(19)

LE GOUVERNEMENT
DU GRAND-DUCHÉ DE LUXEMBOURG
Ministère de l'Économie

(11) N° de publication : **LU506246**

(12) **BREVET D'INVENTION** **B1**

(21) N° de dépôt: LU506246

(22) Date de dépôt: 29/01/2024

(51) Int. Cl.:
G01R 29/26, G01R 25/00, G01R 23/167

(30) Priorité:
19/07/2023 CN 202310886241.X

(43) Date de mise à disposition du public: 30/07/2024

(47) Date de délivrance: 30/07/2024

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(54) **METHOD FOR FULLY-DIGITAL PHASE NOISE MEASUREMENT WITH MILLIHERTZ-LEVEL FREQUENCY RESOLUTION.**

(57) The present disclosure relates to the field of phase noise measurement, and provides a method for fully-digital phase noise measurement with millihertz-level frequency resolution. The method includes: converting a measured signal into a digital signal, generating two digital signals with orthogonal phases and no phase truncation spur based on a frequency control word of the digital signal, mixing the two digital signals with the digital signal separately, and performing multi-segment decimation and filtering, phase unwrapping, frequency offset elimination, power spectrum estimation, and segmental splicing. The method uses only one analog-to-digital converter (ADC) to measure phase noise of a signal, which has a simple implementation principle and is easy to achieve a miniaturized system. The multi-segment decimation and filtering can reduce a data rate after mixing while filtering out a high-frequency component.

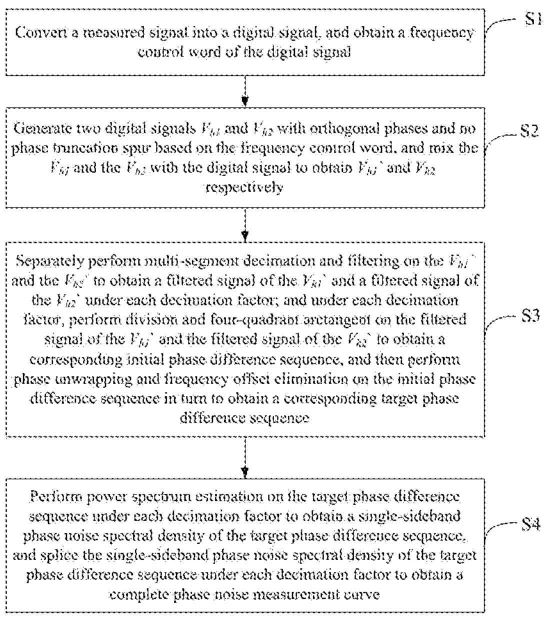

FIG. 1

DESCRIPTION

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METHOD FOR FULLY-DIGITAL PHASE NOISE MEASUREMENT WITH MILLIHERTZ-LEVEL FREQUENCY RESOLUTION

TECHNICAL FIELD

[0001] The present disclosure relates to the field of phase noise measurement, and more specifically, to a method for fully-digital phase noise measurement with millihertz-level frequency resolution.

BACKGROUND

[0002] In some special occasions, a signal in a 10k Hz to 500k Hz frequency band has a measurement requirement for millihertz-level frequency resolution. For example, in the "Tianqin Project", a plurality of key technologies need to be implemented in order to detect cosmic gravitational waves in a 0.001 Hz to 0.1 Hz frequency band. A capacitive displacement sensor is one of key components, and its sensing performance plays an important role in sensitivity of gravitational wave detection. In the capacitive displacement sensor, one carrier needs to be injected into a plurality of capacitive plates around a test mass (TM). If a carrier used for modulation has poor frequency stability, a millihertz-level low-frequency differential capacitive signal may be covered by phase noise of the carrier. Therefore, it is of great research significance to measure phase noise extremely close-in to the carrier (millihertz frequency offset).

[0003] However, no phase noise instruments currently on the market can achieve millihertz-level measurement for phase noise extremely close-in to a signal in the 10k Hz to 500k Hz frequency band.

SUMMARY

[0004] In view of the above defects or improvement requirements in the prior art, the present disclosure provides a fully-digital phase noise measurement method with millihertz-level frequency resolution, to resolve a technical problem that an existing phase noise instrument cannot achieve millihertz-level measurement for phase noise extremely close-in to a signal in a 10k Hz to 500k Hz.

[0005] In order to achieve the above objective, according to a first aspect of the present disclosure, a method for fully-digital phase noise measurement with millihertz-level measurement frequency resolution is provided, including:

[0006] S1: converting a measured signal into a digital signal, and obtaining a frequency control word of the digital signal;

[0007] S2: generating two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur based on the frequency control word, and mixing the V_{h1} and the V_{h2} with the digital signal to obtain V_{h1}\` and $V_{h2}$\` respectively;

[0008] S3: separately performing multi-segment decimation and filtering on the V_{h1}\` and the $V_{h2}$\` to obtain a filtered signal of the V_{h1}\` and a filtered signal of the $V_{h2}$\` under each decimation factor; and under each decimation factor, performing division and four-quadrant arctangent on the filtered signal of the V_{h1}\` and the filtered signal of the $V_{h2}$\` to obtain a corresponding initial phase difference sequence, and then performing phase unwrapping and frequency offset elimination on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence; and

[0009] S4: performing power spectrum estimation on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and splicing the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor to obtain a complete phase noise measurement curve.

[0010] According to a second aspect of the present disclosure, a device for fully-digital phase noise measurement with millihertz-level measurement frequency resolution is provided, including:

[0011] a first processing module configured to convert a measured signal into a digital signal, and obtain a frequency control word of the digital signal;

[0012] a second processing module configured to generate two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur based on the frequency control word, and mix the V_{h1} and the V_{h2} with the digital signal to obtain V_{h1}` and $V_{h2}$` respectively;

[0013] a third processing module configured to: separately perform multi-segment decimation and filtering on the V_{h1}` and the $V_{h2}$` to obtain a filtered signal of the V_{h1}` and a filtered signal of the $V_{h2}$` under each decimation factor; and under each decimation factor, perform division and four-quadrant arctangent on the filtered signal of the V_{h1}` and the filtered signal of the $V_{h2}$` to obtain a corresponding initial phase difference sequence, and then perform phase unwrapping and frequency offset elimination on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence; and

[0014] a fourth processing module configured to perform power spectrum estimation on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and splice the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor to obtain a complete phase noise measurement curve.

[0015] According to a third aspect of the present disclosure, a system for fully-digital phase noise measurement with millihertz-level measurement frequency resolution is provided, including a computer-readable storage medium and a processor, where

[0016] the computer-readable storage medium is configured to store an executable instruction; and

[0017] the processor is configured to read the executable instruction stored in the computer-readable storage medium to execute the method described in the first aspect.

[0018] According to a fourth aspect, a computer-readable storage medium is provided, where the computer-readable storage medium stores a computer instruction, and the computer instruction is configured to enable a processor to execute the method in the first aspect.

