Research and design of a low phase-noise dielectric resonator oscillator for X-band radar applications

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ABSTRACT

This paper presents the research, design, and implementation of a low phase-noise dielectric resonator oscillator (DRO) operating at 9.4 GHz for X-band radar applications. A high-quality-factor dielectric resonator operating in the $TE_{10\delta}$ *mode was employed to enhance frequency stability and suppress phase noise. The oscillator was realized using a microstrip coupling structure, with a GaAs FET transistor serving as the active element in the feedback network. Comprehensive electromagnetic and circuit-level simulations were conducted in CST and ADS to optimize the resonator coupling and feedback configuration, ensuring stable oscillation and effectively mitigating spurious modes. Compared with conventional oscillator designs, the proposed DRO demonstrates superior frequency stability and reduced phase noise, while maintaining a compact form factor. These characteristics render the proposed design highly suitable for modern X-band radar front-ends as well as other advanced microwave communication systems requiring low-noise and high-stability local oscillators.*

Keywords: Local oscillator; X-band radar; Phase noise figure; Dielectric resonator.

1. INTRODUCTION

With the rapid development of science and technology, increasingly higher requirements are being placed on the components that make up radar systems. The local oscillator is an indispensable part of modern radar systems. Its function is to generate a signal with a specific frequency, used for mixing with the received or transmitted signals in the radar system. However, ensuring that the local oscillator operates with high frequency stability, low phase noise, and compact size is a complex problem. Currently, many research results on local oscillators have been published in reputable journals, using various solutions to improve performance—such as dielectric resonators, phase-locked loops (PLLs), and bandwidth optimization of loop filters. In [1–3], the optimal solution involves selecting active components with high output power; however, the generated frequency deviation is large. In [4–12], the approach focuses on designing dielectric resonators with compact size and relatively low phase noise, but the resulting frequency deviation is still significant—up to 23.8% compared to the center frequency. Such large frequency deviations in local oscillators significantly affect the processing and operational quality of radar systems. Therefore, in this paper, the authors propose a local oscillator design with a frequency-tunable resonant element and optimized active components to minimize phase noise. The proposed local oscillator ensures compact size, high frequency stability, and low phase noise.

2. PROBLEMS

2.1. Design of the resonant element and tuning of the resonator's central resonant frequency

A dielectric resonator (DR) is coupled to a microstrip line as illustrated in figure 1. The dielectric resonator is placed on the substrate surface at a distance *d* from the microstrip line. The distance *d*, together with the physical characteristics of the DR, determines the coupling level between the two elements. To minimize radiation losses, a metallic enclosure is used around the structure, which helps enhance the quality factor (Q) of the resonator. In this configuration, the $TE_{10\delta}$ mode (figure 2) is excited in the DR by the electromagnetic field generated by the microstrip line. In turn, the DR reflects RF energy at its resonant frequency, resulting in a high-Q resonator.

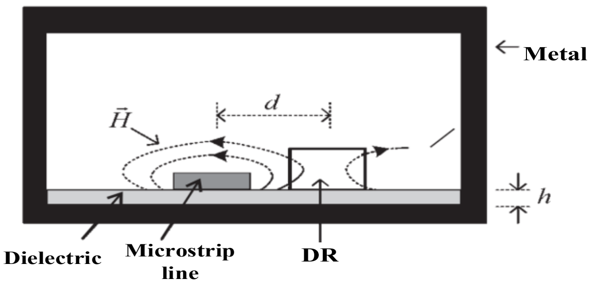


Figure 1. *Coupling between the dielectric resonator and the microstrip line.*

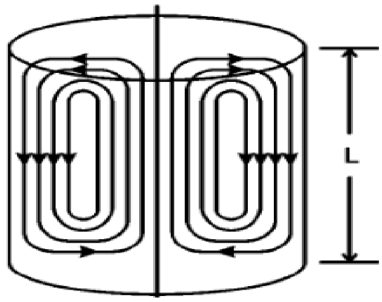


Figure 2. *Electromagnetic field distribution of the $TE_{10\delta}$ mode.*

The resonant frequency of an unshielded dielectric resonator can be approximately calculated using the following equation:

$$f_{\mathrm{GHz}} = \frac{34}{a\sqrt{\varepsilon_r}}\left(\frac{a}{L} + 3.45\right) \quad (1)$$

where a is the radius of the dielectric resonator (mm), L is its height (mm), and ε_r is the dielectric constant of the DR. To ensure that the $TE_{10\delta}$ resonant mode is not affected by other resonance modes, the height L of the DR should be between 35% and 45% of the element's diameter. If it falls outside this range, frequency overlap with other resonant modes may occur. Table 1 presents the basic dimensions of the dielectric resonator.

