

Article

Simulation of Heterodyne Signal for Science Interferometers of Space-Borne Gravitational Wave Detector and Evaluation of Phase Measurement Noise

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Abstract

Interferometric signals in space-borne Gravitational Wave Detectors are measured by digital phasemeters. The phasemeter processes signals generated by multiple interferometers, with its primary function being micro-radian level phase measurements. The Science Interferometer is responsible for inter-spacecraft measurements, including relative ranging, absolute ranging, laser communication, and clock noise transfer. Since the scientific interferometer incorporates multiple functions and various signals are simultaneously coupled into the heterodyne signal, establishing a suitable evaluation environment is a crucial foundation for achieving micro-radian level phase measurement during ground testing and verification. This paper evaluates the phase measurement noise of the science interferometer by simulating the heterodyne signal and establishing a test environment. The experimental results show that when the simulated heterodyne signal contains the main beat-note, upper and lower sideband beat-notes, and PRN modulation simultaneously, the phase measurement noise of the main beat-note, upper and lower sideband beat-notes all reach $2\pi \mu\text{rad}/\text{Hz}^{1/2}@ (0.1 \text{ mHz} - 1 \text{ Hz})$, meeting the requirements of the space gravitational wave detection mission. An experimental verification platform and performance reference benchmark have been established for subsequent research on the impact of specific noise on phase measurement performance and noise suppression methods.

Keywords: space gravitational wave detection; science interferometer; phasemeter



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1. Introduction

Currently, several ultra-long-baseline space-based gravitational wave detection missions have been proposed internationally, all based on the principle of heterodyne laser interferometry. These missions typically adopt a triangular satellite constellation comprising three spacecraft to meet the requirements of space gravitational wave detection missions. Representative examples include the Laser Interferometer Space Antenna (LISA), led by the European Space Agency, as well as China's proposed Taiji and TianQin [1–3]. In these missions, the interferometric measurement system generally includes three distinct types of interferometers: the Test Mass Interferometer, which measures the attitude variations in the in-satellite test mass relative to the optical bench; the Reference Interferometer, which characterizes the intrinsic noise of the optical bench system; and the Science Interferometer,

which is responsible for inter-spacecraft measurements, including relative ranging, absolute ranging, laser communication, and clock noise transfer. Consequently, the heterodyne signal contains information such as main beat-note, upper and lower sideband beat-notes, Pseudo-Random Noise (PRN) modulation, and coupling noises [4]. The payload that processes the beat-note signals generated by these three types of interferometers is called phasemeter, whose main function is micro-radian level phase measurement based on a digital phase-locked loop (DPLL), as well as multiple auxiliary functions. Given the complexity of signals from scientific interferometers, establishing a realistic and controllable simulation environment during ground testing is a key foundation for verifying the precision phase measurement capability.

Currently, the research teams of the Taiji and TianQin missions employ signal generators to produce MHz-range sine waves as simulated main beat-notes and pilot tones, thereby establishing test environments for evaluating phase measurement noise [5–7]. Iouri Bykov from the Albert Einstein Institute (AEI) designed a signal simulation system capable of mimicking LISA-like heterodyne signals, although the detailed design has not been publicly disclosed. The LISA phasemeter uses this device as the signal source for testing its electrical signal processing performance, demonstrating a phase measurement noise level of $6 \mu\text{rad}/\text{Hz}^{1/2}$ at some fixed frequency points [8].

Unlike these prior works, including our earlier study in Reference [9] which focused mainly on the main beat-note generated by a signal generator, this paper proposes a heterodyne interference signals simulation system that can simultaneously generate composite signals containing the main beat-note, upper and lower sideband beat-notes, and PRN modulation. This system-level simulation platform enables a more comprehensive and realistic evaluation of the phasemeter's phase measurement noise performance under mission-representative signal conditions.

Experimental results show that when the heterodyne signal simulation system simultaneously includes the main beat-note, upper and lower sideband beat-notes, and PRN modulation, the phase measurement noise of the main beat-note and sidebands reaches $2\pi \mu\text{rad}/\text{Hz}^{1/2}$ in the frequency range of 0.1 mHz to 1 Hz [10–12], meeting the requirements of space gravitational wave detection missions. To further carry out investigations on the effects of Doppler shift, shot noise, laser frequency noise, Quadrant Photodiode (QPD) noise, and temperature noise on phase measurement performance, as well as the development of noise suppression methods, an experimental validation platform and performance reference benchmarks have been established. This work has significant research relevance and engineering application value for advancing ground testing and evaluation of both the main and auxiliary functions of the phasemeter.

2. Scheme for Heterodyne Signal Simulation System of the Science Interferometer

In the design schemes of the LISA and Taiji missions, the gravitational wave measurement system employs a Michelson interferometer formed by a triangular constellation of three satellites separated by millions of kilometers, performing pairwise heterodyne interferometric measurements. The principle of laser heterodyne interferometry between two satellites is illustrated in Figure 1 [13]. This measurement system enables inter-satellite ranging and communication, as well as clock noise transfer. The laser signals exchanged between the satellites mainly consist of the main beat-note, upper sideband beat-note, lower sideband beat-note, and PRN modulation. The PRN modulation is applied to the phase of the main beat-note using an Electro-Optic Modulator (EOM). To minimize the impact on gravitational wave detection signals and effectively suppress higher-order sideband components, the modulation depth of the PRN modulation is limited to 0.1 rad (accounting

for 1% of the total power), while the sideband modulation depth must not exceed 0.45 rad, with the upper and lower sideband beat-notes together accounting for no more than 10% of the total power [14,15].

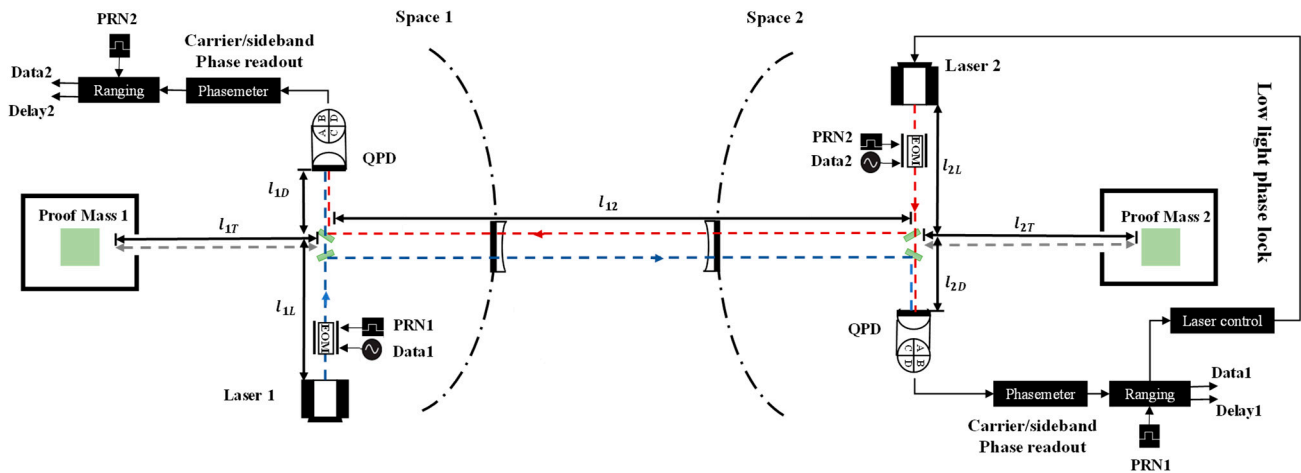


Figure 1. Space gravitational wave interferometric measurement system.

According to the working principle illustrated in Figure 1, the laser emitted from Satellite 1 interferes with the local laser at Satellite 2, and the resulting heterodyne signal is shaped and filtered by QPD2 to complete photoelectric conversion. Weak-light phase locking is applied to phase-lock Satellite 2's local laser to the received laser. Subsequently, the laser emitted from Satellite 2 interferes with the local laser at Satellite 1, and the resulting heterodyne optical signal is photoelectrically converted after being shaped and filtered by QPD1. This heterodyne electrical signal contains information about the distance variations between the test masses of the two satellites. Finally, the phasemeter on Satellite 1 measures the phase variation in the heterodyne signal, from which the inter-satellite distance changes caused by gravitational waves can be inferred.