[0019] In general, compared with the prior art, the above technical solutions conceived by the present disclosure can achieve following beneficial effects:

[0020] 1. The method provided in the present disclosure uses only one analog-to-digital converter (ADC) to measure phase noise of a signal, which has a simple implementation principle and is easy to achieve a miniaturized measurement system, thereby facilitating a site test. Multi-segment decimation and filtering can reduce a data rate after mixing while filtering out a high-frequency component. This reduces a quantity of operation points required for power spectrum estimation, thus achieving measurement with millimeter-level frequency resolution for phase noise extremely close-in to the signal. For example, phase noise extremely close-in to a signal in a 10k Hz to 500k Hz frequency band can be measured, making up for a deficiency that a commercial instrument cannot achieve the millimeter-level frequency resolution for the signal in the frequency band.

[0021] 2. The method provided in the present disclosure can generate two mutually orthogonal signals with no phase truncation spur by using a low-spurious orthogonal digital signal synthesis technology, which can reduce measurement interference caused by spurs.

[0022] 3. The method provided in the present disclosure can reduce an impact caused by a slight frequency difference between a measured signal and a synthetic signal by using phase unwrapping and frequency offset elimination algorithms.

BRIEF DESCRIPTION OF THE DRAWINGS

[0023] FIG. 1 is a flowchart of a method for fully-digital phase noise measurement with millihertz-level frequency resolution according to an embodiment of the present disclosure;

[0024] FIG. 2 is a schematic diagram of fully-digital phase noise measurement according to according to an embodiment of the present disclosure;

[0025] FIG. 3 is a schematic diagram of equal-precision frequency measurement;

[0026] FIG. 4 is a schematic diagram of a low-spurious orthogonal digital signal synthesis method;

[0027] FIG. 5 is a schematic diagram of comparing a single-sideband phase noise spectral density of a 500 kHz signal according to an embodiment of the present disclosure; and

[0028] FIG. 6 is a schematic diagram of splicing a single-sideband phase noise spectral density of a 500 kHz signal according to an embodiment of the present disclosure.

DETAILED DESCRIPTION OF THE EMBODIMENTS

[0029] In order to make the objectives, technical solutions, and advantages of the present disclosure clearer, the present disclosure is further described in detail below with reference to the accompanying drawings and embodiments. It should be understood that the specific embodiments described herein are only intended to explain the present disclosure and are not intended to limit the present disclosure. Further, the technical features involved in the various implementations of the present disclosure described below may be combined with each other as long as they do not constitute a conflict with each other.

[0030] In a plurality of phase noise measurement techniques, a method of directly using a spectrum analyzer is affected by background noise of the spectrum analyzer itself and is not suitable for measuring a signal with high frequency stability. A beat method is difficult to achieve measurement with millihertz-level frequency resolution extremely close-in to the signal due to a limitation of a counter. A frequency discrimination method has poor sensitivity when measuring near-end phase noise.

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[0031] A phase discrimination method has many physical devices and a large-volume phase discrimination system, and needs to perform compensation for a loss of a low frequency. A fully-digital method is free from the limitations above, and has advantages of a simple implementation principle, easy implementation of a miniaturized system, and convenient field testing.

[0032] In view of this, an embodiment of the present disclosure provides a method for fully-digital phase noise measurement with millihertz-level signal frequency resolution. As shown in FIG. 1, the method includes following steps.

[0033] S1: A measured signal is converted into a digital signal, and a frequency control word of the digital signal is obtained.

[0034] Specifically, a to-be-measured signal (for example, a sine wave) is sampled by an ADC and converted into a digital signal, and a frequency control word of the digital signal is obtained by using an equal-precision frequency measurement method, a direct measurement method, or a periodic measurement method.

[0035] When equal-precision frequency measurement is adopted, the frequency control word of the to-be-measured signal can be expressed by count values of a standard signal and the to-be-measured signal.

[0036] S2: Two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur are generated based on the frequency control word, and are mixed with the digital signal to obtain V_{h1}` and $V_{h2}$` respectively.

[0037] Preferably, in the step S2, the V_{h1} and the V_{h2} are generated by using a phase truncation–free frequency synthesis method, a ROM table compression frequency synthesis method, or a coordinate rotation digital computer (CORDIC) frequency synthesis method.

[0038] Specifically, based on the frequency control word generated through the equal-precision frequency measurement, a low-spurious orthogonal digital signal synthesis technology can be used to generate two low-spurious digital signals that have orthogonal phases and whose frequencies are similar to a frequency of the to-be-measured signal, which can reduce interference of a spur on phase noise measurement of the to-be-measured signal.

[0039] S3: Multi-segment decimation and filtering is separately performed on the V_{h1}\` and the $V_{h2}$\` to obtain a filtered signal of the V_{h1}\` and a filtered signal of the $V_{h2}$\` under each decimation factor; and under each decimation factor, division and four-quadrant arctangent are performed on the filtered signal of the V_{h1}\` and the filtered signal of the $V_{h2}$\` to obtain a corresponding initial phase difference sequence, and then phase unwrapping and frequency offset elimination are performed on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence.

[0040] Specifically, the two signals obtained through low-spurious orthogonal digital signal synthesis are separately mixed with the to-be-measured signal in a digital domain. Through the multi-segment decimation and filtering, a high-frequency component after the mixing can be filtered out, and a low-frequency signal after the mixing can be obtained. In addition, multi-segment decimation can achieve multi-resolution measurement, which can reduce a quantity of data points required to achieve millihertz-level frequency resolution.

[0041] S4: Power spectrum estimation is performed on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor is spliced to obtain a complete phase noise measurement curve.

[0042] Specifically, the division and the four-quadrant arctangent are performed on the two signals obtained after the multi-segment decimation and filtering to obtain the initial phase difference sequence. The phase unwrapping is performed on the phase difference sequence to recover a jump of the phase difference sequence at $\pm\pi$.

[0043] The digital signal generated by the low-spurious orthogonal digital signal synthesis technology has a slight frequency difference from the measured signal. In order to eliminate a measurement error caused by the frequency difference, the frequency offset elimination needs to be performed on a phase obtained after the phase unwrapping sequence.

[0044] Preferably, the step S4 includes: segmenting a frequency range offset from the carrier frequency, and selecting different sampling frequencies for different biased carrier frequency bands to achieve the power spectrum estimation for the frequency bands.

[0045] The method provided in the present disclosure will be explained in detail below with reference to FIG. 2.

[0046] As shown in FIG. 2, the method for fully-digital phase noise measurement with millihertz-level frequency resolution provided in the present disclosure mainly includes following steps: collection by an ADC, equal-precision frequency measurement, low-spurious orthogonal digital signal synthesis, multi-segment decimation and filtering, four-quadrant arctangent, phase unwrapping, frequency offset elimination, power spectrum estimation, and segmental splicing.

[0047] Step 1: A measured signal is converted into a digital signal by using the ADC.

[0048] Step 2: A frequency control word of the digital signal is obtained by using an equal-precision frequency measurement method, to provide a variable frequency reference for next-stage low-spurious orthogonal digital signal synthesis.

[0049] Step 3: Two digital signals with a phase difference of $\pi/2$ and no phase truncation spur are generated by using a low-spurious orthogonal digital signal synthesis method based on the frequency control word provided in the step 2. The two signals will be mixed with the measured signal in a digital domain.