Table 1. *Basic dimensions of the dielectric resonator.*

№	Parameters	Value (mm)
1	L	2.6
2	a	2.75
3	d	2.25
4	h	2.6
5	ε_r	38

The resonant frequency of the DR may not exactly match the simulation results, since any variation in the radius, height, or dielectric constant of the DR can alter the resonant frequency. Therefore, it is essential to minimize fabrication errors to ensure the highest possible accuracy. To maintain the desired resonant frequency, appropriate tuning techniques must be applied.

A DR resonant element with the capability to tune its center frequency using a mechanical tuning screw is shown in figure 3. In this configuration, the resonant frequency of the DR can be adjusted over a narrow frequency range through the use of a metallic shield combined with the tuning screw. In principle, the tuning screw changes the effective height h, which directly affects the resonant frequency. Using this method, the tuning range of the resonant frequency can reach from 0.1% to 1% relative to the original frequency value.

In addition to mechanical tuning, in this paper, the resonant frequency of the DR is also adjusted electrically. The varactor-tuned dielectric resonator is shown in figure 4. In this configuration, the varactor diode is coupled to the DR, forming two coupled circuits. The capacitance of the diode can be varied by applying an appropriate DC voltage, thereby changing the resonant frequency of the DR. This electrical tuning method can provide a resonant frequency adjustment range of approximately 1%.

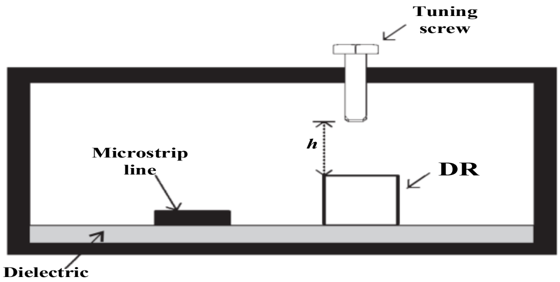


Figure 3. *Tuning the resonant frequency of the DR using a mechanical tuning screw.*

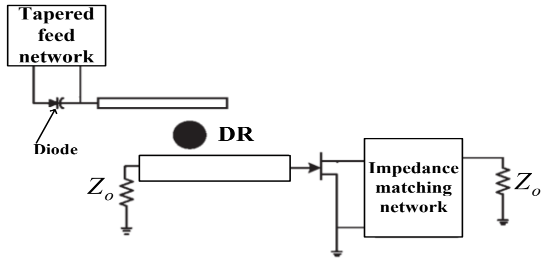


Figure 4. *Tuning the resonant frequency of the DR using a varactor diode.*

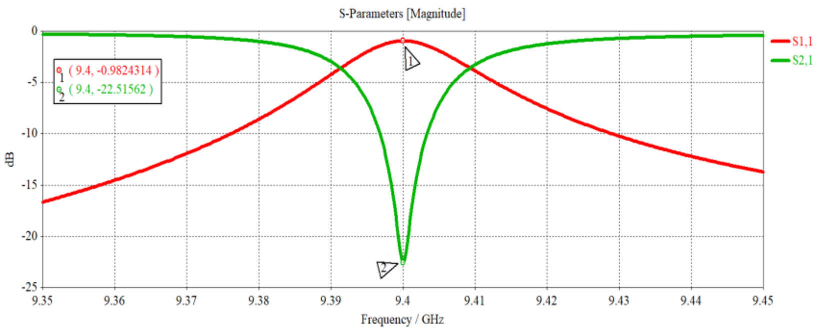


Figure 5. *S-parameters of the dielectric resonator.*

2.2. Design of the active element in the dielectric resonator oscillator

The negative-resistance method is commonly used in the design of microwave oscillators. This approach ensures a high success rate by accounting for the nonlinear characteristics of the active element employed. Calculations and design can be based on the scattering parameters of the components in the oscillator circuit. A model of an oscillator designed using the negative-resistance method is illustrated in figure 6, comprising three main components: the resonator, the active element, and the matching network.

