

Coherent Dual-Band Dual-Chirp Microwave Waveform Generation Based on a Polarization-Multiplexed Optoelectronic Oscillator

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Abstract—We propose and experimentally demonstrate a coherent dual-band dual-chirp microwave waveform generation scheme based on a polarization-multiplexed optoelectronic oscillator (OEO). A dual-polarization binary phase shift-keying (DP-BPSK) modulator is employed to construct orthogonal signal paths, with a dual-passband filter integrated into the OEO loop. Mutual injection locking enables stable dual-frequency oscillations with low phase noise and a high side-mode suppression ratio (SMSR). The generated dual-frequency signal serves as a local oscillator (LO) to modulate an externally injected baseband linear frequency-modulated (LFM) waveform, producing coherent dual-band dual-chirp LFM signals. The proof-of-concept experiments demonstrate waveform generation centered at 10 and 16 GHz with tunable bandwidths, duty cycles, and repetition periods. Compared to a free-running OEO, the proposed scheme improves the SMSR by 36 dB and enhances the signal-to-interference ratio (SIR) of the generated LFM signals by 31 dB. Furthermore, relative to direct waveform synthesis using an arbitrary waveform generator (AWG), it achieves a 29-dB SIR improvement and significantly reduces spectral linewidth. The output bandwidth can be flexibly extended up to 12 GHz by adjusting the baseband chirp, demonstrating high configurability and spectral scalability. Owing to its compact architecture, high coherence, and excellent spectral purity, the proposed approach is well suited for advanced radar systems requiring high resolution and robust interference suppression.

Index Terms—Coherent signal generation, dual-band dual-chirp linear frequency-modulated (LFM), mutual injection locking, optoelectronic oscillator (OEO), signal-to-interference ratio (SIR).

I. INTRODUCTION

THE chirped microwave signals are widely used in modern radar systems due to their capability for pulse

compression, which enhances range resolution and target detection accuracy [1], [2], [3]. However, conventional single-chirp signals suffer from range-Doppler coupling, limiting simultaneous high resolution in both domains because of the knife-edge shape of the ambiguity function [4], [5]. In contrast, dual-chirp signals, consisting of complementary up- and down-chirps within a single period, can mitigate this coupling effect and improve resolution for both range and velocity measurements [6], [7], [8]. Meanwhile, dual-band radar systems further provide advantages over single-band systems, especially in complex electromagnetic environments. Operating across distinct frequency bands improves the detection of stealth or low-RCS targets, enhances anti-jamming capability, and leverages complementary scattering characteristics of each band. For example, X-band supports high-resolution imaging, while Ku-band offers superior penetration and target recognition. This enables integrated imaging, tracking, and through-wall detection, thereby improving overall sensing performance [9], [10], [11], [12]. When signals across multiple bands are generated coherently, they enable precise data fusion, high-resolution imaging, and reliable multitarget tracking [13], [14], [15]. Therefore, generating coherent dual-band dual-chirp microwave waveforms is of critical importance for the next-generation dual-band radar platforms.

Conventional electronic solutions, such as voltage-controlled oscillators (VCOs) and direct digital synthesizers (DDSs), are commonly used to generate chirped signals. However, these methods face inherent limitations in achievable signal-to-interference ratio (SIR), modulation bandwidth, and phase noise at high frequencies [16], [17]. In contrast, microwave photonics offers a compelling alternative with its wide operating bandwidth, low transmission loss, and strong immunity to electromagnetic interference [18].

To address these challenges, various photonic schemes have been proposed for dual-band linear chirp waveform generation [19], [20], [21], [22], [23]. In [20], a mode-locked laser (MLL) combined with multiple optical filters is used to extract distinct optical carriers, which are then modulated and combined to produce dual-band radar signals. However, the signal bandwidth in this approach is constrained by the MLL repetition rate. External modulation schemes

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have been proposed to extend the bandwidth and enable flexible waveform synthesis [21], [22], [23]. For instance, Yu et al. [21] demonstrated a switchable dual-band dual-chirp signal generation method using a dual-drive dual-parallel Mach–Zehnder modulator (DD-DPMZM). Two RF signals and a baseband linear frequency-modulated (LFM) waveform are injected into the DD-DPMZM, and by tuning its bias point, dual-band dual-chirp signals with tunable center frequencies and wide bandwidths are obtained. Although these approaches allow flexible waveform generation, they rely on external microwave sources and frequency multiplication, which degrade phase noise performance at high frequencies. Optoelectronic oscillator (OEO) represents a promising solution for low-phase-noise signal generation [24], [25], [26], [27], [28], [29], [30], [31], with the state-of-the-art phase noise performance reaching -163 dBc/Hz at a 6-kHz offset [26]. Using an OEO-generated low-phase-noise microwave signal as the local oscillator (LO) to modulate and mix with a baseband chirped signal can significantly improve the phase noise of the resulting LFM waveform [32], [33]. However, overcoming gain competition in the loop to achieve phase-coherent dual-frequency signal generation—which is required as the LO for baseband LFM modulation—remains a challenge [30]. In [31], a dual-polarization binary phase shift-keying (DP-BPSK)-based architecture was proposed, in which an actively mode-locked OEO generates a coherent optical frequency comb in one polarization, while the orthogonal polarization modulates the carrier with a baseband chirp signal. After polarization recombination and photodetection, a dual-band dual-chirp signal is obtained. Nevertheless, actively mode-locked OEOs are susceptible to supermode noise due to intracavity gain competition, especially under harmonic mode-locking, which can degrade waveform coherence and radar detection performance. Insufficient side-mode suppression further limits the SIR of the generated waveform.

In this work, we propose and experimentally demonstrate a coherent dual-band dual-chirp microwave waveform generation scheme based on a polarization-multiplexed OEO incorporating a DP-BPSK modulator. The X-polarized arm forms an OEO loop, where dual-frequency oscillation is enabled by a dual-band bandpass filter (DB-BPF) and stabilized via mutual injection locking. A microwave signal matching the frequency spacing between the two oscillation modes is injected to facilitate nonlinear coupling and suppress gain competition, resulting in a highly coherent dual-frequency signal with high side-mode suppression ratio (SMSR) and low phase noise. This dual-tone signal, together with an externally injected baseband LFM waveform, is applied to the Y-polarized arm biased at the minimum transmission point, enabling carrier-suppressed modulation to generate dual-band dual-chirp microwave signals upon photodetection. Experimental results show waveform generation centered at 10 and 16 GHz, with flexible control over bandwidth, duty cycle, and period. Compared to a free-running OEO, the proposed scheme improves the SMSR by 36 dB and the SIR of the generated LFM waveform by 31 dB. Compared to conventional arbitrary waveform generator (AWG)-based synthesis,

it further improves the SIR by 29 dB, demonstrating its effectiveness for high-fidelity radar applications.