[0050] The low-spurious orthogonal digital signal synthesis technology in the step 3 is a synthesis technology that can generate two digital signals with orthogonal phases and no phase truncation spur.

[0051] Step 4: The multi-segment decimation and filtering is performed on a mixed signal to filter out a high-frequency component in the mixed signal. Decimation can reduce a data rate after the mixing, and can reduce a quantity of operation points required for the power spectrum estimation in a step 8, thus achieving phase noise measurement with millimeter-level frequency resolution.

[0052] Step 5: After a quotient of data output from upper and lower channels (in other words, I and Q channels) is obtained, the four-quadrant arctangent is performed to obtain an initial phase difference sequence.

[0053] Step 6: The phase unwrapping is performed on the initial phase difference sequence to recover a jump of the phase difference sequence at $\pm\pi$.

[0054] Step 7: A frequency offset elimination algorithm is applied to a phase difference sequence obtained after the phase unwrapping to eliminate an impact from a slight frequency difference, such that a low-error phase difference sequence is obtained.

[0055] The phase unwrapping and the frequency offset elimination that are performed for the initial phase difference sequence the steps 6 and 7 work together to reduce a measurement error caused by the slight frequency difference.

[0056] Step 8: The power spectrum estimation is performed on a phase difference sequence obtained after the frequency offset elimination to obtain a phase fluctuation spectral density, and then a single-sideband phase noise spectral density of the measured signal is further obtained. A single-sideband phase noise spectral density of each segment in a multi-segment decimation filter can be spliced to obtain a complete phase noise measurement curve.

[0057] To achieve a trade-off between a quantity of Fast Fourier Transform (FFT) points and resolution (higher resolution leads to a larger quantity of FFT points required), a frequency range segmentation method is adopted, and different sampling frequencies are selected for different frequency bands (a higher sampling multiple leads to a lower sampling rate) to achieve the power spectrum estimation for the frequency bands. For example, for an extremely close frequency band (a frequency band close to an origin of a coordinate system) that requires millisecond-level resolution (frequency band I shown in FIG. 5), a decimation factor is high, a sampling rate is low, and the quantity of FFT points is small, making it easy to implement.

[0058] It can be understood that when the multi-segment decimation and filtering is performed in the step 4, due to a plurality of decimation factors, the steps 5 to 7 are separately performed under each decimation factor, that is, after the steps 5 to 7, a phase difference sequence obtained after the frequency offset elimination under each decimation factor is obtained. For example, when the decimation factors are 16, 64, 256, and 2048, phase difference sequences that are obtained after the frequency offset elimination and respectively correspond to the decimation factors 16, 64, 256, and 2048 are obtained in the step 7. Accordingly, in the step 8, single-sideband phase noise spectral densities of phase difference sequences obtained after the frequency offset elimination under the decimation factors are spliced.

[0059] For example, when the measured signal is a sine wave, the measured signal sampled by the ADC is as shown in formula (1): LU506246

$$V(n) = (A_0 + \varepsilon(n)) \times \sin(w_0 n + \Delta\varphi(n)) \quad (1)$$

[0060] In the above formula, $w_0 = 2\pi f_0$. A_0, w_0, and f_0 respectively represent an ideal amplitude, angular frequency, and frequency of the signal, and $\varepsilon(n)$ and $\Delta\varphi(n)$ respectively represent a random amplitude fluctuation and a phase fluctuation of the signal.

[0061] A principle of the equal-precision frequency measurement is shown in FIG. 3. Two counters are used to count the measured signal and a standard signal respectively. When a rising edge of a first measured signal is detected after a preset gate is a high level, an actual gate is enabled and the two counters start counting. After a period of time, the preset gate is lowered. When the measured signal is detected to rise, the actual gate is set to a low level, and the counters stop counting. Count values of the standard signal and the measured signal are respectively denoted as N_0 and N_x.

[0062] A principle of the low-spurious orthogonal digital signal synthesis is shown in FIG. 4. N represents a bit width of a phase accumulator, and the frequency control word K is as shown in formula (2):

$$K = 2^N N_x / N_0 \quad (2)$$

[0063] A reference frequency source uses an internal clock with a frequency of f_{clk}. Therefore, a relationship between generated signal frequency f_x and the f_{clk} is as follows:

$$f_x = f_{clk} \frac{K}{2^N}$$

[0064] The two signals generated by the low-spurious orthogonal digital signal synthesis technology are respectively as shown in formulas (3) and (4).

$$V_{h1}(n) = A_{h1} \sin\left(w_{h1} n + \varphi_{h1} + \Delta\varphi_{h1}(n)\right) \quad (3)$$

$$V_{h2}(n) = A_{h2} \cos\left(w_{h2} n + \varphi_{h2} + \Delta\varphi_{h2}(n)\right) \quad (4)$$

[0065] Ideally, amplitudes, angular frequencies, initial phases, and random phase fluctuations of the two signals satisfy $A_{h1} = A_{h2}$, $w_{h1} = w_{h2}$, $\varphi_{h1} = \varphi_{h2}$, and $\Delta\varphi_{h1}(n) = \Delta\varphi_{h2}(n)$ respectively.

[0066] After the mixing and the decimation and filtering, the high-frequency component in the mixing can be filtered out to obtain signals I and Q as shown in formulas (5) and (6). LU506246

$$I(n)=\left(A_0+\varepsilon(n)\right)A_{h1}\cos\left((w_0-w_{h1})n-\varphi_{h1}+\varphi_0-\Delta\varphi_{h1}(n)+\Delta\varphi(n)\right) \quad (5)$$

$$Q(n)=\left(A_0+\varepsilon(n)\right)A_{h2}\sin\left((w_0-w_{h2})n-\varphi_{h2}+\varphi_0-\Delta\varphi_{h2}(n)+\Delta\varphi(n)\right) \quad (6)$$

[0067] The division and the four-quadrant arctangent are performed on the $Q(n)$ and the $I(n)$ to obtain initial phase difference sequence $\theta(n)$ between the measured signal and each of the two orthogonal signals, as shown in formula (7).

$$\theta(n)=(w_0-w_{h2})n-\varphi_{h2}+\varphi_0-\Delta\varphi_{h2}(n)+\Delta\varphi(n) \quad (7)$$

[0068] There is a slight frequency difference between each of the two digital signals $V_{h1}(n)$ and $V_{h2}(n)$ generated by a numerically controlled oscillator (NCO) and the measured signal, namely, $w_0 \neq w_{h2}$. In addition, because the four-quadrant arctangent limits the $\theta(n)$ to be within $[-\pi,\pi]$, the $\theta(n)$ is in a sawtooth shape, which will distort phase noise when the power spectrum estimation is directly performed on the $\theta(n)$.

[0069] Therefore, before the frequency offset elimination is performed on the $\theta(n)$, it is necessary to perform the phase unwrapping on the $\theta(n)$. Specific steps are as follows:

[0070] 1) Starting from a second phase difference sequence, a phase difference between current data and previous data is calculated.