The signals exchanged between Satellite 1 and Satellite 2 can be expressed as follows [16–18].

$$S_1 = Ee^{-i[2\pi f_1 t + m_{sb} \cos(2\pi f_{USO1} t + n_1 \phi_{clk1}(t))]} \quad (1)$$

$$S_2 = Ee^{-i[2\pi f_2 t + m_{sb} \cos(2\pi f_{USO2} t + n_2 \phi_{clk2}(t))]} \quad (2)$$

In the equation, f_1 and f_2 represent the initial laser frequencies; m_{sb} is the modulation depth of the sideband clock noise; f_{USO1} and f_{USO2} denote the frequencies of the Ultra-Stable Oscillators (USOs); $\phi_{clk1}(t)$ and $\phi_{clk2}(t)$ represent the clock noise measured by the two satellites; and n_1 and n_2 are the clock multiplication factors. (To reduce the phase noise requirements during clock noise transfer, frequency multiplication is employed).

Along the propagation path described above, after transmission through the optical channel, the signal S_1 transmitted from Satellite 1 and the local signal S_2 at Satellite 2 are converted into S'_1 and S'_2 , respectively, before interference.

$$S'_1 = E_s e^{-i[2\pi f_1 t + \frac{2\pi}{\lambda}(l_{1L} + l_{12} + 2l_{2T} + l_{2D}) + m_{sb} \cos(2\pi f_{USO1} t + n_1 \phi_{clk1}(t))]} \quad (3)$$

$$S'_2 = E_{LO} e^{-i[2\pi f_2 t + \frac{2\pi}{\lambda}(l_{2L} + l_{2D}) + m_{sb} \cos(2\pi f_{USO2} t + n_2 \phi_{clk2}(t))]} \quad (4)$$

In the equation, E_s and E_{LO} represent the amplitudes of the received light and the local oscillator light, respectively; l_{1L} , l_{12} , l_{2T} , l_{2L} , and l_{2D} , as shown in Figure 1, denote the respective propagation paths of the laser before interference.

The two light waves interfere coherently.

$$S_{12} = S'_1 + S'_2 \quad (5)$$

Let $\varphi_1 = \frac{2\pi}{\lambda}(l_{1L} + l_{12} + 2l_{1T} + l_{1D})$ and $\varphi_2 = \frac{2\pi}{\lambda}(l_{2L} + l_{2D})$. The heterodyne signal received by the QPD can be expressed as:

$$P = S_{12}S_{12}^* = E_s^2 + E_{LO}^2 + 2E_sE_{LO}\text{Re}\left\{e^{i[2\pi(f_1-f_2)t+\varphi_1-\varphi_2]} \cdot e^{i(m_{sb}\cos(2\pi f_{USO1}t+n_1\phi_{clk1}(t))+m_{sb}\cos(\pi-2\pi f_{USO2}t+n_2\phi_{clk2}(t)))}\right\} \quad (6)$$

In the equation, m_{sb} denotes the modulation depth of the sideband clock noise, with a known modulation index not exceeding 0.45 rad. Under this modulation index, the Bessel function expansion retains only the J_0 and J_1 terms. Let $f_1 - f_2 = \Delta f$, $\varphi_1 - \varphi_2 = \varphi_{12}$, $f_{USO1} - f_{USO2} = \Delta f_{USO}$, and $n_1\phi_{clk1}(t) - n_2\phi_{clk2}(t) = \Delta\phi$. Then P is expressed as:

$$P = S_{12}S_{12}^* = E_s^2 + E_{LO}^2 + 2E_sE_{LO}\text{Re}\left\{e^{i[2\pi\Delta ft+\varphi_{12}]}[J_0^2(m_{sb}) - J_1^2(m_{sb})](e^{i(\pi+2\pi\Delta f_{USO}t+\Delta\phi)} + e^{-i(\pi+2\pi\Delta f_{USO}t+\Delta\phi)}) + e^{i(\pi-2\pi(\Delta f_{USO1}+\Delta f_{USO2})t-\Delta\phi)} + e^{-i(\pi-2\pi(\Delta f_{USO1}+\Delta f_{USO2})t-\Delta\phi)}]\right\} \quad (7)$$

In Equation (7), $f_{USO1} + f_{USO2}$ exceeds the response range of the QPD. Therefore, the following expression can be obtained:

$$P = S_{12}S_{12}^* = E_s^2 + E_{LO}^2 + 2E_sE_{LO} \cdot J_0^2(m_{sb}) \cos(2\pi\Delta ft + \varphi_{12}) + 2E_sE_{LO} \cdot J_1^2(m_{sb}) \cos(2\pi(\Delta f + \Delta f_{USO})t + \Delta\phi + \varphi_{12}) + 2E_sE_{LO} \cdot J_1^2(m_{sb}) \cos(2\pi(\Delta f - \Delta f_{USO})t - \Delta\phi + \varphi_{12}) \quad (8)$$

It can be seen that the optical intensity signal received by QPD2 contains a DC component and three beat-notes with different frequencies. Among them, $J_1^2(m_{sb}) \cos(2\pi(\Delta f \pm \Delta f_{USO})t \pm \Delta\phi + \varphi_{12})$ represents the case where clock noise is modulated onto the upper and lower side-band beat-notes, and is subsequently eliminated using TDI (Time Delay Interferometry) techniques [18–20]. The term $J_0^2(m_{sb}) \cos(2\pi\Delta ft + \varphi_{12})$ corresponds to the main beat-note, whose phase is measured by Phasemeter 2 and locked into the phase of Laser 2. The laser is then transmitted to Satellite 1, and the signal can be expressed as:

$$S_3 = Ee^{-i[2\pi f_2t+\varphi_{12}+m_{prn}\sum_{n=-\infty}^{\infty}c_nP(t-nT_c)+m_{sb}\cos(2\pi f_{USO2}t+n_2\phi_{clk2}(t))]} \quad (9)$$

In the equation, m_{prn} denotes the modulation depth of the PRN modulation; C_n represents the PRN code, whose pulse waveform is $p(t)$ with a period of T_c .

The signal S_3 is transmitted to Satellite 1 and interferes again with the Laser 1 signal. Before interference, the two signals are like those in Equations (5) and (6):

$$S'_3 = E_se^{-i[2\pi f_2t+\frac{2\pi}{\lambda}(l_{2L}+l_{12}+2l_{1T}+l_{1D})+\varphi_{12}+m_{prn}\sum_{n=-\infty}^{\infty}c_nP(t-nT_c)+m_{sb}\cos(2\pi f_{USO2}t+n_2\phi_{clk2}(t))]} \quad (10)$$

$$S'_4 = E_{LO}e^{-i[2\pi f_2t+\frac{2\pi}{\lambda}(l_{1L}+l_{1D})+m_{sb}\cos(2\pi f_{USO1}t+n_1\phi_{clk1}(t))]} \quad (11)$$

Let $\varphi_3 = \frac{2\pi}{\lambda}(l_{2L} + l_{12} + 2l_{1T} + l_{1D})$, $\varphi_4 = \frac{2\pi}{\lambda}(l_{1L} + l_{1D})$, $\varphi_0 = \varphi_3 + \varphi_{12} - \varphi_4$ and $\varphi_0 = \frac{4\pi}{\lambda}(l_{12} + l_{1T} + l_{2T})$. Similarly, the heterodyne signal is:

$$P = E_s^2 + E_{LO}^2 + 2E_sE_{LO} \cdot J_0^2(m_{sb}) \cos(2\pi\Delta ft + \varphi_0 + m_{prn}\sum_{n=-\infty}^{\infty}c_nP(t-nT_c)) + 2E_sE_{LO} \cdot J_1^2(m_{sb}) \cos(2\pi(\Delta f + \Delta f_{USO})t + \Delta\phi + \varphi_0) + 2E_sE_{LO} \cdot J_1^2(m_{sb}) \cos(2\pi(\Delta f - \Delta f_{USO})t - \Delta\phi + \varphi_0) \quad (12)$$

From Equation (12), it can be seen that φ_0 represents the phase variation caused by gravitational waves considering clock noise. The corresponding distance variation can be derived through phase transformation as follows:

$$\delta L = \frac{\lambda}{4\pi} \delta \varphi \quad (13)$$

In the equation, $\delta \varphi$ denotes the phase variation, and δL represents the distance variation between the test masses of the two satellites. The derivation of the gravitational wave signal above includes factors such as clock noise, signal modulation, and the QPD response bandwidth. However, in practical scenarios, the signals are often influenced by additional factors.

