples, but they are distributed all along the time interval, which is utilized for reconstructing the frequency sampling value (FFT). Therefore it is not possible to associate the main contribution of the noise to the main samples and there is no additional advantage

for higher modes, as in the case of time rotation.

[0166] The twisted waves add independent communication channels, one for each mode, and the information capacity increases with respect to the one of the single channel associated with the main mode.

[0167] The above is valid for both frequency twist and time twist, but it is very interesting to analyze the similarities and the differences in order to optimize the use of the two twisting processes according to the overall system conditions.

[0168] In general terms, it is possible to perform a system comparison on the basis of what presented in the foregoing and in PCT/FR2013/052636. In particular, the following TABLE I presents a comparison between time twisted waves and frequency twisted waves at system level, wherein supplementary references are provided for single carrier case and OFDM case.

TABLE I

Parameter assuming as reference S/N _{thermal} =13.9 dB QPSK	Time Twist (2 modes)	Frequency Twist (2 modes)	Single carrier	OFDM (parameters from LTE)
Linearity (HPA output back-off - dB)	1.5	4.5	1.5	4.5
S/N with interference (dB)	10	12.7	10.5	12.7
Bandwidth increase (with respect to 1/T _s , including roll-off)	33%	0%	12.5%	0%
Time increase (with respect to 1/B _s)	0%	5%	0%	7-25%
Self-interference (dB)	-28	-28	-35	-35
Additional thermal noise	0.5	0	0	0
BER for main mode (no code)	6*10 ⁻³	1.2*10 ⁻⁴	3*10 ⁻³	1.2*10 ⁻⁴
BER for higher modes (no code)	8*10 ⁻⁴	1.2*10 ⁻⁴	NA	NA
Spectral efficiency (bit/s/Hz)	2.5	3.1-2.5	1.8	1.9-1.5
Shannon limit (one channel - bit/s/Hz)	3.5	5.4	3.6	5.4
Shannon limit for twist	5.5	6.9	5.8	6.9

[0169] In summary, time twist operates better in those cases in which the amplifiers work closer to the saturation, while frequency twist operates better when linearity can be preserved. That is a general condition for standard transmission too. In fact, it is well known that, in the case of LTE, on the Forward link (i.e., from the Base Station to mobile device) OFDM is used, while on the Return link (i.e., from the mobile device to the Base Station) Single Channel FDMA is preferred.

[0170] Figure 34 schematically shows (link) spectral efficiency (bit/s/Hz) as a function of the energy per symbol over noise density (E_{symbol}/N_0) for frequency twisted waves, on the assumption that there are present thermal noise and self-interference, which is the noise due to the interference between side-by-side frames.

[0171] In particular, Figure 34 shows that the behavior (in terms of bit/s/Hz) of frequency twisted wave modes presents values that are always lower than the main mode, due to the absence of the noise reduction for twisted modes.

[0172] In the time domain the noise signal samples are not associated with the main signal samples, which are essentially complex values of frequency samples. They are distributed all along the time interval, which is utilized for reconstructing the frequency sampling value (FFT). Therefore, it is not possible to associate the main contribution of the noise with the main samples and there is no additional advantage for higher modes, as in the case of time rotation.

[0173] As far as the transmitter according to the present invention is concerned, the generation of transmission signals is based on the transformation of a symbol serial time flow frames into a parallel flow for each frame, which is equivalent to the generation of a number of sinusoidal signal in the time window frame.

[0174] This process is known and used for OFDM (or OFDMA) architecture; it is equivalent to an IFFT operation. In the case of OFDM, the frequency samples are spaced according to the sampling theorem applied to the frequency domain.

[0175] When using frequency twisted waves, it is necessary to oversampling the overall frequency band via the introduction of additional frequency samples, spaced as previously defined for each mode.

[0176] In this respect, it is worth recalling that generic modes $\pm N$ are generated repeating the same symbol at 2^{N+1}

different frequencies $\frac{1}{T} \left(\frac{2^N - 1}{2^N} + k \right)$, changing each time their phases according to $e^{\pm jk \frac{\pi}{2^N}}$, with $k=0,1,\dots,2^{N+1}-1$.

1. This means that the associated IFFT is the sum of $4N$ decimated IFFT, having only 1 row for each sample and each

$$\frac{1}{N+1} e^{\pm jk \frac{\pi}{2^N}}$$

5 one is weighted as

[0177] A preferred use of frequency twisted waves is inside an OFDM-OFDMA architecture. Taking into consideration that an OFDM structure includes a very large number of frequencies, a possible architecture is proposed here below.

10 **[0178]** On the assumption that modes up to $N=\pm 2$ are used, the band occupied by this configuration is given by $2^{2+1}+1=9$ frequency slots. This section is called Twisted Frequency frame Unit (TFU) and to this TFU corresponds a Twisted Time frame Unit (TTU).

$$\frac{0.5}{B_S}$$

15 **[0179]** The inclusion of the TFU cyclic prefix increases the time frame by $\frac{0.5}{B_S}$ and, thence, for the TTU there results

$$\frac{9.5}{B_S}$$

20 **[0180]** The cyclic prefix is used for each TFU present in the full OFDM-OFDMA bandwidth (and, as previously explained, is physically the same one used for OFDM, but used for each TFU).

[0181] As previously explained, the OFDM-twisted frequency has two hierarchical levels:

- the former is related to the TTUs structure, which depends on the number of modes chosen and the number of frequency slots adopted; and
- the latter is related to the assembly of the TFUs constituting the OFDM structure.

25 **[0182]** Again as previously explained, both the levels have the same time duration and the difference is given by the component frequency blocks:

- each TFU corresponds to a number of samples defined by the twisted frequency structure, which introduces additional frequency samples located between the main frequency samples;
- the super frame structure is a set of TTUs, centered at the proper frequency, and a set of traditional OFDM frequency samples, if wished;
- the standard frequency samples can simplify the process of synchronization and phasing.

35 **[0183]** The generation of the main mode signal and of the twisted mode signals via this process is called, as previously explained, Generalized Inverse Fast Fourier Transform (GIFFT).

40 **[0184]** For the sake of simplicity, it is assumed to use twisted modes ± 1 and ± 2 . This implies, as previously explained, the presence of 9 frequency slots related to the main mode; the frequency rotation requires a time interval slightly larger than the minimum one required by the sampling theorem in order to avoid ambiguities, due to the determination of the frame boundary; this implies that, for instance, every 18 frequency band one additional one is necessary; therefore, there are two time reference window: one defined by the clock T_{cl} and one defined by the symbol T_{sym} . The relationship between T_{cl} and T_{sym} is given by:

$$T_{sym} = \frac{2^{k+2} + 3}{2^{k+2} + 2} T_{cl}$$

where

$$\frac{1}{2^{k+2} + 2} T_{cl}$$

is equivalent to the cyclic prefix.

55 **[0185]** In the present case,

- $T_{sym} = \frac{19}{18}T_{cl}$, and
- $\frac{1}{18}T_{cl}$ is the cyclic prefix.

[0186] In order to create the correct reference between the real and the imaginary signals, it is important to avoid possible ambiguities on the zero of the reference system. In fact, this system shall be used as the reference system of the principal value of the Cauchy integral.

[0187] Therefore, it is important to have a sampling rate slightly larger than the minimum possible for the symbol rate associated with the plane wave mode.

[0188] The frequency representation is shown in Figure 35, which schematically illustrates frequency slots with the additional one for ambiguity resolution (always on the assumption that twisted modes up to ± 2 are used). This additional frequency slot implies that the bandwidth of each symbol is smaller than the maximum one given by Nyquist criterion, and that the additional time duration creates the repetition of part of the sinusoidal at the beginning of the time slot (as in the case of "traditional" cyclic prefix).

[0189] The main mode has the same structure of the traditional IFFT, but the sampling is performed 19 times instead of 18.

[0190] Higher-order modes are generated considering that each of them can be derived considering a set of frequency pulses properly shifted in frequency and phased (either $1; \pm j; -1; \mp j$).

[0191] The frequency pulses of each mode are properly positioned via a transformation algorithm, which is very similar to the IFFT, having in mind the fact that the starting frequency is properly positioned on the frequency axis and that more samples are associated with the same symbol, as explained in the foregoing.

[0192] In order for the operation of the present invention to be better understood, reference is made to Figure 36, which shows a functional block diagram of a transmitting system (denoted as whole by 7) according to an illustrative embodiment of the present invention.

[0193] In particular, the transmitting system 7 shown in Figure 36 is designed to generate frequency twisted waves up to modes ± 3 , and comprises:

- a symbol generation section 70 configured to generate and output a digital symbol stream;
- a frequency twisted mode generation unit 700 based on GIFFT, said frequency twisted mode generation unit 700 being coupled with the symbol generation section 70 to receive the digital symbol stream outputted by the latter, and being configured to generate and output, for each sequence of S_{TOT} digital symbols received from the symbol generation section 70 (in particular, in the example shown in Figure 36, $S_{TOT}=31$), a respective digital time signal obtained by transforming from frequency domain to time domain
 - main mode frequency samples carrying M_{MFS} of said S_{TOT} received digital symbols (in particular, in the example shown in Figure 36, $M_{MFS}=17$) via a frequency main mode (preferably, as previously explained, the main mode frequency samples are OFDM/OFDMA-type frequency samples), and
 - twisted mode frequency samples carrying the other $S_{TOT}-M_{MFS}$ received digital symbols via frequency twisted modes, wherein, in the example shown in Figure 36, the twisted mode frequency samples include
 - frequency samples which are related to frequency twisted mode +1 and which carry four respective digital symbols via the frequency twisted mode +1,
 - frequency samples which are related to frequency twisted mode -1 and which carry four respective digital symbols via the frequency twisted mode -1,
 - frequency twisted mode which are related to frequency twisted mode +2 and which carry two respective digital symbols via the frequency twisted mode +2,
 - frequency twisted mode which are related to frequency twisted mode -2 and which carry two respective digital symbols via the frequency twisted mode -2,
 - frequency twisted mode which are related to frequency twisted mode +3 and which carry one respective digital symbol via the frequency twisted mode +3, and
 - frequency twisted mode which are related to frequency twisted mode -3 and which carry one respective digital symbol via the frequency twisted mode -3; and
- an RF transmission section 7000 which is coupled with the frequency twisted mode generation unit 700 to receive the digital time signals outputted by the latter, and which is configured to transmit at predefined radio frequencies

the received digital time signals by means of a single antenna or a plurality of antennas/antenna elements (not shown in Figure 36 for the sake of illustration simplicity).

[0194] Conveniently, the aforesaid predefined radio frequencies can range from a few KHz to hundreds of GHz depending on the specific application for which the transmitting system 7 is designed.

[0195] Preferably, the transmitting system 7 is a device/system for wireless communications based on OFDM and/or OFDMA, or, more preferably, on LTE and/or WiMAX.

[0196] Conveniently, the symbol generation section 70 is designed to generate the digital symbol stream by performing several operations, such as the following operations (not necessarily all performed and not necessarily performed in the following sequence): information encoding (conveniently by performing one or more signal modulations), one or more frequency shifting operations, one or more analog-to-digital conversion operations, and one or more filtering operations.

[0197] Again conveniently, the RF transmission section 7000 can be designed to transmit at the predefined radio frequencies the digital time signals by performing several operations, such as the following operations (not necessarily all performed and not necessarily performed in the following sequence): frequency up-shifting (in particular from Intermediate Frequency (IF) up to RF), one or more filtering operations, one or more digital-to-analog conversion operations, and power amplification.

[0198] More in detail, as shown in Figure 36, the frequency twisted mode generation unit 700 includes:

- a frequency main mode generation module 701, which, in use,
 - determines, for each of said $M_{MFS}=17$ digital symbols, a corresponding symbol complex value $a_p e^{j\varphi_p}$ (with $p=1,2, \dots, M_{MFS}$) which represents said digital symbol,
 - allocates each of the M_{MFS} symbol complex values to a respective frequency p/T , or (since $T = 1/B_S$) $p \cdot B_S$ (as in the case of traditional OFDM/OFDMA technique), thereby obtaining M_{MFS} main mode frequency samples, and
 - performs an IFFT of all the M_{MFS} main mode frequency samples, thereby generating a digital time signal related to the frequency main mode;

- a frequency twisted mode +1 generation module 702, which, in use,
 - determines, for the first of the four respective digital symbols, a corresponding symbol complex value $a_{+1,1} e^{j\varphi_{+1,1}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{+1,1} e^{j\varphi_{+1,1}}$ to four respective frequencies $B_S \left(\frac{1}{2} + k \right)$ (with $k=0,1,2,3$)

changing, for each frequency sample, the respective phase according to $e^{+jk\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value $a_{+1,1} e^{j\varphi_{+1,1}}$

by a respective complex coefficient $\frac{e^{+jk\frac{\pi}{2}}}{2}$), thereby obtaining four twisted mode frequency samples which are related to the frequency twisted mode +1 and which carry said first respective digital symbol via said frequency twisted mode +1,

- determines, for the second of the four respective digital symbols, a corresponding symbol complex value $a_{+1,2} e^{j\varphi_{+1,2}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{+1,2} e^{j\varphi_{+1,2}}$ to four respective frequencies $B_S \left(\frac{1}{2} + k \right)$ (with $k=4,5,6,7$)

changing, for each frequency sample, the respective phase according to $e^{+j(k-4)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value

$a_{+1,2} e^{j\varphi_{+1,2}}$ by a respective complex coefficient $\frac{e^{+j(k-4)\frac{\pi}{2}}}{2}$), thereby obtaining four further twisted mode

frequency samples which are related to the frequency twisted mode +1 and which carry said second respective digital symbol via said frequency twisted mode +1,

- determines, for the third of the four respective digital symbols, a corresponding symbol complex value $a_{+1,3}e^{j\varphi_{+1,3}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{+1,3}e^{j\varphi_{+1,3}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with $k=8,9,10,11$)

changing, for each frequency sample, the respective phase according to $e^{+j(k-8)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value

$a_{+1,3}e^{j\varphi_{+1,3}}$ by a respective complex coefficient $\frac{e^{+j(k-8)\frac{\pi}{2}}}{2}$), thereby obtaining four further twisted mode frequency samples which are related to the frequency twisted mode +1 and which carry said third respective digital symbol via said frequency twisted mode +1,

- determines, for the fourth of the four respective digital symbols, a corresponding symbol complex value $a_{+1,4}e^{j\varphi_{+1,4}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{+1,4}e^{j\varphi_{+1,4}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with

$k=12,13,14,15$) changing, for each frequency sample, the respective phase according to $e^{+j(k-12)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol

complex value $a_{+1,4}e^{j\varphi_{+1,4}}$ by a respective complex coefficient $\frac{e^{+j(k-12)\frac{\pi}{2}}}{2}$), thereby obtaining four final twisted mode frequency samples which are related to the frequency twisted mode +1 and which carry said fourth respective digital symbol via said frequency twisted mode +1, and

- performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode +1, thereby generating a digital time signal related to the frequency twisted mode +1;

- a frequency twisted mode -1 generation module 703, which, in use,

- determines, for the first of the four respective digital symbols, a corresponding symbol complex value $a_{-1,1}e^{j\varphi_{-1,1}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{-1,1}e^{j\varphi_{-1,1}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with $k=0,1,2,3$)

changing, for each frequency sample, the respective phase according to $e^{-jk\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value $a_{-1,1}e^{j\varphi_{-1,1}}$

by a respective complex coefficient $\frac{e^{-jk\frac{\pi}{2}}}{2}$), thereby obtaining four twisted mode frequency samples which are related to the frequency twisted mode -1 and which carry said first respective digital symbol via said frequency twisted mode -1,

- determines, for the second of the four respective digital symbols, a corresponding symbol complex value $a_{-1,2}e^{j\varphi_{-1,2}}$ which represents said digital symbol,

5 - allocates said symbol complex value $a_{-1,2}e^{j\varphi_{-1,2}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with $k=4,5,6,7$)

changing, for each frequency sample, the respective phase according to $e^{-j(k-4)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value

10
$$\frac{e^{-j(k-4)\frac{\pi}{2}}}{2}$$

$a_{-1,2}e^{j\varphi_{-1,2}}$ by a respective complex coefficient $\frac{e^{-j(k-4)\frac{\pi}{2}}}{2}$), thereby obtaining four further twisted mode frequency samples which are related to the frequency twisted mode -1 and which carry said second respective digital symbol via said frequency twisted mode -1,

15 - determines, for the third of the four respective digital symbols, a corresponding symbol complex value $a_{-1,3}e^{j\varphi_{-1,3}}$ which represents said digital symbol,

20 - allocates said symbol complex value $a_{-1,3}e^{j\varphi_{-1,3}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with $k=8,9,10,11$)

changing, for each frequency sample, the respective phase according to $e^{-j(k-8)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol complex value

25
$$\frac{e^{-j(k-8)\frac{\pi}{2}}}{2}$$

$a_{-1,3}e^{j\varphi_{-1,3}}$ by a respective complex coefficient $\frac{e^{-j(k-8)\frac{\pi}{2}}}{2}$), thereby obtaining four further twisted mode frequency samples which are related to the frequency twisted mode -1 and which carry said third respective digital symbol via said frequency twisted mode -1,