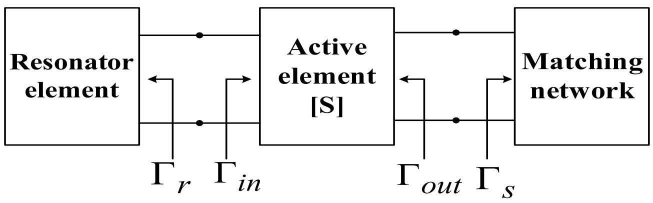


Figure 6. *Two-port oscillator model using the negative-resistance method.*

When designing an oscillator using the negative-resistance method, the parameters of the elements in the oscillator can be calculated from the oscillation conditions using the following formulas:

$$\Gamma_r \Gamma_{in} = 1; \tag{2}$$

$$|\Gamma_{in}| = \frac{1}{|\Gamma_r|} \tag{3}$$

$$\angle\Gamma_r = -\angle\Gamma_{in} \tag{4}$$

$$\Gamma_{out} = S_{22} + \frac{S_{21}S_{12}\Gamma_r}{1 - S_{11}\Gamma_r} \tag{5}$$

Since the reflection coefficient of the resonator is always less than or equal to 1 ($|\Gamma_r| \leq 1$) due to the passive nature of the element, from (2) it follows that $|\Gamma_{in}|$ must be greater than 1. If the input reflection coefficient is greater than 1, then the input impedance becomes negative, which is referred to as the negative-resistance method.

With the output reflection coefficient Γ_{out} calculated from (3), Γ_s can then be determined to satisfy the oscillation condition at the output, $\Gamma_{out}\Gamma_s = 1$. However, to maximize the output power of the oscillator circuit, the impedance of the matching network Z_s must satisfy the condition [12]:

$$|Z_S| = \frac{-|Z_{OUT}|}{3}\ ;\ \angle Z_S = -\angle Z_{OUT} \quad (6)$$

Series feedback in a dielectric resonator oscillator

Either series or parallel feedback can be used to induce instability in a transistor. In this paper, the series-feedback method is applied due to its simple structure. A transmission line is connected to the Source terminal of the transistor, as shown in figure 7 - the dielectric resonator oscillator model with a series-feedback configuration. The transmission line sections coupling elements TL1 to TL5 are microstrip lines with a characteristic impedance of 50 Ω. By adjusting the lengths of transmission lines TL2 and TL3, negative resistance can be achieved, inducing the required instability for the transistor.

First, the dielectric resonator element, after being simulated and optimized in CST software, is modeled in ADS software using its scattering parameters Γ_r. From the scattering parameters of the resonator, the reflection coefficient can be easily calculated. The length of TL3 is adjusted to satisfy condition (3), while the length of TL2 is adjusted to satisfy condition (4). However, to generate oscillations, the active element must operate in the unstable region [12]. Therefore, the adjustments of TL2 and TL3 to satisfy conditions (3) and (4) must also ensure that the transistor operates in this unstable region. If this condition is not met, the coupling coefficient of the dielectric resonator must be modified to achieve the optimal system performance.

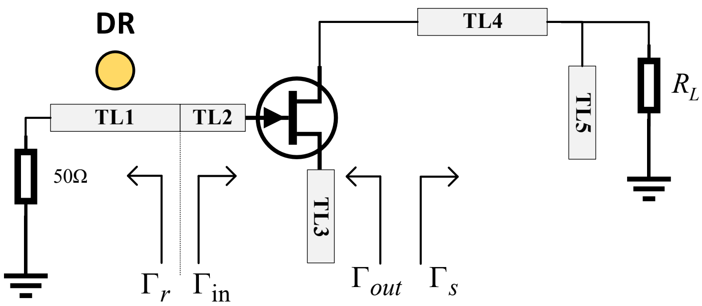


Figure 7. *Series-feedback dielectric resonator oscillator structure.*

In this paper, the authors selected the HEMT transistor FHX35LG from Fujitsu as the active element for the dielectric resonator oscillator. The microstrip circuit is designed on a Rogers 4350B substrate with a thickness of 0.76 mm and a dielectric constant of $\varepsilon_r = 3.66$. After optimization through simulation, the obtained values are: the reflection coefficient of the dielectric resonator, $\Gamma_r = 0.78\angle 59.45°$, and the input reflection coefficient required for the transistor, $\Gamma_{in} = 1.282\angle -59.45°$. With these reflection coefficient values, the transmission lines with a characteristic impedance of 50 Ω (TL2 and TL3) were determined to have a width of $w = 1.76$ mm and lengths of $d_2 = 18.21$ mm and $d_3 = 9.05$ mm, respectively. Figure 8 shows the simulated calculation results of Γ_{in}, and figure 9 illustrates that the transistor operates in the unstable region. Within the simulated frequency range, the Mu parameter is consistently less than 1 [12], and at the target design frequency of 9.4 GHz, Mu equals −0.275, ensuring that the transistor operates in the unstable region.