The above derivation of the gravitational wave interferometric measurement process yields a physical model of the heterodyne signal. For ground testing of the phasemeter, this paper establishes a heterodyne signal simulation system based on Equation (12). The system includes the main beat-note, upper and lower sideband beat-notes, PRN modulation, and BPSK modulation, satisfying the following requirements:

- (1) The main beat-note, accounting for 89% of the total power, with adjustable amplitude and frequency, featuring a frequency dynamic range from 5 MHz to 25 MHz;
- (2) Sideband beat-notes, each accounting for 5% of the total power, with adjustable amplitude and frequency within a dynamic range of 4 MHz to 26 MHz, forming approximately ± 1 MHz frequency offsets relative to the main beat-note;
- (3) Generation, coupling, and modulation of PRN and data codes, with an adjustable modulation depth accounting for 1% of the total power.

2.1. Software System Design

To realize the heterodyne signal simulation system, the main beat-note signal is generated by a Direct Digital Synthesizer (DDS), while the upper and lower sideband beat-notes are provided by commercial signal sources. The working principle of the DDS is illustrated in Figure 2 [16]. It mainly consists of a Phase Accumulator (PA), a phase-to-amplitude conversion module (typically a sine lookup table ROM), a Digital-to-Analog Converter (DAC), and an output filter. The system is driven by a fixed clock f_{clk} . The input frequency control word Δp is added to the phase register at each clock cycle to form a linearly increasing phase value, i.e., phase accumulation. During each clock cycle, the phase register is incremented to achieve linear phase growth. The resulting phase value is converted via the lookup table into corresponding waveform amplitudes and output as an analog signal by the DAC. This signal is then smoothed by a low-pass filter to generate a continuous sine wave or other desired waveform. The DDS output frequency is related to the frequency control word Δp and the system clock frequency f_{clk} by the following relationship:

$$f_{out} = \frac{\Delta p \times f_{clk}}{2^N} \quad (14)$$

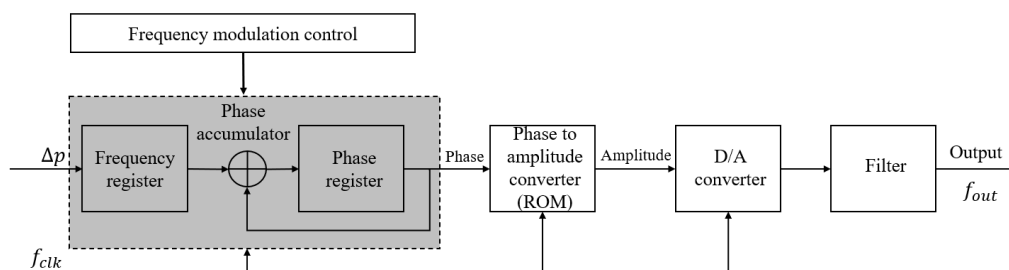


Figure 2. Principle block diagram of the DDS.

In the equation, N represents the bit width of the phase register.

In addition, the DDS also supports modulation functions, as shown in Figure 3 [21], enabling various modulation controls such as frequency, phase, and amplitude to meet the phase modulation requirements in gravitational wave detection. During operation, the input frequency control word Δp is added to the frequency modulation control signal and then fed into the phase accumulator to generate the corresponding phase information. This phase information is further combined with the phase modulation control signal to achieve phase modulation. The modulated phase value is then input to the phase-to-amplitude conversion module to obtain the corresponding waveform amplitude data.

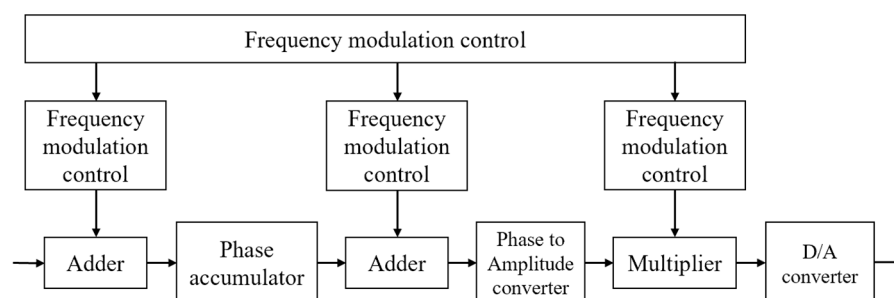


Figure 3. Modulation-capable DDS block diagram.

The above components enable the implementation of various functions of the heterodyne signal simulation system. The overall design parameters are listed in Table 1 [15].

Table 1. Parameters of the heterodyne signal simulation system.

Parameters	Value	Optical Power Ratio
System	80 MHz	
Main beat-note	5 MHz~25 MHz	89%
Sideband beat-note	4 MHz~26 MHz	$2 \times 5\%$
Sideband modulation index	0.45 rad	
Communication modulation index	0.1 rad	1%
PRN rate	1 Mbps	
Data rate	15.625 Kbps	

2.2. Hardware System Design

The hardware architecture of the system is shown in Figure 4. An FPGA (XC7K325T-2FFG900I) generates the main beat-note, PRN modulation, and coupling noise signals and integrates them for output. The FPGA controls an LTC1668, a high-performance 16-bit DAC with a maximum update rate of 50 Msps, ensuring the signal accuracy required by the heterodyne signal simulation system.

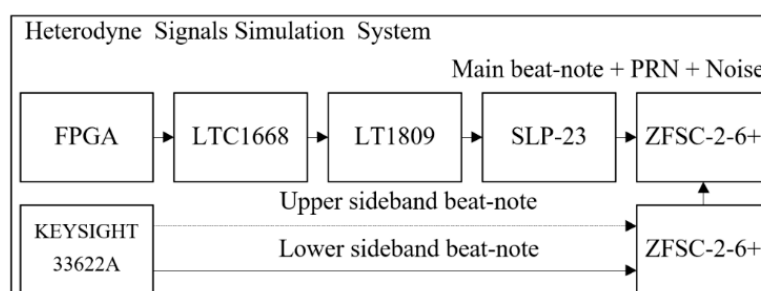


Figure 4. Hardware block diagram of the heterodyne signal simulation system.

The resulting analog signal is amplified by an LT1809, which provides a 320 MHz bandwidth (gain = 1) and a 350 V/ μ s slew rate, fully meeting high-performance demands for signal transmission and processing.

Depending on application requirements, a low-pass filter is selected to further process the amplified signal. In this paper, an SLP-23 low-pass filter from Mini-Circuits, with a -3 dB cutoff frequency of 25 MHz, low insertion loss, and excellent VSWR, is used to suppress out-of-band interference and attenuate high-frequency noise.

The upper and lower sideband beat-notes are generated by a KEYSIGHT 33622A signal generator and combined with the main beat-note, PRN modulation, and coupling noise signals using a Mini-Circuits ZFSC-2-6+ power splitter/combiner.

3. Evaluation Scheme for Phase Measurement Noise

This paper establishes a phase measurement noise evaluation platform for heterodyne signals, composed of a heterodyne signal simulation system, a Sampling Timing Jitter Noise Suppression System (STJNSS), and a phasemeter. The heterodyne signal simulation system is responsible for generating simulated signals containing the main beat-note, upper and lower sideband beat-notes, and PRN modulation, with noise coupling as needed to emulate heterodyne signal outputs under different signal-to-noise ratio conditions. The STJNSS provides an ultra-stable system clock signal (80 MHz) and a low phase noise pilot tone signal (37.5 MHz), adjusting amplitude through amplifiers and suppressing high-frequency noise via low-pass filters to ensure signal quality meets the phasemeter's input requirements. The phasemeter, using the system clock as a reference, performs synchronous sampling and phase extraction of the input heterodyne simulation signals and pilot tone, achieving high-precision phase measurement.

Figure 5 shows the system's block diagram, clearly illustrating the connections between functional modules and the signal transmission paths. Subsequent sections will provide detailed descriptions of the design schemes and performance implementations of each subsystem. The overall system workflow is as follows: the heterodyne signal simulation system generates the main beat-note, upper sideband beat-note, lower sideband beat-note, PRN modulation, and noise signals, which are combined according to preset proportions. The combined signals are then fed into the phasemeter. Meanwhile, the STJNSS supplies system clock and pilot tone to phasemeter.