II. PRINCIPLE

Fig. 1(a) illustrates the schematic of the proposed system. A continuous-wave optical signal from a laser diode (LD) passes through a polarization controller (PC) and enters a DP-BPSK modulator. This device integrates two parallel Mach–Zehnder modulators (MZM-x, MZM-y), a polarization rotator (PR), and a polarization beam combiner (PBC). The orthogonally polarized outputs are combined at the PBC and separated by a polarization beam splitter (PBS). The X-polarized branch from MZM-x propagates through a length of single-mode fiber (SMF) and is converted to an electrical signal by photodetector PD1 and amplified by an electrical amplifier (EA1). A DB-BPF selects two oscillation modes, producing a high-coherence dual-frequency signal. This signal is split: one path serves as the LO for MZM-y, while the other is combined with a single-tone signal at the mode spacing frequency and reinjected into MZM-x to enable mutual injection locking. In parallel, the LO signal and a baseband LFM waveform from an AWG are combined and applied to MZM-y. The Y-polarized arm is biased at the minimum transmission point to achieve carrier-suppressed double-sideband modulation. The modulated optical signal is separated by another PBS, detected by PD2, amplified, and delivered as the final output. Fig. 1(b) illustrates the mutual injection-locking mechanism. In a free-running OEO, gain competition sustains only a single oscillation mode f_2 . By injecting a signal at f_3 , nonlinear mixing within the loop generates a new frequency component $f_1 = f_2 + f_3$. This component injection-locks a longitudinal mode in the second passband, enabling oscillation at f_1 . Simultaneously, the oscillation at f_1 mixes with the injected signal at f_3 to reproduce $f_2 = f_1 - f_3$, reinforcing the original mode. This bidirectional frequency conversion and energy coupling between f_1 and f_3 effectively suppress gain competition, allowing stable, phase-locked dual-frequency oscillation within the OEO cavity.

The microwave signal injected into the MZM-x can be expressed as follows:

$$V_{\text{MZM-x}}(t) = V_3 \cos(\omega_3 t + \varphi_3) + V_1 \cos(\omega_1 t + \varphi_1) + V_2 \cos(\omega_2 t + \varphi_2) \quad (1)$$

where V_3, V_1 , and V_2 denote the electrical amplitudes of the injected signal and the two oscillation signals, and φ_3, φ_1 , and φ_2 represent their corresponding initial phases. After a single-loop propagation within the OEO cavity, the voltage of the electrical signal applied to the RF port of the MZM-x can be expressed as follows:

$$V_{\text{out-x}}(t) = V_{\text{ph}} \{1 + \cos \pi [V_{\text{MZM-x}}(t) / V_{\pi} + V_{\text{DC-x}} / V_{\pi-x}]\} \quad (2)$$

where $V_{\text{ph}} = 0.5\alpha_0 G_A R \Re P_c$, α_0 denotes the loss of the fiber, G_A is the gain of the electrical amplifier, R and $\Re$ denote the impedance and responsivity of the PD, respectively, P_c is the input optical power, and $V_{\text{DC-x}}$ and $V_{\pi-x}$ are the dc bias voltage and the half-wave voltage of the MZM-x, respectively.