[0071] 2) If the phase difference is greater than π or less than –π, phase difference data between the point and each piece of subsequent data is added or subtracted by 2π.

[0072] 3) The step 2) is repeated until all data is traversed.

[0073] A simple linear regression model is used to fit a linear trend error after the phase unwrapping.

[0074] A fitted linear error is subtracted from the phase difference sequence after the phase unwrapping to obtain the phase difference sequence obtained after the frequency offset elimination. The power spectrum estimation is performed on the phase difference sequence, and the single-sideband phase noise spectral density is calculated.

[0075] Phase noise of a 500 kHz signal is measured by using the method provided in the present disclosure. In the multi-segment decimation and filtering, a four-segment decimation filter with decimation factors of 16, 64, 256, and 2048 is used. A quantity of operation points, frequency resolution, and a splicing range for each segment are shown in Table 1. For each frequency band, data with different sampling frequencies (the sampling multiple is inversely proportional to the sampling frequency, and an original sampling frequency/the decimation factor= current sampling frequency) are selected to achieve the power spectrum estimation. For example, for a frequency band with a frequency range of 10k Hz to 100k Hz offset from the carrier frequency, data with a decimation factor of 16 is selected, and 2^{12} is used as the quantity of operation points for the power spectrum estimation. Similarly, for a frequency band with a frequency range of 1k Hz to 10k Hz offset from the carrier frequency, data with a decimation factor of 64 is selected, and 2^{11} is used as the quantity of operation points for the power spectrum estimation.

Table 1 Segmental decimation and splicing range of a 500 kHz signal

Decimation factor	Frequency range offset from the carrier frequency [Hz]	Quantity of operation points	Frequency resolution [Hz]	Splicing range [Hz]
16	(0,312500]	2^{12}	152.6	10k-100k
64	(0,39062]	2^{11}	38.1	1k-10k
256	(0,9765]	2^{13}	2.4	100-1k
2048	(0,1220]	2^{11}	1.2	10-100
		2^{14}	0.15	1-10
		2^{16}	0.04	0.1-1
		2^{19}	0.005	0.01-0.1
		2^{22}	0.0006	0.0006-0.01

[0076]

[0077] FIG. 5 shows a phase noise test result of splicing in this example. A curve labeled "A" shows a data measurement result of the algorithm in the present disclosure, and a curve labeled "B" shows a measurement result of a commercial instrument. As shown in FIG. 5, compared with the commercial instrument, the present disclosure can measure phase noise within [0.0006 Hz, 1 Hz], and achieve an extremely close frequency resolution of 0.0006 Hz. FIG. 6 is a schematic diagram of the splicing in the example in FIG. 5. Target phase difference sequences a_{I}, a_{II}, a_{III}, and a_{IV}, and frequency bands I, II, III, and IV are included. Corresponding decimation factors of the frequency bands I, II, III, and IV are 2048, 256, 64, and 16 respectively. For the extremely close frequency band, namely, the frequency band I, to save computational resources, five segments are obtained through segmentation based on the quantity of FFT points, and then phase noise power spectral densities are spliced to obtain a phase noise power spectral density with a frequency range of [0.0006 Hz, 100 Hz] offset from the carrier frequency. Finally, an image obtained by splicing phase noise power spectral densities with a frequency range of [0.0006 Hz,100k Hz] offset from the carrier frequency is obtained.

[0078] An embodiment of the present disclosure provides a device for fully-digital phase noise measurement with millihertz-level measurement frequency resolution, including:

[0079] a first processing module configured to convert a measured signal into a digital signal, and obtain a frequency control word of the digital signal;

[0080] a second processing module configured to generate two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur based on the frequency control word, and mix the V_{h1} and the V_{h2} with the digital signal to obtain V_{h1}` and $V_{h2}$` respectively;

[0081] a third processing module configured to: separately perform multi-segment decimation and filtering on the V_{h1}` and the $V_{h2}$` to obtain a filtered signal of the V_{h1}` and a filtered signal of the $V_{h2}$` under each decimation factor; and under each decimation factor, perform division and four-quadrant arctangent on the filtered signal of the V_{h1}` and the filtered signal of the $V_{h2}$` to obtain a corresponding initial phase difference sequence, and then perform phase unwrapping and frequency offset elimination on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence; and

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[0082] a fourth processing module configured to perform power spectrum estimation on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and splice the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor to obtain a complete phase noise measurement curve.

[0083] An embodiment of the present disclosure provides a system for fully-digital phase noise measurement with millihertz-level measurement frequency resolution, including a computer-readable storage medium and a processor.

[0084] The computer-readable storage medium is configured to store an executable instruction.

[0085] The processor is configured to read the executable instruction stored in the computer-readable storage medium to execute the method described in any one of the above embodiments.

[0086] An embodiment of the present disclosure provides a computer-readable storage medium. The computer-readable storage medium stores a computer instruction, and the computer instruction is configured to enable a processor to perform the method described in any one of the above embodiments.

[0087] It is easy for those skilled in the art to understand that the above-mentioned contents are merely the preferred embodiments of the present disclosure, and are not intended to limit the present disclosure. Any modifications, equivalent substitution, and improvements made within the spirit and principles of the present disclosure should fall within the protection scope of the present disclosure.

CLAIMS

1. A method for fully-digital phase noise measurement with millihertz-level measurement frequency resolution, comprising:

S1: converting a measured signal into a digital signal, and obtaining a frequency control word of the digital signal;

S2: generating two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur based on the frequency control word, and mixing the V_{h1} and the V_{h2} with the digital signal to obtain V_{h1}\` and $V_{h2}$\` respectively;

S3: separately performing multi-segment decimation and filtering on the V_{h1}\` and the $V_{h2}$\` to obtain a filtered signal of the V_{h1}\` and a filtered signal of the $V_{h2}$\` under each decimation factor; and under each decimation factor, performing division and four-quadrant arctangent on the filtered signal of the V_{h1}\` and the filtered signal of the $V_{h2}$\` to obtain a corresponding initial phase difference sequence, and then performing phase unwrapping and frequency offset elimination on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence; and

S4: performing power spectrum estimation on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and splicing the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor to obtain a complete phase noise measurement curve.

2. The method according to claim 1, wherein in the step S1, the frequency control word of the digital signal is obtained by using an equal-precision frequency measurement method, a direct measurement method, or a periodic measurement method.

3. The method according to claim 1, wherein in the step S2, the V_{h1} and the V_{h2} are generated by using a phase truncation–free frequency synthesis method, a ROM table compression frequency synthesis method, or a coordinate rotation digital computer (CORDIC) frequency synthesis method.

4. The method according to claim 1, wherein the step S4 comprises: segmenting a frequency range offset from the carrier frequency, and selecting different sampling frequencies for different frequency bands to achieve the power spectrum estimation for the frequency bands.