30 - determines, for the fourth of the four respective digital symbols, a corresponding symbol complex value $a_{-1,4}e^{j\varphi_{-1,4}}$ which represents said digital symbol,

35 - allocates said symbol complex value $a_{-1,4}e^{j\varphi_{-1,4}}$ to four respective frequencies $B_S\left(\frac{1}{2} + k\right)$ (with k

$=12,13,14,15$) changing, for each frequency sample, the respective phase according to $e^{-j(k-12)\frac{\pi}{2}}$ and weighting each frequency sample by 1/2 (i.e., multiplying, for each of the four respective frequencies, the symbol

40
$$\frac{e^{-j(k-12)\frac{\pi}{2}}}{2}$$

complex value $a_{-1,4}e^{j\varphi_{-1,4}}$ by a respective complex coefficient $\frac{e^{-j(k-12)\frac{\pi}{2}}}{2}$), thereby obtaining four final twisted mode frequency samples which are related to the frequency twisted mode -1 and which carry said fourth respective digital symbol via said frequency twisted mode -1, and

45 - performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode -1, thereby generating a digital time signal related to the frequency twisted mode +1;

50 • a frequency twisted mode +2 generation module 704, which, in use,

- determines, for the first of the two respective digital symbols, a corresponding symbol complex value $a_{+2,1}e^{j\varphi_{+2,1}}$ which represents said digital symbol,

55 - allocates said symbol complex value $a_{+2,1}e^{j\varphi_{+2,1}}$ to eight respective frequencies $B_S\left(\frac{3}{4} + k\right)$ (with $k=0,1,\dots,7$)

changing, for each frequency sample, the respective phase according to $e^{+jk\frac{\pi}{4}}$ and weighting each frequency

5 sample by $\frac{1}{\sqrt{8}}$ (i.e., multiplying, for each of the eight respective frequencies, the symbol complex value

10 $a_{+2;1}e^{j\varphi_{+2;1}}$ by a respective complex coefficient $\frac{e^{+jk\frac{\pi}{4}}}{\sqrt{8}}$), thereby obtaining eight twisted mode frequency samples which are related to the frequency twisted mode +2 and which carry said first respective digital symbol via said frequency twisted mode +2,

- determines, for the second of the two respective digital symbols, a corresponding symbol complex value $a_{+2;2}e^{j\varphi_{+2;2}}$ which represents said digital symbol,

15 - allocates said symbol complex value $a_{+2;2}e^{j\varphi_{+2;2}}$ to eight respective frequencies $B_S\left(\frac{3}{4} + k\right)$ (with $k=8,9,\dots,15$)

20 changing, for each frequency sample, the respective phase according to $e^{+j(k-8)\frac{\pi}{4}}$ and 1 weighting each

frequency sample by $\frac{1}{\sqrt{8}}$ (i.e., multiplying, for each of the eight respective frequencies, the symbol complex

25 value $a_{+2;2}e^{j\varphi_{+2;2}}$ by a respective complex coefficient $\frac{e^{+j(k-8)\frac{\pi}{4}}}{\sqrt{8}}$), thereby obtaining eight further twisted mode frequency samples which are related to the frequency twisted mode +2 and which carry said second respective digital symbol via said frequency twisted mode +2, and

- performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode +2, thereby generating a digital time signal related to the frequency twisted mode +2;

• a frequency twisted mode -2 generation module 705, which, in use,

35 - determines, for the first of the two respective digital symbols, a corresponding symbol complex value $a_{-2;1}e^{j\varphi_{-2;1}}$ which represents said digital symbol,

40 - allocates said symbol complex value $a_{-2;1}e^{j\varphi_{-2;1}}$ to eight respective frequencies $B_S\left(\frac{3}{4} + k\right)$ (with $k=0,1,\dots,7$)

changing, for each frequency sample, the respective phase according to $e^{-jk\frac{\pi}{4}}$ and weighting each frequency

45 sample by $\frac{1}{\sqrt{8}}$ (i.e., multiplying, for each of the eight respective frequencies, the symbol complex value

50 $a_{-2;1}e^{j\varphi_{-2;1}}$ by a respective complex coefficient $\frac{e^{-jk\frac{\pi}{4}}}{\sqrt{8}}$), thereby obtaining eight twisted mode frequency samples which are related to the frequency twisted mode -2 and which carry said first respective digital symbol via said frequency twisted mode -2,

- determines, for the second of the two respective digital symbols, a corresponding symbol complex value $a_{-2;2}e^{j\varphi_{-2;2}}$ which represents said digital symbol,

55 - allocates said symbol complex value $a_{-2;2}e^{j\varphi_{-2;2}}$ to eight respective frequencies $B_S\left(\frac{3}{4} + k\right)$ (with $k=8,9,\dots,15$)

changing, for each frequency sample, the respective phase according to $e^{-j(k-8)\frac{\pi}{4}}$ and weighting each

5 frequency sample by $\frac{1}{\sqrt{8}}$ (i.e., multiplying, for each of the eight respective frequencies, the symbol complex

10 value $a_{-2,2}e^{j\varphi_{-2,2}}$ by a respective complex coefficient $\frac{e^{-j(k-8)\frac{\pi}{4}}}{\sqrt{8}}$), thereby obtaining eight further twisted mode

frequency samples which are related to the frequency twisted mode -2 and which carry said second respective digital symbol via said frequency twisted mode -2, and

15 - performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode -2, thereby generating a digital time signal related to the frequency twisted mode -2;

• a frequency twisted mode +3 generation module 706, which, in use,

20 - determines, for the respective digital symbol, a corresponding symbol complex value $a_{+3}e^{j\varphi_{+3}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{+3}e^{j\varphi_{+3}}$ to sixteen respective frequencies $B_S\left(\frac{7}{8} + k\right)$ (with $k=0,1,\dots,15$)

25 changing, for each frequency sample, the respective phase according to $e^{+jk\frac{\pi}{8}}$ and weighting each frequency sample by 1/4 (i.e., multiplying, for each of the sixteen respective frequencies, the symbol complex value $a_{+3}e^{j\varphi_{+3}}$

30 by a respective complex coefficient $\frac{e^{+jk\frac{\pi}{8}}}{4}$), thereby obtaining sixteen twisted mode frequency samples which are related to the frequency twisted mode +3 and which carry said respective digital symbol via said frequency twisted mode +3, and

35 - performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode +3, thereby generating a digital time signal related to the frequency twisted mode +3;

• a frequency twisted mode -3 generation module 707, which, in use,

40 - determines, for the respective digital symbol, a corresponding symbol complex value $a_{-3}e^{j\varphi_{-3}}$ which represents said digital symbol,

- allocates said symbol complex value $a_{-3}e^{j\varphi_{-3}}$ to sixteen respective frequencies $B_S\left(\frac{7}{8} + k\right)$ (with $k=0,1,\dots,15$)

45 changing, for each frequency sample, the respective phase according to $e^{-jk\frac{\pi}{8}}$ and weighting each frequency sample by 1/4 (i.e., multiplying, for each of the sixteen respective frequencies, the symbol complex value $a_{-3}e^{j\varphi_{-3}}$

50 by a respective complex coefficient $\frac{e^{-jk\frac{\pi}{8}}}{4}$), thereby obtaining sixteen twisted mode frequency samples which are related to the frequency twisted mode -3 and which carry said respective digital symbol via said frequency twisted mode -3, and

55 - performs an IFFT of all the sixteen twisted mode frequency samples related to the frequency twisted mode -3, thereby generating a digital time signal related to the frequency twisted mode -3; and

• a combining module 708, which, in use, combines (namely, adds together) the digital time signals outputted by the

frequency main mode generation module 701 (i.e., the digital time signal related to the frequency main mode), by the frequency twisted mode +1 generation module 702 (i.e., the digital time signal related to the frequency twisted mode +1), by the frequency twisted mode -1 generation module 703 (i.e., the digital time signal related to the frequency twisted mode -1), by the frequency twisted mode +2 generation module 704 (i.e., the digital time signal related to the frequency twisted mode +2), by the frequency twisted mode -2 generation module 705 (i.e., the digital time signal related to the frequency twisted mode -2), by the frequency twisted mode +3 generation module 706 (i.e., the digital time signal related to the frequency twisted mode +3), and by the frequency twisted mode -3 generation module 707 (i.e., the digital time signal related to the frequency twisted mode -3), thereby generating an overall digital time signal.

[0199] Linearity of the frequency twisted mode generation unit 700 is important, due to the presence of a wide multi-carrier architecture.

[0200] Conveniently, the frequency twisted mode generation unit 700 carries out all the aforesaid operations of by

using an overall complex transmission matrix $[[GIFFT]]$ designed to implement, in a combined way and at one and the same time, all the aforesaid operations so that, when applied to a sequence of S_{TOT} digital symbols received from the symbol generation section 70, time samples of the corresponding digital time signal are automatically computed by the frequency twisted mode generation unit 700.

[0201] Preferably, for each digital time signal generated and outputted by the combining module 708, the frequency twisted mode generation unit 700 is further designed to insert, at the beginning of said digital time signal, a respective cyclic prefix which is a replica of an end portion of said digital time signal (in accordance with what was previously explained).

[0202] Let us now consider the operation of the present invention at reception side, and, in this respect, reference is made to Figure 37, which shows a functional block diagram of a receiving system (denoted as whole by 8) according an illustrative embodiment of the present invention.

[0203] In particular, as shown in Figure 37, the receiving system 8 comprises:

- an RF reception section 8000, which is designed to receive the RF signals transmitted at the predefined radio frequencies by the transmitting system 7 (in particular, by the RF transmission section 7000); said RF reception section 8000 being designed to receive the RF signals by means of a single antenna or a plurality of antennas/antenna elements (not shown in Figure 37 for the sake of illustration simplicity), and to process the received RF signals so as to obtain, on the basis of said received RF signals, an incoming digital signal;
- a symbol extraction unit 800 based on GFFT, which is coupled with said RF reception section 8000 to receive the incoming digital signal therefrom, and which is designed to
 - process said incoming digital signal so as to extract the digital symbols carried by said incoming digital signal, and
 - output a stream of extracted digital symbols; and
- a symbol processing section 80, which is coupled with said symbol extraction unit 800 to receive the stream of extracted digital symbols outputted by the latter, and which is designed to process said stream of extracted digital symbols.

[0204] The aforesaid predefined radio frequencies coincide with the radio frequencies used in transmission by the transmitting system 7, in particular by the RF transmission section 7000. Conveniently, as already said, the predefined radio frequencies can range from a few KHz to hundreds of GHz depending on the specific application which the overall radio communications system comprising the transmitting system 7 and the receiving system 8 is designed for.

[0205] Preferably, the receiving system 8 is a device/system for wireless communications based on OFDM and/or OFDMA, or, more preferably, on LTE and/or WiMAX.

[0206] Conveniently, the RF reception section 8000 is designed to obtain the incoming digital signal by performing several operations upon the received RF signals, such as the following operations (not necessarily all performed and not necessarily performed in the following sequence): low-noise amplification, one or more frequency down-shifting operations (in particular from RF down to IF), one or more filtering operations, and one or more analog-to-digital conversion operations.

[0207] Again conveniently, the symbol processing section 80 is designed to process the stream of extracted digital symbols by performing several operations, such as the following operations (not necessarily all performed and not necessarily performed in the following sequence): one or more filtering operations, one or more digital-to-analog conversion operations, one or more frequency shifting operations, and information decoding (conveniently by performing

one or more signal demodulations).

[0208] At the reception side the parallel signal flow is to be considered, as in the case of OFDM (or OFDMA) , and a reception matrix $\mathbb{[GFFT]}$ is used by the symbol extraction unit 800 to extract the digital symbols carried by the incoming digital signal.

[0209] The main difference with respect to the standard OFDM signal is that OFDM exploits Hermitian matrices, while in the case of frequency twisted waves the transmission matrix $\mathbb{[GIFFT]}$ is rectangular and, thence, in order for the reception matrix $\mathbb{[GFFT]}$ to be obtained, pseudo-inverse approach is exploited. The use of such a procedure is called Generalized Fast Fourier Transform (GFFT) and is somewhat similar to the Generalized Matched Filter used for the time twisted waves and described in PCT/FR2013/052636.

[0210] In use, the symbol extraction unit 800 processes the incoming digital signal by using a time window T_{Sym} (including the cyclic prefix) so as to process successive, non-overlapped portions of the incoming digital signal each having a time duration equal to T_{Sym} , and to extract the digital symbols respectively carried by each incoming digital signal portion.

[0211] The input sequence in the time window T_{Sym} is oversampled with the same law of the transmission flow; on the assumption that modes up to $\pm N$ are used, the size of the reception matrix is given by:

- the number of unknowns (i.e., symbol complex values) $S_{TOT} = 2^{N+2}-1$ in a frequency frame of $M_{MFS} = 2^{N+1}+1$ main mode frequency pulses; and
- the number of equations, which represents also the overall number M_{TS} of the samples in time domain, which is given by

$$M_{TS} = \left(1 + \frac{N(N+1)}{2}\right) 2^{N+1} + 1 .$$

[0212] More in detail, in order to solve the equation system at the reception side, a reception matrix $\mathbb{[GFFT]}$ is used by the symbol extraction unit 800, which reception matrix $\mathbb{[GFFT]}$ is derived from the transmission matrix $\mathbb{[GIFFT]}$ through a generalized inversion technique, such as the pseudo-inverse technique.

[0213] In mathematical terms, given the transmission matrix $\mathbb{[GIFFT]}$ with $M_{TS} \times S_{TOT}$ complex coefficients, and given also the vector $[S]$ of the S_{TOT} symbol complex values to be transmitted, at transmission side there results that:

$$\mathbb{[GIFFT]} [S] = [TTU]$$

where $[TTU]$ denote the vector of the M_{TS} complex values of the time samples of a digital time signal outputted by the frequency twisted mode generation unit 700.

[0214] Let us now consider the reception side, where it is useful to use a generalized inversion technique, such as the pseudo-inverse technique, to invert the foregoing matrix equation:

$$\mathbb{[GIFFT]}^T \mathbb{[GIFFT]} [S] = \mathbb{[GIFFT]}^T [TTU],$$

and thence

$$[S] = \left(\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}\right)^{-1} \mathbb{[GIFFT]}^T [TTU], \quad (1)$$

where $\mathbb{[GIFFT]}^T$ denotes the transpose of the matrix $\mathbb{[GIFFT]}$, and $\left(\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}\right)^{-1}$ denotes

the operation of inversion of the square matrix resulting from the multiplication $\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}$.

[0215] In particular, at reception side $\mathbf{[S]}$ becomes the vector of the S_{TOT} unknown symbol complex values to be determined by the symbol extraction unit 800, and $\mathbf{[TTU]}$ becomes the vector of the M_{TS} complex values of the time samples determined by the symbol extraction unit 800 on the basis of an incoming digital signal portion.

[0216] Condition for the existence of a set of solutions for the unknown vector $\mathbf{[S]}$ is that the square matrix resulting from the multiplication $\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}$ has a determinant different than zero, i.e., in mathematical terms,

$$\det(\mathbb{[A]}^T \mathbb{[A]}) \neq 0. \quad (2)$$

[0217] Therefore, if the transmission matrix $\mathbb{[GIFFT]}$ is designed so as to satisfy the condition (2), then the S_{TOT} unknown symbol complex values can be determined by the symbol extraction unit 800 by solving the equation system resulting from the matrix equation (1).

[0218] Thence, the reception matrix $\mathbb{[GFFT]}$, which is a non-Hermitian matrix, can be defined as:

$$\mathbb{[GFFT]} = (\mathbb{[GIFFT]}^T \mathbb{[GIFFT]})^{-1} \mathbb{[GIFFT]}^T.$$

[0219] In this respect, Figure 38 schematically shows an example of square matrix resulting from the multiplication $\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}$ on the assumption that modes up to ± 1 are used. In particular, the matrix shown in Figure 38 includes cells which are blank or grey, wherein the grey cells represent the matrix cells actually occupied by coefficients, while the blank cells represent the matrix cells not occupied by any coefficient. This representation of the matrix stems from the fact that the Figure 38 is mainly intended to show the matrix structure (and not the matrix coefficients).

[0220] The condition (2) is satisfied more easily in the frequency twist case than in the time twist one, as it can be inferred by looking at the shape of the time signals. The main reason for such a behaviour is based on the fact that a frequency function is intrinsically complex, while a time signal is real. In other words, the square matrix resulting from the multiplication

$\mathbb{[GIFFT]}^T \mathbb{[GIFFT]}$ is much more robust than the similar matrix obtained in the case of time twisted waves.

[0221] It is important to note that the determinant is well sized and does not require an increase of the bandwidth as in the time twisted wave case. In fact, changing from 19 to 18 samples the determinant relative value changes from 1 to about 0.1, which are both values valid for the matrix inversion.

[0222] Ideally, the use of the cyclic prefix allows interference level to be limited close to zero. This is true when a large number of side lobes are present outside the useful bandwidth. For the TFUs the bandwidth is limited to the first side-lobe of the frequency pattern, therefore the interference level is of the order of about -30/-35 dB. This can be considered self-generated noise due to interframe interference.

