

Design of an RTD-Integrated Slot Antenna for THz Oscillators

Master's thesis in Wireless, Photonics and Space Engineering

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DEPARTMENT OF MICRO and NANO TECHNOLOGY

CHALMERS UNIVERSITY OF TECHNOLOGY

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Abstract

Terahertz (THz) oscillators are key components for future high-speed wireless communication systems, imaging, spectroscopy, and sensing applications. Resonant tunneling diodes (RTDs) are attractive THz sources due to their compact size, simple structure, and ability to generate oscillations at frequencies beyond the limits of conventional transistor technologies. To enable compact and easily integrated terahertz sources, RTDs are commonly combined with antenna structures that directly radiate the generated THz signals.

This thesis presents the design and analysis of an RTD-integrated slot antenna oscillator operating near 650 GHz. An equivalent-circuit-based methodology is first employed to estimate the antenna dimensions from the RTD admittance characteristics. The slot antenna is then analysed using full-wave electromagnetic simulations, and its admittance is combined with a nonlinear RTD model. Oscillation conditions are evaluated through small-signal admittance analysis, while harmonic balance and transient simulations are used to verify large-signal oscillation behaviour. A complete oscillator model including parasitic elements and finite substrate effects is subsequently developed, followed by geometry optimisation of the slot antenna.

The results show that the substrate primarily affects the real part of the oscillator admittance, while the imaginary part remains largely unchanged. Consequently, the oscillation frequency is preserved, whereas the output power is reduced due to decreased radiation efficiency. The optimised design achieves stable oscillation at approximately 646 GHz with an estimated radiated power of approximately 80 μ W using a 0.49 μm^2 RTD. The slot antenna exhibits a radiation efficiency of approximately 62%, and a peak directivity of approximately 6.8 dB without the use of a dielectric lens. The predicted radiated power is within the range reported for RTD oscillators operating in this frequency regime, although below the highest power reported for multi-element array implementations.

Keywords: resonant tunneling diode, terahertz oscillator, slot antenna, nonlinear circuit simulation, THz source

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Xiuting Shi, Gothenburg, June 2026

List of Acronyms

Below is the list of acronyms used throughout this thesis, listed in alphabetical order:

EM	Electromagnetic
FFT	Fast Fourier Transform
HB	Harmonic Balance
MIM	Metal–Insulator–Metal
NDC	Negative Differential Conductance
PEC	Perfect Electric Conductor
PML	Perfectly Matched Layer
RTD	Resonant Tunnelling Diode
S-parameter	Scattering Parameter
TDS	Time-Domain Solution
VNA	Vector Network Analyzer
DC	Direct Current

Nomenclature

Below is the nomenclature of symbols and parameters used throughout this thesis.

q	Electron charge
m^*	Effective electron mass
k_B	Boltzmann constant
$\hbar$	Reduced Planck constant
E_F	Emitter Fermi energy
E_r	Resonant energy level
Γ	Resonance width
$T(E, V)$	Transmission coefficient
J	Current density
Y	Admittance
G	Conductance
B	Susceptance
Y_{RTD}	RTD admittance
G_{RTD}	RTD conductance
B_{RTD}	RTD susceptance
C_{RTD}	RTD capacitance
Y_a	Antenna admittance
G_a	Antenna conductance

B_a	Antenna susceptance
G_r	Radiation conductance
G_l	Ohmic loss conductance
S	RTD mesa area
d_{eff}	Effective thickness of the RTD active region
ε_0	Vacuum permittivity
ε_r	Relative permittivity
f_{osc}	Fundamental Oscillation frequency
P_{out}	Output power
V_{bias}	Bias voltage
V_{out}	Amplitude of the fundamental Oscillation voltage amplitude
η_{rad}	Radiation efficiency

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1

Introduction

The Terahertz (THz) frequency range, typically defined between 0.3 THz and 10 THz, has attracted significant attention due to its unique properties and wide range of applications. These include high-speed wireless communication, spectroscopy, imaging, and sensing. In particular, THz radiation is capable of penetrating many non-metallic materials while maintaining relatively high spatial resolution, owing to its low absorption in many dielectric media and its shorter wavelength compared to microwave radiation [1]. In addition, measurements of molecular spectral lines at sub-millimetre wavelengths provide valuable information for atmospheric science, chemistry, and astronomy [2]. Despite this strong application potential, the development of compact and efficient THz systems remains a major challenge.

One of the main limitations is the lack of practical THz sources, often referred to as the “terahertz gap” [2]. Existing approaches to generate THz radiation can broadly be divided into photonic and electronic methods. Photonic sources, such as quantum cascade lasers (QCLs), can achieve high output power and operate at very high frequencies due to their intersubband transition mechanism and cascade structure, which allow flexible frequency design and efficient photon generation [3]. However, these systems often require complex fabrication processes and, in many cases, cryogenic cooling to achieve optimal performance, which increases system complexity and size [3]. Another photonic approach is based on photomixing in uni-travelling-carrier photodiodes (UTC-PDs), where two optical tones are combined to generate a THz signal [4]. These devices offer wide frequency tunability and can be readily integrated with optical fibre systems. However, they typically require external laser sources and additional optical components, which increases the overall system complexity and limits the level of integration achievable in compact THz transmitters [4].

Electronic approaches, such as transistor-based frequency multipliers and oscillators, offer better compatibility with integrated circuits and room temperature operation. However, their performance degrades at very high frequencies due to carrier transit time limitations and parasitic effects [5]. This restricts their operation in the upper THz range and motivates the exploration of alternative device technologies.

Resonant tunneling diodes (RTDs) are considered promising candidates for compact THz sources. Due to their negative differential conductance (NDC) and their fast dynamics, RTDs can operate as active devices in oscillator circuits at very high frequencies [6, 7]. Experimental studies have demonstrated room-temperature oscillation of RTDs at frequencies approaching 2 THz, highlighting their potential for compact terahertz sources [7]. In addition, RTD oscillators can be designed with a simple planar structure and are well suited for on-chip integration [8].

In RTD-based oscillators, the antenna plays a dual role. It not only radiates electromagnetic energy but also forms part of the resonant circuit. From a circuit perspective, the antenna contributes to the total admittance seen by the RTD. Its conductive component represents radiation loss, which must be compensated by the negative differential conductance of the RTD, while its reactive component participates in forming the resonance condition of the oscillator [9]. As a result, the performance of the oscillator strongly depends on the interaction between the device and the antenna, and as such they must be co-designed [10].

Among various antenna types, the choice of radiator plays a critical role in determining the performance and integration capability of THz sources. High-gain antennas, such as parabolic reflectors, can significantly improve radiation directivity and output power. However, they are inherently bulky and require precise alignment, making them unsuitable for compact and scalable on-chip implementations [11].

Classical dipole antennas represent one of the earliest approaches for integrating antennas with RTD-based sources. Printed microstrip dipoles can be directly fabricated on semiconductor substrates and combined with RTDs to form integrated antenna systems [12]. In such configurations, the dipole antenna can be modeled as a series RLC resonator, allowing the overall RTD-antenna system to be analyzed using equivalent circuit methods [12]. However, dipole antennas typically exhibit low radiation gain and require dimensions comparable to the wavelength, which limits their efficiency and scalability at terahertz frequencies.

Planar antennas offer superior compatibility with integrated circuits and have been widely explored for THz sources. For example, RTD-based active antenna arrays employing integrated patch antennas have demonstrated improved radiation power and directivity through coherent power combining [13]. However, such patch antenna array architectures typically require multiple antenna elements and coupling networks, increasing design complexity and system overhead.

In contrast, slot antennas are particularly suitable for integration with RTDs due to their planar geometry and compatibility with standard microfabrication processes. They can be directly implemented in metal layers and easily integrated with transmission lines and device structures [14, 15]. From a circuit perspective, the slot antenna behaves as a resonant element whose admittance can be tailored through its geometry, enabling control of the effective load seen by the RTD [16].

Recent studies have demonstrated continuous progress in improving the output power of RTD-based THz oscillators through antenna optimization and coherent power-combining techniques. Reported output powers range from several tens of microwatts for single-device oscillators to the milliwatt level for coherent RTD arrays. A summary of representative RTD THz oscillators reported in the literature is given in Table 1.1.

While large-scale coherent arrays can achieve significantly higher output power, single-device oscillators remain attractive due to their compact size, simple structure, and ease of integration. Consequently, improving the interaction between the RTD and the antenna remains an important research topic for compact THz sources.

Despite previous studies demonstrating the feasibility of RTD-based oscillators with slot antennas [10], several challenges remain. Most reported designs focus on relatively lower frequencies, typically below 500 GHz. At higher frequencies, such as

Table 1.1: Representative RTD-based THz oscillators reported in the literature.

Ref.	Antenna	No. of RTDs	Frequency	Output Power
[13]	Patch Array	72	0.45 THz	11.8 mW
[16]	Slot	1	0.368 THz	0.082 mW
[17]	Slot	2	0.62 THz	0.61 mW
[18]	CPW + Patch	1	0.69 THz	0.01 mW
[19]	Slot Array	2	0.53 THz	0.77 mW
This thesis	Slot	1	0.646 THz	0.08 mW*

*Estimated radiated power obtained from simulation.

above 600 GHz, the wavelength becomes significantly shorter, and the device and antenna dimensions approach deep sub-wavelength scales. In this regime, small geometric variations can lead to large changes in the antenna admittance and therefore directly affect the oscillation behavior [16]. Furthermore, the interaction between antenna parameters, device admittance, and oscillation conditions is not yet fully understood, and systematic design approaches are still limited.

This thesis focuses on the design and analysis of a slot antenna integrated with a resonant tunneling diode (RTD) for terahertz oscillation around 650 GHz. The main objective is to investigate how the geometry of the slot antenna, in particular its length and width, influences the admittance seen by the RTD and, consequently, the oscillation behavior of the system. Based on this understanding, the goal is to achieve stable oscillation with practical output power at the target frequency.

The overall oscillator topology adopted in this thesis is based on the symmetrical slot-antenna RTD oscillator reported by Nguyen et al. [16]. This work serves as the starting point for the present study. Building upon this design, the antenna geometry, electromagnetic model, parasitic structures, and oscillator performance are systematically re-investigated and optimized for operation around 650 GHz. To achieve this, a multi-domain modeling approach is adopted, combining electromagnetic simulations, device-level modeling, and circuit-level analysis. Full-wave simulations are first used to extract the antenna admittance, while the RTD is modeled based on its current-voltage characteristics. These models are then integrated into circuit-level simulations to evaluate the oscillation condition and steady-state behaviour.

In addition, more realistic device structures are considered by introducing the device geometry and bias network. The parasitic effects introduced by these structures are incorporated into the model and analyzed in terms of their impact on the overall admittance and oscillator performance. The influence of key antenna parameters, such as slot length and width, on both the admittance and radiation performance is systematically studied.

This thesis is structured in such a way that Chapter 2 presents the theoretical background, including the electrical characteristics of resonant tunneling diodes and the oscillation condition of negative resistance circuits. Chapter 3 describes the modeling approach and design methodology for the antenna and oscillator system. Chapter 4 presents the simulation results and analysis. Finally, Chapter 5 summarizes the conclusions and discusses possible directions for future work.

2

Theory

2.1 Resonant Tunneling Diode

2.1.1 Structure of Resonant Tunneling Diodes

A resonant tunneling diode (RTD) is a quantum device based on a double-barrier heterostructure [20], typically consisting of two thin potential barriers separated by a quantum well region [8], as illustrated in Fig. 2.1. The layer structure used in this work follows the RTD design reported by Asada et al. [8]. This structure enables resonant tunneling of carriers through discrete energy states confined in the quantum well.