5. A device for fully-digital phase noise measurement with millihertz-level measurement frequency resolution, comprising:

a first processing module configured to convert a measured signal into a digital signal, and obtain a frequency control word of the digital signal;

a second processing module configured to generate two digital signals V_{h1} and V_{h2} with orthogonal phases and no phase truncation spur based on the frequency control word, and mix the V_{h1} and the V_{h2} with the digital signal to obtain V_{h1}\` and $V_{h2}$\` respectively;

a third processing module configured to: separately perform multi-segment decimation and filtering on the V_{h1}\` and the $V_{h2}$\` to obtain a filtered signal of the V_{h1}\` and a filtered signal of the $V_{h2}$\` under each decimation factor; and under each decimation factor, perform division and four-quadrant arctangent on the filtered signal of the V_{h1}\` and the filtered signal of the $V_{h2}$\` to obtain a corresponding initial phase difference sequence, and then perform phase unwrapping and frequency offset elimination on the initial phase difference sequence in turn to obtain a corresponding target phase difference sequence; and

a fourth processing module configured to perform power spectrum estimation on the target phase difference sequence under each decimation factor to obtain a single-sideband phase noise spectral density of the target phase difference sequence, and splice the single-sideband phase noise spectral density of the target phase difference sequence under each decimation factor to obtain a complete phase noise measurement curve.

6. A system for fully-digital phase noise measurement with millihertz-level measurement frequency resolution, comprising a computer-readable storage medium and a processor, wherein

the computer-readable storage medium is configured to store an executable instruction; and

the processor is configured to read the executable instruction stored in the computer-readable storage medium to execute the method according to any one of claims 1 to 4.

7. A computer-readable storage medium, wherein the computer-readable storage medium stores a computer instruction, and the computer instruction is configured to enable a processor to execute the method according to any one of claims 1 to 4.

PATENTANSPRÜCHE

1. Methode für die volldigitale Messung des Phasenrauschens mit Messfrequenzauflösung im Millihertzbereich, umfassend die folgenden Schritte:

S1: Umwandeln eines gemessenen Signals in ein digitales Signal und Erhalten eines Frequenzsteuerworts für das digitale Signal;

S2: Erzeugen zweier digitaler Signale V_{h1} und V_{h2} mit orthogonalen Phasen und ohne Phasenkürzungssporn basierend auf dem Frequenzsteuerwort, und Mischen der Signale V_{h1} und V_{h2}, mit dem digitalen Signal, um die Signale V_{h1}\` bzw. $V_{h2}$\` zu erhalten;

S3: getrenntes Durchführen der Multisegment-Dezimierung und Filterung auf den Signalen V_{h1}\` und $V_{h2}$\`, um ein gefiltertes Signal des V_{h1}\` und ein gefiltertes Signal des $V_{h2}$\` unter jedem Dezimierungsfaktor zu erhalten; dann unter jedem Dezimierungsfaktor Durchführen der Division und des Arkustangens für alle vier Quadranten auf dem gefilterten Signal des V_{h1}\` und dem gefilterten Signal des $V_{h2}$\`, um eine entsprechende anfängliche Phasendifferenzsequenz zu erhalten, und dann abwechselndes Durchführen der Phasenentpackung und der Frequenzoffset-Beseitigung auf der anfänglichen Phasendifferenzsequenz, um eine entsprechende Phasendifferenz-Zielfrequenz zu erhalten; und

S4. Durchführen der Leistungsspektrum-Abschätzung auf der Phasendifferenz-Zielfrequenz unter jedem Dezimierungsfaktor, um eine Phasenrauschen-Spektraldichte für ein einzelnes Seitenband auf der Phasendifferenz-Zielfrequenz zu erhalten, und Spleißen der Phasenrauschen-Spektraldichte für das einzelne Seitenband auf der Phasendifferenz-Zielfrequenz unter jedem Dezimierungsfaktor, um eine vollständige Phasenrauschen-Messkurve zu erhalten.

2. Methode nach Anspruch 1, wobei in Schritt S1 das Frequenzsteuerwort des digitalen Signals durch eine gleichpräzise Frequenzmessmethode, eine direkte Messmethode und eine periodische Messmethode erhalten wird.

3. Methode nach Anspruch 1, wobei in Schritt S2 die Signale V_{h1} und V_{h2} durch eine Phasenfrequenz-Synthesemethode ohne Trunkierung, eine Kompressionsfrequenz-Synthesemethode der ROM-Tabelle oder eine CORDIC (Coordinate Rotation Digital Computer)-Frequenzsynthese-Methode erzeugt werden.

4. Methode nach Anspruch 1, wobei Schritt S4 Folgendes umfasst: Segmentieren eines Frequenzbereich-Offsets der Trägerfrequenz und Wählen unterschiedlicher Abtastfrequenzen für verschiedene Frequenzbänder, um die Leistungsspektrum-Abschätzung für die Frequenzbänder zu erzielen. LU506246

5. Vorrichtung für die volldigitale Messung des Phasenrauschens mit Messfrequenzauflösung im Millihertzbereich, umfassend:

ein erstes Verarbeitungsmodul, das konfiguriert ist, um ein gemessenes Signal in ein digitales Signal umzuwandeln und ein Frequenzsteuerwort für das digitale Signal zu erhalten;

ein zweites Verarbeitungsmodul, das konfiguriert ist, um zwei digitale Signale V_{h1} und V_{h2} mit orthogonalen Phasen und ohne Phasenkürzungssporn basierend auf dem Frequenzsteuerwort zu erzeugen und die Signale V_{h1} und V_{h2} mit dem digitalen Signal zu mischen, um die Signale V_{h1}\` bzw. $V_{h2}$\` zu erhalten;

ein drittes Verarbeitungsmodul, das konfiguriert ist, um: die Multisegment-Dezimierung und Filterung auf den Signalen V_{h2}\` und $V_{h1}$\`getrennt durchzuführen, um ein gefiltertes Signal des V_{h2}\` und ein gefiltertes Signal des $V_{h1}$\` unter jedem Dezimierungsfaktor zu erhalten; dann unter jedem Dezimierungsfaktor die Division und den Arkustangens für alle vier Quadranten auf dem gefilterten Signal des V_{h1}\` und dem gefilterten Signal des $V_{h2}$\` durchzuführen, um eine entsprechende anfängliche Phasendifferenzsequenz zu erhalten, und dann abwechselnd die Phasenentpackung und die Frequenzoffset-Beseitigung auf der anfänglichen Phasendifferenzsequenz durchzuführen, um eine entsprechende Phasendifferenz-Zielfrequenz zu erhalten; und

ein viertes Verarbeitungsmodul, das konfiguriert ist, um die Leistungsspektrum-Abschätzung auf der Phasendifferenz-Zielfrequenz unter jedem Dezimierungsfaktor durchzuführen, um eine Phasenrauschen-Spektraldichte für ein einzelnes Seitenband auf der Phasendifferenz-Zielfrequenz zu erhalten, und um die Phasenrauschen-Spektraldichte für das einzelne Seitenband auf der Phasendifferenz-Zielfrequenz unter jedem Dezimierungsfaktor zu spleißen, um eine vollständige Phasenrauschen-Messkurve zu erhalten.

6. Vorrichtung für die volldigitale Messung des Phasenrauschens mit Messfrequenzauflösung im Millihertzbereich, wobei:

das computerlesbare Steuermedium konfiguriert ist, um eine ausführbare Anweisung zu speichern; und

der Prozessor konfiguriert ist, um die im computerlesbaren Speichermedium gespeicherte ausführbare Anweisung zu lesen, um die Methode nach einem der Ansprüche 1 bis 4 auszuführen.