[0223] The presence of self-generated noise produces a limitation on the information transmission capacity when the E_{symbol}/N_0 is very high (namely, higher than 40 dB), as shown in Figure 34. This limitation is present in all the cases where the physical resource is reused (e.g. orthogonal polarization limit 35-40 dB).

[0224] Sizing and configuration of a transmitter and a receiver using frequency twisted waves according to the present invention can be considered an innovative updating of the OFDM, OFDMA and COFDM (i.e., Coded OFDM) architectures.

[0225] In this respect, Figure 39 schematically shows a multilayer architecture wherein the frame structure of the frequency twisted waves is embedded in a traditional OFDM architecture.

[0226] The proposed architecture allows the OFDM basic structure to be used with the additional layer of frequency twisted waves (including the size and the references of the frequency twisted wave frame, i.e., the frequency slot positions, their phases and the association of these frequencies with the transmitted symbols, as described in detail in the foregoing).

[0227] The advantage of frequency twisted waves with respect to time twisted waves is evident in this aspect; in fact, for frequency twisted waves there is no need to build up a dedicated "space reference system" as in the case of time twist and this is due to the important consideration that OFDM and similar transmission techniques are already "block signal transmission" techniques (i.e., implement simultaneous transmission of signals using IFFT).

[0228] The size of transmission block is increased by a factor related to the number of frequency twisted modes used. In fact, as previously explained, each twisted frequency mode $\pm n$ exploits a sequence of 2^{N+1} additional frequency carriers (where $\pm N$ are the highest modes used) positioned at

$$B_S \left(\frac{2^n - 1}{2^n} + k \right),$$

5 where $0 \leq k \leq 2^{N+1}-1$.

[0229] The receiver (in particular, the symbol extraction unit 800) handles a number of unknowns S_{TOT} smaller than the number of equations M_{TS} (as previously explained), this implies the use of the pseudo-inverse technique and an increase of the computational complexity with respect to the usual OFDM block signal computation, which, as is known, produces square matrixes $L_{SF} \times L_{SF}$, where L_{SF} is the length of the super frame.

[0230] Instead, in the case of frequency twisted waves, assuming the same super frame length L_{SF} and a frame length L_{TFU} equal to $2^{N+1}+1$, the number of equation for said super frame is:

$$15 \quad M_{TS} \frac{L_{SF}}{2^{N+1} + 1} = \left[\left(1 + \frac{N(N+1)}{2} \right) 2^{N+1} + 1 \right] \frac{L_{SF}}{2^{N+1} + 1},$$

where, if $N=2$ and $L_{SF}=2016$, the number of equations is 7392 and the number of the unknowns is given by

$$20 \quad \frac{2^{N+2} + 1}{2^{N+1} + 1} L_{SF} = 3808 .$$

[0231] The increased complexity factor of the matrix operations is given by the ratio between the number of operations of twisted waves and the number of OFDM operations:

$$25 \quad \frac{\left(\frac{2^{N+2} + 1}{2^{N+1} + 1} L_{SF} \right) \times \left[\left(1 + \frac{N(N+1)}{2} \right) 2^{N+1} + 1 \right] \frac{L_{SF}}{2^{N+1} + 1}}{(L_{SF})^2} = 6.1 .$$

[0232] In this respect, Figure 40 shows computational complexity of the present invention and frequency reuse according to the latter as a function of N (where, as previously said, $\pm N$ are the highest twisted frequency modes used). From Figure 40 it is evident that a reasonable compromise is obtained with $N=2$, with a very good frequency reuse and a limited increase in computational complexity.

[0233] The introduction of an additional layer in the super frame organization is somewhat limiting the possibility of OFDM adapting itself to the channel characteristics; i.e., in a traditional OFDM frame the adaptability is given by the single carrier bandwidth B_S , while for the frequency twisted wave case the adaptability is given by $(2^{N+1}+1)B_S$. This is a limitation on the system adaptability and has the same trend of the computational complexity. In this respect, Figure 41 schematically illustrates flexibility in using OFDM modularity, complex equation number and implementation criticality of the present invention as a function of N (where, as previously said, $\pm N$ are the highest twisted frequency modes used).

[0234] Some features of the present invention are briefly summarized here below:

- 45 • due the structure of the OFDM signal there is no additional noise due to the introduction of frequency twisted waves;
- the OFDM cyclic prefix includes the equivalent cyclic prefix necessary for frequency twisted waves; anyway, it is clear that, if the cyclic prefix is fully used for OFDM, it should be accordingly increased; and
- there is no practical advantage of performance in using frequency twisted mode higher than ± 3 , but, on the contrary, computational complexity grows quite rapidly.

[0235] As explained in the foregoing, the implementation of the frequency twisted waves according to the present invention can be regarded as an approximation of the frequency Hilbert transform. This fact implies, on one side, a bandwidth increase, and, on the other side, the presence of an absolute limitation on the increase in frequency reuse, which is lower than two. In this respect, the following TABLE II lists some features related to the use of frequency twisted waves according to the present invention.

TABLE II

Parameter	Parameter value (considering using up to modes $\pm N$)	Parameter approximate value for N=2
Frequency reuse	$\frac{2^{N+2} - 1}{2^{N+1} + 1}$	1.67
Vestigial time interval increase	$\frac{2^{N+2} + 3}{2^{N+2} + 2}$	1.056
Total frame bandwidth/symbol bandwidth	$\frac{B_F}{B_S} = \frac{2^{N+2} + 3}{2}$	9.5
Super Frame loss	<1%	0.99
Additional bandwidth noise (dB)	$10 \log \left(\frac{2^{N+3} + 3}{2^{N+3} + 2} \right) \text{ dB}$	0.25 dB
Digitalization noise < -30 dB phase error	$\frac{N}{2^N - 1}$	9 bits
Maximum interframe interference (dB)	<-19-3N dB	< -25 dB

[0236] For N=3, the frame length is smaller than 32 symbols, the necessary number of bits is about 10, the increase of the thermal noise is close to 0 dB, and the frequency reuse close to 1.7.

[0237] As far as practical implementation of the present invention is concerned, the frequency twisted mode generation unit 700 based on GIFFT and the symbol extraction unit 800 based on GFFT are preferably implemented by means of Field-Programmable Gate Array (FPGA), Application-Specific Integrated Circuit (ASIC), and Software Defined Radio (SDR) technologies.

[0238] From the foregoing, it may be immediately appreciated that the present invention allows to increase frequency reuse and transmission capacity by exploiting an original application of the Hilbert transform in frequency domain.

[0239] The present invention can be considered very interesting and almost revolutionary to develop a new theory for digital communications beyond the classical approach based on analytical signals.

[0240] In particular, as previously explained in detail, according to the present invention radio vorticity is considered as a way to approximate the Hilbert transform and is applied in frequency domain so as to generate independent radio channels within one and the same bandwidth. These channels have an available bandwidth decreasing with the radio vorticity mode number and the total bandwidth advantage is growing as $1/2^N$, limited by 2, which represents the maximum possible use of the imaginary channel of the Hilbert transform.

[0241] From a mathematical (and physical) perspective, this Hilbert-transform-based approach is very similar to an interferometry measurement performed in frequency instead of in geometrical space.

[0242] The present invention can be advantageously exploited, in general, in all kinds of radio communications, and, in particular, in radio communications based, in general, on OFDM and/or OFDMA, and, specifically, on LTE and/or WiMAX.

[0243] Finally, it is worth noting that a combined use of frequency twist according to the present invention and time twist according to PCT/FR2013/052636 is particularly advantageous in asymmetrical radio communications systems, such as mobile radio communications systems, for example based on LTE and/or WiMAX. In fact, in such a scenario, frequency twist according to the present invention can be advantageously applied to the Forward channel from a Base Station to a mobile device, while time twist according to PCT/FR2013/052636 can be advantageously applied to the Return channel from a mobile device to a Base Station.

[0244] In conclusion, it is clear that numerous modifications and variants can be made to the present invention, all falling within the scope of the invention, as defined in the appended claims. function of N (where, as previously said, $\pm N$ are the highest twisted frequency modes used).

[0245] Some features of the present invention are briefly summarized here below:

- due the structure of the OFDM signal there is no additional noise due to the introduction of frequency twisted waves;
- the OFDM cyclic prefix includes the equivalent cyclic prefix necessary for frequency twisted waves; anyway, it is clear that, if the cyclic prefix is fully used for OFDM, it should be accordingly increased; and
- there is no practical advantage of performance in using frequency twisted mode higher than ± 3 , but, on the contrary, computational complexity grows quite rapidly.

[0246] As explained in the foregoing, the implementation of the frequency twisted waves according to the present invention can be regarded as an approximation of the frequency Hilbert transform. This fact implies, on one side, a bandwidth increase, and, on the other side, the presence of an absolute limitation on the increase in frequency reuse, which is lower than two. In this respect, the following TABLE II lists some features related to the use of frequency twisted waves according to the present invention.

TABLE II

Parameter	Parameter value (considering using up to modes $\pm N$)	Parameter approximate value for N=2
Frequency reuse	$\frac{2^{N+2} - 1}{2^{N+1} + 1}$	1.67
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Total frame bandwidth/symbol bandwidth	$\frac{B_F}{B_S} = \frac{2^{N+2} + 3}{2}$	9.5
Super Frame loss	<1%	0.99
Additional bandwidth noise (dB)	$10 \log \left(\frac{2^{N+3} + 3}{2^{N+3} + 2} \right) \text{ dB}$	0.25 dB
Digitalization noise < -30 dB phase error	$\frac{N}{2^N - 1}$	9 bits
Maximum interframe interference (dB)	<-19-3N dB	< -25 dB

[0247] For N=3, the frame length is smaller than 32 symbols, the necessary number of bits is about 10, the increase of the thermal noise is close to 0 dB, and the frequency reuse close to 1.7.

[0248] As far as practical implementation of the present invention is concerned, the frequency twisted mode generation unit 700 based on GIFFT and the symbol extraction unit 800 based on GFFT are preferably implemented by means of Field-Programmable Gate Array (FPGA), Application-Specific Integrated Circuit (ASIC), and Software Defined Radio (SDR) technologies.

[0249] From the foregoing, it may be immediately appreciated that the present invention allows to increase frequency reuse and transmission capacity by exploiting an original application of the Hilbert transform in frequency domain.

[0250] The present invention can be considered very interesting and almost revolutionary to develop a new theory for digital communications beyond the classical approach based on analytical signals.

[0251] In particular, as previously explained in detail, according to the present invention radio vorticity is considered as a way to approximate the Hilbert transform and is applied in frequency domain so as to generate independent radio channels within one and the same bandwidth. These channels have an available bandwidth decreasing with the radio vorticity mode number and the total bandwidth advantage is growing as $1/2^N$, limited by 2, which represents the maximum possible use of the imaginary channel of the Hilbert transform.

[0252] From a mathematical (and physical) perspective, this Hilbert-transform-based approach is very similar to an interferometry measurement performed in frequency instead of in geometrical space.

[0253] The present invention can be advantageously exploited, in general, in all kinds of radio communications, and,

in particular, in radio communications based, in general, on OFDM and/or OFDMA, and, specifically, on LTE and/or WiMAX.

[0254] Finally, it is worth noting that a combined use of frequency twist according to the present invention and time twist according to PCT/FR2013/052636 is particularly advantageous in asymmetrical radio communications systems, such as mobile radio communications systems, for example based on LTE and/or WiMAX. In fact, in such a scenario, frequency twist according to the present invention can be advantageously applied to the Forward channel from a Base Station to a mobile device, while time twist according to PCT/FR2013/052636 can be advantageously applied to the Return channel from a mobile device to a Base Station.

REFERENCES CITED IN THE DESCRIPTION

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Patent documents cited in the description

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- WO 2012084039 A1 [0005] [0013]
- WO IB2012056804 A [0015] [0016] [0017] [0018] [0019] [0020] [0033] [0049] [0050] [0056]
- US 2013235885 A1 [0051] [0052] [0053]
- FR 2013052636 W [0059] [0061] [0062] [0063] [0064] [0065] [0066] [0067] [0068] [0069] [0070] [0071] [0072] [0073] [0074] [0075] [0076] [0077] [0095] [0096] [0097] [0098] [0168] [0209] [0243] [0254]

Non-patent literature cited in the description

- Orbital Angular Momentum in Radio - A System Study. **MOHAMMADI S. M. et al.** IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION. IEEE SERVICE CENTER, 01 February 2010, vol. 58, 565-572 [0005]
- **TAMBURINI F. et al.** *Encoding many channels in the same frequency through radio Vorticity: first experimental test*, 12 July 2011 [0005]
- The application of a short-time DFT to a Hilbert transformer. **MASAHICHI KISHI.** Electronics and Communications in Japan (Part I - Communications). Wiley, 01 May 1992, vol. 75, 13-22 [0054]
- Feature Extraction of Pulse Signal Based on Hilbert-Huang Transformation and Singular Value Decomposition. **JUN JING et al.** Proceedings of ICB- BE 2007. IEEE, 01 July 2007, 1007-1010 [0055]

Patentkrav

1. Radiokommunikationsfremgangsmåde omfattende udførelse, via en sender (7), af følgende trin:

- 5 a) tilvejebringelse af et digitalt tidssignal som bærer S_{TOT} digitale symboler, som skal sendes, hvor det digitale tidssignal er tidsbegrænset og er baseret på en transformation, fra frekvensdomæne til tidsdomæne, af frekvens-samples, som bærer nævnte S_{TOT} digitale symboler; og
- b) transmission af et radiofrekvenssignal, som bærer nævnte digitale tidssignal;
- 10 hvilken fremgangsmåde yderligere omfatter udførelse, via en modtager (8), af følgende trin:
- c) modtagelse af radiofrekvenssignalet sendt via senderen (7);
- d) bearbejdning af det modtagne radiofrekvenssignal for at opnå et tilsvarende indgående digitalt signal; og
- 15 e) udledning, fra det indgående digitale signal, af de S_{TOT} digitale symboler båret af nævnte indgående digitale signal;
- hvor frekvensprøverne inkluderer M_{MFS} -hovedmodusfrekvens-samples, som:
- vedrører en frekvenshovedmodus;
 - 20 • er placeret, i frekvensdomæne, ved M_{MFS} -hovedmodus-frekvenser, som er anbragt med afstand via et forudbestemt frekvensmelletrum og et område fra B_s til M_{MFS} gange B_s , hvor B_s angiver nævnte forudbestemte frekvensmelletrum; og
 - bærer M_{MFS} af nævnte S_{TOT} digitale symboler, som skal sendes;
- 25 **kendetegnet ved, at** frekvens-samples yderligere inkluderer drejningsmodusfrekvensprøver, som:
- vedrører $2N$ frekvens-drejnings-modi vedrørende en approksimation af Hilbert-transformationen i frekvensdomæne;
 - er placeret, i frekvensdomæne, ved drejningsmodusfrekvenser forskellige
 - 30 fra M_{MFS} -hovedmodusfrekvenserne; og
 - bærer $S_{TOT}-M_{MFS}$ digitale symboler ikke båret af M_{MFS} -hovedmodus-frekvens-samples;
- hvor S_{TOT} er lig med $2^{N+2}-1$, M_{MFS} er lig med $2^{N+1}+1$, og N angiver et heltal større end nul;

5 hvor, for hver frekvensdrejningsmodus, de 2^{N+1} respektive drejningsmodus-frekvens-samples ved 2^{N+1} tilsvarende drejningsmodusfrekvenser bærer det $2^{N-|n|}$ respektive digitale symbol(er), hvor de 2^{N+1} tilsvarende drejningsmodusfrekvenser er anbragt med afstand via nævnte forudbestemte frekvensmellemlum og er placeret, i frekvensdomæne, ved

$$B_S \left(\frac{2^{|n|} - 1}{2^{|n|}} + k \right),$$

hvor

- 10 • n angiver et heltalindeks, som identificerer nævnte frekvensdrejningsmodus, er omfattet mellem $-N$ og $+N$ og er forskellig fra nul, og
- k angiver et heltal i området fra nul til $2^{N+1}-1$, eller fra en til 2^{N+1} ; hvor hvert af nævnte S_{TOT} digitale symboler, som skal sendes, er repræsenteret af en respektiv kompleks symbolværdi;

15 hvor, for hver frekvensdrejningsmodus, de 2^{N+1} respektive drejningsmodusfrekvens-samples omfatter, for hver af de $2^{N-|n|}$ respektive digitale symbol(er), $2^{|n|+1}$ frekvens-samples, som:

- bærer nævnte digitale symbol;
- er ved frekvenser, som er placeret, i frekvensdomæne, ved

$$B_S \left[\frac{2^{|n|} - 1}{2^{|n|}} + \left(k^* + i \cdot 2^{|n|+1} \right) \right],$$

20 hvor k^* angiver et heltal i området fra nul til $2^{|n|+1}-1$, eller fra en til $2^{|n|+1}$, og hvor i er et indeks, som identificerer nævnte digitale symbol og er omfattet mellem nul og $2^{N-|n|}-1$; og

- har hver en respektiv kompleks værdi opnået ved at multiplicere den komplekse symbolværdi, som repræsenterer nævnte digitale symbol, med en respektiv kompleks faktor, som:

25 - hvis n er større end nul, er lig med

$$\frac{e^{+jk^* \frac{\pi}{2^{|n|}}}}{2^{\frac{|n|+1}{2}}}$$

- eller, hvis n er mindre end nul, er lig med

$$\frac{e^{-jk^* \frac{\pi}{2^{|n|}}}}{2^{\frac{|n|+1}{2}}}$$

hvor j angiver den imaginære enhed;

hvorved hver frekvensdrejningsmodus er en kompleks harmonisk modus,

5 som er ortogonal på frekvenshovedmodusen og på en hvilken som helst anden anvendt frekvensdrejningsmodus;

hvor trin a) inkluderer tilvejebringelse af det digitale tidssignal ved anvendelse af en foruddefineret transmissionsmatrix, som forbinder

- S_{TOT} digitale symboler, som skal sendes,

10 • til tids-samples af det digitale tidssignal

- gennem koefficienter vedrørende transformationen, fra frekvensdomæne til tidsdomæne, af hovedmodusfrekvens-samples og af drejningsmodus-frekvens-samples;

hvor det digitale tidssignal omfatter et antal af tids-samples lig med

15 $\left[1 + \frac{N(N+1)}{2}\right] 2^{N+1} + 1;$

hvor den foruddefinerede transmissionsmatrix omfatter

$M_{TS} \times S_{TOT}$ koefficienter, M_{TS} angiver nævnte antal af tids-samples af det digitale tidssignal;

hvor den foruddefinerede transmissionsmatrix er således, at matrixen

20 resulterende fra multiplikationen af den transponerede af nævnte

foruddefinerede transmissionsmatrix og nævnte foruddefinerede

transmissionsmatrix har en determinant forskellig fra nul;

og hvor trin e) inkluderer udledning af S_{TOT} digitale symboler båret af det

indgående digitale signal ved anvendelse af en modtagelsesmatrix afledt fra

25 den foruddefinerede transmissionsmatrix gennem en pseudo-invers teknik.