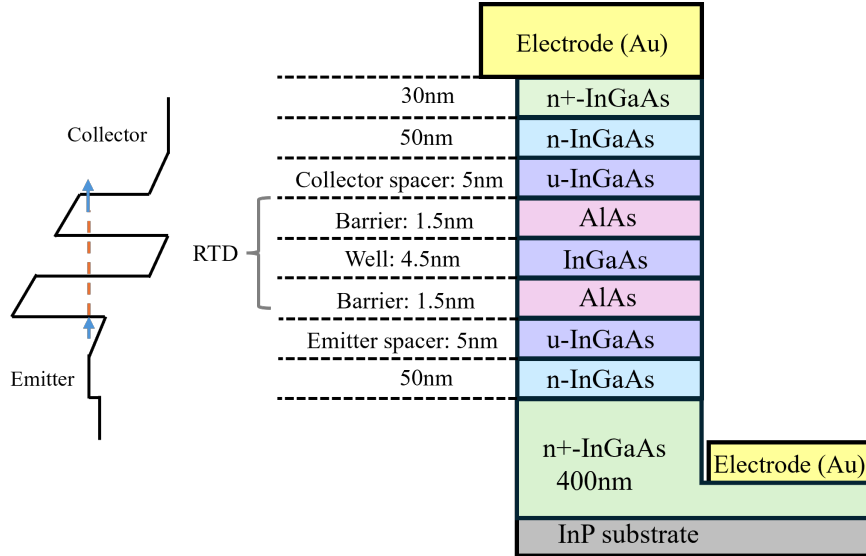


Figure 2.1: Schematic layer structure of the InGaAs/AlAs resonant tunneling diode (RTD) used in this work, following the same structure in Asada group.

When a bias voltage is applied, electrons incident from the emitter can tunnel through the double-barrier structure when their energy aligns with the quantized energy level in the quantum well. Under this resonant condition, constructive interference of the electron wave function inside the well leads to a significantly enhanced transmission probability [21]. This energy-selective transport is a key feature of RTDs and distinguishes them from conventional semiconductor devices such as pn-junction diodes or field-effect transistors, where carrier transport is primarily governed by drift and diffusion mechanisms rather than quantum tunneling. The

formation of quantized energy states in the quantum well arises from spatial confinement at nanometer-scale dimensions, which can be precisely controlled using epitaxial growth techniques such as molecular beam epitaxy (MBE). Typically, the barrier thickness is on the order of one nanometers, allowing strong quantum confinement and efficient tunneling.

Modern RTDs for terahertz applications are commonly realized in material systems such as InGaAs/AlAs or InAs/AlSb, which offer high electron mobility and favorable band alignment for high-speed operation [22], while GaN-based RTDs have also been investigated due to their potential for higher power handling and improved thermal robustness. However, their demonstrated oscillation frequencies remain much lower than those of InGaAs-based RTD oscillators [23]. The choice of material system significantly influences key device parameters, including peak current density and peak-to-valley current ratio (PVCR), which are critical for oscillator performance. In addition to the material system, the structural design of the RTD strongly influences its electrical and high-frequency characteristics [8]. Parameters such as the quantum well width, barrier thickness, and indium composition determine the quantized energy levels and band alignment, thereby affecting the resonant tunneling condition and the peak-to-valley current ratio (PVCR). The resonant energy levels inside the quantum well are approximately determined by the well width according to

$$E_n = \frac{n^2 h^2}{8m^* L_w^2}, \quad (2.1)$$

where L_w is the quantum well width and m^* is the effective electron mass. A narrower quantum well increases the separation between the quantized energy levels and can influence the resonant tunneling characteristics of the device.

Spacer layers are introduced between the active region and the highly doped contact layers to reduce carrier scattering and leakage currents. The collector spacer thickness also affects the depletion capacitance and carrier transit time. Increasing the spacer thickness generally reduces the geometrical capacitance, which can increase the oscillation frequency. However, excessively thick spacer layers increase the carrier transit time and may degrade the high-frequency performance. The lateral area of the RTD also plays an important role in determining the device capacitance and current handling capability. The RTD capacitance mainly arises from charge storage across the double-barrier heterostructure and the depletion region of the active layers. A first-order approximation is the geometrical capacitance,

$$C_0 = \frac{\varepsilon S}{d}, \quad (2.2)$$

where S is the RTD area and d is the effective thickness of the active region. This relation shows that reducing the RTD area lowers the capacitance and therefore enables higher oscillation frequencies. However, smaller RTD areas also reduce the output current and output power. In addition to the geometrical capacitance, the effective RTD capacitance also includes contributions associated with charge accumulation in the quantum well and tunneling-delay effects inside the double-barrier structure [8]. Ohmic contacts and highly doped contact layers introduce additional parasitic resistance and capacitance, which strongly influence the oscillation frequency, impedance matching, and high-frequency performance of terahertz RTD

oscillators. Therefore, careful optimization of both the RTD structure and parasitic elements is essential for achieving efficient terahertz operation.

2.1.2 Current–Voltage Characteristics and Negative Differential Conductance

The current–voltage (I–V) characteristic is one of the most important properties of a resonant tunneling diode (RTD). The negative differential conductance (NDC) arising from resonant tunneling has enabled the use of RTDs as high-frequency oscillators since the early demonstrations of quantum-well oscillators [24]. A typical RTD I–V curve contains a peak current, a valley current, and a negative differential conductance (NDC) region, as shown in Fig. 2.2. In the NDC region, the current decreases as the applied voltage increases. The differential conductance is:

$$G = \frac{dI}{dV}. \quad (2.3)$$

A negative conductance is the key device property that enables RTDs to compensate circuit losses and sustain oscillation.

The origin of the RTD I–V characteristic can be understood from quantum tunneling through the double-barrier quantum well. A physics-based analytical expression for the RTD current was proposed by Schulman et al. [25]. The model starts from the effective-mass expression for the current density through a resonant tunneling structure,

$$J = \frac{qm^*k_BT}{2\pi^2\hbar^3} \int_0^\infty T(E, V) \ln \left[\frac{1 + \exp\left(\frac{E_F - E}{k_BT}\right)}{1 + \exp\left(\frac{E_F - E - eV}{k_BT}\right)} \right] dE, \quad (2.4)$$

where q is the electron charge, m^* is the effective electron mass, k_B is the Boltzmann constant, T is the temperature, $\hbar$ is the reduced Planck constant, E_F is the emitter Fermi energy, and $T(E, V)$ is the energy- and voltage-dependent transmission coefficient. Near a resonant state, the transmission coefficient can be approximated by a Lorentzian function,

$$T(E, V) = \frac{(\Gamma/2)^2}{\left[E - \left(E_r - \frac{eV}{2}\right)\right]^2 + (\Gamma/2)^2}, \quad (2.5)$$

where E_r is the resonant energy level and Γ is the resonance width. The Lorentzian form represents the resonant shape of the transmission coefficient, where electrons tunnel with the highest probability when their energy is aligned with the quasi-bound state in the quantum well.

For a narrow resonance, the transmission coefficient is significant only close to the resonant energy. This allows the slowly varying logarithmic term in Eq. (2.4) to be evaluated near the resonant energy and taken outside the integral. After integration of the Lorentzian term, Schulman et al. obtained a compact expression for the resonant tunneling current density. In practical circuit modeling, the physical

quantities are written as fitting parameters, giving

$$J_1(V) = A \ln \left[\frac{1 + \exp \left(\frac{(B-C+n_1V)e}{k_B T} \right)}{1 + \exp \left(\frac{(B-C-n_1V)e}{k_B T} \right)} \right] \left[\frac{\pi}{2} + \tan^{-1} \left(\frac{C - n_1V}{D} \right) \right]. \quad (2.6)$$

In Eq. (2.6), $J_1(V)$ is the resonant tunneling contribution. The parameter A is a scaling factor related to the tunneling strength and device area. The parameters B and C are related to the emitter Fermi energy and the resonant energy position, respectively. The parameter D represents the resonance broadening, while n_1 describes how the applied voltage shifts the resonant level. The logarithmic factor describes the difference in carrier occupation between the emitter and collector, while the arctangent term comes from the Lorentzian transmission probability of the quasi-bound state. The resonant component $J_1(V)$ produces the peak current and the NDC region, but it does not describe the increasing valley current at higher voltage. To include this contribution, Schulman et al. introduced a background current term in the form

$$J_2(V) = H \left[\exp \left(\frac{n_2 e V}{k_B T} \right) - 1 \right], \quad (2.7)$$

where H controls the magnitude of the background current and n_2 is a voltage scaling factor. This term represents non-resonant transport mechanisms, such as tunneling through higher-energy channels or thermally assisted transport, which become more important after the valley region. The total current density is then written as

$$J(V) = J_1(V) + J_2(V). \quad (2.8)$$

For a device with active area S , the total current can be obtained as

$$I(V) = S J(V). \quad (2.9)$$

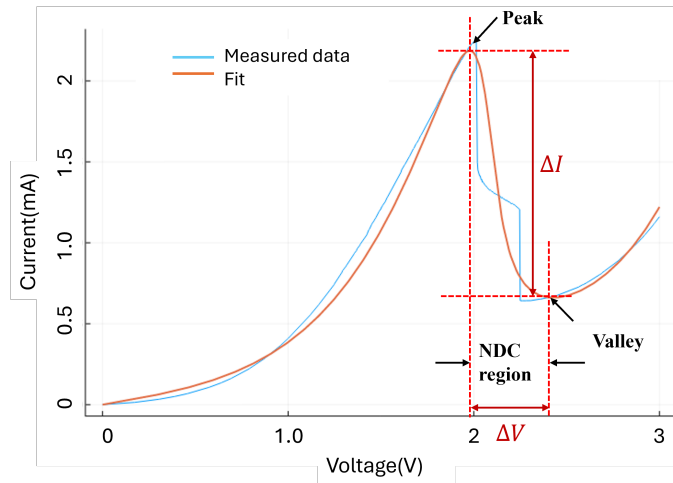


Figure 2.2: Simulated current–voltage (I–V) characteristic of the RTD based on the physics-based analytical model using in this work.

In this thesis, the same functional form is used to fit a measured RTD I–V characteristic in order to extract the NDC region for circuit-level oscillator modeling. The simulated I–V curve used in this work is shown in Fig. 2.2. When the RTD is biased in the NDC region and integrated with an antenna or resonant circuit, the negative differential conductance can compensate for circuit then enable oscillation.

The maximum output power under the impedance-matching condition can be estimated as [9]

$$P_{\text{out,max}} = \frac{3}{16} \Delta I \Delta V, \quad (2.10)$$

where ΔI and ΔV are the current and voltage widths between the peak and valley points of the RTD I–V curve.

2.2 Antenna

In high-frequency circuit design, antennas can be described not only as radiating elements but also as frequency-dependent circuit components. In this work, the antenna is directly connected to the RTD and forms part of the oscillator circuit. From a circuit perspective, an antenna is characterized by its input impedance $Z = R + jX$, or equivalently by its admittance $Y = G + jB$ [14, 15]. The conductance, G , represents power dissipation including both radiation and ohmic losses and must be compensated by the negative differential conductance of the RTD. The susceptance, B , corresponds to reactive energy stored in the near field and contributes to the resonant condition. Therefore, the antenna is treated as a frequency-dependent load whose admittance plays a key role in determining the oscillation behavior.

2.2.1 Slot Antenna Characteristics

A slot antenna consists of an aperture etched in a conductive surface and operates as a resonant radiating structure. According to Babinet’s principle, a slot antenna can be considered complementary to a dipole antenna, with similar radiation characteristics but different current distributions [15]. A typical slot antenna consists of a narrow rectangular aperture etched in a metallic layer, as illustrated in Fig. 2.3. When excited, the slot supports resonant electromagnetic fields and radiates electromagnetic waves into free space.

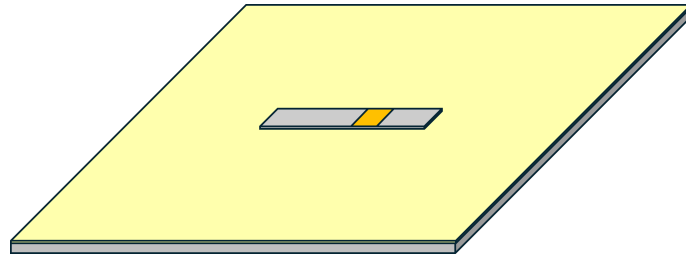


Figure 2.3: Basic structure of a slot antenna consisting of a rectangular aperture in a conductive surface.

From a circuit point of view, the slot antenna can be modeled as a frequency-dependent admittance connected to the RTD oscillator circuit [16]. The input admittance of the slot antenna is influenced by geometric parameters such as the slot length and width, as well as by the effective wavelength in the surrounding medium. Based on the analytical expressions reported in [16], the antenna conductance and susceptance exhibit different dependencies on the slot dimensions. Both the conductance and susceptance vary with the slot dimensions. Therefore, variations in the slot dimensions modify the admittance presented to the RTD.