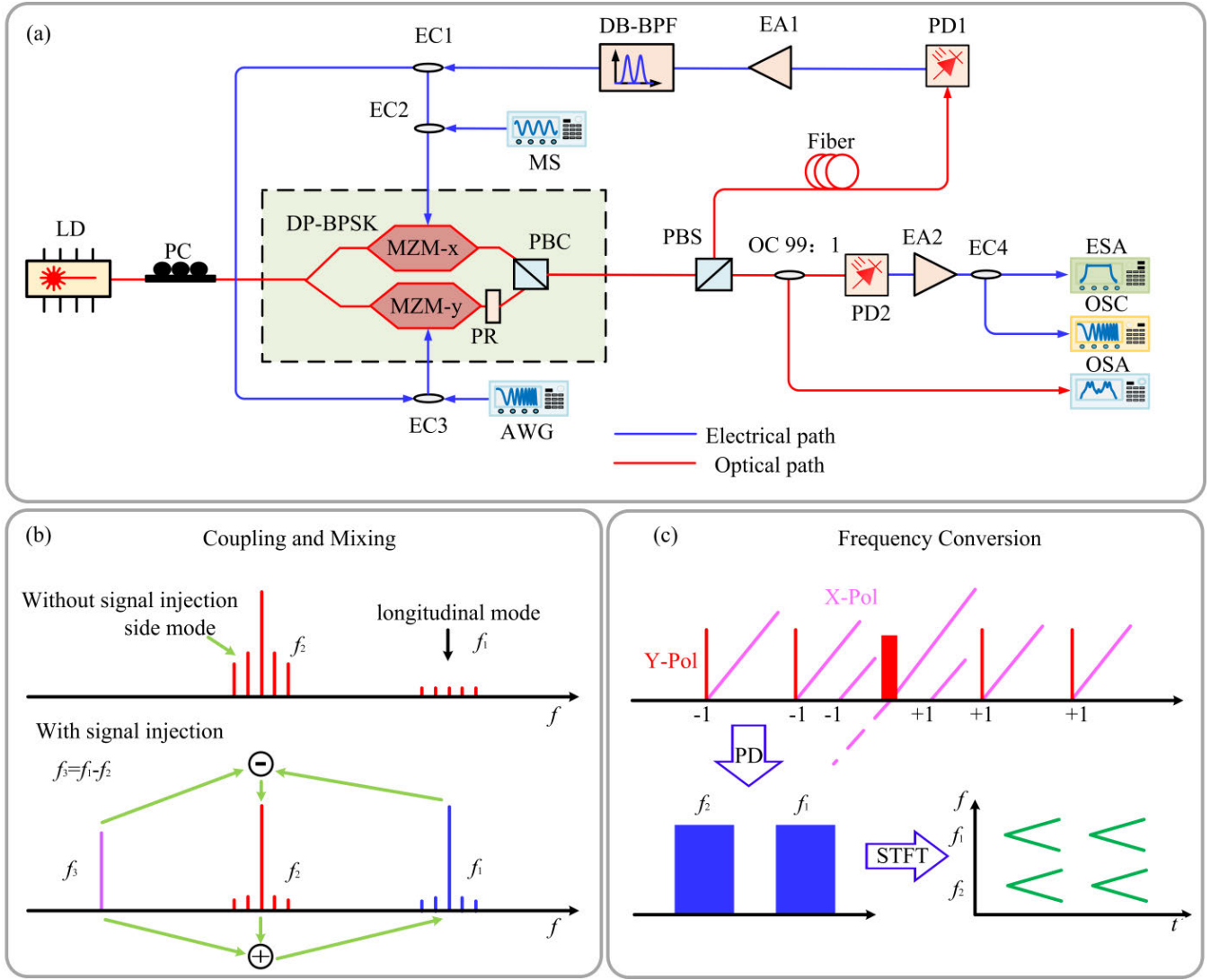


Fig. 1. (a) Schematic of the proposed coherent dual-band dual-chirp microwave waveform generation system based on a phase-locked dual-frequency OEO. (b) Mutual injection-locking process between oscillation modes enabled by nonlinear mixing. (c) Frequency conversion and chirp synthesis process. LD: laser diode; PC: polarization controller; DP-BPSK: dual-polarization binary phase-shift-keying; MZM: Mach-Zehnder modulator; PR: polarization rotator; PBC: polarization beam combiner; PBS: polarization beam splitter; OC: optical coupler; PD: photodetector; EA: electrical amplifier; EC: electrical coupler; DB-BPF: dual-band bandpass filter; AWG: arbitrary waveform generator; ESA: electrical spectrum analyzer; OSC: oscilloscope; OSA: optical spectrum analyzer.

Consequently, after passing through the DB-BPF, the resulting output signal can be expressed as follows:

$$\begin{aligned}
 V_{\text{DB-BPF}}(t) &\propto V_{\text{ph}} \left\{ J_0(\beta_2) J_0(\beta_3) J_1(\beta_1) \sin\left(\frac{\pi V_{\text{DC-x}}}{V_{\pi-x}}\right) \cos(\omega_1 t + \varphi_1) \right. \\
 &+ J_0(\beta_3) J_0(\beta_1) J_1(\beta_2) \sin\left(\frac{\pi V_{\text{DC-x}}}{V_{\pi-x}}\right) \cos(\omega_2 t + \varphi_2) \\
 &- J_0(\beta_2) J_1(\beta_3) J_1(\beta_2) \cos\left(\frac{\pi V_{\text{DC-x}}}{V_{\pi-x}}\right) \\
 &\times \cos[(\omega_2 + \omega_3)t + \varphi_2 + \varphi_3] \\
 &- J_0(\beta_1) J_1(\beta_3) J_1(\beta_1) \cos\left(\frac{\pi V_{\text{DC-x}}}{V_{\pi-x}}\right) \\
 &\times \cos[(\omega_1 - \omega_3)t + \varphi_1 - \varphi_3] \left. \right\} \quad (3)
 \end{aligned}$$

where $J_n(*)$ is the n th Bessel function and $\beta_i = \pi V_i / V_{\pi-x}$ ($i = 1, 2, 3$) is the modulation index. As shown in (3), when the frequency of the injected signal matches the spacing between the two oscillation modes, nonlinear mixing in the loop generates components at the same frequencies as the dual tones, thereby enabling injection locking. Under mutual injection locking, continuous energy coupling and frequency conversion between the two modes suppress gain competition, resulting in stable, phase-coherent dual-frequency oscillation, as shown in Fig. 1(b). Consequently, the SMSR of the generated signal is significantly enhanced compared to that of free-running oscillation. As indicated by (3), the injected signal acting on the free-running OEO modes ω_1 and ω_2 consists of mixed-frequency components generated by the nonlinear interaction between the dual-frequency oscillation signal and the externally injected signal. These mixed terms correspond to the third and fourth terms in (3), which effectively provide

frequency coupling between the two oscillation modes. The locking range of the OEO under injection can be described by [34], [35]

$$\Delta f_{\max} = \frac{\pi f_{\text{osc}}}{Q} \cdot \frac{E_{\text{osc}}}{E_{\text{inj}}} = \frac{\Delta f_{\text{FSR}}}{2} \cdot \frac{E_{\text{inj}}}{E_{\text{osc}}} \quad (4)$$

where f_{osc} corresponds to the oscillation frequency, E_{osc} is the voltage amplitude of the oscillation microwave signal, Q denotes the quality factor of the loop, and E_{inj} represents the amplitude of the mixed-frequency signal that participates in the injection-locking process. FSR refers to the free spectral range of the OEO. The phase noise of the mutually injection-locked dual-frequency signal can be expressed as follows:

$$S_{\phi,i}(f) = \frac{1}{1 + (f/\Delta f_{\max})^2} S_{\phi_{\text{mix},i}}(f) + \frac{1}{1 + (\Delta f_{\max}/f)^2} S_{\phi_{\text{free},i}}(f), (i = 1, 2) \quad (5)$$

where $S_{\phi_{\text{mix},i}}(f)$ and $S_{\phi_{\text{free},i}}(f)$ denote the phase-noise power spectral densities of the mixed-frequency component at ω_i and the free-running OEO mode at the same frequency, respectively. From (5), it can be observed that the phase noise of the mutually coherent dual-frequency signals is determined by the phase noise of the injection-locked mixed-frequency component in the near-offset region, while in the far-offset region, it is dominated by the intrinsic noise of the free-running OEO.