7. Computerlesbares Speichermedium, wobei das computerlesbare Speichermedium eine Computeranweisung speichert und die Computeranweisung konfiguriert ist, um es einem Prozessor zu ermöglichen, die Methode nach einem der Ansprüche 1 bis 4 auszuführen.

FIGURES

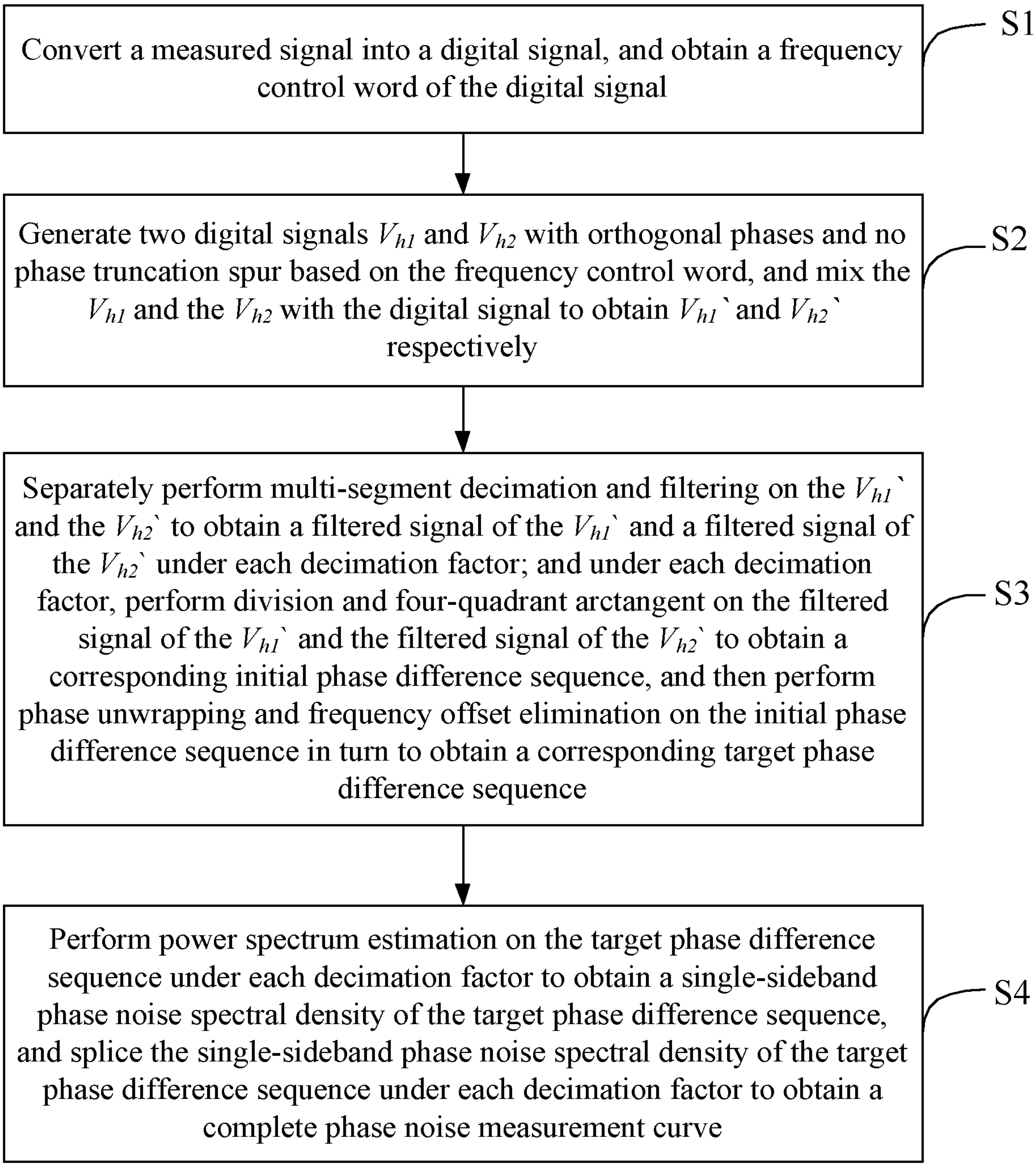

FIG. 1

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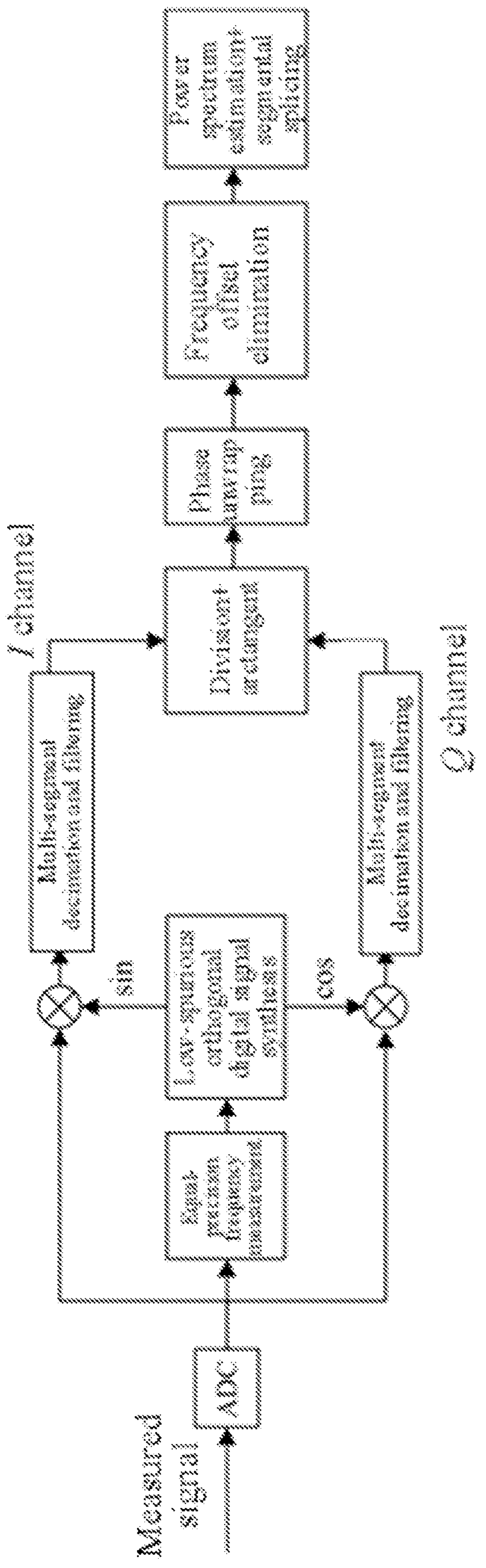