As will be described in section 2.3, the oscillation frequency of an RTD-based oscillator is determined by the balance between the antenna and RTD susceptance, and when it reaches the resonant condition. Consequently, the slot dimensions shift the frequency at which the oscillation occurs. In addition to that, the slot antenna also determines the radiation characteristics of the oscillator, including the radiation efficiency, directivity, and gain. Although the antenna admittance is critical for satisfying the oscillation condition, efficient radiation is also required to effectively couple the generated power into free space. Therefore, both the input admittance and radiation performance must be considered in the design of terahertz RTD slot antennas.

2.3 Oscillation Condition and Equivalent Circuit

An oscillator is a circuit that generates a periodic signal without an external RF input. In an RTD-based oscillator, the active device is the resonant tunneling diode, which provides negative differential conductance when it is biased in the NDC region. This negative conductance can compensate for the losses of the surrounding resonant circuit and sustain oscillation [9, 16]. The use of negative-resistance devices for microwave and millimeter-wave oscillators has been studied since the early development of tunnel-diode oscillators [26].

From an admittance point of view, oscillation is determined by the total admittance seen by the RTD. For steady-state oscillation, the reactive part of the total admittance must be cancelled, while the real part must satisfy the loss compensation condition.

In this thesis, the oscillator is analyzed using this admittance-based approach, which the antenna is represented by its input admittance, while the RTD is represented by its negative differential conductance and capacitance. This representation is convenient because both the antenna and RTD are connected in parallel in the small-signal equivalent circuit.

2.3.1 Simplified Equivalent Circuit

A simplified small-signal equivalent circuit of the RTD-integrated slot antenna oscillator is shown in Fig. 2.4.

In this model, the RTD is represented by a negative conductance and a capacitance, while the antenna is represented by its input admittance,

$$Y_a = G_a + jB_a. \quad (2.11)$$

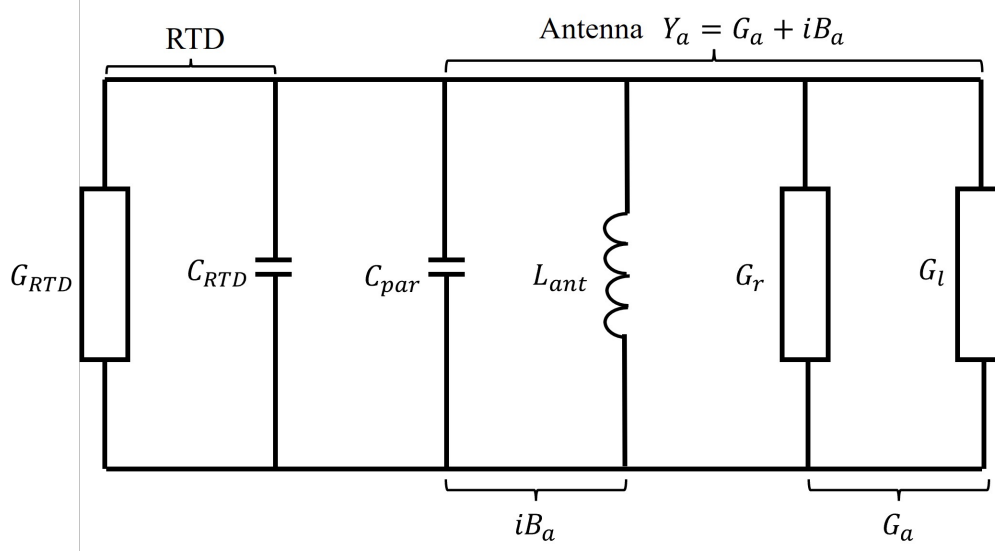


Figure 2.4: Simplified equivalent circuit of an RTD-integrated slot antenna oscillator. The RTD is modeled by its negative conductance and capacitance, while the antenna is represented by a frequency-dependent admittance consisting of conductance and susceptance components.

The real part of the antenna admittance, G_a , represents the total antenna loss. It consists of the radiation conductance G_r and the ohmic loss conductance G_l ,

$$G_a = G_r + G_l. \quad (2.12)$$

Here, G_r describes the power radiated into free space, while G_l represents the power dissipated as conductor and material losses.

The imaginary part of the antenna admittance, B_a , represents the reactive energy stored by the antenna structure. In the simplified circuit, this reactive part is described by an antenna inductance L_{ant} and a parasitic capacitance C_{par} . The capacitance C_{par} represents the electric-field storage associated with the antenna structure, such as the field between the slot edges and between the metal layer and the surrounding substrate. These lumped elements are not separate physical components, but a simplified equivalent representation of the antenna reactive behavior near the operating frequency.

Using the small-signal RTD model introduced in Section 2.1.1, the RTD admittance can be written as

$$Y_{\text{RTD}} = G_{\text{RTD}} + j\omega C_{\text{RTD}}, \quad (2.13)$$

where G_{RTD} is the RTD differential conductance, which becomes negative in the NDC region, and C_{RTD} is the effective RTD capacitance. The RTD capacitance originates from charge storage across the double-barrier heterostructure and contributes to the reactive part of the oscillator admittance.

The simplified equivalent circuit therefore captures the essential interaction between the RTD and the antenna: the RTD provides negative conductance, while the antenna contributes radiation loss and reactive loading.

2.3.2 Oscillation Condition

Based on the simplified equivalent circuit in Fig. 2.4, the RTD and the antenna are connected in parallel. Using the admittance definitions introduced in the previous section, the total admittance of the oscillator can be written as

$$Y_{\text{tot}} = Y_{\text{RTD}} + Y_a. \quad (2.14)$$

With equation 2.11 and equation 2.13, the total admittance becomes

$$Y_{\text{tot}} = (G_a + G_{\text{RTD}}) + j(B_a + \omega C_{\text{RTD}}). \quad (2.15)$$

The real part of the total admittance determines whether the negative conductance of the RTD is sufficient to overcome the antenna losses. Therefore, the start-up condition for oscillation can be written as

$$\text{Re}(Y_{\text{tot}}) = G_a + G_{\text{RTD}} < 0. \quad (2.16)$$

The imaginary part determines the resonance frequency, which occurs as the susceptance of the RTD and antenna cancel each other. This condition is given by

$$\text{Im}(Y_{\text{tot}}) = B_a + \omega C_{\text{RTD}} = 0. \quad (2.17)$$

In this admittance-based interpretation, the antenna geometry is designed to provide an appropriate conductance and susceptance at the target frequency, while the RTD provides the negative differential conductance required for oscillation.

As the oscillation amplitude increases, the voltage across the RTD swings through the nonlinear NDC region. Due to the highly nonlinear shape the negative conductance changes, it can be regarded as bias dependant. The oscillation amplitude is limited by the finite NDC region and by the circuit losses. Nevertheless, the small-signal admittance conditions above are useful for identifying the frequency range where oscillation can start and for guiding the slot antenna design.

The oscillation condition can be visualized by comparing the antenna admittance with the RTD admittance, as shown in Fig. 2.5. An oscillation point is expected near the frequency where the real-part condition and the imaginary-part cancellation condition are satisfied simultaneously.

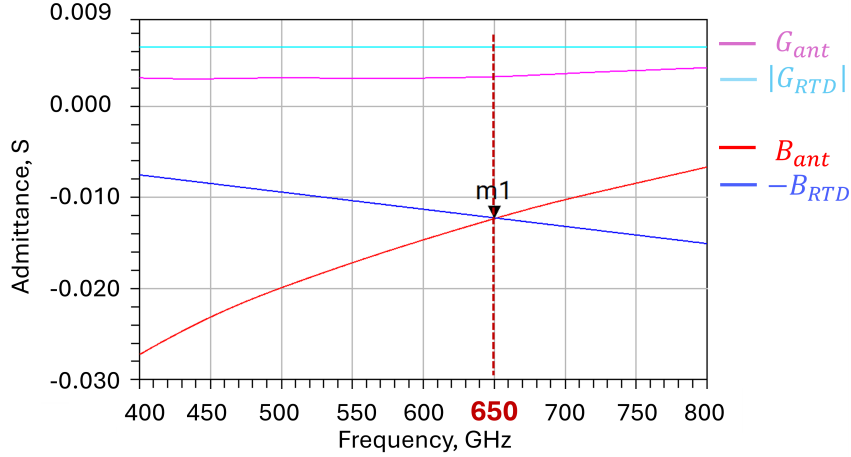


Figure 2.5: Graphical illustration of the oscillation condition using the extracted antenna admittance and RTD admittance. An oscillation point is expected near the frequency where the real-part condition $G_a + G_{RTD} < 0$ and the imaginary-part cancellation condition $B_a + \omega C_{RTD} = 0$ are simultaneously satisfied.

2.3.3 Extended Equivalent Circuit Including Parasitics

In practical implementations, additional parasitic elements are introduced by the device structure, interconnects, and fabrication layout. A more complete equivalent circuit of the RTD-integrated antenna oscillator is shown in Fig. 2.6.

Compared to the simplified model, the extended circuit includes parasitic capacitances and resistances that arise from the contact layers, overlap regions, and connecting structures between the RTD and the antenna. These elements form an additional admittance contribution, which can be grouped as

$$Y_p \approx G_p + jB_p, \quad (2.18)$$

where G_p represents the parasitic loss and B_p represents the additional reactive loading.

The total admittance of the circuit can therefore be written as

$$Y_{\text{tot}} = Y_{RTD} + Y_a + Y_p. \quad (2.19)$$

The parasitic capacitance C_c mainly originates from the overlap between metal layers and contact regions, while the series resistances R_c and R_{spr} represent the finite conductivity of the metal interconnects and spreading resistance in the semiconductor layers. These parasitic elements modify both the real and imaginary parts of the total admittance.

In particular, the additional susceptance introduced by the parasitic capacitance shifts the imaginary part of the total admittance, which changes the frequency at which the condition

$$\text{Im}(Y_{\text{tot}}) = 0 \quad (2.20)$$

is satisfied. Depending on their configuration, parasitic resistances can modify the effective conductance of the system, but in all cases they introduce additional losses that must be compensated by the RTD negative differential conductance.

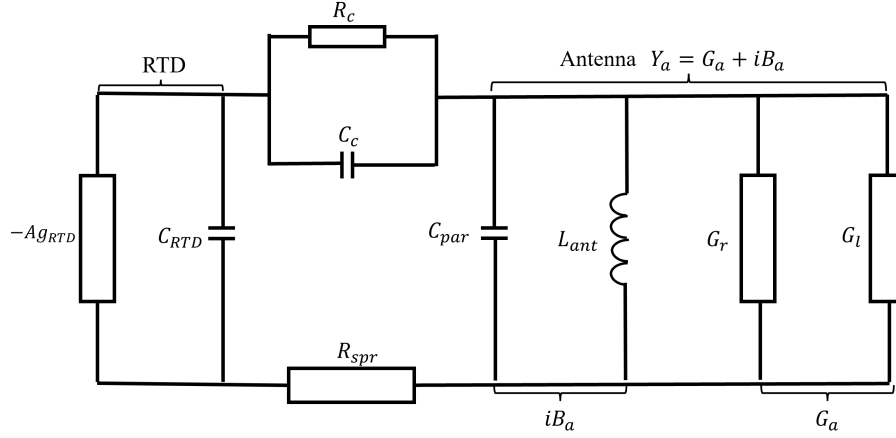


Figure 2.6: Extended equivalent circuit of the RTD-integrated slot antenna oscillator including parasitic elements. The parasitic capacitance and resistance introduce additional admittance that modifies the oscillation condition.

As a result, parasitic elements can significantly affect both the oscillation frequency and the start-up condition of the oscillator. Therefore, they must be included in the circuit model for accurate prediction of the oscillation behavior.

However, due to the complex three-dimensional geometry of the RTD-integrated slot antenna oscillator, including the metal interconnects, dielectric layers, air bridge, and semiconductor contact regions, analytical approximations alone are not sufficient to accurately describe the electromagnetic behavior and parasitic contributions of the structure. Therefore, full-wave electromagnetic simulations are employed in this work using Ansys HFSS to extract the admittance characteristics and parasitic effects of the oscillator structure. The extracted results are then used for the subsequent nonlinear circuit analysis in ADS.

3

Methods

3.1 Overall design methodology

The design of the RTD-integrated slot antenna oscillator was carried out using a multi-stage methodology combining analytical estimation, electromagnetic simulation, and nonlinear circuit analysis. The overall workflow used in this work is illustrated in Fig. 3.1.