The coherent dual-frequency signal generated by the OEO and a baseband single-chirp waveform $\cos(kt^2)$ with a chirp rate of k are applied to MZM-y. The microwave signal injected into the MZM-y can be expressed as follows:

$$V_{\text{MZM-y}}(t) = V_4 \cos(kt^2) + V_1 \cos(\omega_1 t + \varphi_1) + V_2 \cos(\omega_2 t + \varphi_2) \quad (6)$$

where V_4 represents the electrical amplitude of the baseband single-chirp waveform. By introducing $V_{\text{MZM-y}}(t)$ into the transmission function of the MZM (2), when the MZM-y is biased at the minimum transmission point, under small-signal modulation condition, the output electrical signal can be written by

$$\begin{aligned} V_{\text{out-y}}(t) &\propto 4J_1(\beta_1) J_1(\beta_4) [\cos(\omega_1 t + kt^2 + \varphi_1) \\ &\quad + \cos(\omega_1 t - kt^2 + \varphi_1)] \\ &\quad + 4J_1(\beta_2) J_1(\beta_4) [\cos(\omega_2 t + kt^2 + \varphi_2) \\ &\quad + \cos(\omega_2 t - kt^2 + \varphi_2)] \\ &\quad + 2J_1^2(\beta_1) \cos(2\omega_1 t + 2\varphi_1) + 2J_1^2(\beta_2) \cos(2\omega_2 t + 2\varphi_2) \\ &\quad + 4J_1(\beta_1) J_1(\beta_2) \cos((\omega_1 - \omega_2)t + \varphi_1 - \varphi_2) \end{aligned} \quad (7)$$

where $\beta_i = \pi V_i / V_{\pi-y}$ ($i = 1, 2, 4$), and $V_{\pi-y}$ is the half-wave voltage of the MZM-y. As shown in (7) and Fig. 1(c), a dual-band dual-chirp LFM signal centered at f_1 and f_2 is generated. Since the dual-frequency signal used as the LO is phase-coherent and both frequency components modulate the same baseband single-chirp waveform, the resulting signal preserves interband phase coherence [19], [20]. In addition, the key waveform parameters, such as bandwidth, duty cycle,

and repetition period, can be flexibly adjusted by tuning the properties of the baseband chirp signal.

III. EXPERIMENTAL RESULTS

The experimental setup, shown in Fig. 1(a), is designed to validate the proposed dual-band dual-chirp microwave signal generation scheme. A continuous-wave LD (NXDFB-1550-100-MOD) operating at 1550 nm with an output power of 21 dBm serves as the optical carrier source. The output is directed through a PC and fed into a commercial DP-BPSK modulator (Fujitsu FTM7980EDA) with a 3-dB modulation bandwidth of ~ 20 GHz. The two orthogonally polarized outputs from the modulator are separated by a PBS. The X-polarized path forms the OEO loop, which comprises a 2-km SMF, a high-speed photodetector (PD1, DSC40S, 3-dB bandwidth: 18 GHz), and a low-noise amplifier (Marki APM-6849PA) to compensate loop losses and sustain oscillation. A DB-BPF (center frequencies: 10 and 16 GHz; 3-dB bandwidth: 50 MHz) selects the desired oscillation modes. A single-frequency microwave signal from an external source (Agilent E8257D) is combined with the internally generated dual-frequency signal via an RF coupler and injected into the loop through MZM-x, enabling mutual injection locking in a closed-loop configuration. The resulting phase-coherent dual-frequency signal and a baseband single-chirp waveform generated by an AWG (Tektronix AWG700001A) are simultaneously applied to MZM-y via an electrical coupler. The modulated Y-polarized optical signal is separated again by a PBS, detected by PD2, and converted into an electrical waveform. This output is analyzed using a real-time oscilloscope and a phase noise analyzer (R&S FSWP 26.5) to evaluate its temporal and spectral characteristics.

Without external injection, the OEO loop supports only single-frequency oscillation at 10 GHz due to gain competition among cavity modes. When a 6-GHz microwave signal with a power of 4 dBm is injected, nonlinear mixing with the existing 10-GHz tone generates a 16-GHz component via frequency upconversion. This component injection-locks the longitudinal modes within the 16-GHz passband of the dual-band filter, enabling stable dual-frequency oscillation at 10 and 16 GHz through mutual injection locking. Fig. 2(a) shows the electrical spectrum of the dual-frequency output, clearly revealing two distinct tones centered at 10 and 16 GHz. Fig. 2(b) and (c) provides the zoomed-in views around 10 and 16 GHz, respectively, both exhibiting SMSRs exceeding 71 dB, indicating high spectral purity and effective suppression of spurious modes. Compared to the free-running single-frequency case, this represents an improvement of ~ 36 dB in SMSR, as highlighted by the blue dashed line in Fig. 2(b). The insets in Fig. 2(b) and (c) present the measured frequency responses of the dual-band filter obtained using a vector network analyzer (R&S ZNB40), showing ~ 50 -MHz 3-dB bandwidth for each passband. The 2-km SMF in the loop yields an FSR of ~ 100 kHz, much narrower than the filter bandwidth, allowing multiple longitudinal modes within each passband and thus poor SMSR without injection locking.

Fig. 3 depicts the phase noise spectra of the generated dual-frequency signals at 10 and 16 GHz. The red