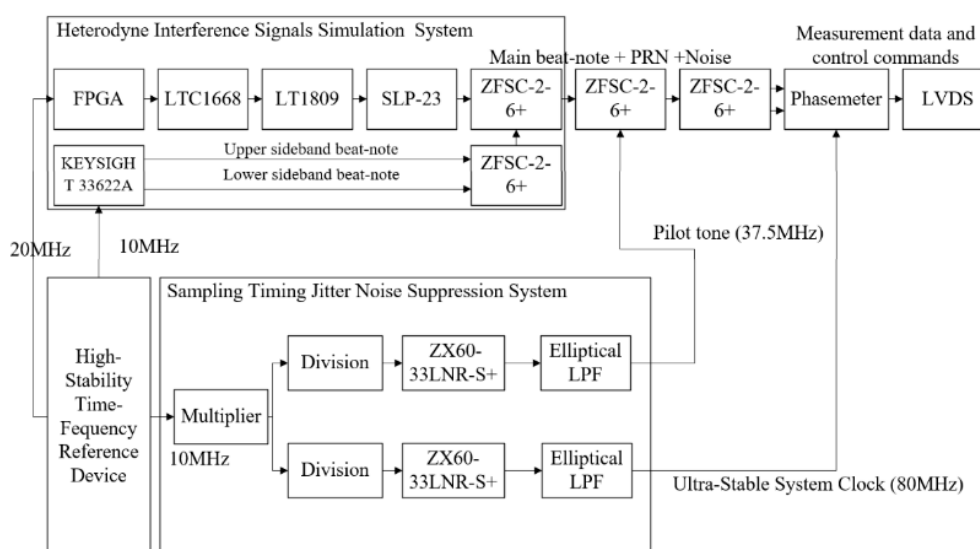


Figure 5. Schematic block diagram of the phase measurement noise evaluation scheme.

3.1. STJNSS

Building upon prior work [9], the STJNSS optimizes the clock signal quality by replacing the original VLFX-80+ low-pass filter with a newly designed elliptic filter before the clock signal enters the phasemeter, thereby further suppressing high-frequency interference in the clock distribution chain. Compared to Butterworth and Chebyshev filters, the elliptic filter exhibits equiripple characteristics in both the passband and stopband, allowing it to achieve the lowest order for the same passband ripple and stopband attenuation requirements. This results in a steeper transition band and superior frequency selectivity, making it well-suited for effectively filtering high-frequency interference in the STJNSS. Since the main beat-note frequency range is 5 MHz to 25 MHz and the system clock remains at 80 MHz, the original pilot tone at 75 MHz would be aliased to 5 MHz after sampling by the system clock, overlapping with the main beat-note frequency and failing to meet mission requirements. Therefore, in this work, the pilot tone frequency was changed to 37.5 MHz.

The design employs a fourth-order passive LC π -type topology, with two filters specifically designed for the pilot tone signal and the system clock signal, respectively. For the 37.5 MHz pilot tone signal, the design parameters are: passband cutoff frequency of 40 MHz, passband ripple not exceeding 0.01 dB, and stopband starting frequency of 80 MHz. For the 80 MHz system clock signal, the parameters are: passband cutoff frequency of 90 MHz, passband ripple not exceeding 0.01 dB, and stopband starting frequency of 180 MHz. The normalized amplitude-frequency response is expressed as:

$$|H(j\Omega)|^2 = \frac{1}{1 + \varepsilon^2 R_n^2(\xi)} \quad (15)$$

Here, ε denotes the passband ripple magnitude, and $R_n^2(\xi)$ represents the n th-order elliptic rational function. The filter order n can be estimated using elliptic integral functions. In this design, the parameters were automatically calculated using MATLAB's `ellipord` and `ellip` functions, then mapped to actual component values through frequency and impedance scaling. After completing the design, frequency-domain simulations were conducted on the MATLAB 2021b platform, with results shown in Figures 6 and 7. The simulation results demonstrate that both elliptic low-pass filters effectively suppress high-frequency interference outside the target frequency bands, significantly improving the signal-to-noise ratio of the system clock and pilot tone signals, thereby providing a stable timing reference for the ADC and FPGA modules in the phasemeter.

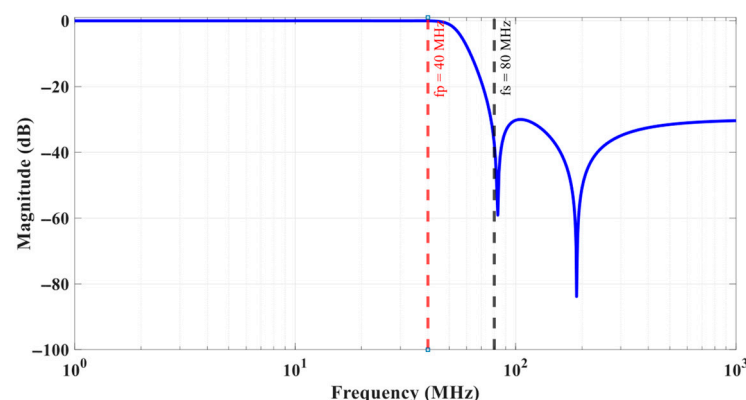


Figure 6. Amplitude response of the elliptic low-pass filter with a 40 MHz passband cutoff frequency.

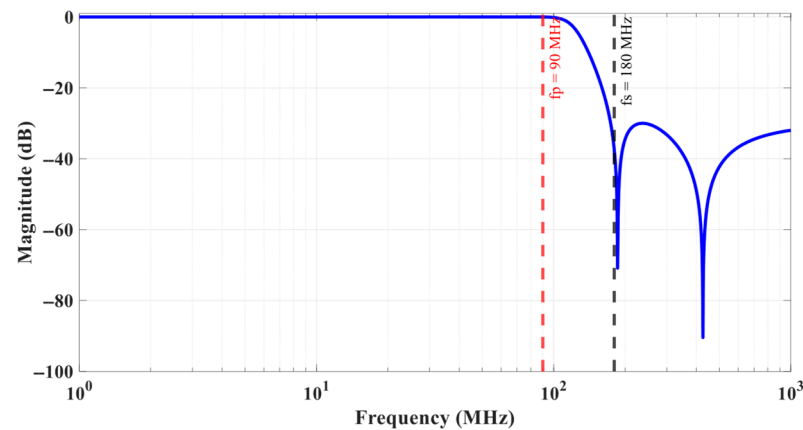


Figure 7. Amplitude response of the elliptic low-pass filter with a 90 MHz passband cutoff frequency.

3.2. Phasemeter

The phasemeter system architecture used in this paper is shown in Figure 8. The overall design is divided into three main parts: the front-end analog board responsible for heterodyne signals acquisition, the digital processing board tasked with data processing and system control, and the high-speed communication interface board connecting the two. The system supports parallel processing of 16 heterodyne signal channels, comprising two analog front-end boards (each supporting 8 channels), one digital processing board, and one interface board.

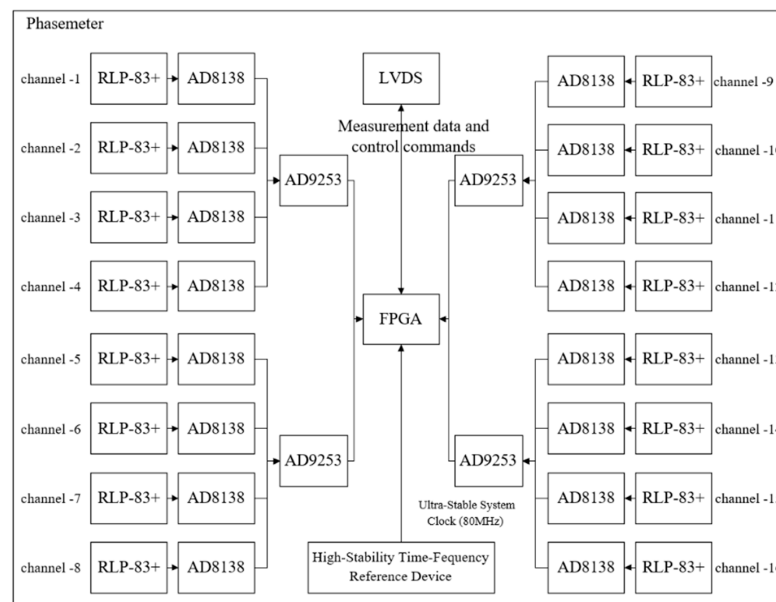


Figure 8. Block diagram of the phasemeter system.