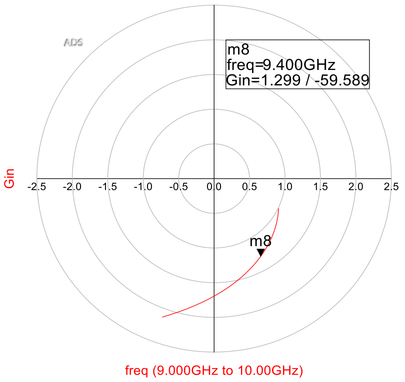


Figure 8. *Optimized simulation results of the input reflection coefficient Γ_{in}.*

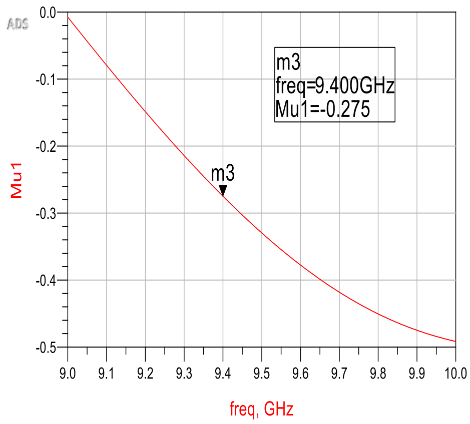


Figure 9. *Transistor stability after adding transmission line sections TL2 and TL3.*

Next, it is necessary to calculate the impedance at the transistor output and determine the required impedance for the matching network. Formula (5) can be applied, or more simply, ADS software can be used to determine the value of Zout. The calculated result is $Z_{out} = -89.5 + j12.5\ \Omega$.

There are several possible structures for designing a suitable impedance-matching network. One simple configuration is a transmission-line-based network combined with an open-circuit stub. The transmission line sections with a characteristic impedance of 50 Ω are labeled TL4 and TL5, as illustrated in figure 7. With an output load of RL = 50 Ω, the physical dimensions of TL4 and TL5 can be calculated or determined using ADS. The resulting dimensions are: width $w = 1.63$ mm, and lengths $d_4 = 9.54$ mm and $d_5 = 1.71$ mm.

3. RESULTS AND DISCUSSION

3.1. Checking the oscillation conditions

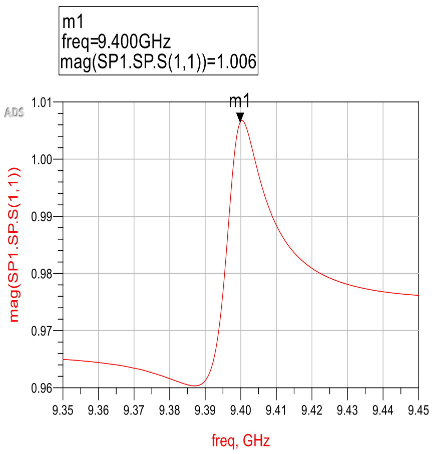


Figure 10. *Amplitude response results using the OscTest tool.*

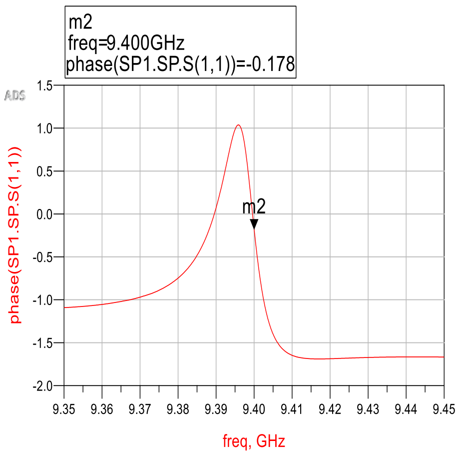


Figure 11. *Phase response results using the OscTest tool.*

In ADS, the OscTest tool can be used to verify how well the oscillator circuit satisfies the oscillation conditions. Figures 10 and 11 show the amplitude and phase responses of S(1,1) obtained from OscTest. At the frequency of 9.4 GHz, the results are: mag(SP1.SP.S(1,1)) = 1.006 ≈ 1 and phase(SP1.SP.S(1,1)) = 0.21 ≈ 0°, satisfying the oscillation startup condition. Figure 12 shows the Nyquist criterion response of the oscillator simulated using OscTest in ADS. The plot encircles the point 1 + j0 clockwise as the frequency varies around the center frequency from 9.35 GHz to 9.45 GHz.