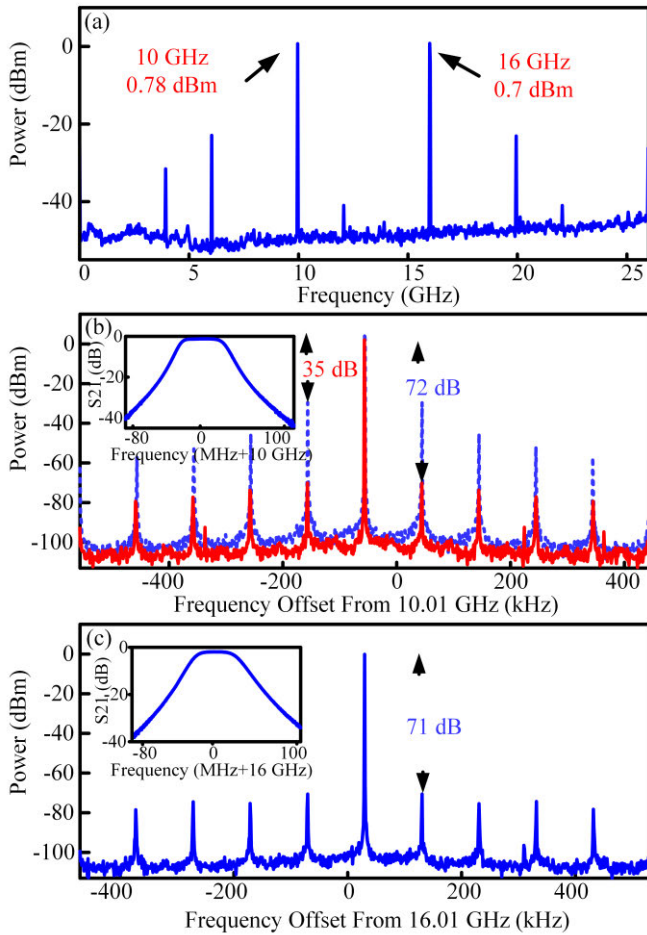


Fig. 2. (a) Electrical spectrum of the generated dual-frequency signal. (b) and (c) Zoomed-in views of the spectral components centered at 10 and 16 GHz, respectively, within a 1-MHz span.

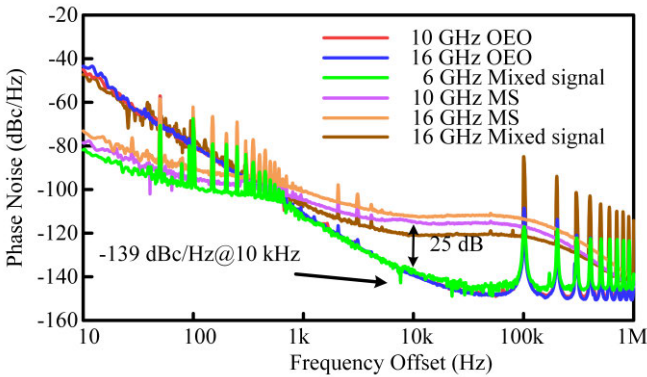


Fig. 3. Phase noise of the dual-frequency signals generated by the proposed OEO (red: 10 GHz; blue: 16 GHz), the mixed signal (green), the 16-GHz signal obtained by mixing the OEO-generated 10-GHz signal with an external 6-GHz source (brown), and by external microwave sources (purple: 10 GHz; orange: 16 GHz).

and blue traces correspond to the OEO-generated 10- and 16-GHz signals, respectively, while the purple and orange traces correspond to signals from commercial microwave sources (Agilent E8257D). The green trace shows the phase noise of the intermediate frequency (IF) signal obtained by mixing the two OEO-generated tones. Both OEO-generated

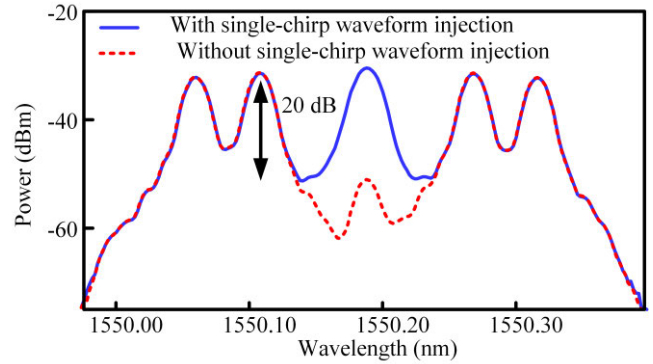


Fig. 4. Measured optical spectra at the output of MZM-y. The red dashed curve corresponds to modulation without the injected baseband LFM waveform, while the blue solid curve represents modulation with the baseband LFM signal applied.

signals exhibit nearly identical phase noise, reaching -139 dBc/Hz at a 10-kHz offset. Relative to the commercial sources, the OEO achieves an improvement of ~ 25 dB in phase noise, confirming its effectiveness in generating low-noise dual-frequency carriers. Such a reduction is critical for improving the phase noise of the generated LFM signals, directly benefiting radar detection accuracy, resolution, and robustness. To verify that the 16-GHz component generated by the OEO is a genuine oscillation mode enabled by mutual injection locking—rather than a simple nonlinear mixing product—we additionally measured the 16-GHz frequency component obtained by mixing the free-running 10-GHz signal with a 6-GHz signal from an external microwave source. The resulting phase noise, shown as the brown trace in Fig. 3, provides a clear comparison. Relative to this mixed 16-GHz signal, the OEO-generated 16-GHz component exhibits significantly lower phase noise at a 10-kHz offset, where the far-offset noise behavior is governed by the intrinsic phase noise of the free-running OEO. Meanwhile, the near-offset phase-noise behavior of the OEO-generated 16-GHz tone closely matches that of the mixed signal, consistent with the prediction of (5). This agreement confirms that the 16-GHz tone in the proposed dual-frequency OEO is produced through injection locking of the corresponding mode by the mixing-induced component, rather than being a mere nonlinear artifact. Furthermore, coherence between the two OEO-generated tones is verified by filtering them separately and applying them to the LO and RF ports of a microwave mixer. The resulting IF signal (green trace) exhibits a ~ 38 -dB phase noise improvement at a 10-Hz offset, demonstrating strong phase correlation between the two carriers.

The dual-frequency signal is fed into MZM-y, biased at the minimum transmission point to perform carrier-suppressed double-sideband modulation. The Y-polarized modulated optical signal is extracted using a PBS, passed through a 99:1 optical coupler (OC), and analyzed using an optical spectrum analyzer (YOKOGAWA AQ6370D). As shown by the red dashed curve in Fig. 4, the optical spectrum exhibits the first-order sidebands at 10 and 16 GHz, with an optical sideband suppression ratio of 20 dB. When the baseband single-chirp waveform generated by the AWG is also applied to

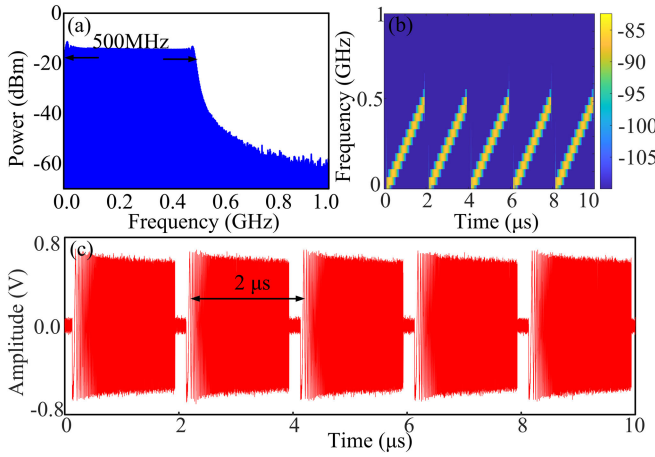


Fig. 5. (a) Electrical spectrum of the baseband single-chirp waveform within a 1-GHz span. (b) Instantaneous frequency-time diagram. (c) Temporal waveform of the baseband single-chirp signal.