The 16 analog signals from the heterodyne signal simulation system pass through low-pass filtering, differential amplification, and analog-to-digital conversion processes, all performed by the front-end analog board to convert analog signals into digital form. The digital signals are then transmitted at high speed via the interface board to the FPGA platform on the digital processing board, where key processing modules such as the digital phase-locked loop (DPLL) and data downsampling are implemented. Ultimately, the system outputs high-precision phase measurement results [6].

This design uses the RLP-83+ low-pass filter from Mini-Circuits, featuring a -3 dB cutoff frequency of 93 MHz. It effectively suppresses out-of-band high-frequency interference, enhancing the system's noise immunity. The front-end analog amplification employs

an active differential driver structure based on the AD8138, which offers a 320 MHz bandwidth and a 1150 V/ μ s slew rate. Compared to transformer-coupled solutions, this design provides better fidelity and noise suppression for low-frequency pilot tone signals [10], while symmetric impedance matching and DC coupling improve channel consistency and phase stability.

The analog-to-digital conversion uses the AD9253, a four-channel 14-bit ADC supporting synchronous sampling up to 125 Msps, ensuring phase consistency across multiple channels. The digital processing board is based on the Kintex-7 FPGA (XC7K325T-2FFG900I), providing abundant logic resources and Block RAM to meet phase processing and data transmission requirements. Data communication and debugging are carried out via a dual-channel LVDS bus operating at 10 Mbps.

3.3. Phase Measurement Performance Evaluation Method

This paper adopts the separated measurement method illustrated in Figure 9, where the heterodyne signal InS generated by the heterodyne signal simulation system is coupled with the 37.5 MHz pilot tone signal PT provided by the STJNSS. The signals are further divided into two pairs: InSa with PTa, and InSb with PTb. Each pair is fed into four independently configured DPLL modules for phase extraction, yielding phase data φ_{InS1} , φ_{PT1} , φ_{InS2} , and φ_{PT2} sequentially. Based on the heterodyne signal frequency f_{InS} and the pilot tone frequency f_{PT} , an error term CORR caused by sampling timing jitter is calculated. This term is used to compensate the original phase measurement results of the heterodyne signal, thereby obtaining the accurate phase φ_{InS} after jitter noise suppression.

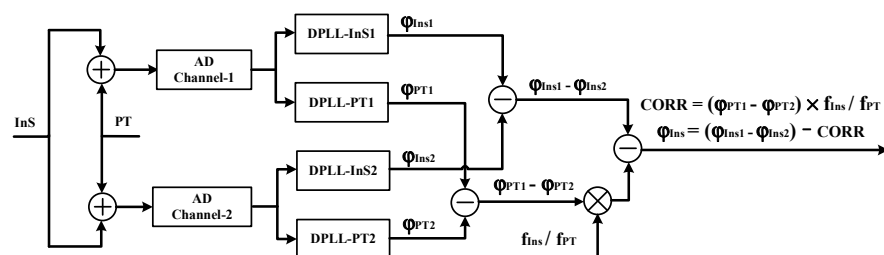


Figure 9. Schematic diagram of the phase measurement performance evaluation method [6].

4. Experimental Verification

4.1. Evaluation Environment Design

The evaluation environment for phase measurement noise is illustrated in Figure 10.

4.1.1. Heterodyne Signal Simulation System

To verify the effectiveness of the system design, two heterodyne signal modes were devised based on the previously described system architecture and signal coupling principle, taking into account the required frequency dynamics of the heterodyne signal. The corresponding parameters are listed in Table 2.

Table 2. Design Parameters of the Heterodyne Signals.

Mode	Signal	Frequency	Optical Power Ratio	Signal Source Configuration
One	Main Beat-note	5/10/15/25 MHz	89%	DDS
	PRN code	1 MHz	1%	PRN Generator
Two	Main Beat-note	20 MHz	89%	DDS
	Up Sideband	21 MHz	5%	RF Signal Generator
	Down Sideband	19 MHz	5%	RF Signal Generator
	PRN code	1 MHz	1%	PRN Generator

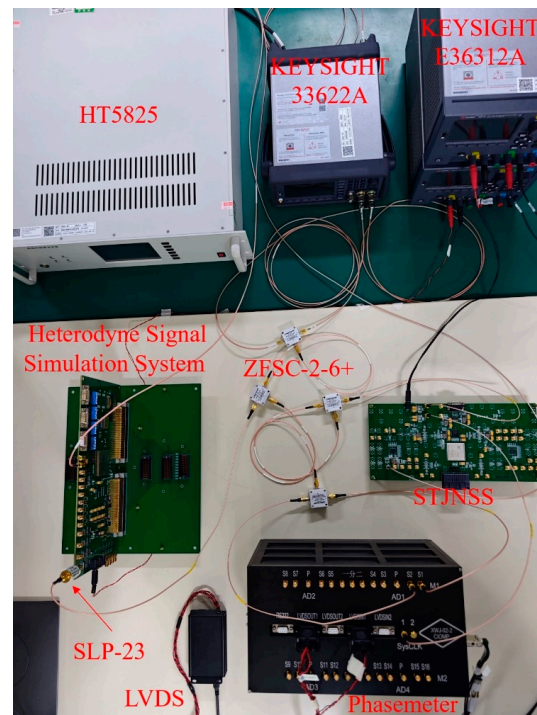


Figure 10. Evaluation environment for phase measurement noise.

As listed in Table 2, the first mode includes main beat-note and PRN code. The frequency of main beat-note is configured at four values: 5 MHz, 10 MHz, 15 MHz, and 25 MHz. The second mode consists of 20 MHz main beat-note, 21 MHz upper sideband beat-note, 19 MHz lower sideband beat-note, and PRN code. According to the system's optical power allocation, the power ratios of main beat-note, upper sideband beat-note, lower sideband beat-note, and PRN code are 89%, 5%, 5%, and 1%, respectively. Given that signal amplitude is proportional to the square root of optical power, the amplitude ratios of upper and lower sideband beat-notes relative to main beat-note are approximately 0.2, while that of PRN code is about 0.1. The measured spectra of the heterodyne signals for both modes are shown in Figures 11 and 12.

Table 3 presents the amplitude characteristics of main beat-note, upper sideband beat-note, and lower sideband beat-note for both signal modes. The amplitudes of the upper and lower sideband beat-notes correspond well with the designed ratios relative to the main beat-note frequency (the calculation formula is provided in the notes of Table 3, where P_{side} , P_{main} , and R_x represent the sideband frequency, main beat-note frequency, and ratio of the total power occupied, respectively), and exhibit symmetrical distribution. The phase modulation state of the ranging communication code has minimal impact on the amplitudes of the main beat-note frequency and sidebands, indicating stable amplitude output of the system.

Table 3. Amplitudes of the Simulated heterodyne Signals.

Mode	Main Beat-Note Frequency	Main Beat-Note Amplitude /dBm	Up Sideband Beat-Note Amplitude /dBm	Down Sideband Beat-Note Amplitude /dBm	Remark
One	5 MHz	4.260	--	--	$P_{side} = 10 \log_{10}(\frac{R_x}{P_{main}})$
	10 MHz	4.523	--	--	
	15 MHz	4.344	--	--	
	25 MHz	−3.898	--	--	
Two	20 MHz	4.656	−7.831	−7.917	