Fig.2

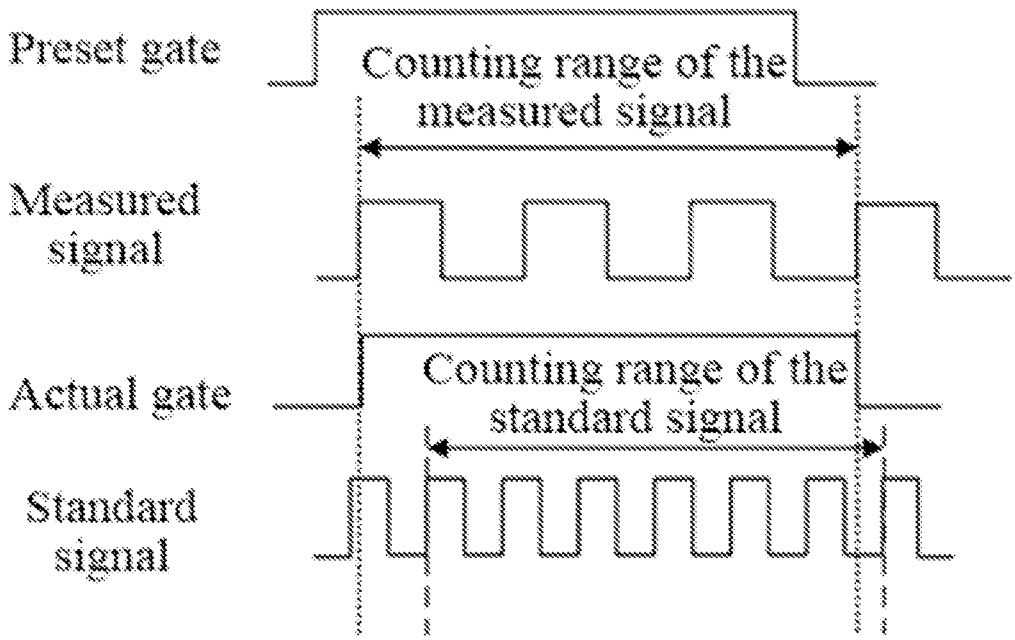
Preset gate
Counting range of the measured signal
Measured signal
Actual gate
Counting range of the standard signal
Standard signal

FIG. 3

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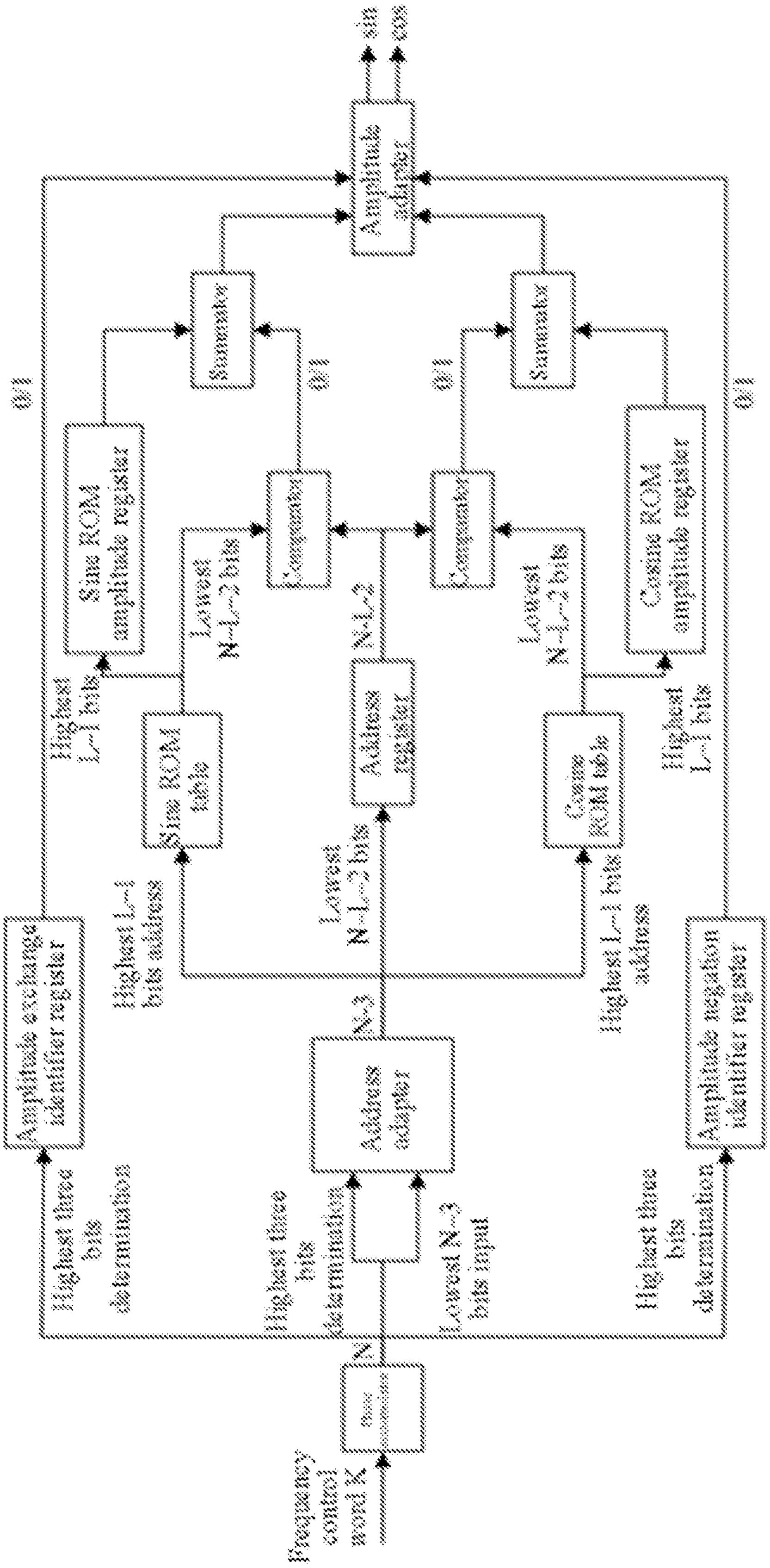