2. Fremgangsmåden ifølge krav 1, hvor modtagelsesmatrixen beregnes på basis af følgende formel:

$$[[GFFT]] = ([[GIFFT]]^T [[GIFFT]])^{-1} [[GIFFT]]^T,$$

30 hvor

$$[[GFFT]]$$

angiver modtagelsesmatrixen,

$$[[GIFFT]]$$

angiver den foruddefinerede transmissionsmatrix,

$$[[GIFFT]]^T$$

5 angiver den transponerede af den foruddefinerede transmissionsmatrix, og

$$([[GIFFT]]^T [[GIFFT]])^{-1}$$

angiver handlingen af inversion af matrixen resulterende fra multiplikationen af den transponerede af den foruddefinerede transmissionsmatrix og den foruddefinerede transmissionsmatrix.

10

3. Radiokommunikationsfremgangsmåde ifølge krav 1 eller 2, hvor hovedmodus-frekvens-samples er frekvens-samples af Orthogonal Frequency-Division Multiplexing (OFDM) typen, eller af Orthogonal Frequency-Division Multiple Access (OFDMA) typen.

15

4. Radiokommunikationsfremgangsmåden ifølge et hvilket som helst af kravene 1-3, hvor nævnte trin a) inkluderer:

- tilvejebringelse af et første digitalt tidssignal resulterende fra hovedmodusfrekvens-samples og drejningsmodusfrekvens-samples; og
 - tilvejebringelse af et andet digitalt tidssignal som inkluderer et cyklisk præfix efterfulgt af det første digitale tidssignal, hvor det cykliske præfix er en kopi af et endefsnit af nævnte første digitale tidssignal;
- og hvor nævnte trin b) inkluderer afsendelse af et radiofrekvenssignal, som bærer det andet digitale tidssignal.

20

25

5. System (7) til radiokommunikation, konfigureret til:

- at tilvejebringe et digitalt tidssignal som bærer S_{TOT} digitale symboler, som skal sendes, hvor det digitale tidssignal er tidsbegrænset og er baseret på en transformation, fra frekvensdomæne til tidsdomæne, af frekvens-samples, som bærer nævnte S_{TOT} digital symboler; og
- at sende et radiofrekvenssignal som bærer nævnte digitale tidssignal; hvor frekvens-samples inkluderer M_{MFS} -hovedmodusfrekvens-samples, som:
- vedrører en frekvenshovedmodus;

30

- er placeret, i frekvensdomæne, ved M_{MFS} -hovedmodusfrekvenser, som er anbragt med afstand via et forudbestemt frekvensmelletrum og et område fra B_S til M_{MFS} gange B_S , hvor B_S angiver nævnte forudbestemte frekvensmelletrum; og
- 5
- bærer M_{MFS} af nævnte S_{TOT} digitale symboler, som skal sendes; **kendetegnet ved, at** frekvens-samples yderligere inkluderer drejningsmodusfrekvens-samples, som:
 - vedrører $2N$ frekvensdrejningsmodi vedrørende en approksimation af Hilbert-transformationen i frekvensdomæne;
- 10
- er placeret, i frekvensdomæne, ved drejningsmodusfrekvenser forskellige fra M_{MFS} -hovedmodusfrekvenserne; og
 - bærer $S_{TOT}-M_{MFS}$ digitale symboler ikke båret af M_{MFS} hovedmodusfrekvens-samples; hvor S_{TOT} er lig med $2^{N+2}-1$, M_{MFS} er lig med $2^{N+1}+1$, og N angiver et heltal større end nul;
- 15
- hvor, for hver frekvensdrejningsmodus, de 2^{N+1} respektive drejningsmodusfrekvens-samples ved de 2^{N+1} tilsvarende drejningsmodusfrekvenser bærer det $2^{N-|n|}$ respektive digitale symbol(er), hvor nævnte 2^{N+1} tilsvarende drejningsmodusfrekvenser er anbragt med
- 20
- afstand via nævnte forudbestemte frekvensmelletrum og er placeret, i frekvensdomæne, ved
- $$B_S \left(\frac{2^{|n|}-1}{2^{|n|}} + k \right),$$
- hvor
- n angiver et heltalindeks, som identificerer nævnte frekvensdrejnings-
- 25
- modus, er omfattet mellem $-N$ og $+N$ og er forskellig fra nul, og
- k angiver et heltal i området fra nul til $2^{N+1}-1$, eller fra en til 2^{N+1} ; hvor hver af nævnte S_{TOT} digitale symboler, som skal sendes, er repræsenteret af en respektiv kompleks symbolværdi; hvor, for hver frekvensdrejningsmodus, de 2^{N+1} respektive

30

drejningsmodusfrekvens-samples omfatter, for hver af de $2^{N-|n|}$ respektive digitale symbol(er), $2^{|n|+1}$ frekvens-samples, som:

 - bærer nævnte digitale symbol;

- er ved frekvenser, som er placeret, i frekvensdomæne, ved

$$B_S \left[\frac{2^{|n|} - 1}{2^{|n|}} + (k^* + i \cdot 2^{|n|+1}) \right],$$

hvor k^* angiver et heltal i området fra nul til $2^{|n|+1} - 1$, eller fra en til $2^{|n|+1}$, og hvor i er et indeks, som identificerer nævnte digitale symbol og er omfattet mellem nul og $2^{N-|n|}-1$; og

5

- har hver en respektiv kompleks værdi opnået ved at multiplicere den komplekse symbolværdi, som repræsenterer nævnte digitale symbol, med en respektiv kompleks faktor, som:

- hvis n er større end nul, er lig med

$$\frac{e^{+jk^* \frac{\pi}{2^{|n|}}}}{2^{\frac{|n|+1}{2}}}$$

10

- eller, hvis n er mindre end nul, er lig med

$$\frac{e^{-jk^* \frac{\pi}{2^{|n|}}}}{2^{\frac{|n|+1}{2}}}$$

hvor j angiver den imaginære enhed;

hvorved hver frekvensdrejningsmodus er en kompleks harmonisk modus, som er ortogonal på frekvenshovedmodusen og på en hvilken som helst anden anvendt frekvensdrejningsmodus;

15

hvor nævnte system (7) er konfigureret til at tilvejebringe det digitale tidssignal ved anvendelse af en foruddefineret transmissionsmatrix, som forbinder

20

- de S_{TOT} digitale symboler, som skal sendes,
- med tids-samples af det digitale tidssignal
- gennem koefficienter vedrørende transformationen, fra frekvensdomæne til tidsdomæne, af hovedmodusfrekvens-samples og af drejningsmodus-frekvens-samples;

25

hvor det digitale tidssignal omfatter et antal af tids-samples lig med

$$\left[1 + \frac{N(N+1)}{2} \right] 2^{N+1} + 1;$$

- hvor den foruddefinerede transmissionsmatrix omfatter $M_{TS \times S_{TOT}}$ koefficienter, M_{TS} angiver nævnte antal tids-samples af det digitale tidssignal;
- 5 og hvor den foruddefinerede transmissionsmatrix er således, at matrixen resulterende fra multiplikationen af den transponerede af nævnte foruddefinerede transmissionsmatrix og nævnte foruddefinerede transmissionsmatrix har en determinant forskellig fra nul.
- 6.** Softwareprogramprodukt omfattende softwarekodeafsnit som:
- 10
- kan eksekveres af bearbejdningsorgan af en indretning/et system til radiokommunikation; og
 - er således at det, når det eksekveres, får nævnte indretning/system til at blive konfigureret som systemet (7) ifølge krav 5.