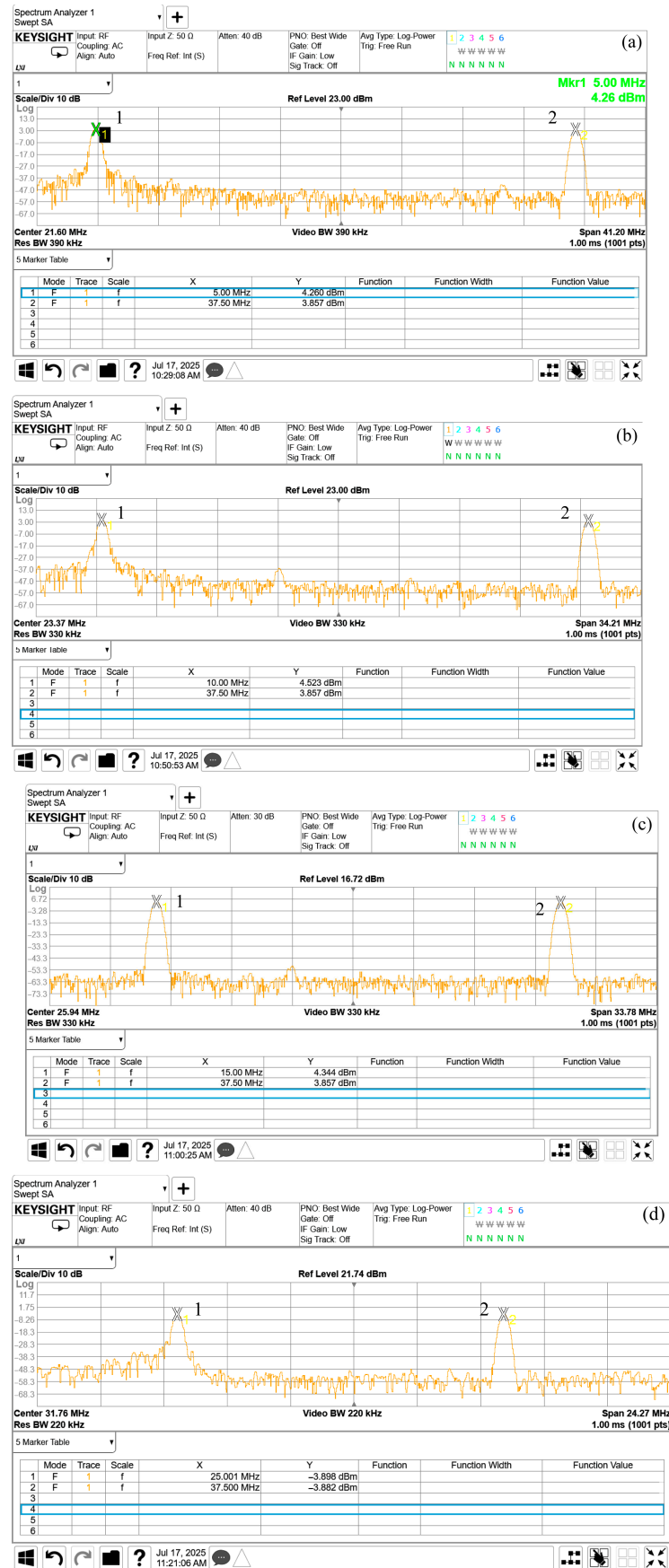


Figure 11. Frequency spectra of the simulated heterodyne signals for the first mode, with (a–d) representing main beat-note with frequencies of 5 MHz, 10 MHz, 15 MHz, and 25 MHz, respectively.

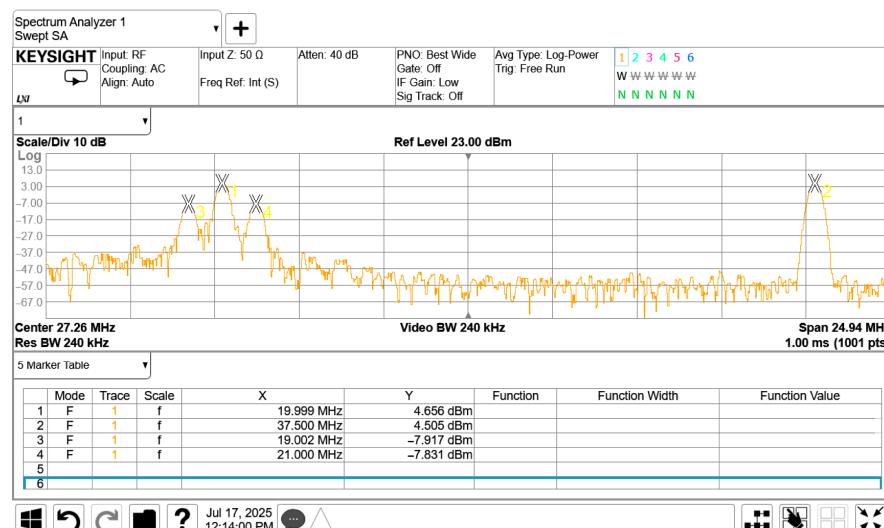


Figure 12. Frequency spectrum of the simulated heterodyne signal for the second mode.

4.1.2. Sampling Timing Jitter Noise Suppression System (STJNSS)

The system supports output of multiple pilot tones and ultra-stable system clock signals. This paper evaluates the quality of the system's 2.4 GHz, 80 MHz, and 37.5 MHz outputs and compares the results with those reported in reference [9].

This paper employed a KEYSIGHT MSOS404A oscilloscope to conduct time-domain characteristic and eye diagram tests. The time-domain waveforms and eye diagrams of the 2.4 GHz, 80 MHz, and 37.5 MHz output signals are shown in Figure 13. As indicated by the results in Table 4, the improved STJNSS exhibits superior time-domain stability. Compared to reference [9], the DC bias of the output signals at all frequencies is more stable, with significantly reduced overshoot and ringing. The eye height is also notably enhanced, indicating a higher signal-to-noise ratio for the system clock and pilot tone signals. These findings further validate the effectiveness of the elliptic filter design in suppressing high-frequency interference in this system.

Table 4. Time domain characterization test results.

	2.4 GHz	80 MHz [9]	80 MHz	75 MHz [9]	37.5 MHz
Mean Value	1.4 mV	5.01 mV	3.4 mV	2.37 mV	5.21 mV
Peak-to-Peak Amplitude	233 mV	2.08 V	2 V	712.11 mV	1.5 V
RMS	82.83 mV	717.4 mV	704 mV	245.72 mV	524 mV
Rise/Fall Time of Signal Edges	120 ps/121 ps	3.66 ns/3.59 ns	3.59 ns/3.62 ns	3.63 ns/4.02 ns	7.61 ns/7.69 ns
Overshoot	0.81%	0.864%	0.48%	0.51%	0.45%
Eye Height	208 mV	1.34 V	1.828 V	608 mV	1.35 V
Eye Width	201 ps	6.52 ns	6.10 ns	6.47 ns	13 ns

This paper employs the KEYSIGHT CXA Signal Analyzer to test and analyze the frequency-domain characteristics and phase noise of the system output signals. During the frequency-domain tests, the video bandwidth was set to 240 Hz and the spectrum span to 25 kHz, with spectrum curves obtained for the 2.4 GHz, 80 MHz, and 37.5 MHz output signals, respectively. For phase noise testing, the frequency offset range was set from 100 Hz to 1 MHz, with the corresponding results shown in Figure 14.

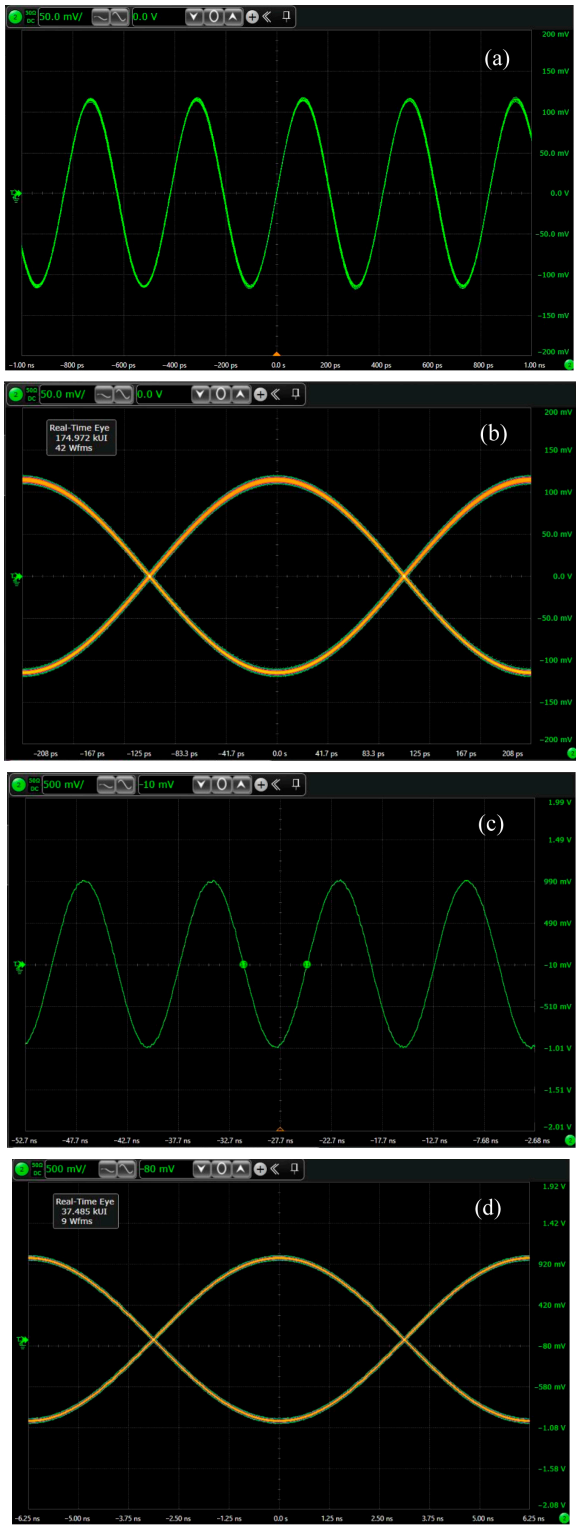


Figure 13. Cont.