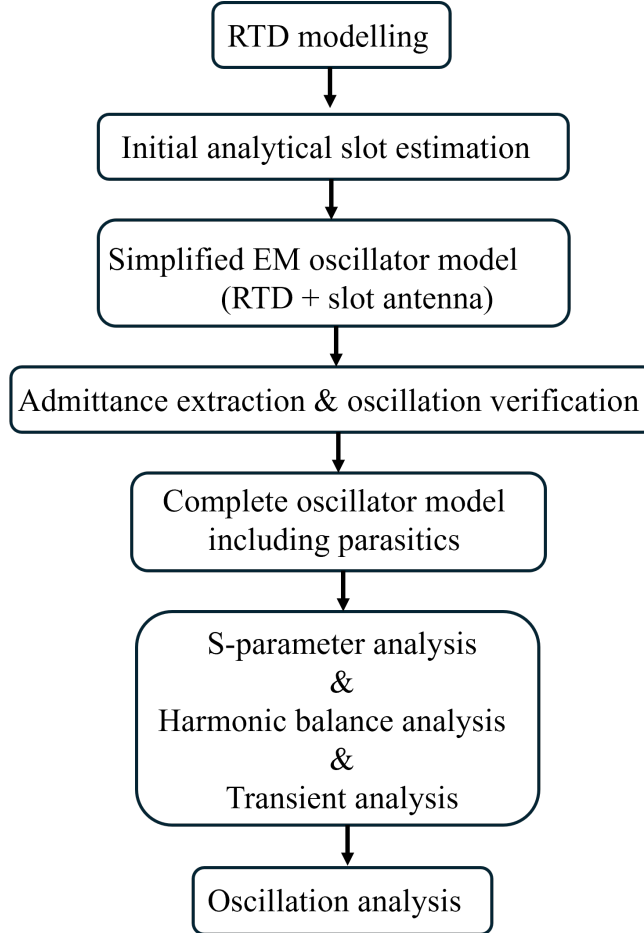


Figure 3.1: Overall design methodology used for the RTD slot-antenna oscillator.

First, the RTD was represented by the nonlinear current–voltage model and the capacitance model introduced in Section 2.1. The model was used to describe the negative differential conductance and capacitive contribution of the RTD, which are

required for the oscillator design.

The initial slot antenna geometry was then estimated using the maximum-output-power design approach reported in [16]. This approach is based on a simplified slot-antenna RTD oscillator model, where the relation between slot dimensions and output power is described through the radiation conductance, ohmic loss, and antenna susceptance. Since the analytical relations are derived under short-slot and simplified parasitic-free assumptions, they were used only to obtain a starting geometry for the subsequent full-wave design.

After the initial geometry was obtained, a simplified oscillator model consisting of the RTD and slot antenna was analysed. This stage was used to evaluate the antenna admittance and to verify whether the basic oscillation requirements could be satisfied before additional parasitic elements were included.

The oscillator model was then extended to include the complete device structure, including the electrode geometry, air bridge, dielectric region, semiconductor layers, and other parasitic contributions. The complete electromagnetic model was combined with circuit-level analysis to evaluate the overall oscillator admittance and nonlinear oscillation behaviour. Then, harmonic balance analysis was performed to determine the oscillation frequency, output power, and bias-dependent oscillator characteristics. This step-by-step methodology allowed the design to progress from an initial analytical estimate to a more realistic full-device model. Finally, transient simulations were performed using the complete oscillator model. Unlike harmonic balance analysis, which directly computes the steady-state solution, transient analysis captures the start-up behaviour of the oscillator in the time domain. The transient results were used to verify the oscillation frequency predicted by harmonic balance simulations and to investigate the possible existence of alternative oscillation modes over a broader frequency range. Since the exported antenna admittance data covered a wide frequency range from 1 THz to 2 THz, the transient analysis also served as an additional validation that no unintended low-frequency or high-frequency oscillation modes dominated the circuit. The agreement between harmonic balance and transient simulations provided further confidence in the predicted oscillation behaviour.

3.2 RTD nonlinear model

3.2.1 I–V curve fitting

The RTD model used in this work was based on the physics-based analytical current–voltage model introduced in Section 2.1.2 and originally proposed by Schulman *et al.* [25].

The fitting parameters used in this work were obtained from experimentally measured RTD characteristics provided by Christoffer Johansson, Gabriel Josefsson at the Terahertz and Millimetre Wave Laboratory. Using the analytical model described in Chapter 2, the measured RTD current–voltage characteristics were fitted to obtain a compact nonlinear model suitable for circuit-level oscillator simulations. The fitted RTD current was expressed as

$$\begin{aligned}
I(V) = 0.001 \left[A \ln \left(\frac{1 + \exp \left((B - C + n_1 V) \frac{q}{kT} \right)}{1 + \exp \left((B - C - n_1 V) \frac{q}{kT} \right)} \right) \right. \\
\times \left(\frac{\pi}{2} + \tan^{-1} \left(\frac{C - n_1 V}{D} \right) \right) \\
\left. + H \left(\exp \left(\frac{n_2 q V}{kT} \right) - 1 \right) \right].
\end{aligned} \tag{3.1}$$

where the fitting coefficients are listed in Table 3.1, and the fitted model includes a scaling factor of 0.001 to express the output current in amperes rather than milliamperes. The RTD mesa area used in this work was $0.49 \mu\text{m}^2$.

Table 3.1: Fitting parameters used for the RTD I–V model.

Parameter	Value
A	1.4360
B	2.228×10^{-4}
C	0.1199
D	0.00475
n_1	0.05744
H	0.00135
n_2	0.05643
T	300 K

The resulting fitted I–V curve is shown in Fig. 2.2. Using the maximum output power estimate introduced in Eq. (2.10), the available output power can be calculated from the extracted peak and valley values as

$$P_{\text{out,max}} \approx 128 \mu\text{W}. \tag{3.2}$$

This value represents the theoretical maximum available output power of the RTD based on the fitted I–V characteristic and serves as an upper bound for the oscillator performance.

3.2.2 RTD model

For the oscillator analysis, the fitted nonlinear I–V model was converted into a bias-dependent admittance model. As introduced in Section 2.3.2, the RTD was represented by a differential conductance in parallel with a junction capacitance. The structure of the RTD used in this work is shown in Fig. 2.1.

The RTD capacitance was estimated using a parallel-plate approximation,

$$C_{\text{RTD}} = \frac{\varepsilon_0 \varepsilon_r S}{d_{\text{eff}}}, \tag{3.3}$$

where ε_0 is the vacuum permittivity, ε_r is the relative permittivity of the RTD material system, S is the RTD mesa area, and d_{eff} is the effective thickness of the active region. The mesa area used in this work was

$$S = 0.49 \text{ } \mu\text{m}^2. \quad (3.4)$$

The effective thickness was taken as the sum of the emitter spacer, double-barrier quantum well region, and collector spacer,

$$d_{\text{eff}} = 17.5 \text{ nm}. \quad (3.5)$$

Using $\varepsilon_r = 12.4$ [14] gives

$$C_{\text{RTD}} \approx 3 \text{ fF}. \quad (3.6)$$

Therefore, a constant capacitance of $C_{\text{RTD}} = 3 \text{ fF}$ was used in the subsequent oscillator simulations.

In oscillator operation, the instantaneous RTD voltage consists of a DC bias component and a time-varying AC component, which can generally be expressed as

$$V(t) = V_{\text{DC}} + V_{\text{AC}} \cos(\omega t), \quad (3.7)$$

where V_{DC} is the applied bias voltage and V_{AC} is the oscillation amplitude. The RTD differential conductance was obtained from the slope of the fitted I–V curve,

$$G_{\text{RTD}}(V) = \frac{dI(V)}{dV}. \quad (3.8)$$

Since the I–V model is nonlinear, G_{RTD} is bias dependent. In the negative differential conductance region, the slope of the I–V curve becomes negative, giving $G_{\text{RTD}} < 0$. This negative conductance is the gain mechanism that compensates for the antenna and circuit losses.

Based on Eq. (2.12), the RTD admittance can therefore be written as

$$Y_{\text{RTD}}(V, \omega) = G_{\text{RTD}}(V) + j\omega C_{\text{RTD}}, \quad (3.9)$$

or, using the capacitance value used in this work,

$$Y_{\text{RTD}}(V, \omega) = \frac{dI(V)}{dV} + j\omega (3 \times 10^{-15}). \quad (3.10)$$

This admittance model was subsequently used in the oscillator analysis and circuit-level simulations.

3.3 Initial Slot Antenna Estimation

3.3.1 Analytical estimation of initial slot dimensions

The initial slot antenna dimensions were estimated using the maximum-output-power design method reported in [16]. This method is based on a simplified symmetrical slot-antenna RTD oscillator model, where the slot antenna is described

by its radiation conductance, ohmic loss conductance, and susceptance. The aim of this step was not to obtain the final antenna geometry directly, but to provide a physically reasonable starting point for the subsequent full-wave electromagnetic optimization.

In the simplified model, the additional parasitic admittance is neglected, so the circuit admittance is approximated by the antenna admittance. The imaginary-part oscillation condition introduced in Eq. (2.16) can then be written as

$$B_a + \omega C_{\text{RTD}} = 0. \quad (3.11)$$

Therefore, the required antenna susceptance at the target frequency is

$$B_{a,\text{target}} = -\omega C_{\text{RTD}}. \quad (3.12)$$

For the target frequency of 650 GHz and $C_{\text{RTD}} = 3$ fF,

$$B_{a,\text{target}} \approx -1.23 \times 10^{-2} \text{ S}. \quad (3.13)$$

The optimization method in [16] is based on maximizing the oscillation amplitude of the RTD by balancing the radiation and ohmic losses of the slot antenna. With the RTD parasitics neglected, the output power can be written in the simplified form

$$P = \frac{1}{2} G_r \Delta V^2 (1 - \gamma) \alpha(\gamma), \quad (3.14)$$

where G_r is the radiation conductance, which is only one part of the total antenna conductance G_a , as defined in Eq. (2.11). ΔV is the voltage separation between the peak and valley points of the RTD I-V curve, and γ describes the reduction of the effective negative conductance under large-signal oscillation. And $\alpha(\gamma)$ is a correction factor introduced to account for the waveform shape under large-signal oscillation conditions.

γ can be expressed as

$$\gamma = \omega \tau_{\text{RTD}} \frac{G_r + G_l}{B_a}, \quad (3.15)$$

where G_l is the ohmic loss conductance and τ_{RTD} is the RTD time constant. The ohmic loss conductance can be approximately represented by an equivalent series loss resistance R_l , giving

$$G_l \approx R_l B_a^2. \quad (3.16)$$

Substituting Eq. (3.16) into Eq. (3.15) gives

$$\gamma = \omega \tau_{\text{RTD}} \left(\frac{G_r}{B_a} + R_l B_a \right). \quad (3.17)$$

The optimum susceptance is obtained by minimizing the term inside the bracket with respect to B_a ,

$$\frac{d}{dB_a} \left(\frac{G_r}{B_a} + R_l B_a \right) = -\frac{G_r}{B_a^2} + R_l = 0. \quad (3.18)$$

This gives

$$B_a^2 = \frac{G_r}{R_l}. \quad (3.19)$$

Since the slot antenna is operated below its resonance, it behaves mainly as an inductive element and B_a is negative,

$$B_a = -\sqrt{\frac{G_r}{R_l}}. \quad (3.20)$$

To relate this condition to the slot dimensions, the analytical approximations for a short slot antenna reported in [16] were used. The radiation conductance is approximated as

$$G_r(l) = \frac{2\pi}{3Z_{\text{eff}}} \left(\frac{l}{\lambda_{\text{eff}}} \right)^2, \quad (3.21)$$

and the antenna susceptance is approximated as

$$B_a(l, w) = -\frac{4}{\pi Z_{\text{eff}}} \frac{\ln(l/w) - 1}{\tan(\pi l / \lambda_{\text{eff}})}. \quad (3.22)$$

Here, l and w are the slot length and width, respectively. The effective impedance and effective wavelength are given by

$$Z_{\text{eff}} = \frac{Z_0}{\sqrt{\varepsilon_{\text{eff}}}}, \quad \lambda_{\text{eff}} = \frac{\lambda_0}{\sqrt{\varepsilon_{\text{eff}}}}, \quad (3.23)$$

with

$$\varepsilon_{\text{eff}} = \frac{\varepsilon_s + 1}{2}. \quad (3.24)$$

For the InP substrate, $\varepsilon_s \approx 12.4$ [27], giving

$$\varepsilon_{\text{eff}} = 6.7, \quad \lambda_{\text{eff}} \approx 178 \text{ } \mu\text{m} \quad \text{at } 650 \text{ GHz}. \quad (3.25)$$