DRAWINGS

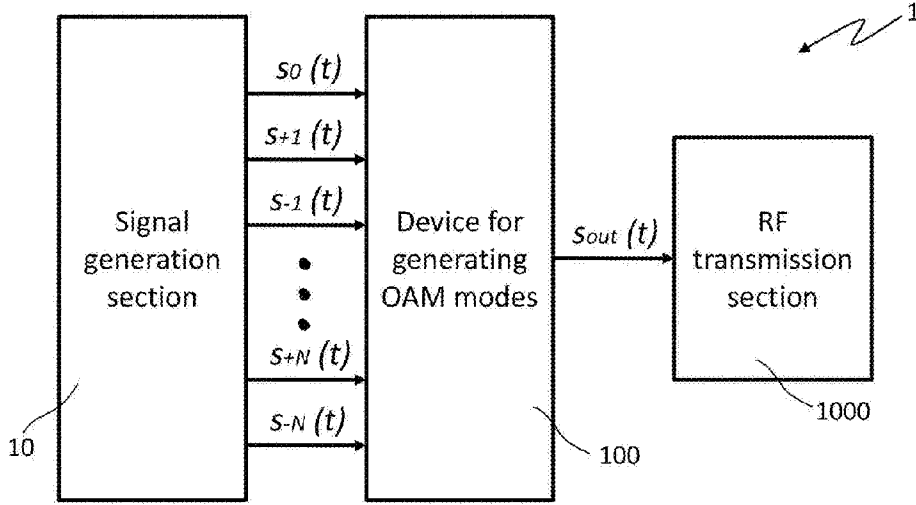


FIG. 1

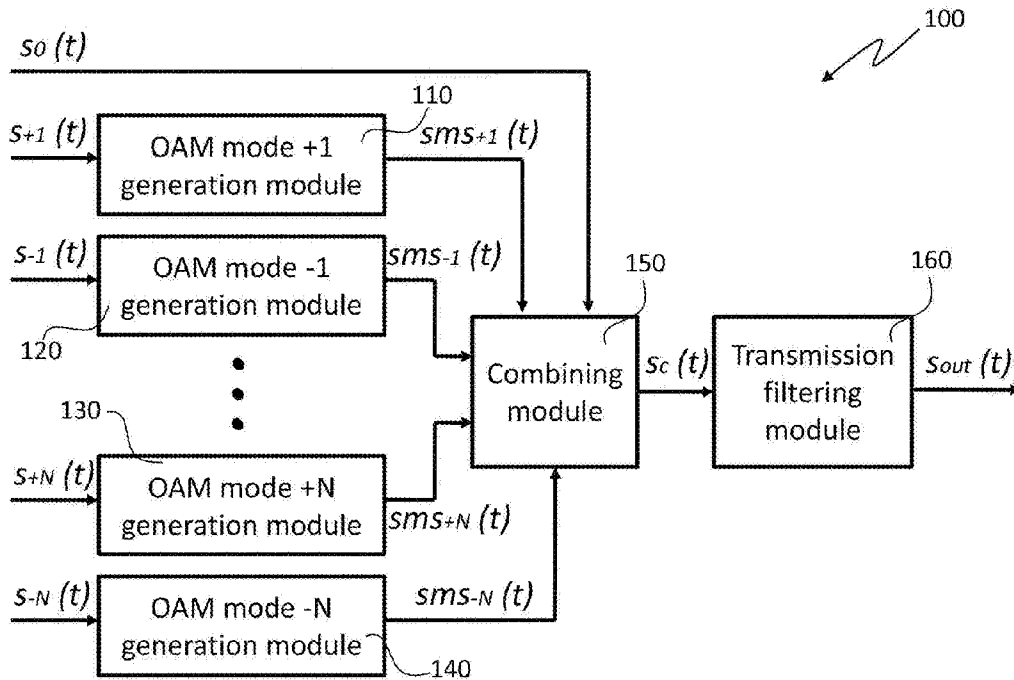


FIG. 2

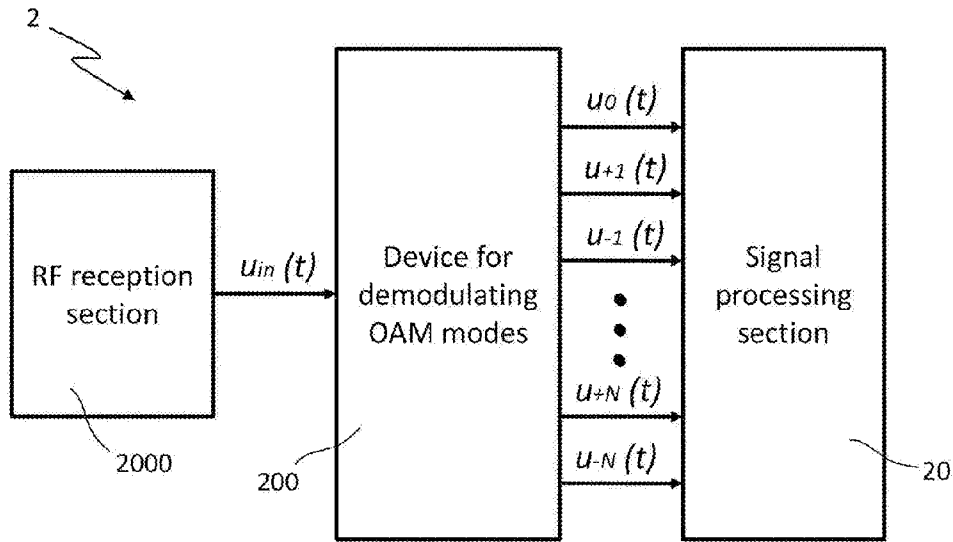


FIG. 3

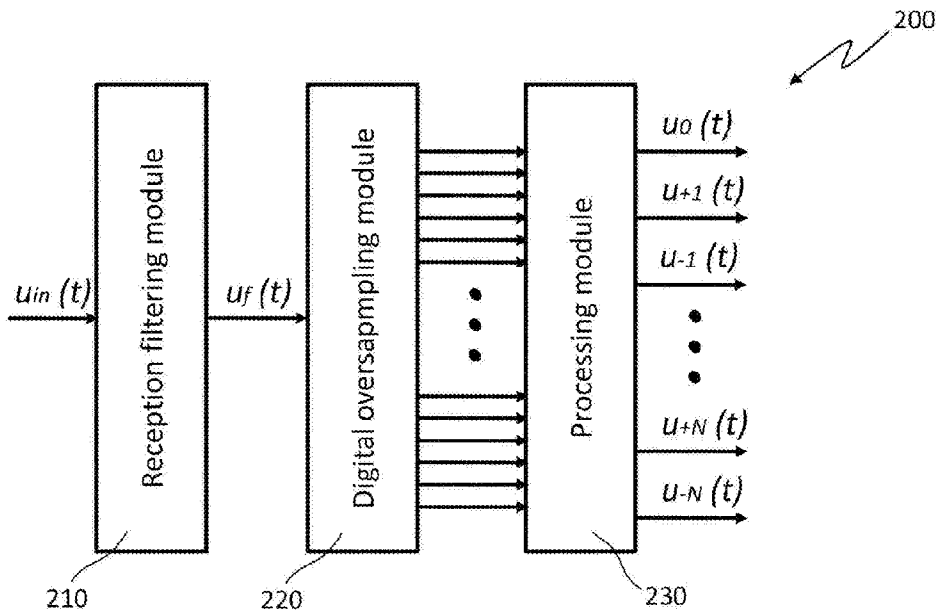


FIG. 4

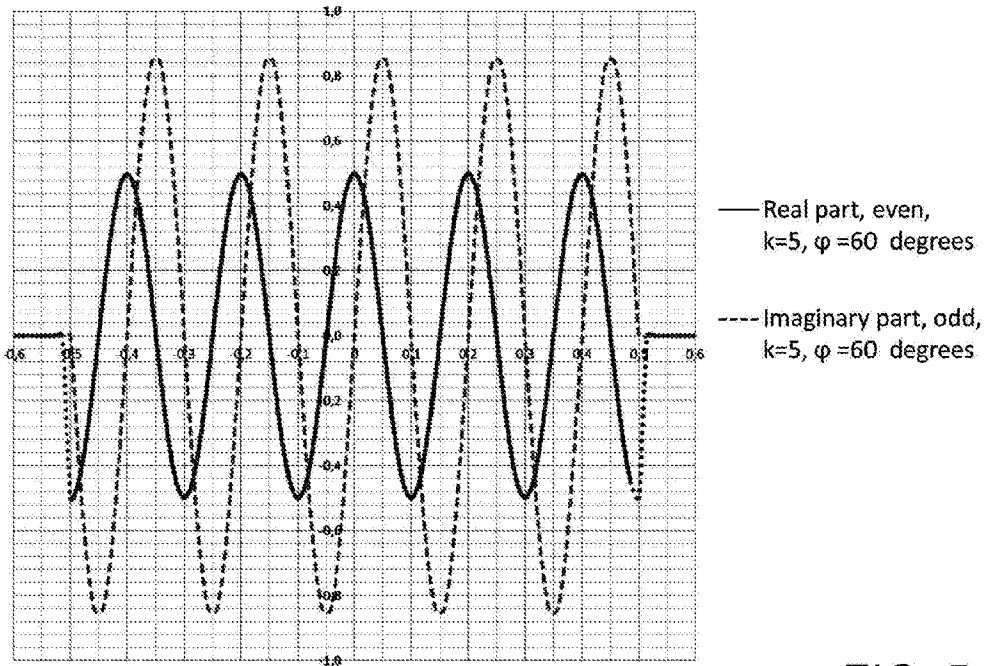


FIG. 5

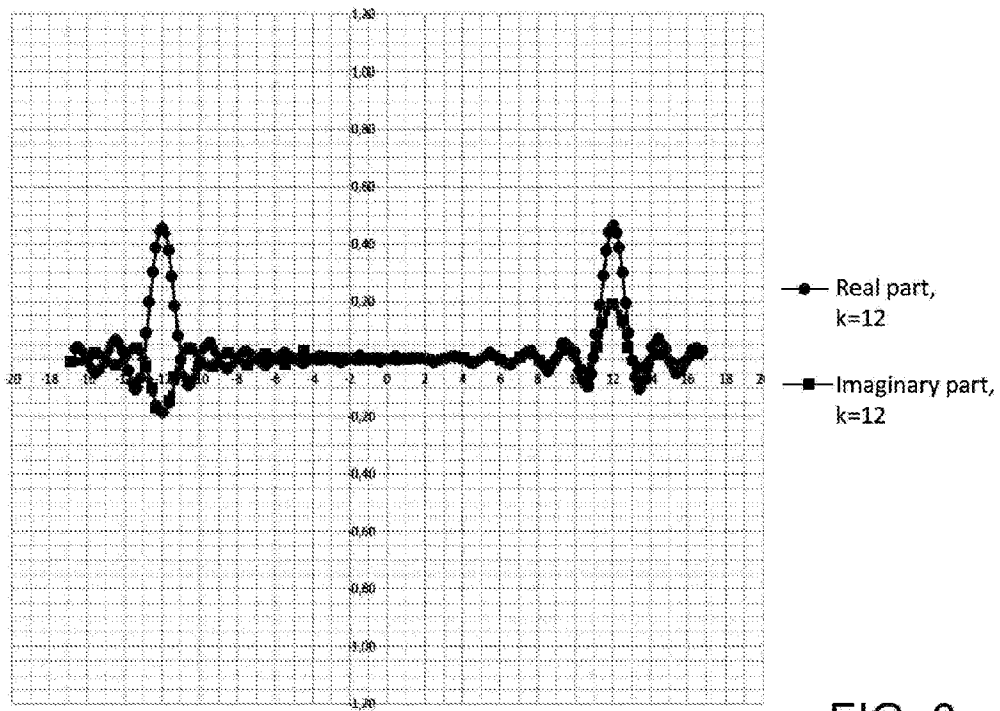
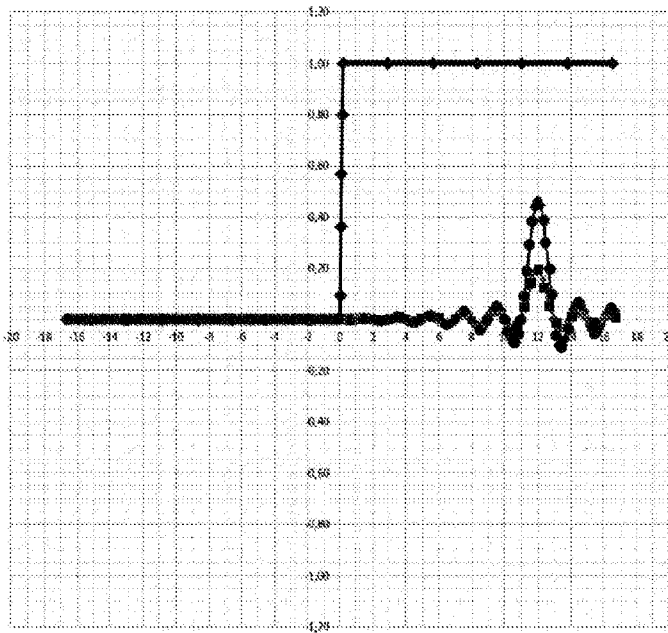
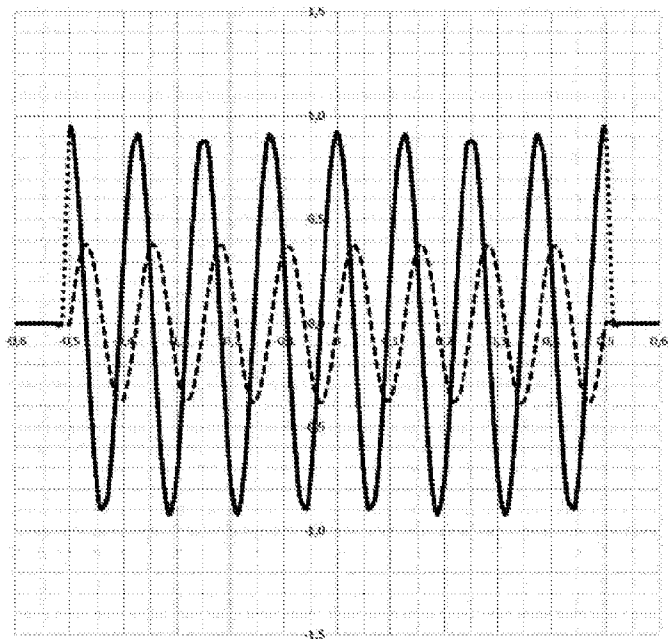


FIG. 6



- Real part, $k=12$
- Imaginary part, $k=12$
- ◆ Analytical signal filter

FIG. 7



- Real part of $x(t)$, even, $k=8$, $\varphi = 22.5$ degrees
- - - Imaginary part of $x(t)$, even, $k=8$, $\varphi = 22.5$ degrees