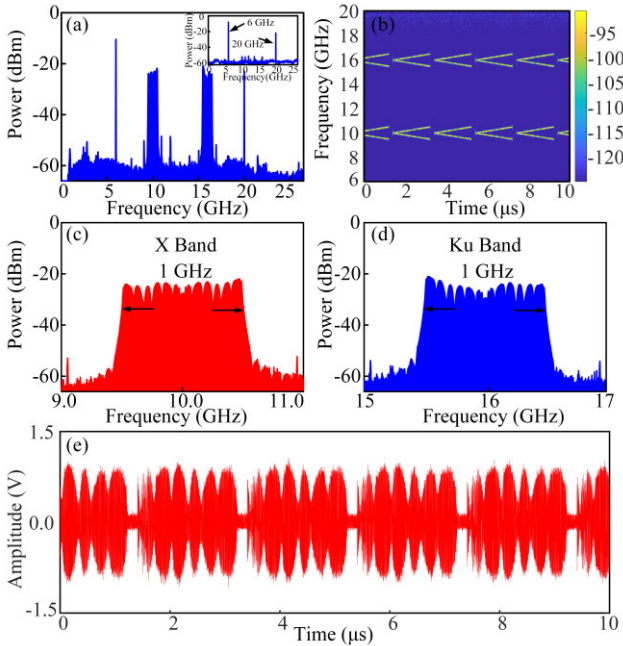


Fig. 6. (a) Electrical spectrum of the generated dual-band dual-chirp LFM signal within a 26-GHz span, (b) instantaneous frequency-time diagram, and (c) temporal waveform of the dual-band dual-chirp LFM signal. (c) and (d) Zoomed-in views of the electrical spectrum within a 2-GHz span.

MZM-y along with the dual-frequency signal, the resulting optical spectrum is shown as the blue curve in Fig. 4. Owing to the low-frequency nature of the baseband chirp, the generated optical sidebands are located closer to the optical carrier.

The injected baseband waveform is an up-chirp LFM signal with a center frequency of 250 MHz, a bandwidth of 500 MHz, a period of 2 μ s, and a duty cycle of 90%. Its measured spectrum, instantaneous frequency-time profile, and temporal waveform are shown in Fig. 5(a)–(c), respectively. Fig. 6(a) presents the spectrum of the generated dual-band dual-chirp LFM signal, clearly showing two wideband components centered at 10 and 16 GHz. The inset in Fig. 6(a) shows the microwave spectrum when only the dual-frequency signal is applied to MZM-y without the baseband injection,

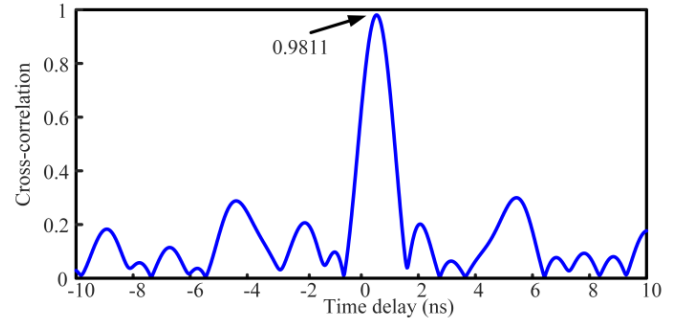


Fig. 7. Normalized cross correlation of the dual-band LFM signals.

corresponding to the red dashed curve in Fig. 4. With MZM-y biased at the minimum transmission point, nonlinear mixing between the 10- and 16-GHz components produces a 6-GHz IF signal and second-order harmonics, as described in (5). Due to the PD bandwidth limitation, only the second harmonic of the 10-GHz component is visible. Low-frequency spurious components are also observed, attributed to the limited extinction ratio of MZM-y. Using a DP-BPSK modulator with a higher extinction ratio is expected to further suppress these undesired components and improve signal quality. The undesired spurious components can also be effectively suppressed by employing a properly designed bandpass filter (6–20 GHz), thereby further improving the output signal quality. Fig. 6(c) and (d) provides the zoomed-in view spectra near each carrier, confirming that both LFM signals occupy $\sim$ 1-GHz bandwidths. Fig. 6(b) and (e) depicts the instantaneous frequency-time diagram and temporal waveform of the dual-band dual-chirp signal, respectively.

To quantitatively evaluate the coherence of the dual-band LFM waveforms, a cross correlation analysis is performed. The dual-band signal is first separated using two bandpass filters centered at 10 and 16 GHz, each with a 1.5-GHz bandwidth, and the filtered components are fed into channels 1 and 2 of an oscilloscope (Tektronix DPO73304SX). Both channels are synchronously acquired at a sampling rate of 50 GS/s under a common trigger, ensuring a consistent temporal reference. Since the two waveforms have different center frequencies, the recorded signals are multiplied by complex exponentials at their respective carrier frequencies to accomplish digital down-conversion to baseband. The resulting cross correlation, shown in Fig. 7, exhibits a peak value of 0.9811, demonstrating the high coherence of the generated dual-band dual-chirp LFM waveforms. The slight offset of the correlation peak from zero delay is mainly attributed to the small group-delay mismatch between the two hardware paths, including differences in filter group delay, unequal electrical lengths of the RF cables, and the inherent interchannel sampling delay of the oscilloscope.

Fig. 8(a) and (c) shows the spectra of the proposed LFM signals at 10 GHz [red curve in Fig. 8(a)] and 16 GHz [blue curve in Fig. 8(c)], both generated using the mutual injection-locked dual-frequency LO. It can be observed that in both frequency bands, the LFM signals exhibit SIRs exceeding 65 dB, indicating that the high-SMSR phase-coherent dual-frequency LO preserves spectral purity during modulation.