The estimation was then carried out as follows. First, the target susceptance was set by Eq. (3.13). Then, the optimum-loss condition in Eq. (3.20) was used to calculate the required radiation conductance,

$$G_{r,\text{target}} = R_l B_{a,\text{target}}^2. \quad (3.26)$$

The equivalent loss resistance R_l was estimated from a preliminary simplified full-wave antenna simulation with $l = 40 \text{ } \mu\text{m}$ and $w = 4 \text{ } \mu\text{m}$, giving approximately

$$R_l \approx 6 \text{ } \Omega. \quad (3.27)$$

Using Eq. (3.21), the slot length can then be obtained from

$$l = \lambda_{\text{eff}} \sqrt{\frac{3Z_{\text{eff}} G_{r,\text{target}}}{2\pi}}. \quad (3.28)$$

After the slot length is found, the slot width is obtained by solving Eq. (3.22) for w . Rearranging Eq. (3.22) gives

$$\ln(l/w) = 1 - B_{a,\text{target}} \frac{\pi Z_{\text{eff}}}{4} \tan\left(\frac{\pi l}{\lambda_{\text{eff}}}\right). \quad (3.29)$$

Thus, the slot width is

$$w = \frac{l}{\exp\left[1 - B_{a,\text{target}} \frac{\pi Z_{\text{eff}}}{4} \tan\left(\frac{\pi l}{\lambda_{\text{eff}}}\right)\right]}. \quad (3.30)$$

Using this calculation procedure, the initial slot dimensions were estimated as

$$l \approx 45 \text{ } \mu\text{m}, \quad w \approx 4 \text{ } \mu\text{m}. \quad (3.31)$$

These dimensions were used only as the initial geometry for electromagnetic simulation. The analytical model assumes a short slot, an infinitely thin ideal conductor which in my design we have a conductance loss, and neglects additional device parasitics. Therefore, the final slot dimensions were expected to deviate from the analytical estimate after full-wave simulation and oscillator-level optimization.

3.3.2 Initial full-wave validation and geometry refinement

After the analytical estimation, the initial slot geometry was implemented in HFSS for full-wave electromagnetic validation. The purpose of this step was to extract the antenna admittance and to refine the slot dimensions so that the antenna susceptance could compensate the capacitive susceptance of the RTD near the target oscillation frequency.

In the HFSS model, the slot antenna was formed by a rectangular aperture in a 100 nm thick gold layer placed on an InP substrate. To reduce artificial reflections from the substrate sidewalls, the lateral size of the substrate was chosen to be sufficiently large compared with the slot dimensions, since in the realistic case our substrate is very wide compared to the oscillator. In this work, both the substrate length and width were set to 600 μm . A larger substrate would further reduce the influence of the side boundaries, but would also significantly increase the mesh size and computational cost. Therefore, 600 μm was used as a compromise between physical accuracy and simulation feasibility.

The substrate thickness was set to 200 μm . To avoid artificial reflections from the simulation boundaries, radiation boundaries were applied directly on the substrate sidewalls and bottom surface. This configuration preserves the physical substrate–air interface at the bottom of the substrate while minimizing the influence of the simulation domain on the antenna admittance. The simulation region was extended above the antenna to ensure sufficient space for the electromagnetic fields. The resulting simplified antenna model is shown in Fig. 3.2.

A small-range sweep of the slot length and width was then performed around the initial analytical design. The extracted antenna admittance is shown in Fig. 3.3. For the geometry with a slot length of 57 μm and a slot width of 4 μm , the antenna admittance at 650 GHz is approximately

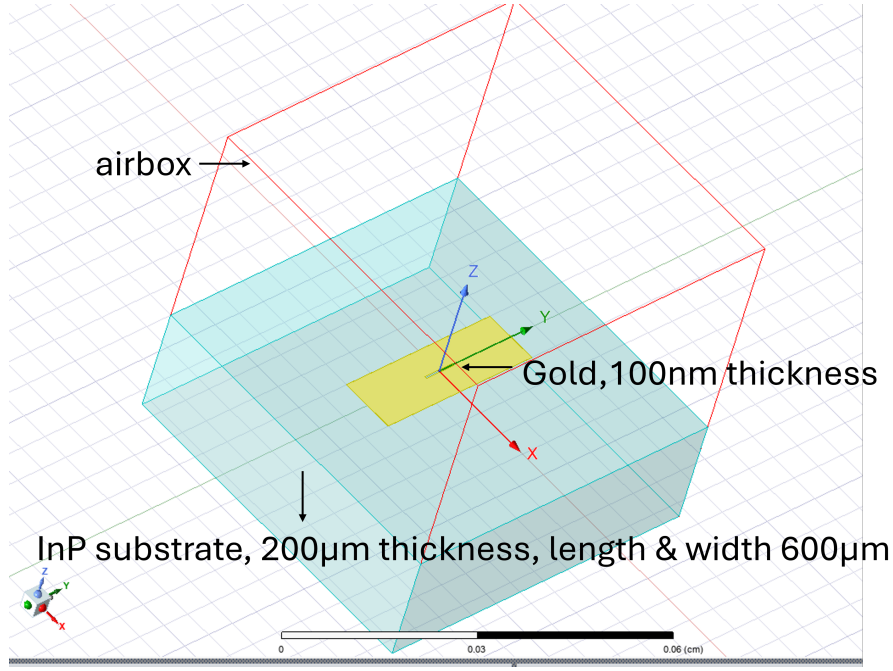


Figure 3.2: Simplified HFSS model used for the initial full-wave validation. The model consists of a 100 nm thick gold slot antenna on a 200 μm thick InP substrate with a lateral size of 600 $\mu\text{m} \times 600 \mu\text{m}$.

$$Y_a \approx 0.003 - j0.012 \text{ S}. \quad (3.32)$$

This result is close to the required admittance for oscillation. The imaginary part of the antenna admittance is approximately -0.012 S , which can compensate the capacitive susceptance of the RTD near the target frequency. Meanwhile, the real part is approximately 0.003 S , which is smaller than the maximum negative differential conductance provided by the RTD, , seen in Figure 3.4.

Around the target operating point near $V_{DC} \approx 2.1 \text{ V}$, the RTD conductance reaches approximately $G_{RTD} \approx -0.009 \text{ S}$, which can compensate much more than antenna loss around 650 GHz. Therefore, the extracted admittance indicates that this geometry can satisfy the small-signal oscillation condition around 650 GHz.

The discrepancy between the analytically estimated dimensions and the full-wave HFSS results can be explained by the assumptions used in the simplified analytical model. In [16], the expressions for G_r and B_a were derived under the assumptions of a narrow short slot formed by an infinitely thin perfect conductor on a dielectric substrate, while RTD parasitic effects were neglected. The authors also noted that the model becomes increasingly qualitative at THz frequencies due to stronger parasitic contributions and the breakdown of the narrow-slot approximation. Furthermore, the experimental validation presented in [16] mainly focused on oscillators operating in the 150–400 GHz range. Since the present work targets operation near 650 GHz, deviations between the analytical estimation and the full-wave electromagnetic response were expected.

Nevertheless, the analytical model still provided a physically meaningful starting point for the antenna design and significantly reduced the parameter search space

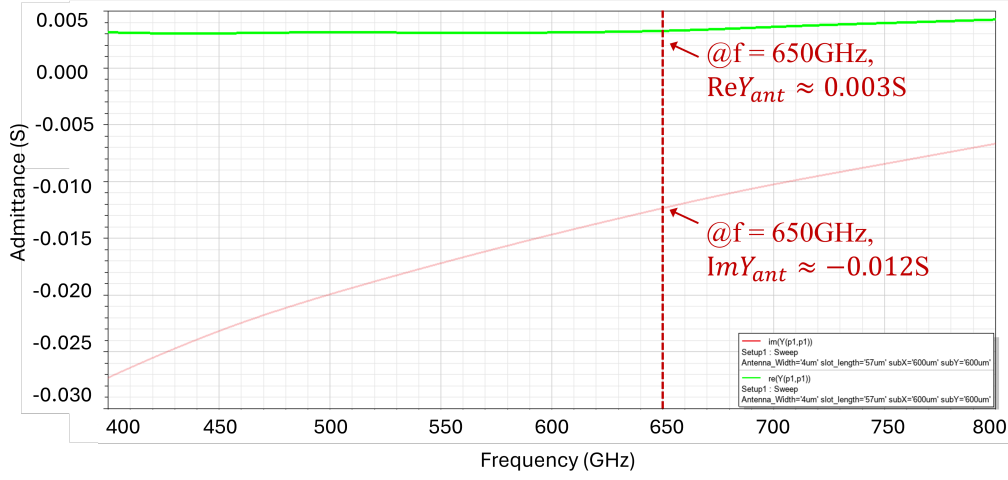


Figure 3.3: Extracted antenna admittance of the simplified slot antenna model. For $l = 57 \mu\text{m}$ and $w = 4 \mu\text{m}$, the antenna admittance at 650 GHz is approximately $Y_a \approx 0.003 - j0.012 \text{ S}$.

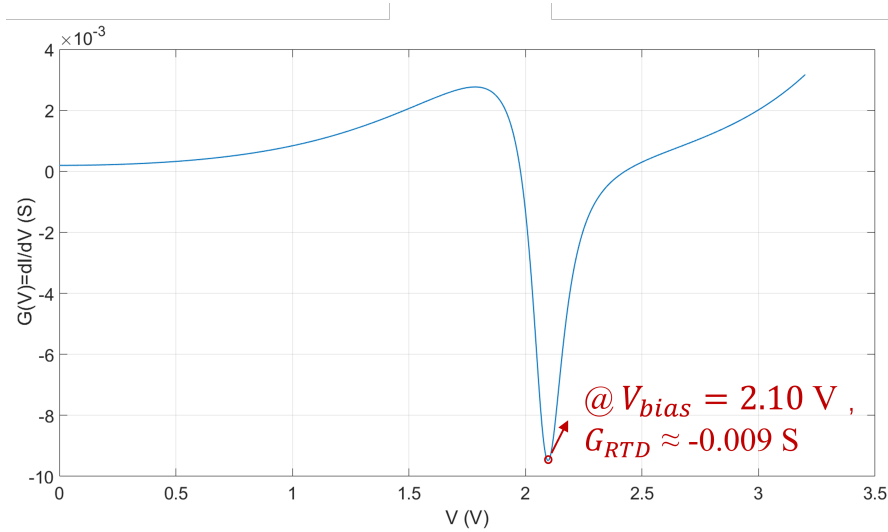


Figure 3.4: Differential conductance extracted from the fitted RTD I-V curve. The minimum conductance near $V_{DC} \approx 2.1 \text{ V}$ corresponds to approximately $G_{RTD} \approx -0.009 \text{ S}$.

for the subsequent HFSS optimization.

3.3.3 Nonlinear Circuit Analysis in ADS

After completing the simplified slot antenna optimization in HFSS, the antenna model was imported into Keysight ADS for nonlinear circuit analysis. The purpose of this analysis was to evaluate the oscillation behavior of the simplified RTD-slot antenna system and identify the RTD bias voltage range capable of supporting oscillation around 650 GHz.

The overall analysis procedure consisted of three steps. First, a small-signal S-parameter analysis was performed to evaluate the oscillation conditions under different RTD bias voltages. Second, harmonic balance (HB) simulations with bias voltage sweeping were carried out to analyze the bias-dependent oscillation frequency and oscillation voltage amplitude. Finally, a fixed-bias HB simulation was performed at the optimum bias point to further evaluate the steady-state oscillation spectrum and output power.

The ADS circuit used for the S-parameter analysis is shown in Fig. 3.5. The slot antenna simulated in HFSS was exported as a one-port Snp block and imported into ADS. The RTD was modeled using the nonlinear current-voltage relation described in Section 2.2 together with a parallel junction capacitance of 3 fF. Two ports were placed at the RTD side and antenna side, respectively, in order to extract the real and imaginary parts of the admittance from both sides of the circuit.