FIG. 8

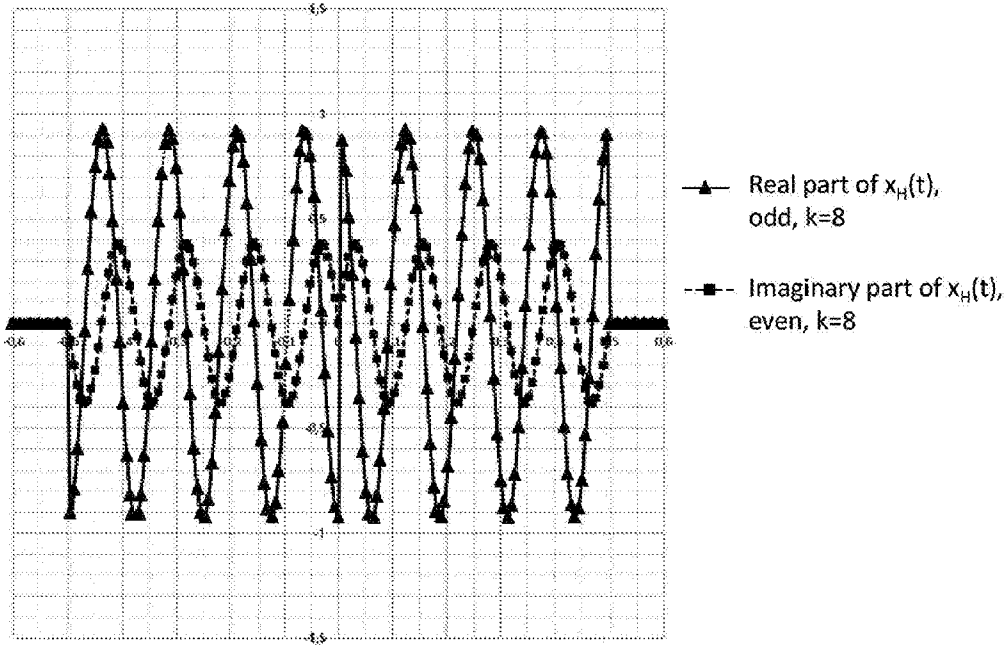


FIG. 9

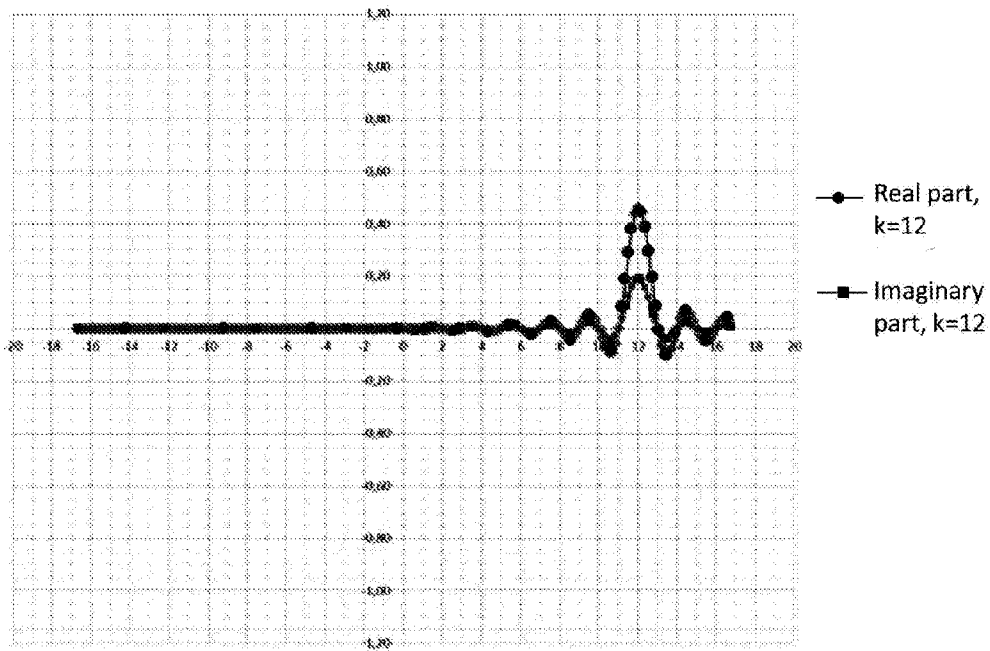


FIG. 11

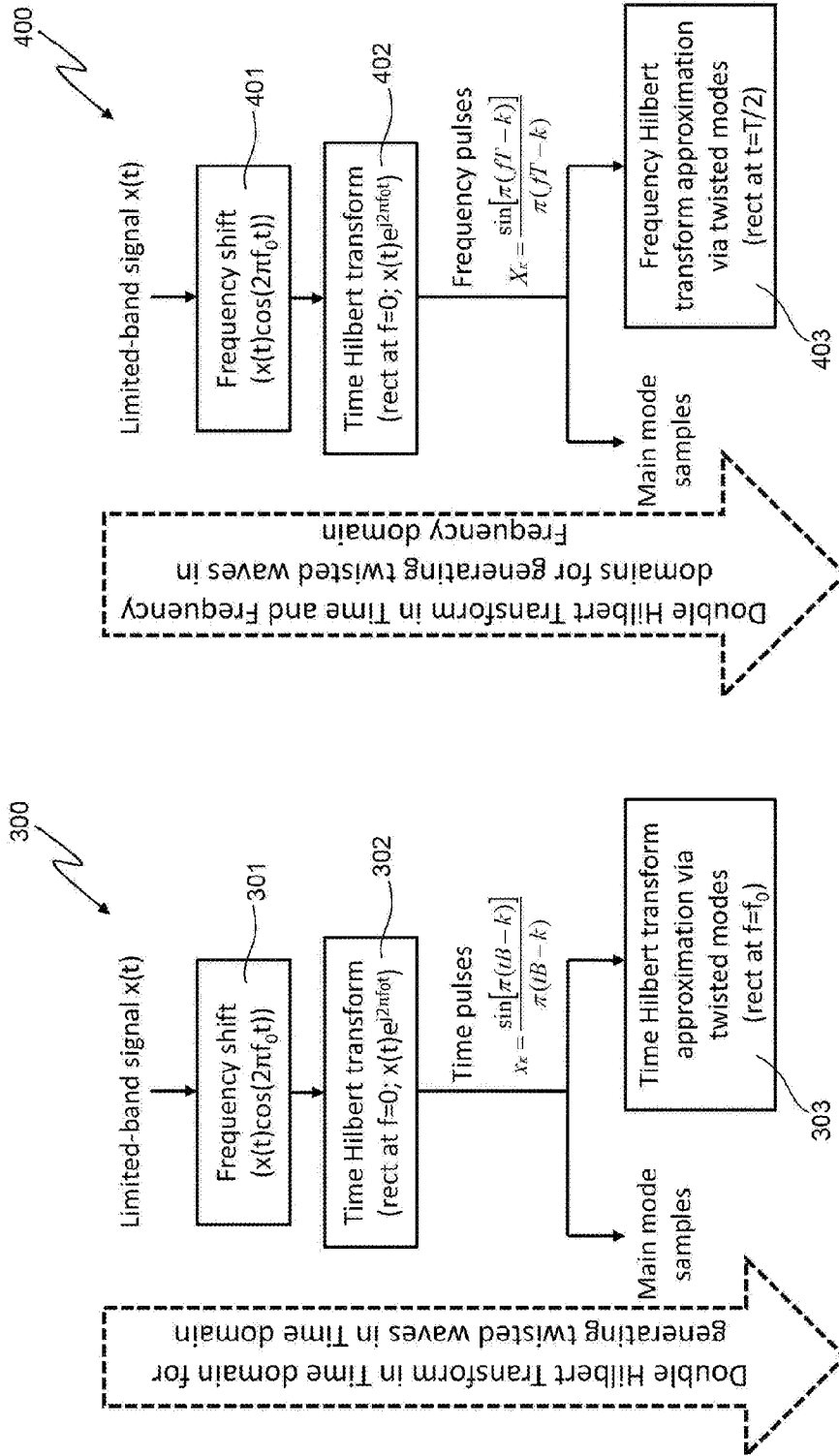


FIG. 10

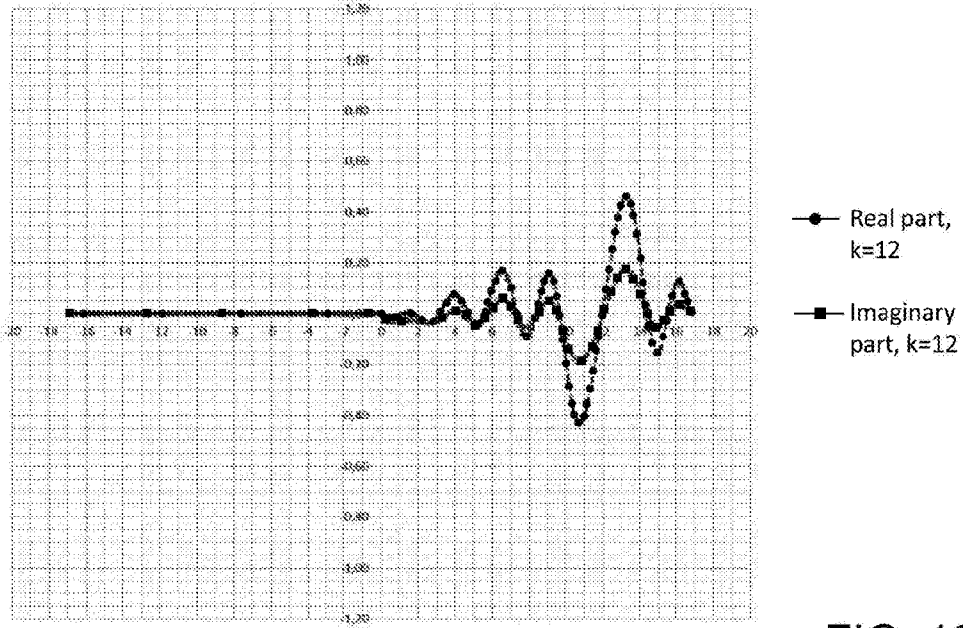


FIG. 12

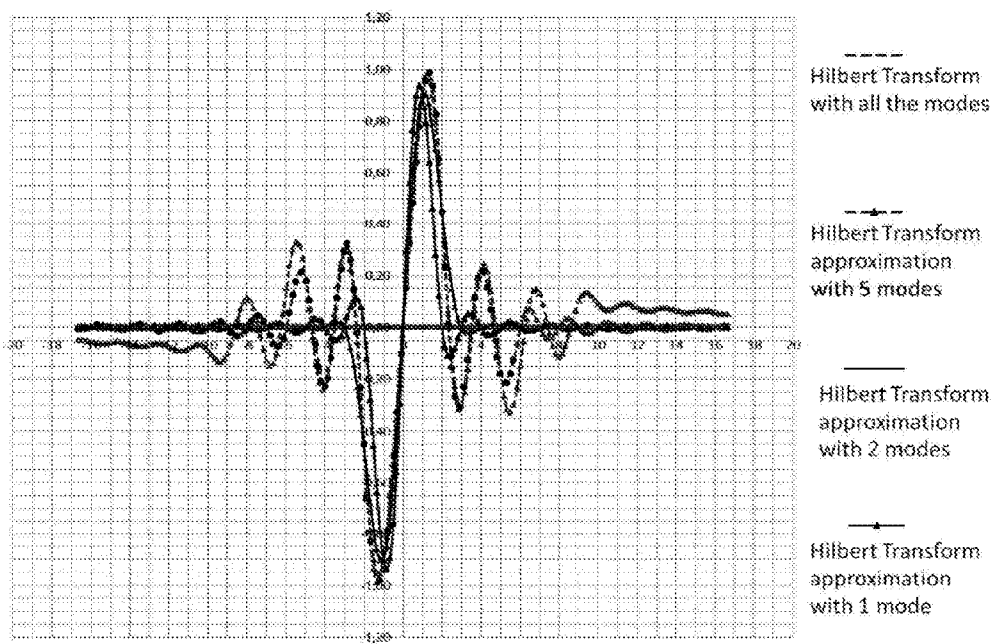
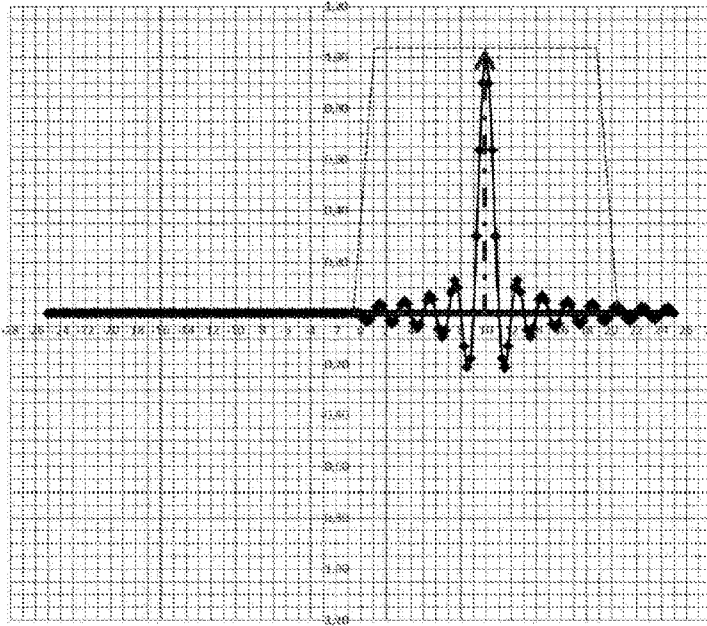
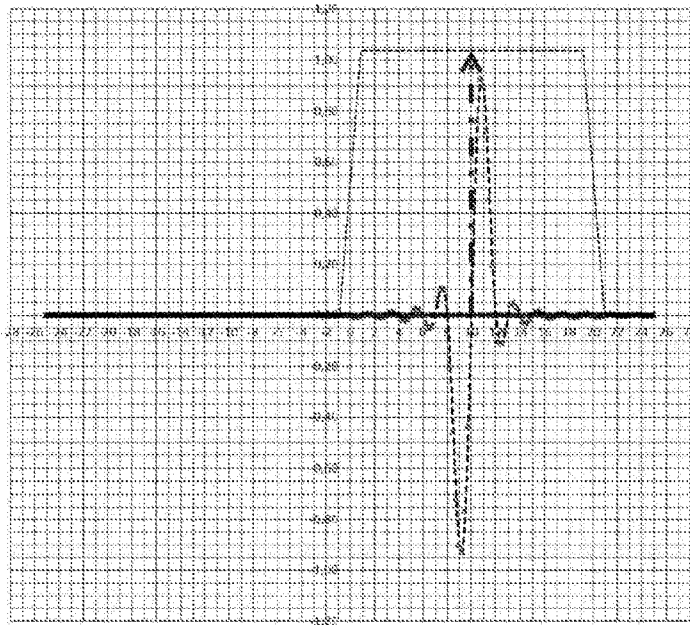


FIG. 13



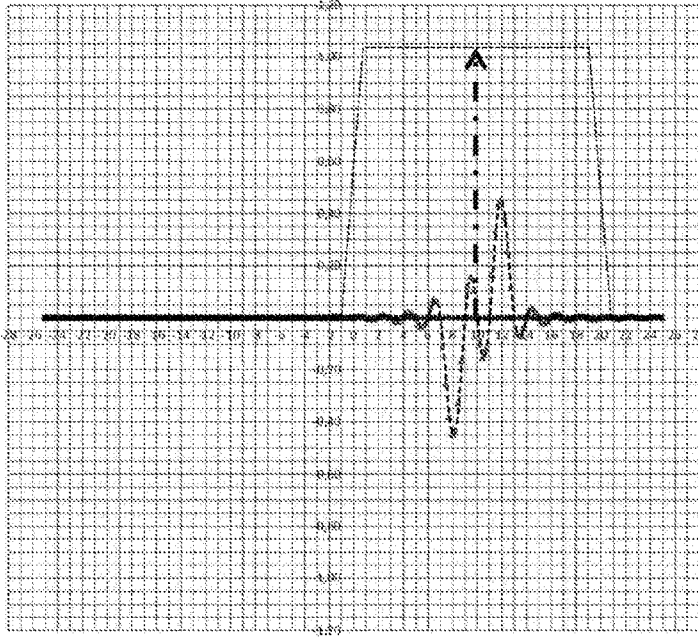
Mode 0 with reference frequency pulse positioned at the middle frequency $f_0=10/T$ (T being the length of the time window)

FIG. 14



Mode 1 with reference frequency pulses positioned at frequencies $f_9=9/T$ and $f_{11}=11/T$ (T being the length of the time window)

FIG. 15



Mode 2 with reference frequency pulses positioned at frequencies $f_8=8/T$ and $f_{12}=12/T$ (T being the length of the time window)