Fig.4

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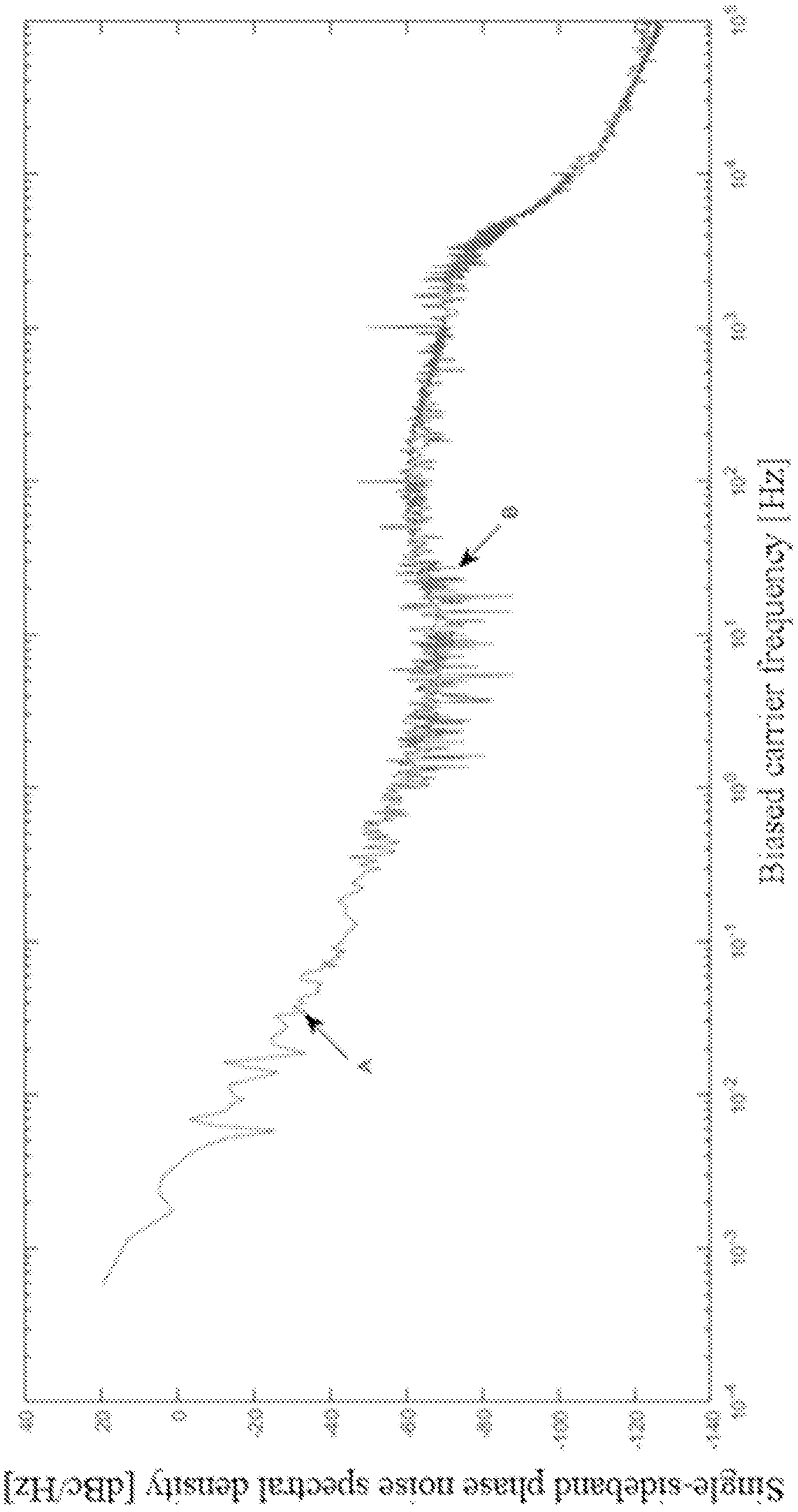

Fig.5

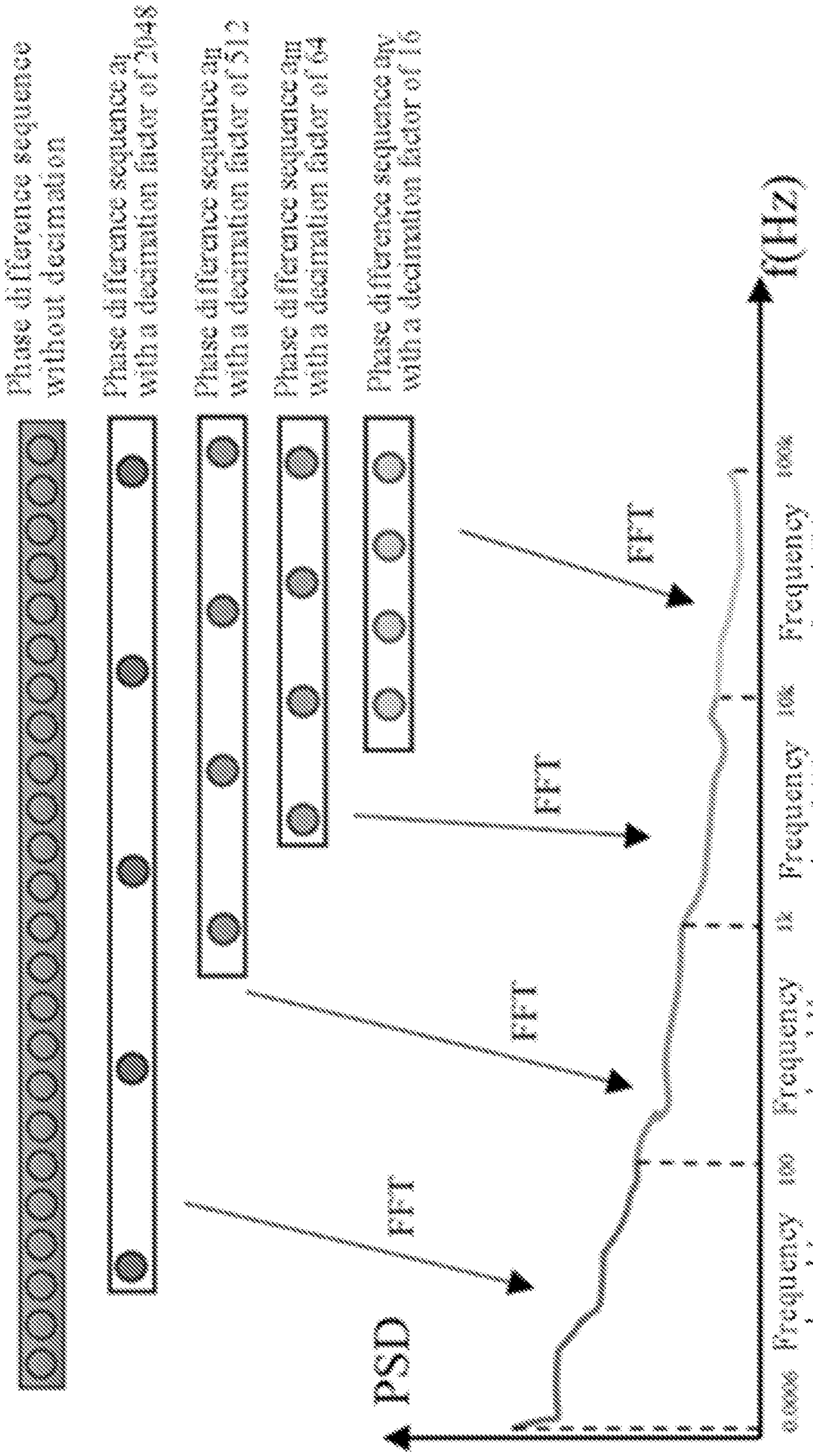
Phase difference sequence without decimation
Phase difference sequence aI with a decimation factor of 2048
Phase difference sequence aII with a decimation factor of 512
Phase difference sequence aIII with a decimation factor of 64
Phase difference sequence aIV with a decimation factor of 16
FFT
FFT
FFT
FFT
PSD
f(HZ)
Frequency band I
Frequency band II
Frequency band III
Frequency band IV

Fig.6