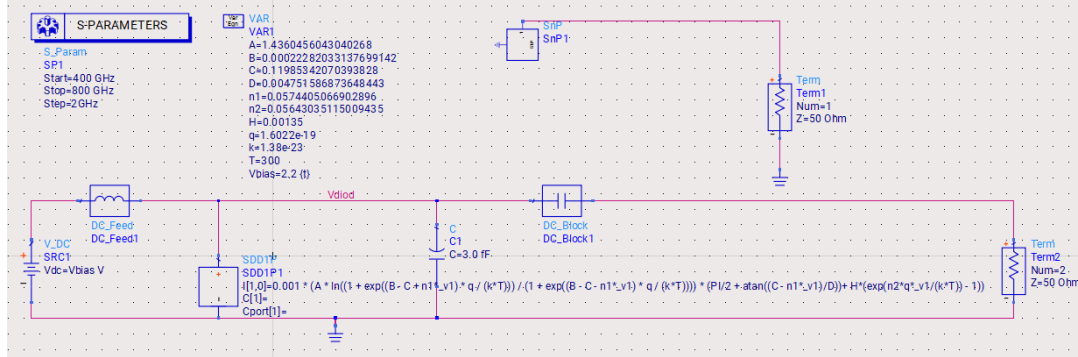


Figure 3.5: ADS setup used for the small-signal S-parameter analysis of the simplified RTD-slot antenna oscillator.

By comparing the real and imaginary parts of the antenna admittance and RTD admittance, the bias-dependent oscillation condition could be evaluated. The S-parameter analysis showed that reactive crossing points started to appear near 650 GHz within the RTD bias voltage range of approximately 2.02–2.2 V, which is shown in Fig. 4.1 - Fig. ?? . This indicates that the simplified RTD-slot antenna structure is capable of supporting oscillation only within a limited negative differential conductance region of the RTD.

After identifying the possible oscillation region from the S-parameter analysis, harmonic balance simulations were carried out using the circuit shown in Fig. 3.6. An OscPort component was introduced into the circuit to allow ADS to determine the

steady-state oscillation frequency self-consistently. The RTD bias voltage was swept from 2.02 V to 2.20 V in order to evaluate the bias dependence of the oscillation behavior.

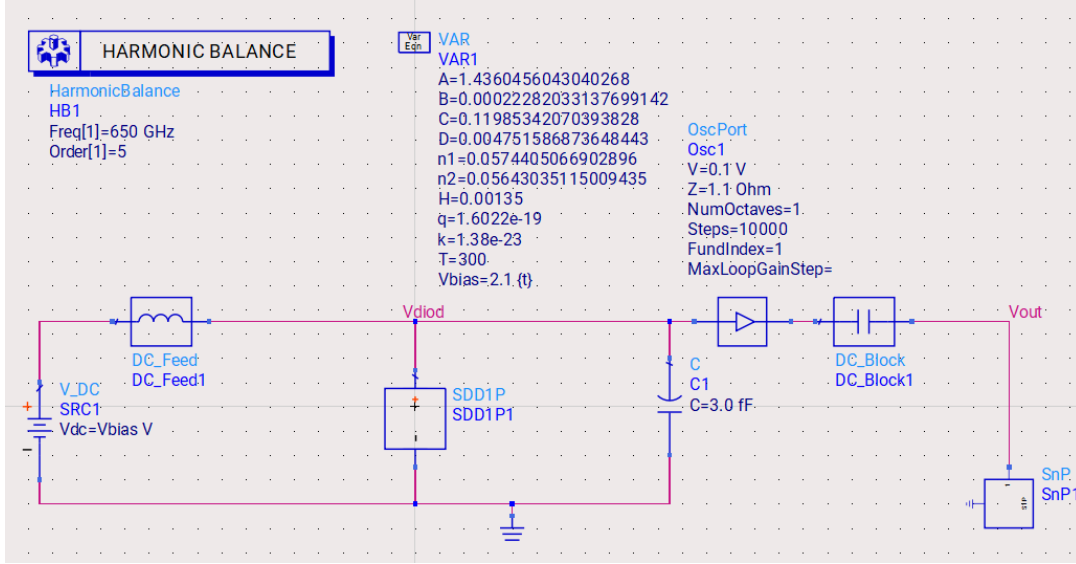


Figure 3.6: ADS harmonic balance setup used for the bias-dependent oscillation analysis.

During the HB analysis, the oscillation frequency f_{osc} and oscillation voltage amplitude V_{out} were monitored under different RTD bias voltages. The quantity V_{out} represents the RF oscillation voltage amplitude across the antenna terminal and is directly related to the available radiated output power. Therefore, a larger V_{out} generally indicates stronger oscillation and higher output power.

The HB simulations showed that the oscillation frequency remained close to 650 GHz throughout the investigated bias range, while the oscillation voltage amplitude varied significantly with the RTD bias voltage.

Finally, a fixed-bias HB simulation was carried out at the selected operating bias voltage in order to further analyze the steady-state oscillation spectrum and estimate the output power.

The detailed oscillation frequency, harmonic spectrum, oscillation voltage amplitude, and output power obtained from the harmonic balance simulations are presented and discussed in Chapter 4.

3.4 Complete Oscillator Structure Including Parasitics

After the simplified slot antenna structure was investigated, a more complete oscillator model was developed in HFSS by including the main parasitic contributions. The purpose of this model was to move from the idealized antenna-only structure toward a more realistic RTD-slot antenna oscillator layout.

The complete electromagnetic structure is shown in Fig. 3.7. Compared with the simplified slot antenna model, additional layers and interconnections were included. From bottom to top, the structure consists of the InP substrate, a highly doped n^+ InGaAs layer, the lower Au electrode, the Si_3N_4 MIM dielectric layer, the upper Au electrode, the air bridge, and the integrated shunt resistor.

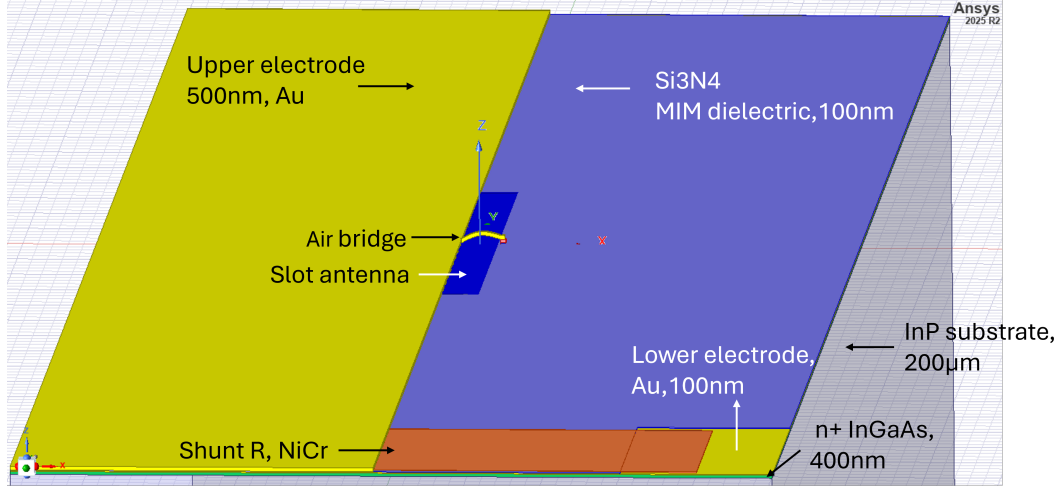


Figure 3.7: Complete RTD-slot antenna oscillator structure including the parasitic layers and integrated shunt resistor used in the full electromagnetic simulation.

The n^+ InGaAs layer was included to represent the highly doped RTD contact layer, which is used to provide an ohmic contact to the device. The original antenna metal layer was retained as the lower electrode, while an additional upper electrode was added above the dielectric layer. The Si_3N_4 layer was used to form MIM capacitor to electrically isolate the upper and lower electrodes while introducing the required capacitive coupling between them. A high breakdown voltage is important for this layer because the MIM dielectric is very thin and can experience a strong electric field between the two metal electrodes. If the dielectric breaks down, leakage current or a short circuit may be introduced between the upper and lower electrodes, which would change the biasing condition and the high-frequency admittance of the oscillator. Therefore, Si_3N_4 is a suitable choice for the MIM dielectric layer because SiN_x thin-film capacitors have been demonstrated for submillimeter-wave integrated circuits and are reported to have high breakdown voltage [28]. A shunt resistor was also included in the complete oscillator structure to suppress unwanted low-frequency oscillations. Although the RTD negative differential conductance is required for oscillation at the designed THz frequency, the same negative conductance can also make the circuit unstable at low frequencies if no damping path is provided. Therefore, the resistor is introduced to provide an additional positive conductance at low frequency and stabilize the bias circuit.

At low frequency, the RTD capacitance can be treated as open and the oscillation condition is mainly determined by the total conductance. To suppress low-frequency oscillation, the conductance of the shunt resistor should be larger than the magnitude of the RTD negative differential conductance,

$$\frac{1}{R_{sh}} > |G_{RTD}|. \quad (3.33)$$

Equivalently, the shunt resistance should satisfy

$$R_{sh} < \frac{1}{|G_{RTD}|}. \quad (3.34)$$

Using the maximum extracted negative differential conductance from the fitted RTD I–V curve, $|G_{RTD}| \approx 9.5 \text{ mS}$, the upper limit of the shunt resistance is approximately

$$R_{sh} < \frac{1}{9.5 \times 10^{-3}} \approx 105 \text{ } \Omega. \quad (3.35)$$

Therefore, a shunt resistance of approximately $95 \text{ } \Omega$ was selected in this work. This value is below the estimated stability limit and provides low-frequency damping while still being close enough to the limit to avoid excessive loading of the intended THz oscillation. NiCr was selected as the resistor material because it is commonly used for integrated thin-film resistors and provides high resistivity and stable sheet resistance [28].

As a first step, an idealized version of the complete oscillator model was constructed. In this model, the airbox was placed directly adjacent to the substrate sidewalls and bottom surface, such that the radiation boundaries coincided with the substrate boundaries. This configuration minimizes substrate-related reflections and cavity resonances, allowing the influence of the oscillator parasitics to be evaluated independently of substrate effects. The corresponding HFSS model is shown in Fig. 3.8. The slot dimensions obtained from the simplified antenna-only model ($l = 57 \text{ } \mu\text{m}$ and $w = 4 \text{ } \mu\text{m}$) were initially adopted in the complete oscillator structure. The extracted circuit admittance is expressed as

$$Y_{\text{circ}} = Y_a + Y_p, \quad (3.36)$$

where Y_a represents the antenna admittance and Y_p denotes the additional parasitic contribution introduced by the surrounding oscillator structure.

The extracted admittance of the complete oscillator structure was compared with the antenna-only admittance obtained previously. The introduction of the parasitic structures produced only a small change in the extracted conductance and susceptance, and the oscillation condition remained close to the target frequency of 650 GHz.

Although the initial geometry satisfied the oscillation condition, maximum radiated power is not necessarily achieved at the same slot dimensions. Therefore, additional parameter sweeps of slot length and slot width were performed using the complete oscillator model. The objective was to identify geometries that simultaneously satisfy the oscillation condition and maximize the achievable output power. The detailed comparison of the extracted admittance, oscillation frequency, voltage amplitude, efficiency, and estimated output power is presented in Chapter 4.