FIG. 16

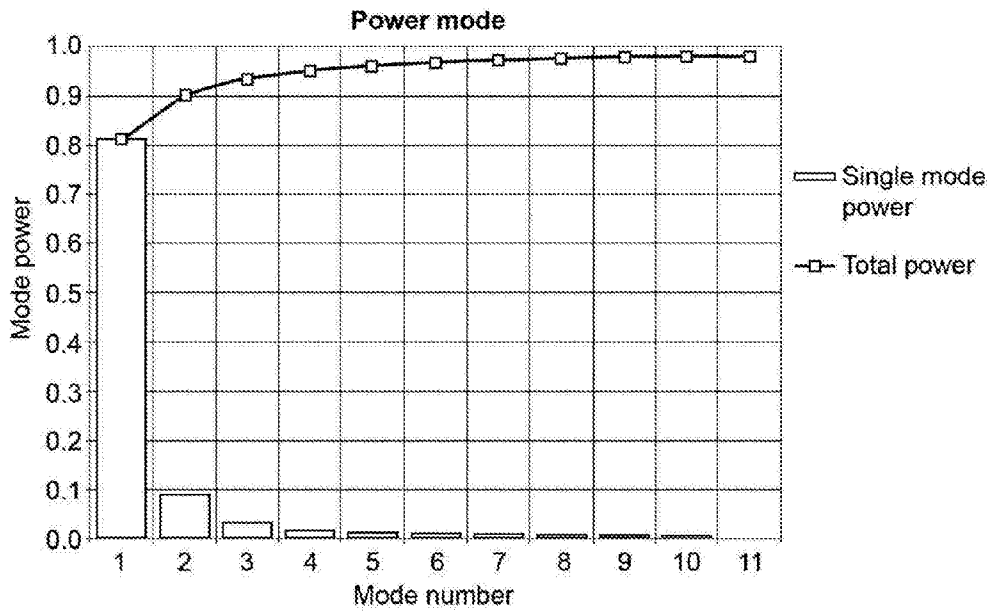


FIG. 17

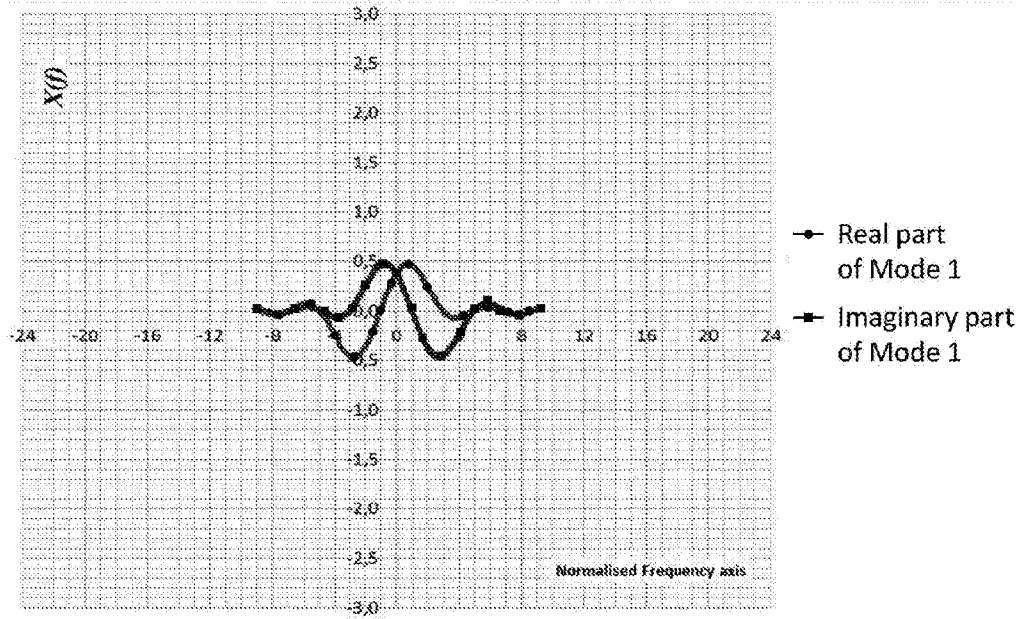


FIG. 18

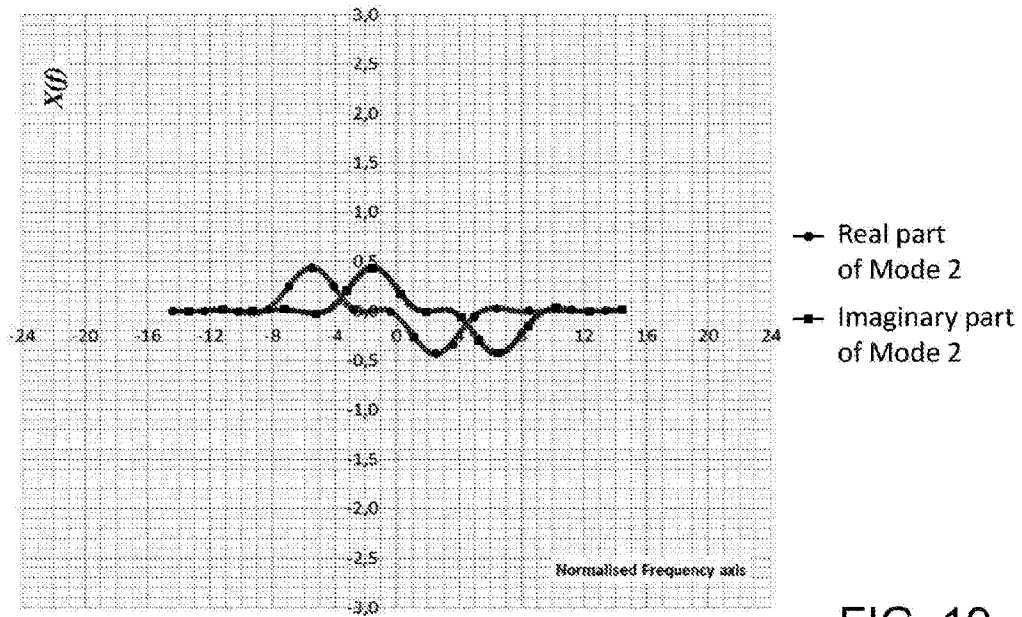


FIG. 19

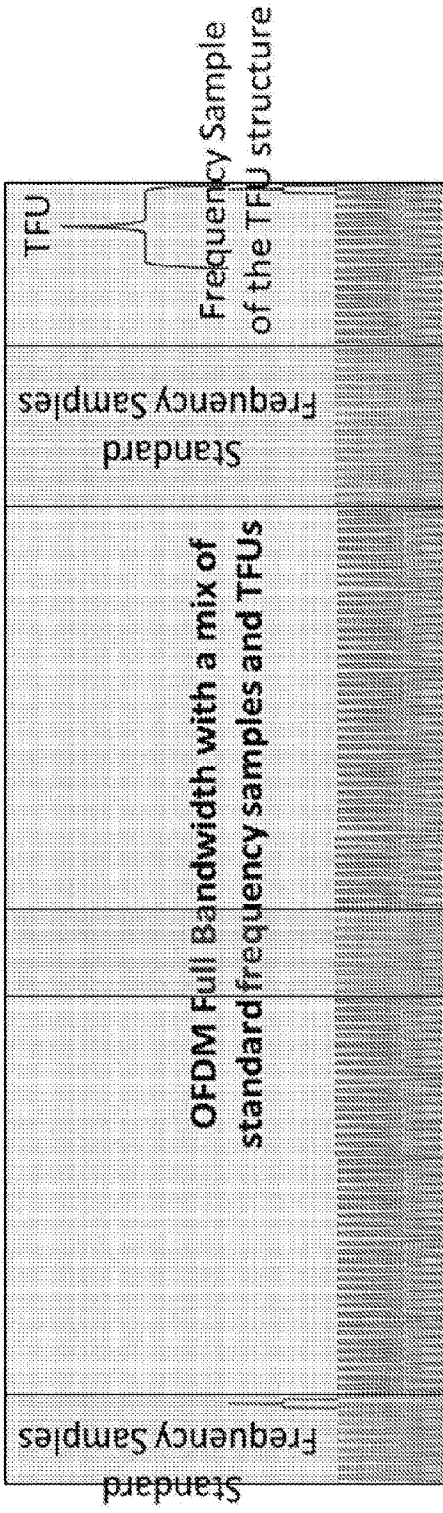


FIG. 20

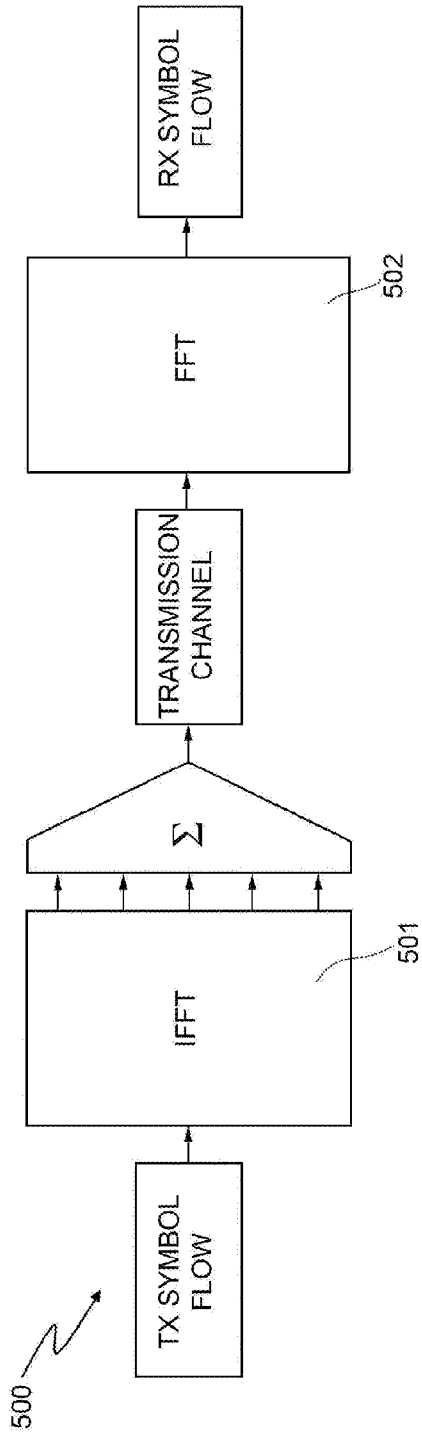


FIG. 22

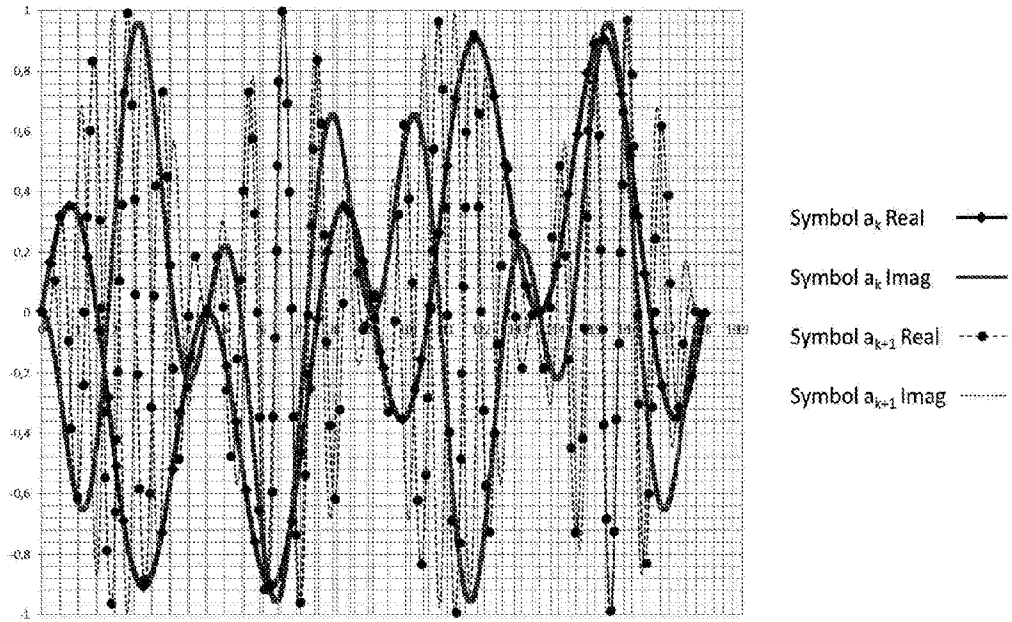


FIG. 24

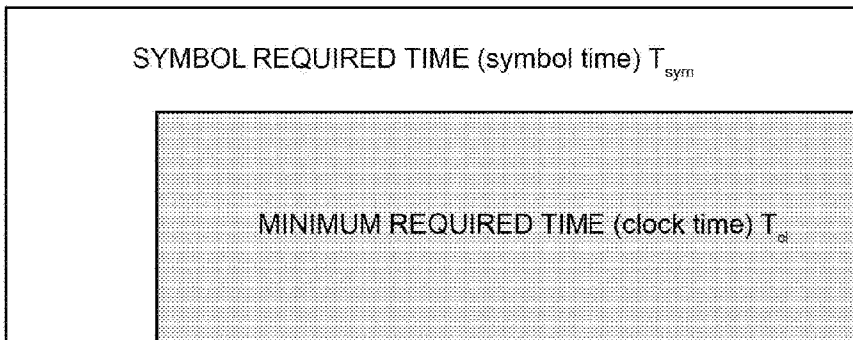


FIG. 25

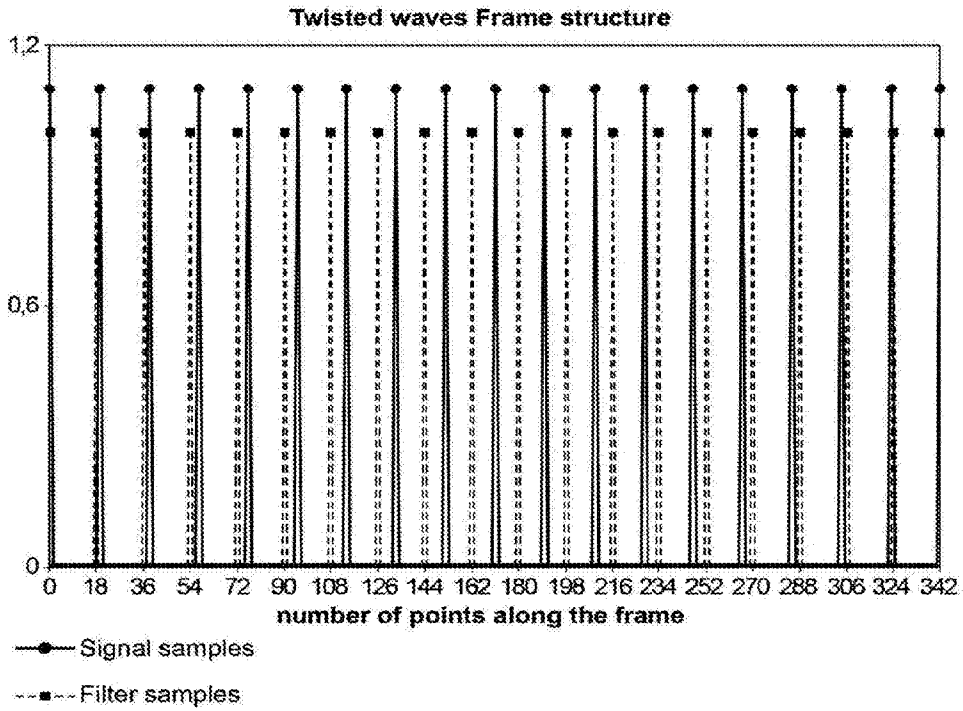


FIG. 26

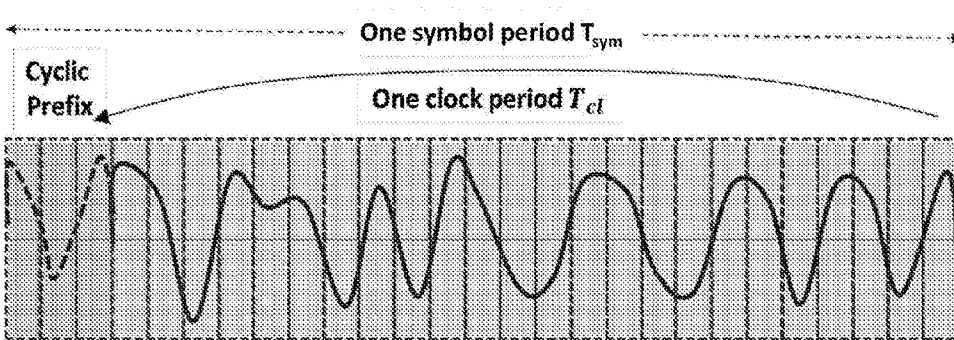


FIG. 27

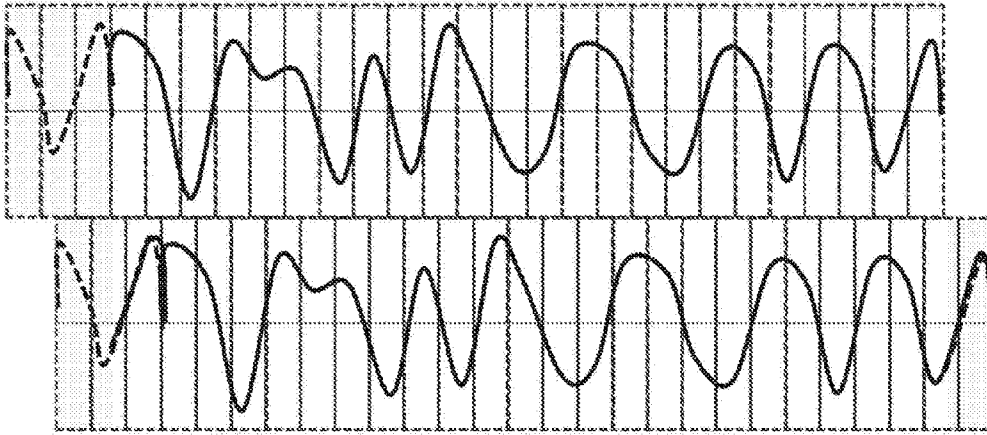
FFT processing window = T_{cl} 

FIG. 28

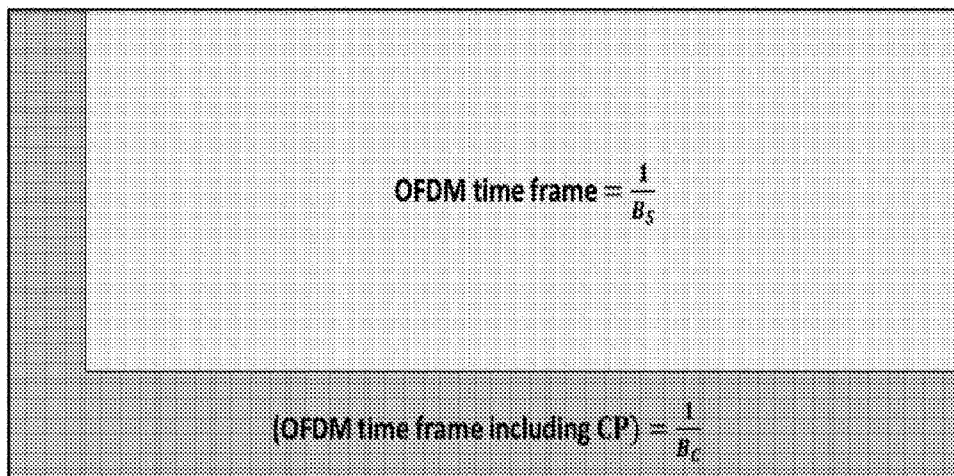


FIG. 29

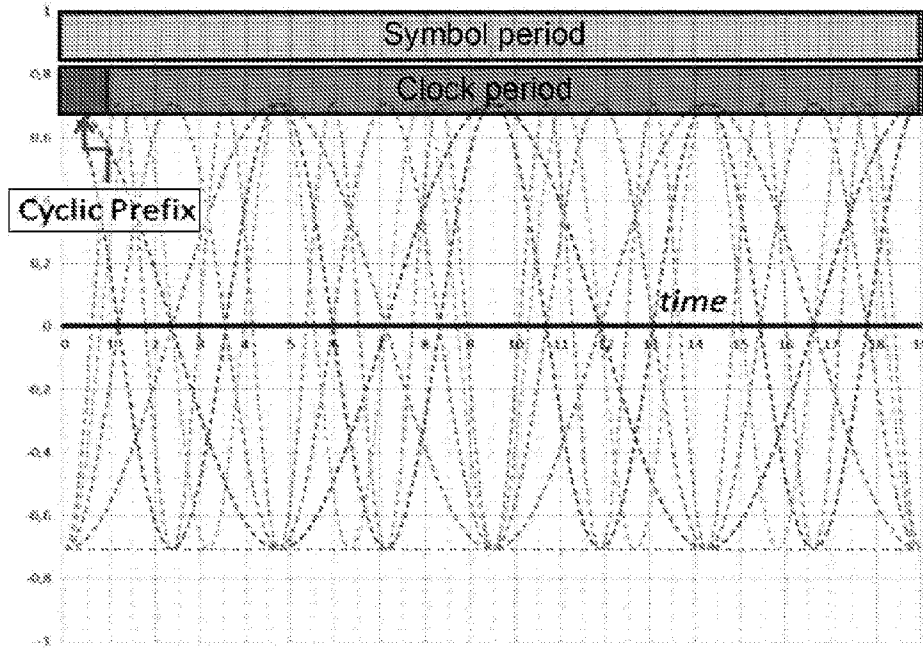


FIG. 30

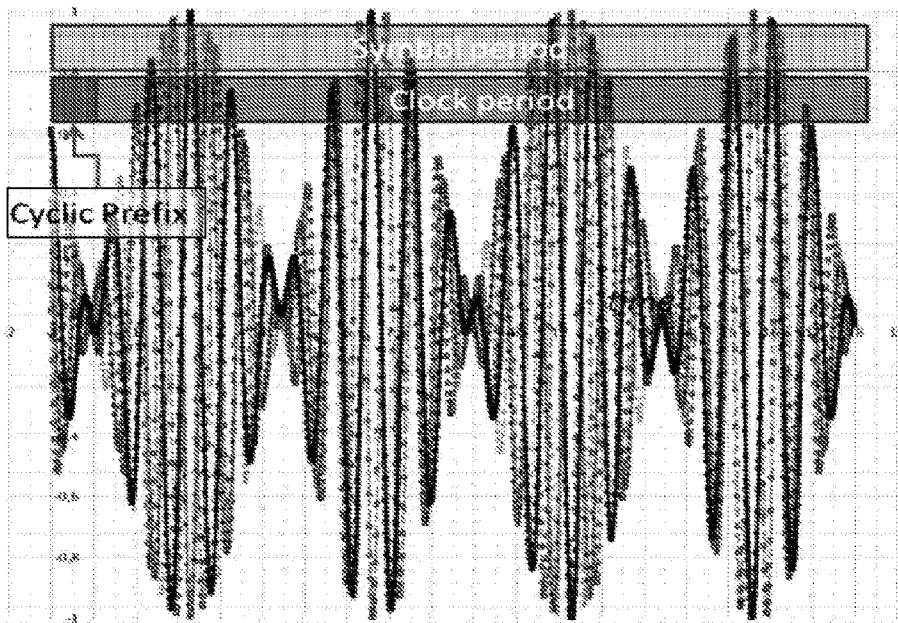


FIG. 31

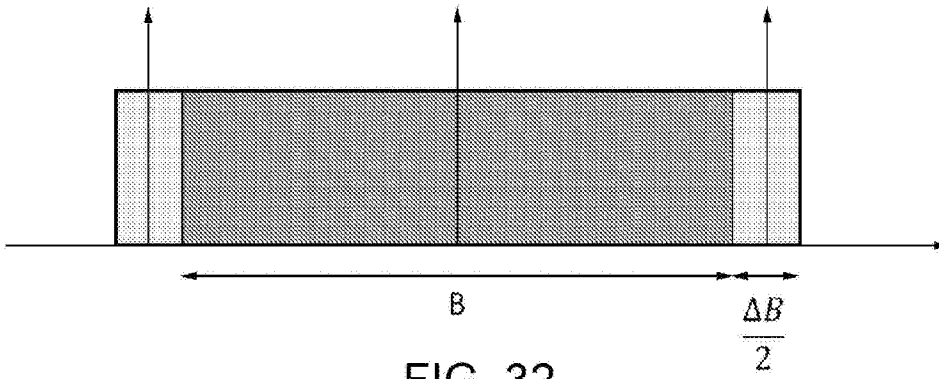