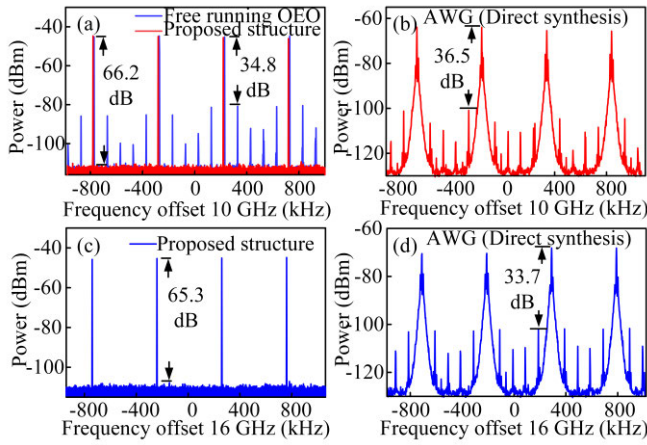


Fig. 8. Spectral details of LFM signals generated by the proposed system and by direct AWG synthesis. (a) X-band LFM signal generated using the mutual injection-locked dual-frequency signal as the LO (red) and using a free-running OEO as the LO (blue). (b) X-band LFM signal directly generated by an AWG. (c) Ku-band LFM signal generated using the mutual injection-locked dual-frequency signal as the LO (blue). (d) Ku-band LFM signal directly generated by an AWG. RBW: 10 Hz.

Notably, the SIR values are even lower than the SMSR of the dual-frequency signals themselves, confirming that the high SMSR of the coherent dual-frequency source does not degrade the SIR of the generated LFM signals. For comparison, the blue curve in Fig. 8(a) shows the spectrum of a 10-GHz LFM signal generated using a free-running OEO, which suffers from spectral degradation with an SIR of only 34.8, 31 dB lower than that achieved with the proposed scheme, highlighting its effectiveness in suppressing spurious components. The SIR is a critical parameter in radar systems, as it directly affects detection accuracy, false alarm rate, and range resolution [36], [37]. Fig. 8(b) and (d) shows the spectra of 10- and 16-GHz LFM signals directly synthesized by an AWG, with SIRs of 36.5 and 33.7 dB, respectively, about 29 dB lower than those of the proposed scheme. Moreover, compared with AWG-generated signals, those from the proposed structure exhibit significantly narrower spectral linewidths, as seen by comparing Fig. 8(a) and (c) with (b) and (d). Narrow-linewidth microwave signals enhance radar range resolution and velocity accuracy, enable precise target detection and localization, reduce false alarms, and improve interference immunity, making them highly desirable for high-precision radar applications in complex environments [38], [39], [40].

Since the dual-band LFM signals are mutually phase-coherent, their phase fluctuations are strongly correlated. In addition, both frequency bands propagate through the same hardware components and therefore experience nearly identical environmental perturbations. To evaluate the stability of the generated signals, the 10-GHz frequency-comb line was selected as a reference, and its frequency drift and power fluctuation were recorded every 30 s over a 10-min interval. As shown in Fig. 9, the blue curve represents the measured frequency drift, while the red curve denotes the corresponding power fluctuation. During the entire 10-min observation period, the frequency drift of the output LFM signal remained below 4.5 kHz, and the power variation was less than

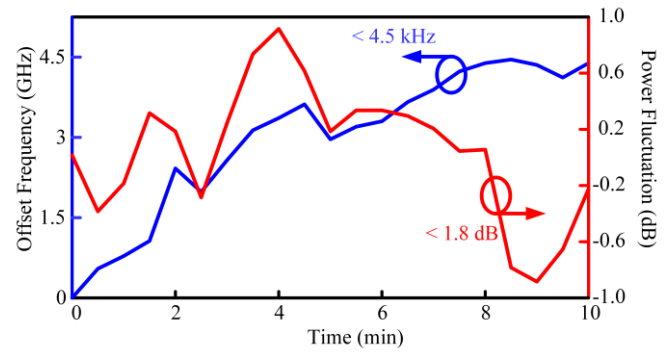


Fig. 9. Measured long-term stability of the generated LFM signal, including frequency drift (blue) and power fluctuation (red) recorded every 30 s over a 10-min interval.

1.8 dB. Such variations primarily arise from thermally induced expansion and contraction of the fiber caused by ambient temperature changes, as well as perturbations to polarization-dependent components, such as the PC, PBS, and PBC. To further suppress environmental impacts, a temperature-controlled enclosure can be employed to thermally stabilize both the optical fiber and the polarization-dependent components, thereby effectively reducing temperature-induced output fluctuations.

An LFM waveform with tunable duty cycle and pulse-repetition period/frequency can be optimized for range resolution, detection capability, and interference suppression, improving system flexibility and adaptability. In the proposed scheme, the characteristics of the high-frequency dual-band dual-chirp LFM signal can be flexibly tailored by adjusting the parameters of the injected baseband LFM signal. For example, by varying the duty cycle of the injected low-frequency LFM signal to 10% and 50%, the time-frequency characteristics of the resulting dual-band dual-chirp LFM signals are obtained, as shown in Fig. 10(a) and (c), with their corresponding time-domain waveforms presented in Fig. 10(b) and (d). In addition, the period of the generated dual-band dual-chirp LFM signal can be controlled by modifying the period of the injected low-frequency LFM signal. Fig. 10(e) and (g) shows the time-frequency diagrams of the output signals corresponding to baseband LFM periods of 10 and 5 μ s, while their time-domain waveforms are given in Fig. 10(f) and (h). These results verify that the proposed system can dynamically control key waveform parameters, such as duty cycle and period, enabling reconfigurable signal generation to meet a wide range of application demands.