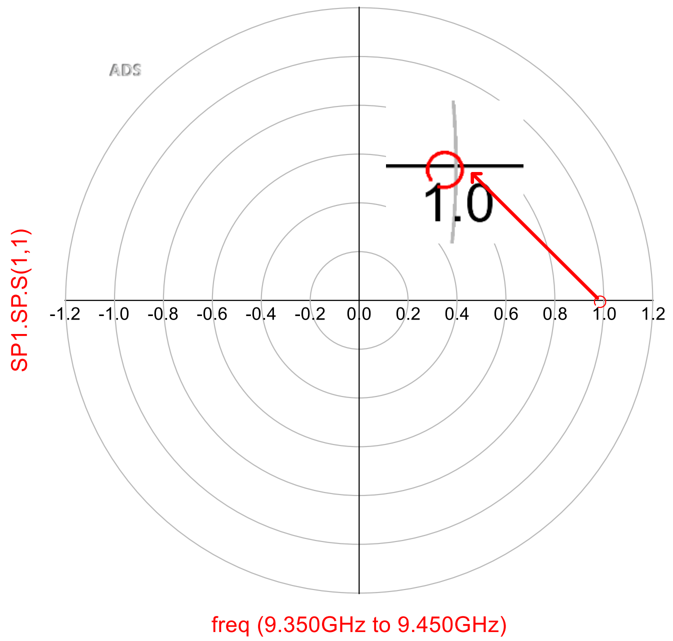
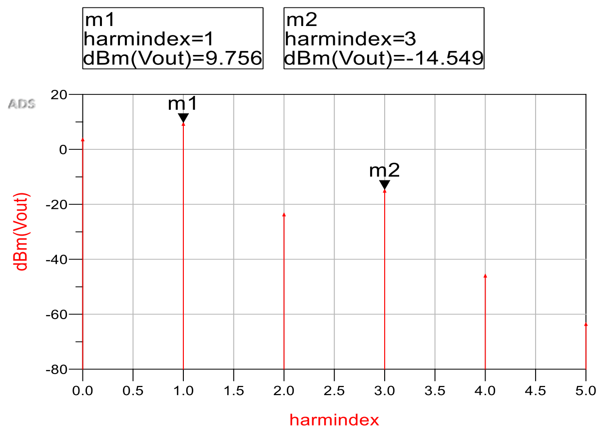


Figure 12. *Nyquist criterion response of the dielectric resonator oscillator.*

3.2. Output signal results: power, frequency, and phase noise

Figure 13 shows the simulated output of the generated oscillation signal. The results indicate that the dielectric resonator oscillator achieves high frequency stability and minimal frequency error, fully meeting the design requirements. Specifically, the output power is 9.756 dBm at a frequency of 9.398 GHz, corresponding to a frequency deviation of 2 MHz. This is characteristic of the negative-resistance method used in the oscillator design calculations. In practice, mechanical or electronic tuning screws are used to adjust the oscillation frequency to the desired target value.



harmindex	freq
0	0.0000 Hz
1	9.398 GHz
2	18.80 GHz
3	28.19 GHz
4	37.59 GHz

(a) (b)

Figure 13. *Simulated output of the generated oscillation signal.*

Figure 14 shows the simulated phase noise characteristic of the generated oscillation signal. The results indicate that the high-Q dielectric resonator oscillator exhibits very low phase noise, fully meeting the design requirements. Specifically, the phase noise is −122.5 dBc/Hz at 10 kHz and −142.4 dBc/Hz at 100 kHz.