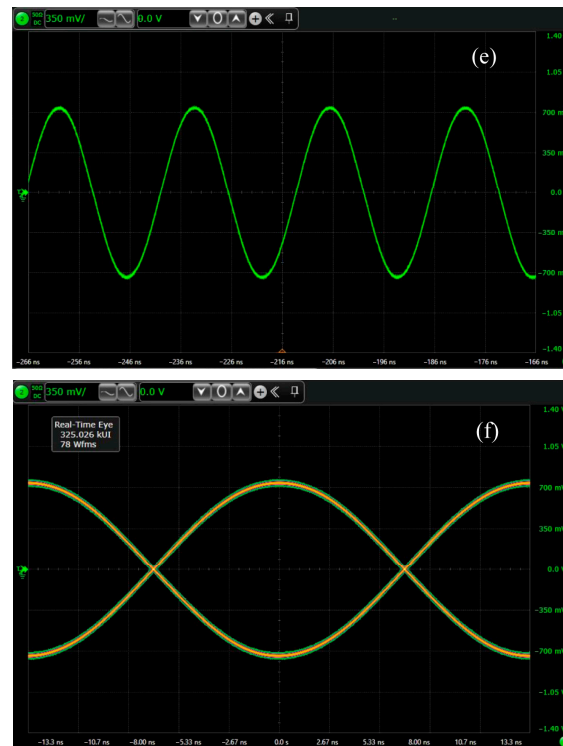


Figure 13. Testing of time-domain characteristics and eye diagrams by MSOS404A oscilloscope ((a) 2.4 GHz time-domain; (b) 80 MHz time-domain; (c) 37.5 MHz time-domain; (d) 2.4 GHz eye diagram; (e) 80 MHz eye diagram; (f) 37.5 MHz eye diagram).

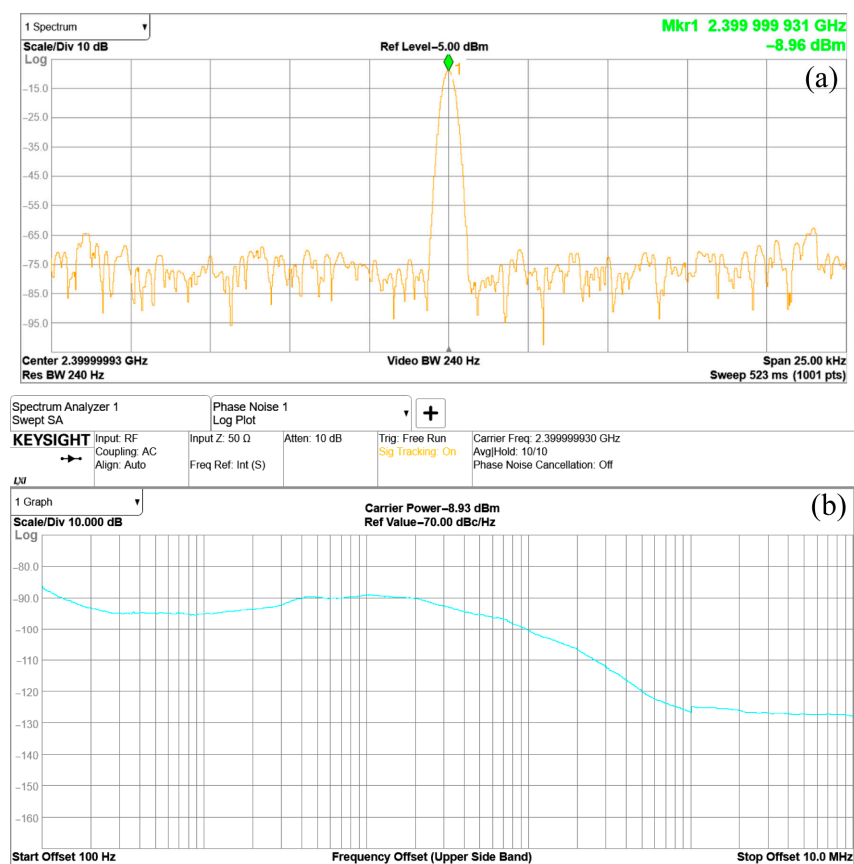


Figure 14. Cont.

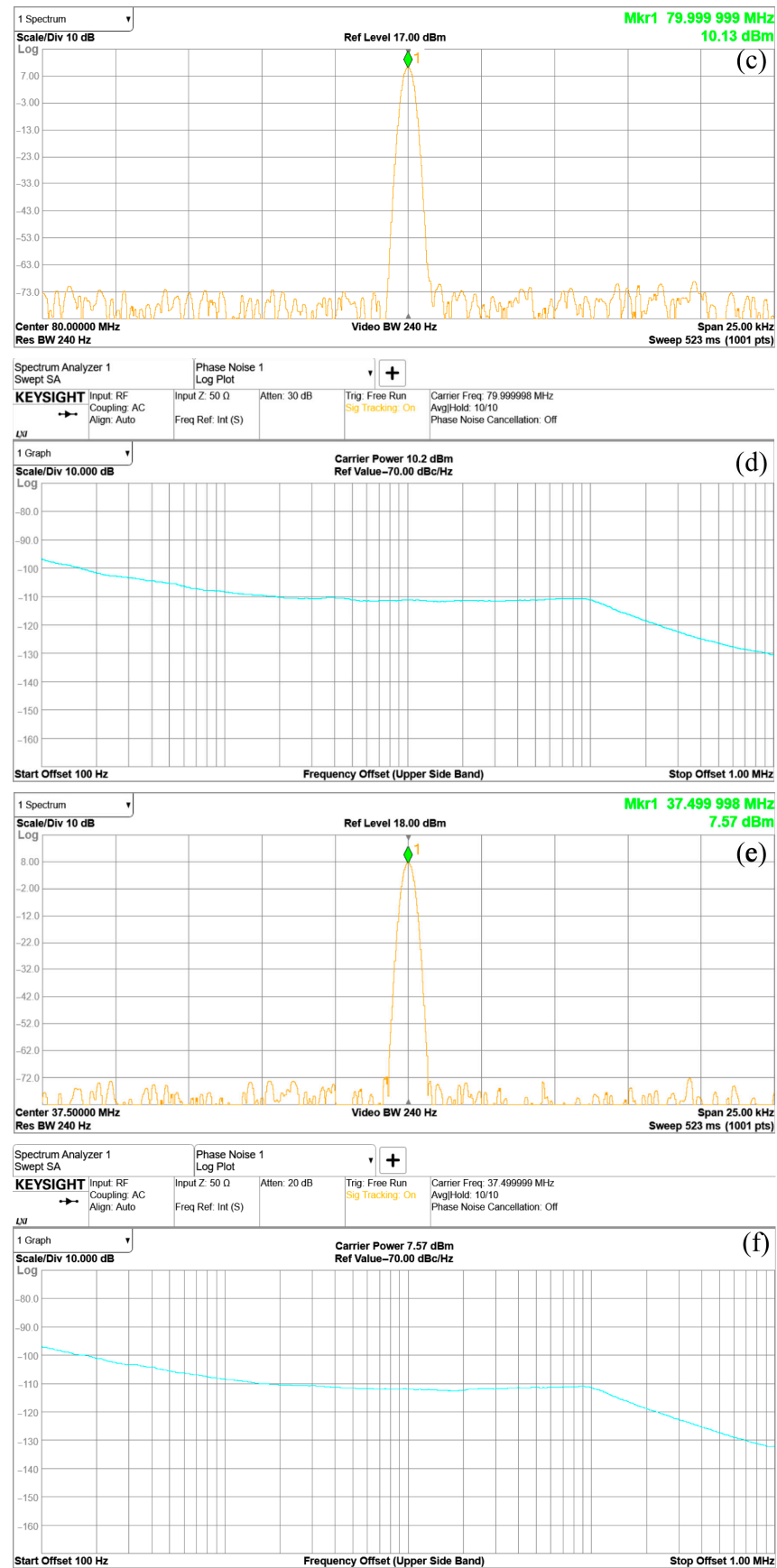


Figure 14. Spectrum and Phase Noise of Signals Tested by CXA Signal Analyzer ((a) 2.4 GHz Spectrum; (b) 2.4 GHz Signal Phase Noise; (c) 80 MHz Spectrum; (d) 80 MHz Signal Phase Noise; (e) 37.5 MHz Spectrum; (f) 37.5 MHz Signal Phase Noise).