Furthermore, the wideband LFM signals are essential for enhancing radar resolution and improving target discrimination [15]. In the proposed system, increasing the bandwidth of the injected baseband chirp directly broadens the output bandwidth in both frequency bands. For instance, when the AWG-generated chirp has a center frequency of 1 GHz and a bandwidth of 2 GHz, the corresponding spectrum, time-frequency diagram, and time-domain waveform of the generated dual-band dual-chirp signal are shown in Fig. 11(a), (c), and (e), respectively, where the LFM bandwidth in each band increases to 4 GHz. When the center frequency is

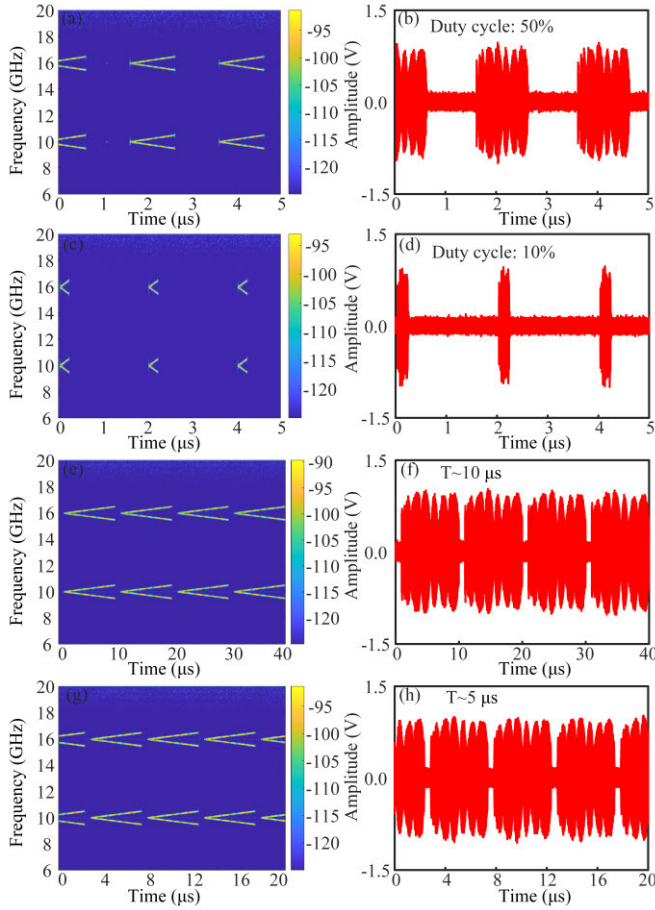


Fig. 10. Demonstration of tunable duty cycle and period in the generated dual-band dual-chirp LFM waveforms. (a) and (c) Instantaneous frequency–time diagrams of waveforms generated with baseband duty cycles of 10% and 50%, respectively. (b) and (d) Corresponding time-domain waveforms. (e) and (g) Instantaneous frequency–time diagrams with baseband periods of 10 and 5 μ s, respectively. (f) and (h) Corresponding time-domain waveforms.

further increased to 1.5 GHz with a bandwidth of 3 GHz, the generated dual-chirp waveform reaches its widest nonoverlapping configuration, achieving a total spectral span of 12 GHz. The corresponding spectrum, time–frequency diagram, and time-domain waveform are shown in Fig. 11(b), (d), and (f), respectively. These results demonstrate that the proposed system supports coherent dual-band waveform generation with reconfigurable bandwidth, significantly improving radar range resolution, target detection capability, and overall system performance.

IV. CONCLUSION

In this work, a coherent dual-band dual-chirp microwave waveform generation scheme based on a polarization-multiplexed OEO has been proposed and experimentally demonstrated. By incorporating a dual-passband filter into the OEO loop and employing mutual injection locking, stable dual-frequency oscillations with low phase noise and high SMSR are achieved. The generated dual-frequency signal is used as an LO to modulate an externally injected baseband chirp waveform, enabling the generation of coherent dual-band dual-chirp LFM signals. The proof-of-concept experiments

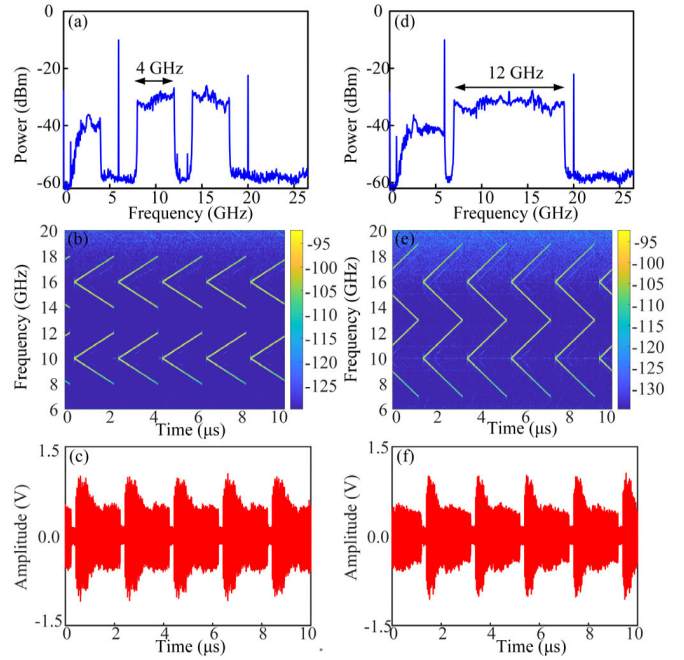


Fig. 11. Demonstration of bandwidth tunability in the generated dual-band dual-chirp LFM waveforms. (a) and (b) Spectral diagrams showing 4-GHz bandwidth per band when a 2-GHz baseband chirp is injected (center frequency: 1 GHz). (c) Corresponding time-domain waveform. (d) and (e) Spectral diagrams showing a 12-GHz total span when a 3-GHz baseband chirp is injected (center frequency: 1.5 GHz). (f) Corresponding time-domain waveform.

verify the successful generation of dual-band LFM waveforms centered at 10 and 16 GHz with tunable bandwidths, duty cycles, and repetition periods. Compared to a free-running OEO, the proposed scheme improves the SMSR by 36 dB and enhances the SIR of the generated LFM signals by 31 dB. Furthermore, relative to conventional waveform generation using an AWG, the proposed approach achieves a 29-dB improvement in SIR and a significantly narrower spectral linewidth. The output bandwidth can be flexibly extended up to 12 GHz by adjusting the baseband chirp, demonstrating high configurability and spectral scalability. With its compact architecture, strong phase coherence, and excellent spectral purity, the proposed method shows great potential value for modern multiband and multifunctional radar systems.

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