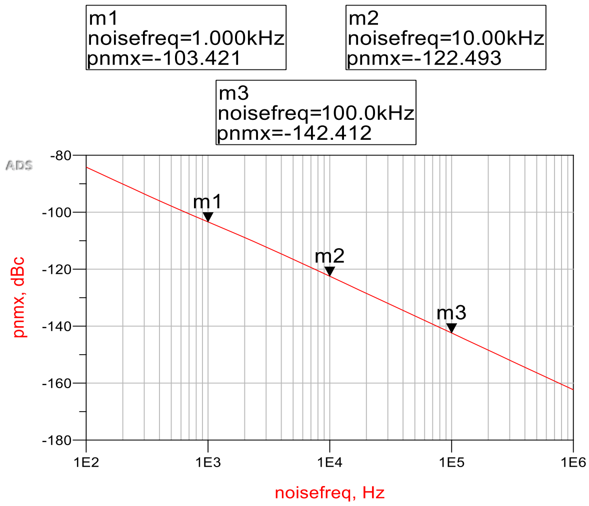


Figure 14. *Simulated phase noise characteristic of the local oscillator.*

Tables 2 and 3 present the calculated results of the local oscillator and compare them with results reported in several articles published in reputable international journals.

Table 2. *Technical parameters of the local oscillator.*

№	Technical parameters	Unit	Required value	Simulation results
1	Output signal frequency	GHz	9.4	9.398
2	Output power	dBm	≥ 5	9.756
3	Phase noise coefficient			
3.1	At 10 kHz	dBc/Hz	≤-95	-122.5
3.2	At 100 kHz	dBc/Hz	≤-110	-142.4

Table 3. *Comparison with selected published local oscillators.*

№	Frequency, GHz	Frequency offset produced, MHz	Output Power, dBm	Phase noise coefficient, dBc/Hz		FOM
				10 kHz	100 kHz	
[1]	10.7	3800	24.975	-	-145.2	-218.3
[2]	10.5	40	35	-	-91.7	-237.2
[3]	9.5	56	32.3	-	-87	-136.2
[4]	8	10	11	-	-	-219.6
[5]	9.8	18.9	8.5	-	-102.2	-219.7
[6]	10	2380	15.6	-	-105.3	-190.5
[7]	12	5	7	-92.5	-112.8	-202.6
[8]	10	610	8.9	-	-102.72	-169.2
[9]	8.5	62	8.1	-	-	-
[10]	9.4	45	18.82	-	-133.5	-216.0
[11]	9.6	43	4.7	-83.8	-123.9	-211.8
Article	**9.4**	**2**	**9.756**	**-122.5**	**-142.4**	**-227.1**

3.3. Analysis of the simulation results

The calculation results show that the technical parameters meet the specified requirements. To provide an objective evaluation, the authors compared the calculated results of the proposed local oscillator with several results reported in reputable international journals. From this comparison, it can be seen that the proposed local oscillator not only significantly reduces frequency error compared to previously published results but also maintains low phase noise and a compact size.

4. CONCLUSIONS

In this paper, the authors have presented a general overview of the design and calculation steps for a local oscillator for X-band radar. The calculation results show that the proposed local oscillator operates with high frequency stability, low phase noise, and a compact size. These results provide a solid basis for fabricating the local oscillator for application in radar systems across different frequency bands.

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TÓM TẮT

Nghiên cứu, thiết kế bộ tạo dao động ngoại sai tạp pha thấp cho ra đa băng X

Bài báo này trình bày kết quả nghiên cứu, thiết kế và chế tạo một bộ tạo dao động cộng hưởng điện môi (DRO) có nhiễu pha thấp, hoạt động ở tần số 9.4 GHz cho các ứng dụng ra đa băng X. Một bộ cộng hưởng điện môi có hệ số phẩm chất cao, hoạt động ở chế độ $TE_{01\delta}$*, đã được sử dụng nhằm đạt được độ ổn định tần số và giảm nhiễu pha. Bộ tạo dao động được thực hiện bằng cấu trúc ghép vi dải, trong đó, một transistor GaAs FET được sử dụng làm phần tử tích cực trong mạch hồi tiếp. Các mô phỏng điện từ trường và mạch đã được tiến hành trên phần mềm CST và ADS để tối ưu hóa ghép cộng hưởng và thiết kế hồi tiếp, đảm bảo dao động ổn định và hạn chế các chế độ ký sinh. So với các thiết kế bộ dao động thông thường, DRO đề xuất đạt được độ ổn định tần số cao hơn và nhiễu pha thấp hơn trong khi vẫn duy trì kích thước nhỏ gọn. Những đặc điểm này khiến thiết kế trở nên rất phù hợp cho các bộ thu phát ra đa băng X hiện đại và các hệ thống siêu cao tần khác đòi hỏi bộ tạo dao động ngoại sai có nhiễu thấp và độ ổn định cao.*

Từ khóa: Dao động ngoại sai; Ra đa băng X; Hệ số tạp pha; Cộng hưởng điện môi.