The summarized test data are presented in Table 5. The results indicate that all pilot tone signals and the ultra-stable clock signals output by the system meet the expected performance in terms of frequency-domain performance. Notably, within the 100 Hz to 1 MHz offset range, the phase noise of all signals is better than -85 dBc/Hz.

Table 5. Frequency domain performance test results of the STJNSS.

	2.4 GHz	80 MHz [9]	80 MHz	75 MHz [9]	37.5 MHz
Carrier Signal	−9 dBm	10.32 dBm	10.13 dBm	0.67 dBm	7.57 dBm
Second Harmonic	−35 dBm	−17.49 dBm	−56.58 dBm	−21.64 dBm	−62.49 dBm
THD	5.01%	4.07%	0.05%	7.66%	0.10%
Phase Noise @100 Hz	−86.32 dBc/Hz	−97.04 dBc/Hz	−96.49 dBc/Hz	−98.13 dBc/Hz	−96.82 dBc/Hz
Phase Noise @1 kHz	−95.14 dBc/Hz	−107.42 dBc/Hz	−108.34 dBc/Hz	−108.33 dBc/Hz	−108.44 dBc/Hz
Phase Noise @10 kHz	−89.09 dBc/Hz	−109.68 dBc/Hz	−111.23 dBc/Hz	−111.84 dBc/Hz	−111.90 dBc/Hz
Phase Noise @100 kHz	−100.55 dBc/Hz	−110.77 dBc/Hz	−111.10 dBc/Hz	−110.84 dBc/Hz	−111.42 dBc/Hz
Phase Noise @1 MHz	−126.11 dBc/Hz	−130.46 dBc/Hz	−130.63 dBc/Hz	−130.6 dBc/Hz	−132.07 dBc/Hz

Compared with the scheme proposed in reference [9], the elliptic low-pass filter introduced in this work demonstrates a significant advantage in suppressing high-frequency interference. It effectively attenuates higher-order harmonics and markedly improves the spectral purity of the output signals, thereby enhancing the overall frequency-domain stability and phase noise performance of the system.

4.2. Phase Measurement Noise Test Results

The phase noise measurement results of the main beat-note in the first mode heterodyne signal are shown in Figure 15, while those for the second mode heterodyne signal are presented in Figure 16. The curves PT1–PT2, InS1–InS2, InUS1–InUS2, and InDS1–InDS2 represent the phase measurement noise of the pilot tone, main beat-note, upper sideband beat-note, and lower sideband beat-note after common-mode noise suppression, respectively. The curves InS-CORR, InUS-CORR, and InDS-CORR correspond to the phase noise after sampling timing jitter noise suppression using the 37.5 MHz pilot tone for the main beat-note, upper sideband beat-note, and lower sideband beat-note, respectively. All reach $2\pi \mu\text{rad}/\text{Hz}^{1/2}$ within the frequency range of 0.1 mHz to 1 Hz. The main beat-note frequency covers 5 MHz to 25 MHz, meeting the requirements for space-based gravitational wave detection.

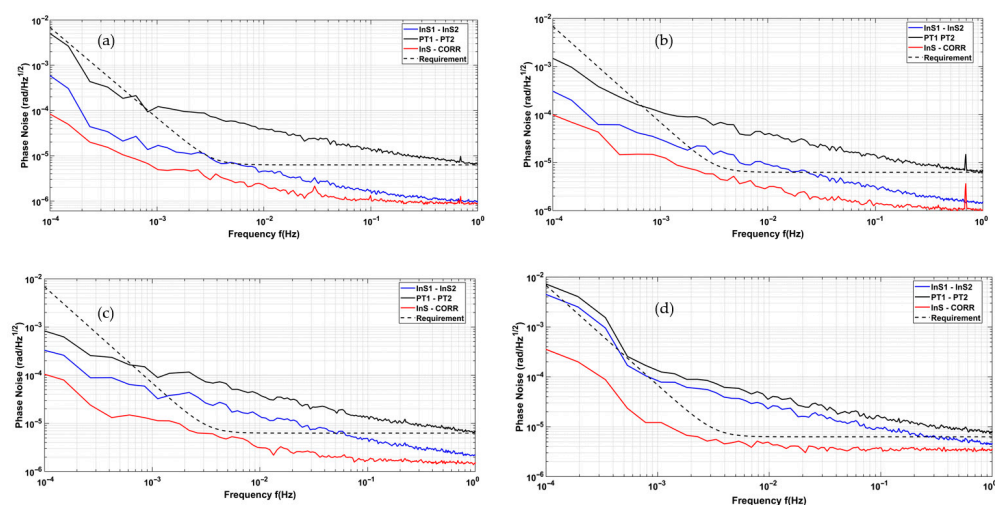


Figure 15. Phase noise of the main beat-note in the first mode ((a) Phase noise of the 5 MHz main beat-note; (b) Phase noise of the 10 MHz main beat-note; (c) Phase noise of the 15 MHz main beat-note; (d) Phase noise of the 25 MHz main beat-note).

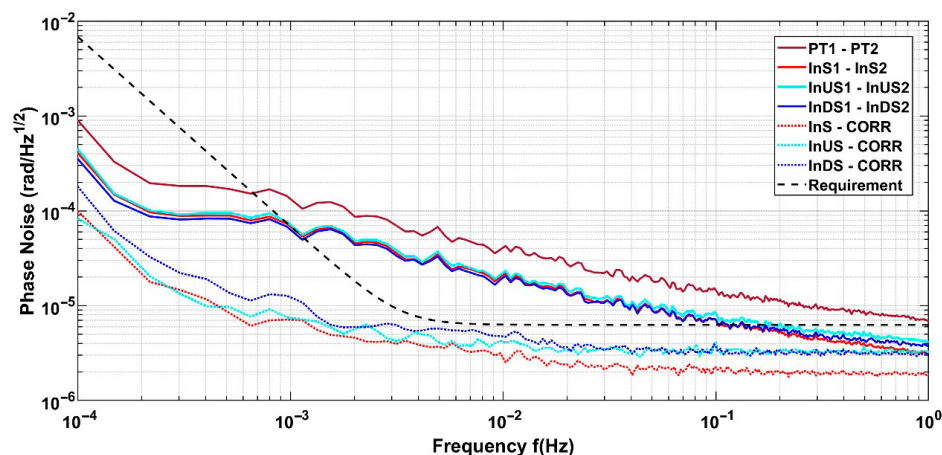


Figure 16. Phase noise of the 20 MHz main beat-note coupled PRN code, 21 MHz upper sideband beat-note, 19 MHz lower sideband beat-note in the second mode.

5. Conclusions

An experimental platform was established in an electronic environment to evaluate the impact of noise on phase measurement and to investigate noise suppression methods. This paper proposes a simulation scheme for the heterodyne signals of a science interferometer, optimizes the filter design of the STJNSS, and establishes a phase measurement noise evaluation environment. Unlike previous studies focusing mainly on single beat-note signals, the proposed heterodyne interference signals simulation system can simultaneously generate the main beat-note, upper and lower sideband beat-notes, and PRN modulation, enabling system-level testing and performance verification. Experimental results demonstrate that when the simulated heterodyne signal simultaneously contains the main beat-note, upper and lower sideband beat-notes, and PRN modulation, the phase measurement noise of the main beat-note and sidebands reaches $2\pi \mu\text{rad}/\text{Hz}^{1/2}$ within the frequency range of 0.1 mHz to 1 Hz, meeting the requirements of space gravitational wave detection missions. For further studies involving simulation of Doppler shifts, shot noise, laser frequency noise, QPD noise, and temperature noise, as well as assessment of their effects on phase measurement sensitivity and the development of noise identification and suppression techniques, the aforementioned experimental validation platform holds significant application value.

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Data Availability Statement: The data presented in this study are not publicly available due to privacy, and access can be requested from [yut@ciomp.ac.cn] upon reasonable request.

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