The extracted HFSS admittance data were subsequently imported into ADS for oscillator analysis. First, a small-signal admittance analysis was performed to identify

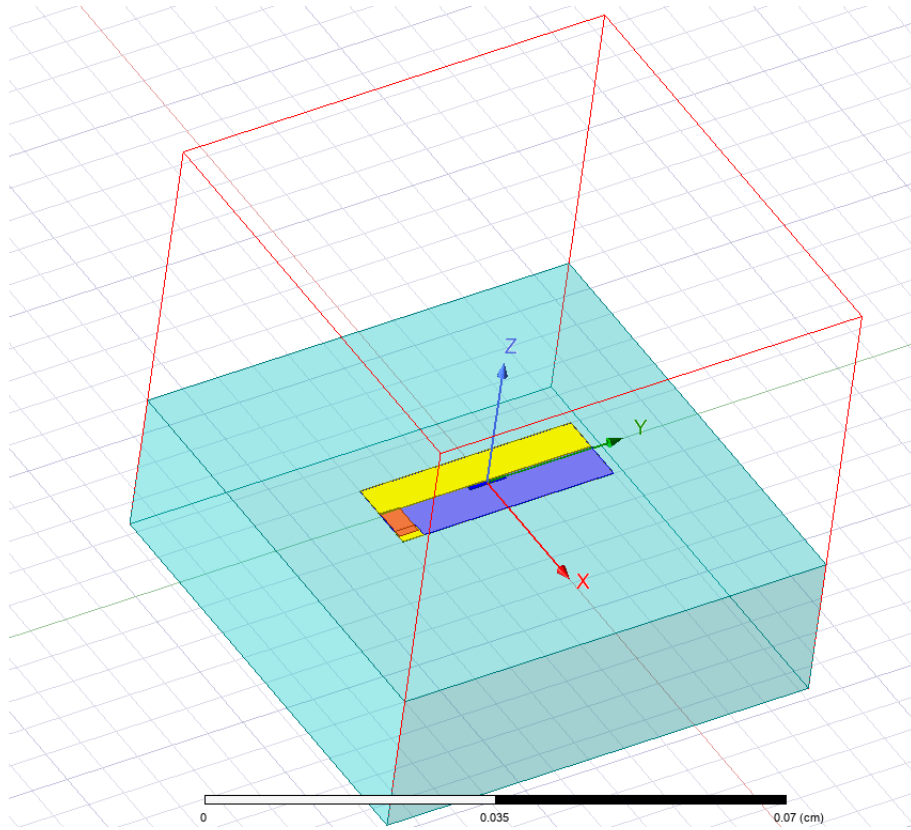


Figure 3.8: Idealized complete oscillator model used for the initial evaluation of the parasitic structure. The substrate sidewalls and bottom surface are directly terminated by the radiation boundary to suppress substrate reflections.

the RTD bias range satisfying the oscillation condition. Harmonic balance simulations were then carried out to determine the oscillation frequency and steady-state voltage amplitude. Finally, for each geometry, the bias point providing the largest oscillation amplitude was selected and used for output-power estimation. To further verify the oscillation frequency predicted by the harmonic balance simulations, transient analysis was also performed in ADS. The HFSS admittance data were exported over a wide frequency range from 0.1 GHz to 2 THz and incorporated into the transient oscillator model. A small current pulse source was introduced to provide an initial perturbation and trigger oscillation startup. Since the pulse is only applied during the startup phase, it does not influence the final steady-state oscillation behaviour. The transient simulation schematic is shown in Fig. 3.9.

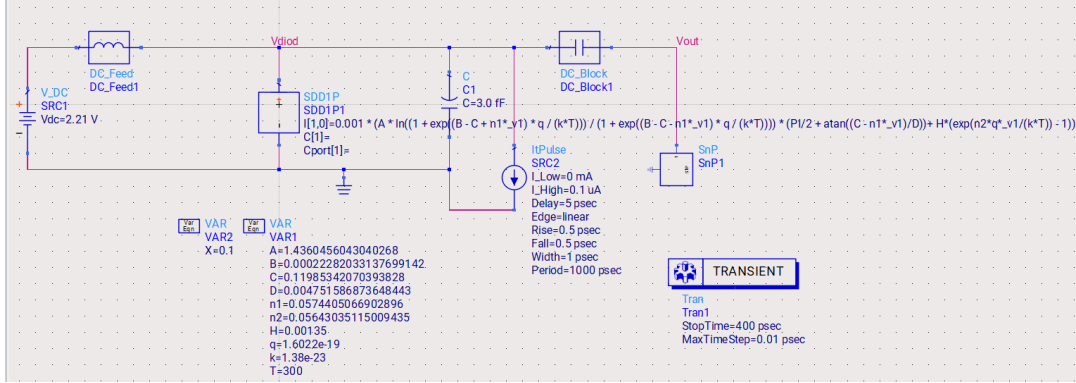


Figure 3.9: ADS transient simulation setup used to verify the oscillation frequency predicted by the harmonic balance simulations. A small current pulse is introduced to initiate oscillation startup.

The transient response was subsequently analysed in both the time and frequency domains. The resulting oscillation frequency was compared with the harmonic balance prediction to verify the consistency of the oscillator model and to investigate the existence of possible competing oscillation modes over a broad frequency range. Finally, a more realistic electromagnetic model was constructed to investigate the influence of substrate reflections. Compared with the idealized model, the substrate thickness was increased to 600 μm and an air region was introduced between the substrate bottom surface and the radiation boundary. This configuration allows reflections from the dielectric–air interface and therefore more closely represents the physical device environment. The corresponding model is shown in Fig. 3.10.

The extracted admittance of the realistic model was compared with that of the idealized model in order to evaluate the sensitivity of the oscillation condition to substrate-related resonances. The resulting differences in conductance, susceptance, oscillation frequency, and output power are discussed in Chapter 4.

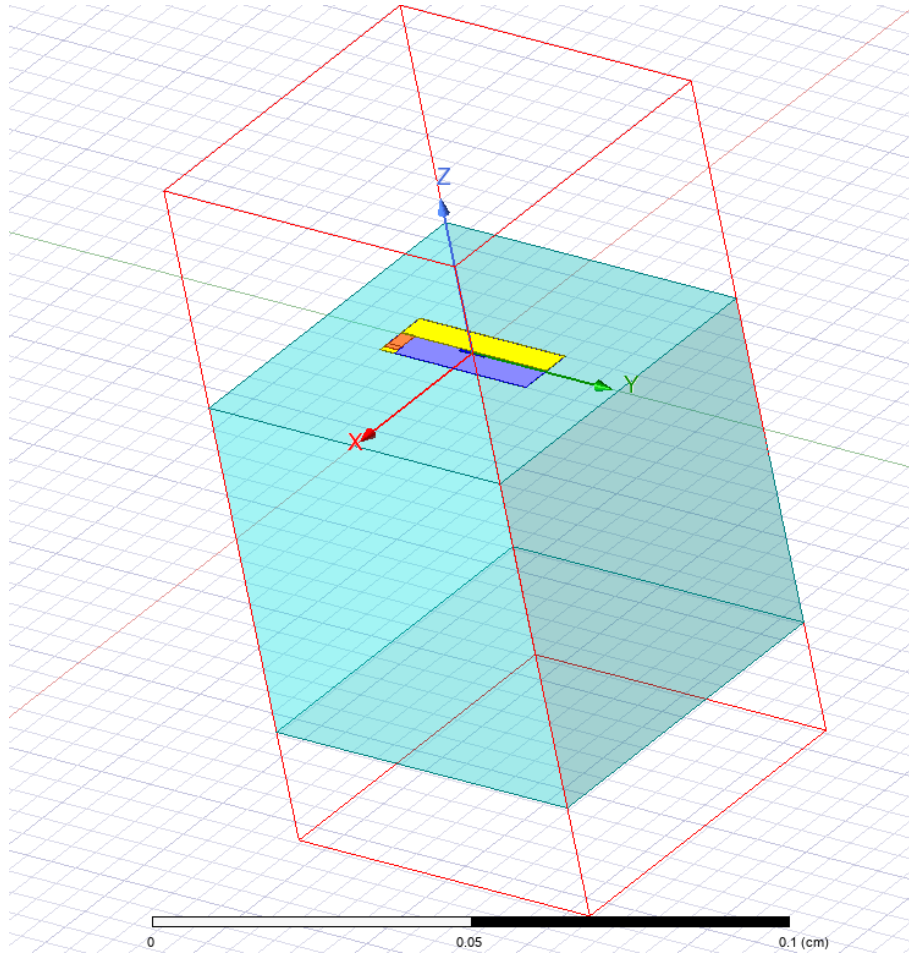


Figure 3.10: Realistic complete oscillator model used to evaluate the influence of substrate reflections. The substrate thickness is increased to $600\text{ }\mu\text{m}$ and an air region is included below the substrate before the radiation boundary.

4

Results and Discussion

4.1 Simplified RTD-Slot Antenna Oscillator

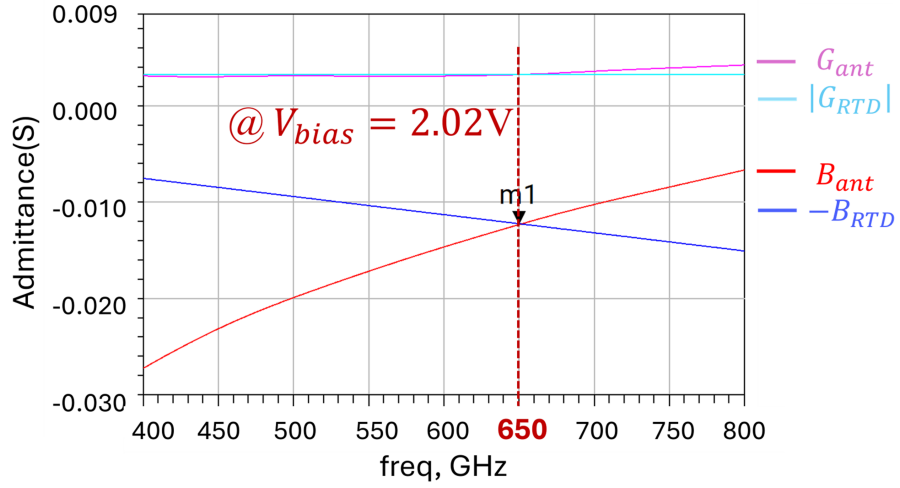
4.1.1 S-parameter Analysis and Oscillation Condition

After importing the simplified RTD-slot antenna model from EM simulation into non linear circuit analysis, a S-parameter analysis was performed to evaluate the oscillation condition under different RTD bias voltages.

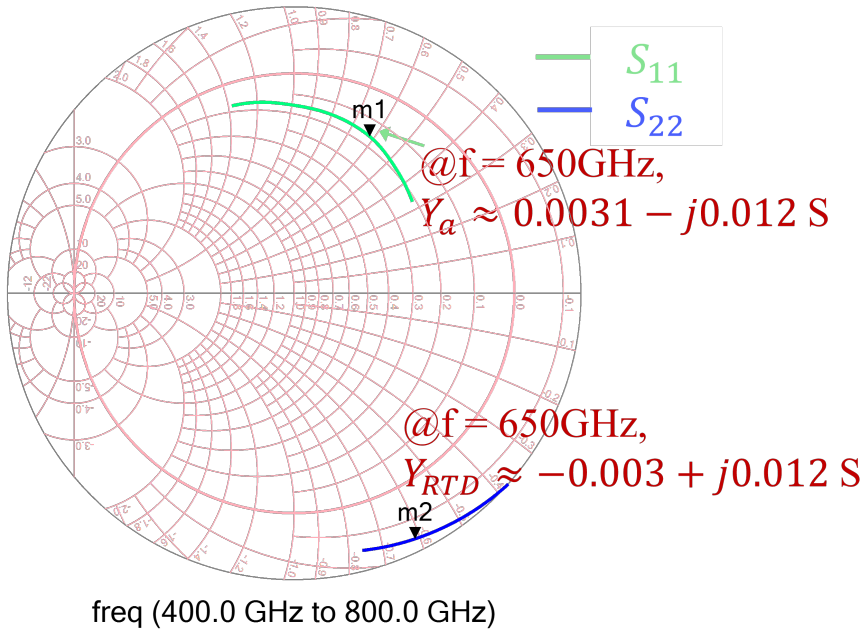
Figure 4.1 shows the evaluation of oscillation condition at a bias voltage of 2.02 V. In Fig. 4.1a, the conductance condition $G_a < |G_{RTD}|$ is just satisfied, while the susceptance curves intersect close to 650 GHz. This is also reflected in Fig. 4.1b, where the antenna and RTD admittances are approximately balanced in their reactive parts at 650 GHz. Since the magnitude of the RTD negative conductance is slightly larger than the antenna conductance, the oscillation condition is marginally satisfied, making 2.02 V the lower boundary of the oscillation region.

Figure 4.2 shows the oscillation condition at a bias voltage of 2.10 V, which is close to the point where the RTD provides its maximum negative differential conductance. Compared with the case at 2.02 V, the magnitude of the RTD negative conductance is much larger, so the conductance margin is clearly increased. Meanwhile, the susceptance curves still intersect close to 650 GHz, meaning that the oscillation frequency is almost unchanged. This behavior is also seen in Fig. 4.2b, at 650 GHz, the reactive parts of the antenna and RTD admittances are still approximately balanced, while the RTD negative conductance is clearly larger than the antenna conductance. Therefore, this bias point also satisfies the oscillation requirements.