FIG. 32

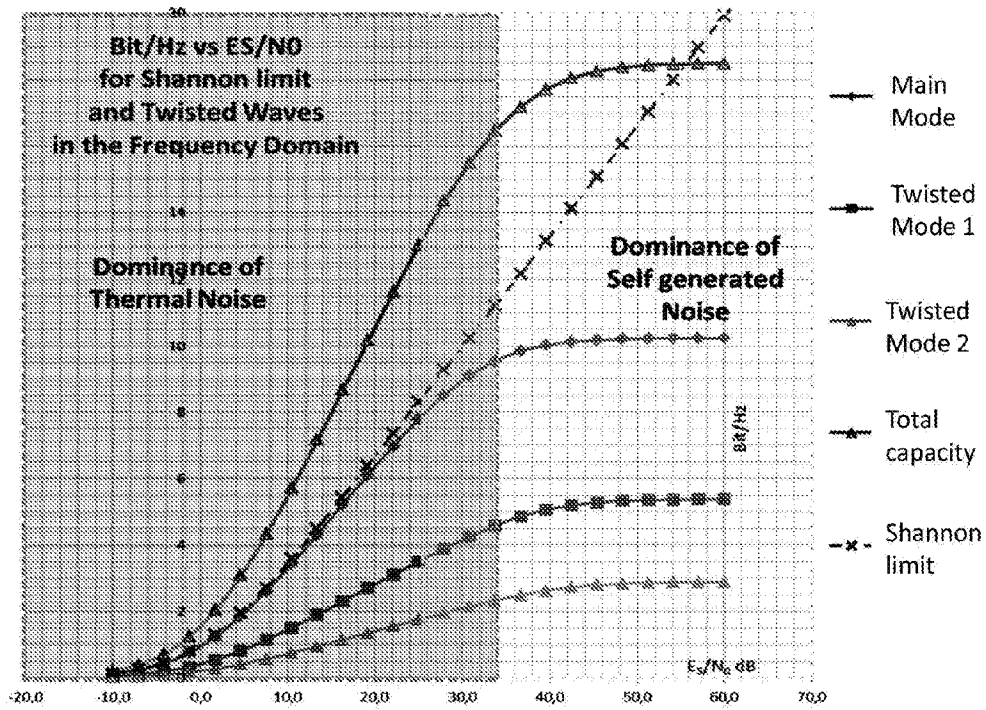


FIG. 34

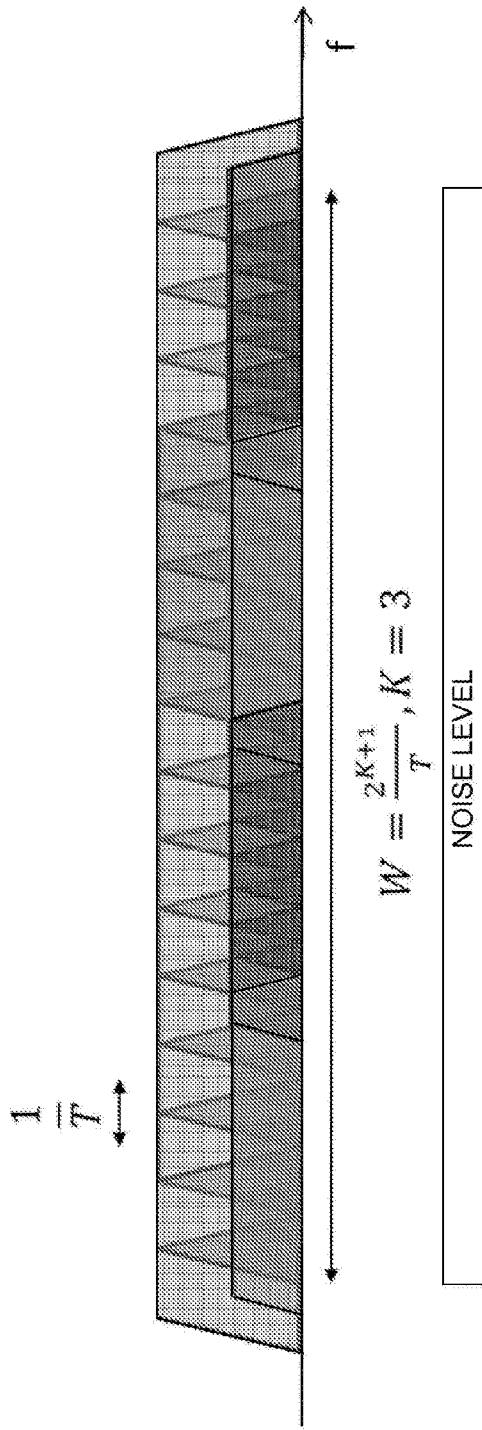


FIG. 33

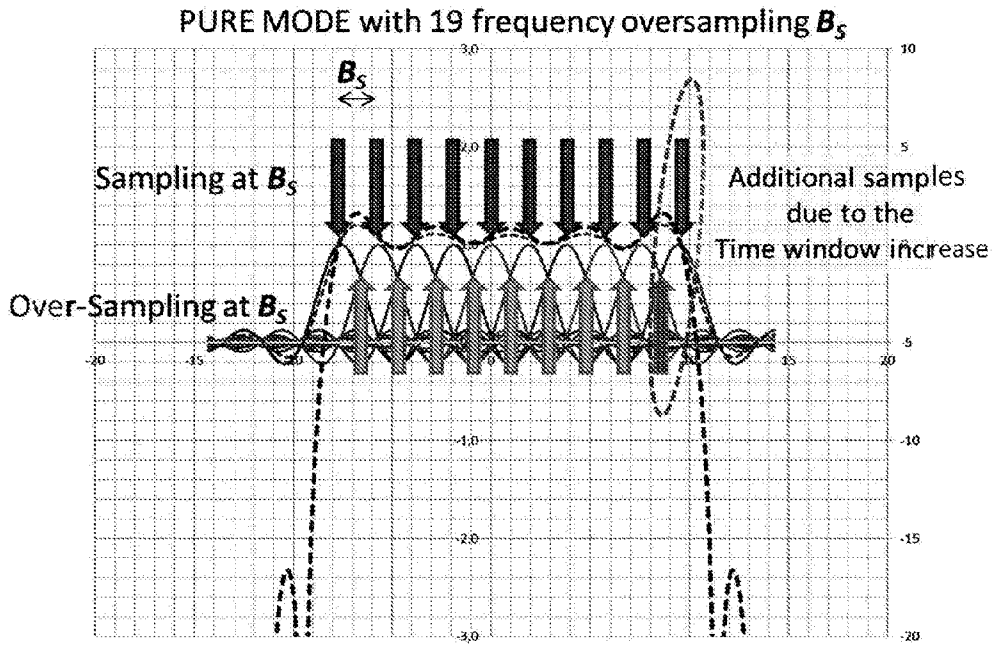


FIG. 35

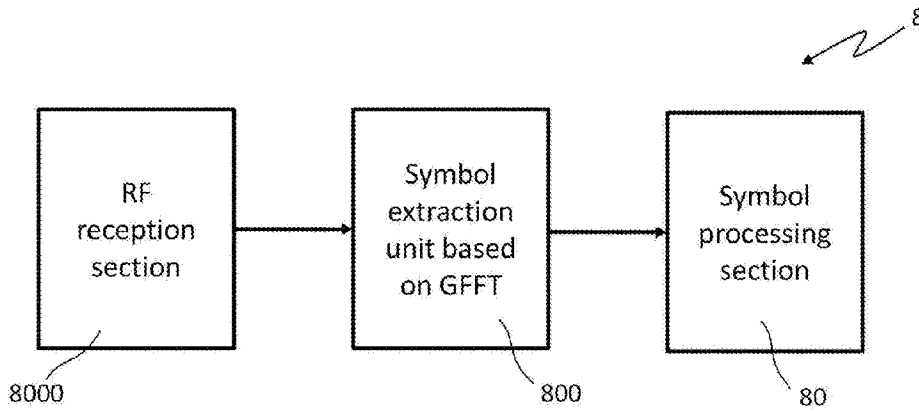


FIG. 37

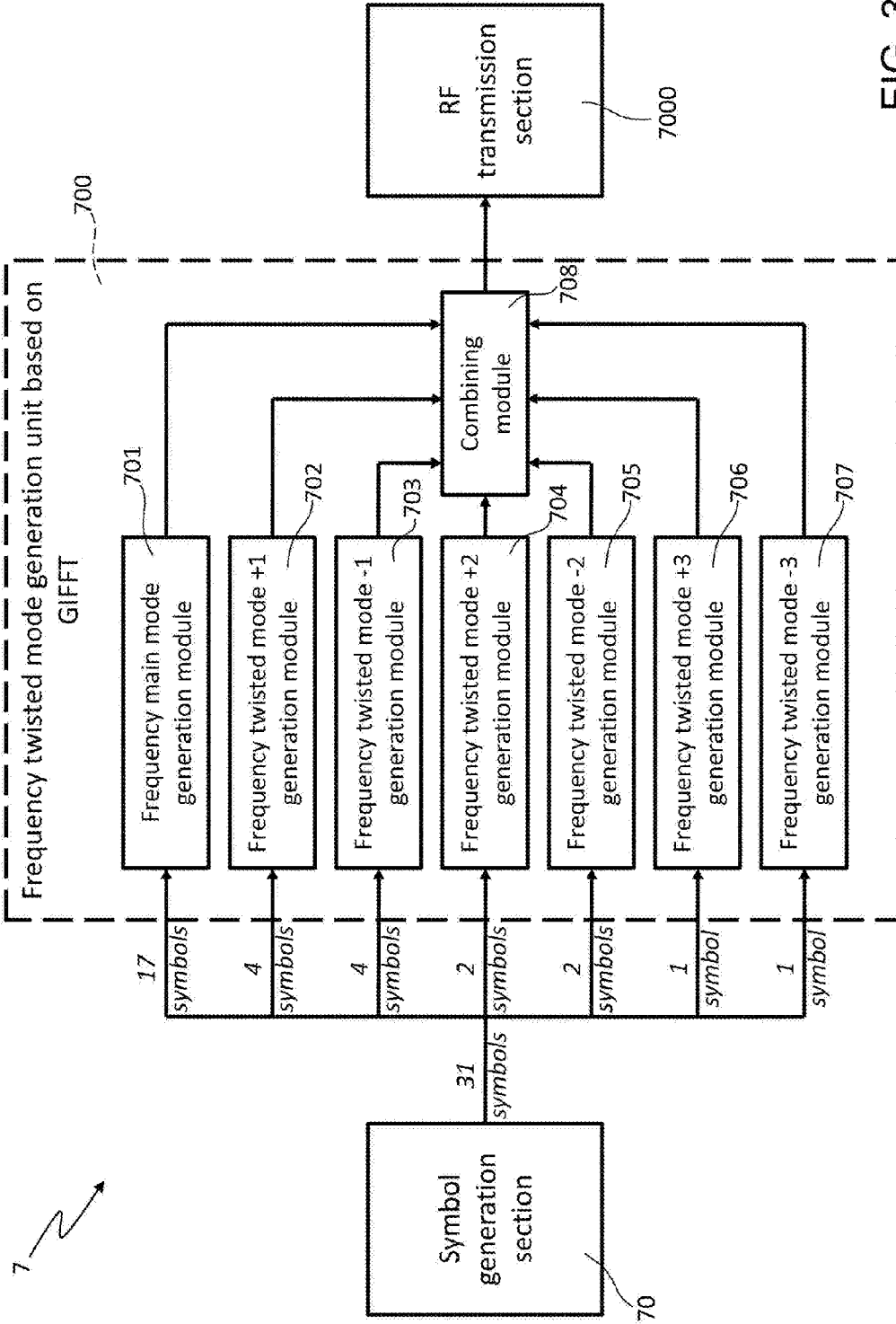


FIG. 36

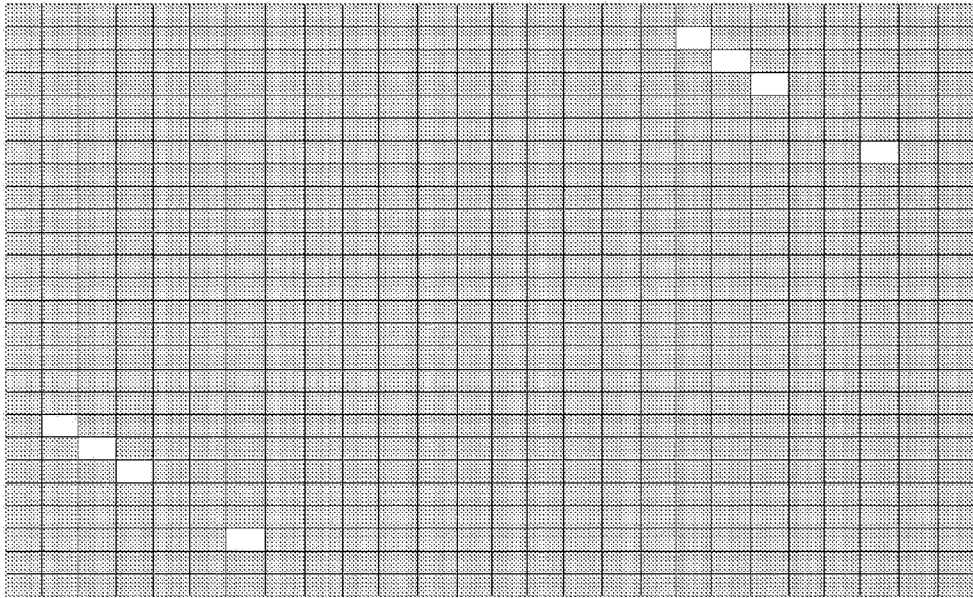


FIG. 38

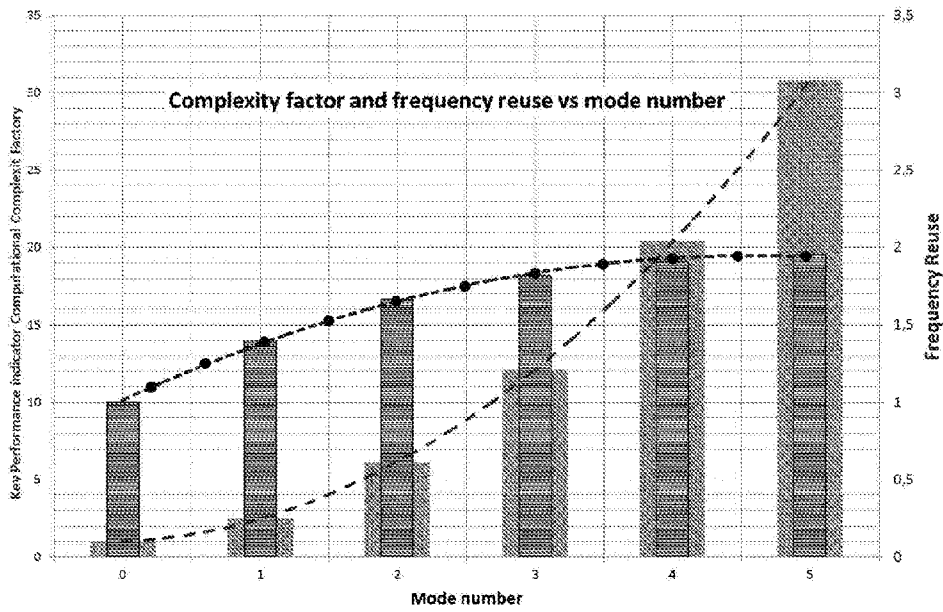


FIG. 40

Super frame including Frequency Twisted waves

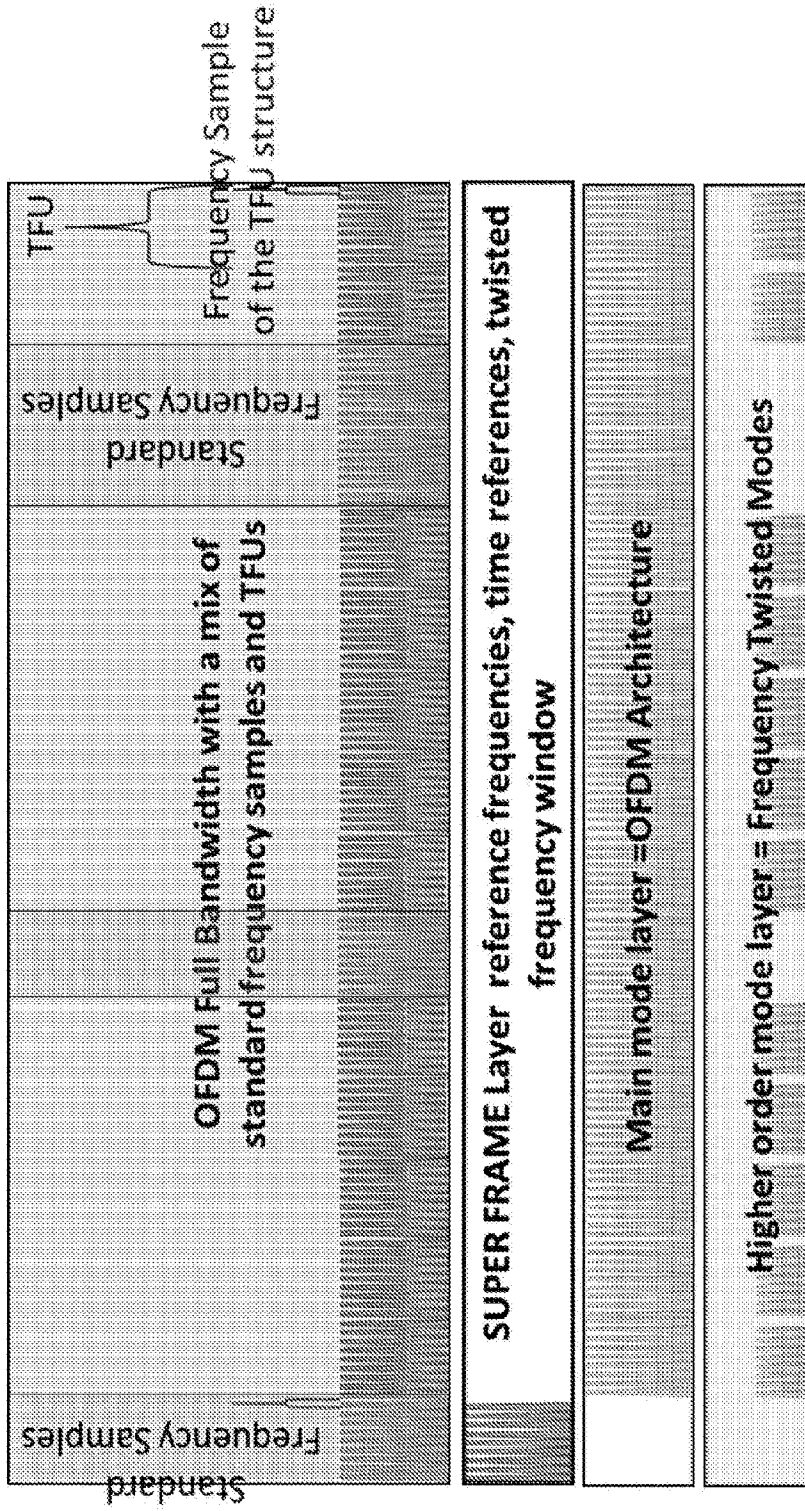
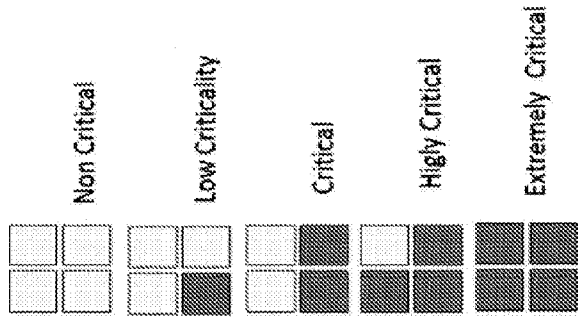


FIG. 39



	0	1	2	3	4	5	6	7	8	9	10
Flexibility in using OFDM modularity											
Complex Equation number	1	9	33	113	353	1,025	2,817	7,425	18,945	47,105	114,689
Implementation Criticality											
MODE NUMBER	0	1	2	3	4	5	6	7	8	9	10

FIG. 41