Figure 4.3 shows the oscillation condition at a bias voltage of 2.15 V. Compared with the case at 2.10 V, the magnitude of the RTD negative conductance has started to decrease. Nevertheless, the conductance condition $G_a < |G_{RTD}|$ is still satisfied, while the susceptance curves continue to intersect close to 650 GHz. The same trend can be observed in Fig. 4.3b. At 650 GHz, the antenna and RTD admittances remain approximately balanced in their reactive parts, indicating that the oscillation frequency is largely unchanged. However, the distance between the RTD and antenna conductances becomes smaller as the bias voltage increases beyond the maximum negative conductance point. This suggests that the available negative conductance margin gradually decreases, although the oscillation condition is still maintained.

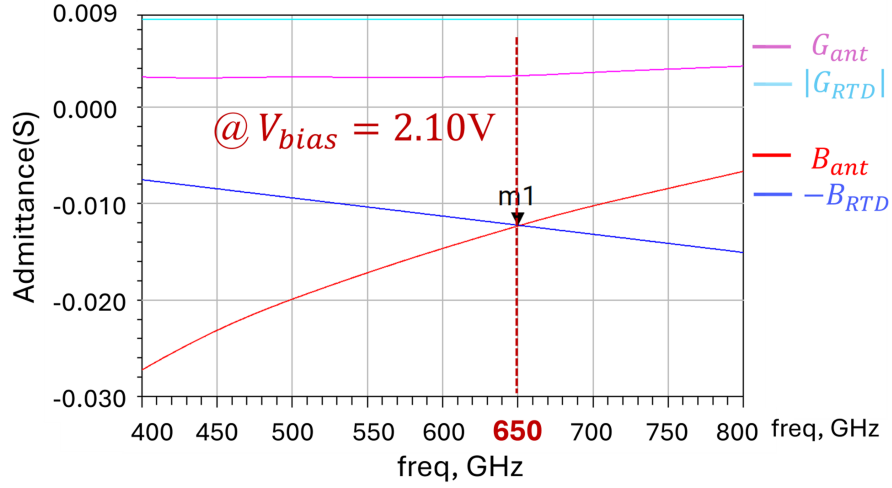
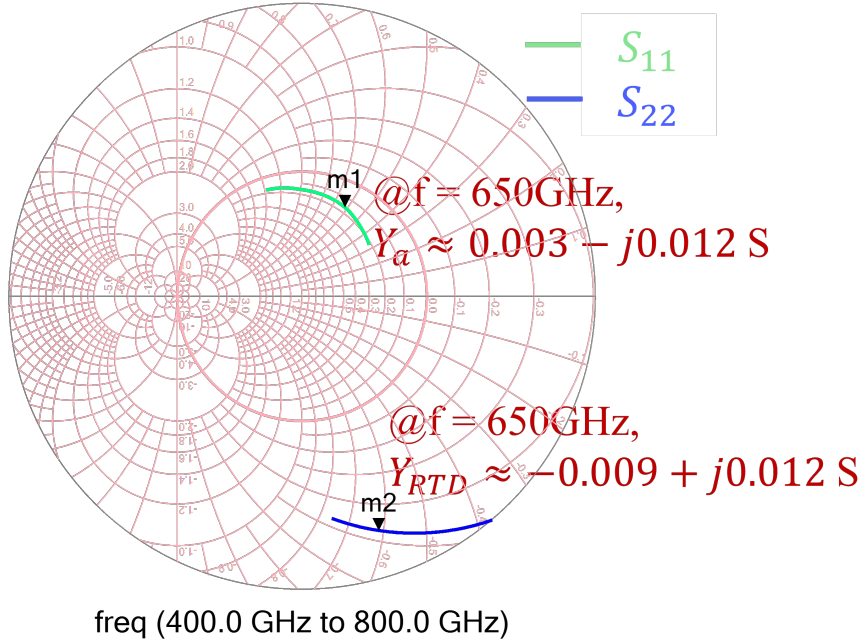


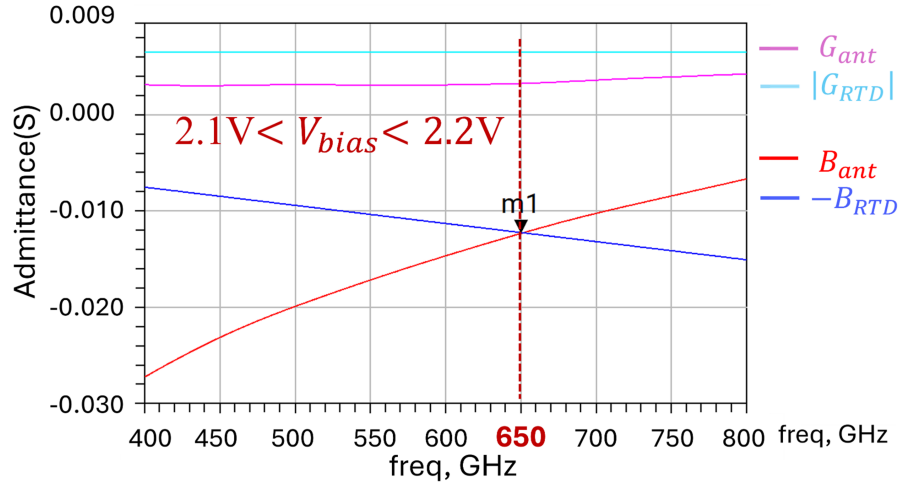
(a) Admittance comparison at $V_{bias} = 2.02\text{ V}$.



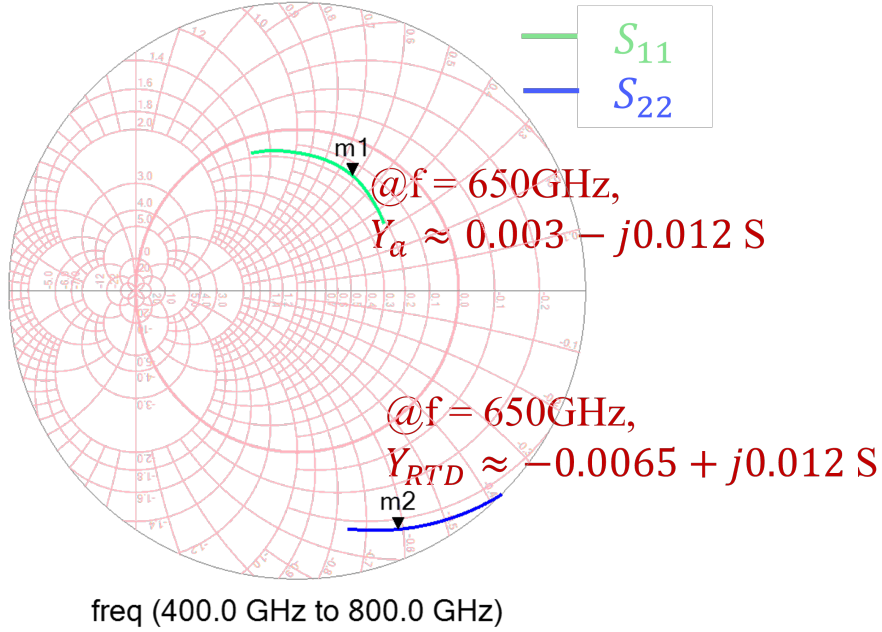
(b) Smith chart at $V_{bias} = 2.02\text{ V}$.

Figure 4.1: Evaluation of the oscillation condition at $V_{bias} = 2.02\text{ V}$: (a) admittance comparison and (b) Smith chart.

(a) Admittance comparison at $V_{bias} = 2.10$ V.(b) Smith chart at $V_{bias} = 2.10$ V.**Figure 4.2:** Oscillation condition at $V_{bias} = 2.10$ V: (a) admittance comparison and (b) Smith chart representation.



(a) Admittance comparison near the upper part of the oscillation region.



(b) Smith chart at $V_{bias} = 2.15 \text{ V}$.

Figure 4.3: Oscillation condition at $V_{bias} = 2.15 \text{ V}$: (a) admittance comparison and (b) Smith chart representation.

4.1.2 Harmonic Balance Analysis

After identifying the possible oscillation region from the S-parameter analysis, harmonic balance (HB) simulations were performed in ADS. The RTD bias voltage was swept from 2.02 V to 2.20 V in order to investigate the bias dependence of the oscillation frequency and oscillation amplitude. The resulting HB analysis is shown in Fig. 4.4 and Fig. 4.5.

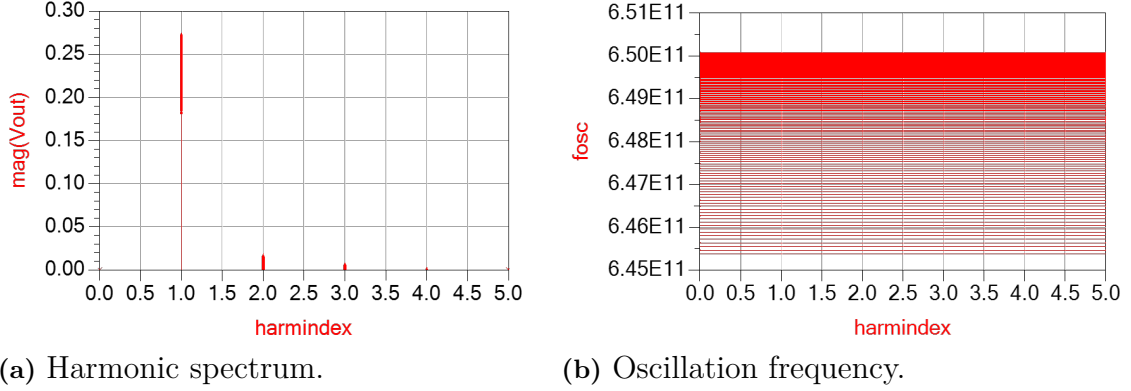


Figure 4.4: Harmonic balance results of the simplified RTD-slot antenna oscillator: (a) harmonic spectrum and (b) Oscillation frequency over the bias range from 2.02 V to 2.20 V.

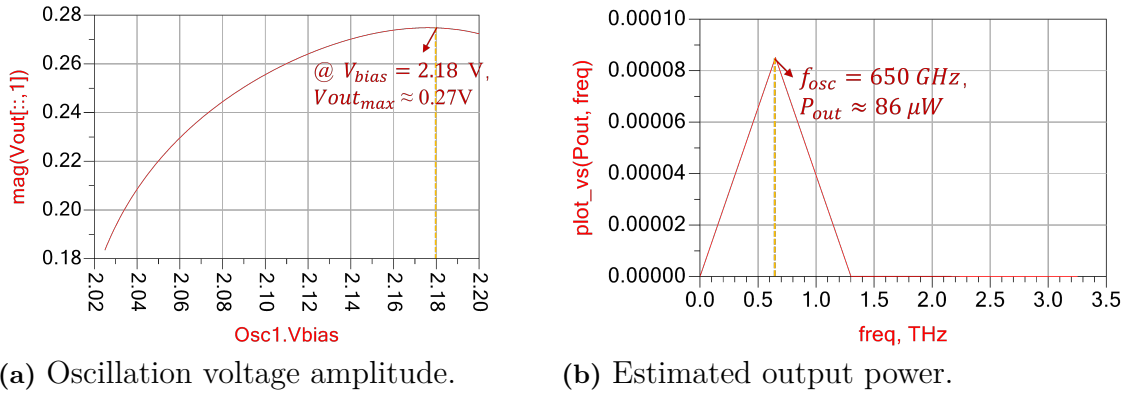


Figure 4.5: Bias-dependent oscillation amplitude and estimated output power: (a) oscillation voltage amplitude and (b) estimated output power.

Figure 4.4a shows the harmonic spectrum of the oscillation voltage. The fundamental component is clearly dominant, while the higher-order harmonics remain much smaller. This indicates that the oscillator operates mainly at a single frequency around the fundamental mode. Figure 4.4b shows the oscillation frequency obtained for different RTD bias voltages. Although a slight variation can be observed, the oscillation frequency remains close to 650 GHz throughout the investigated bias range. This agrees well with the oscillation frequency predicted from the admittance analysis presented in the previous section.

Figure 4.5a shows the oscillation voltage amplitude as a function of bias voltage. As the bias voltage increases, the oscillation amplitude first increases and then decreases. The maximum oscillation amplitude is obtained at approximately $V_{\text{bias}} = 2.18$ V, where $V_{\text{out,max}} \approx 0.27$ V. The corresponding estimated output power is shown in Fig. 4.5b. A clear maximum appears near 650 GHz, corresponding to the oscillation frequency of the RTD-slot antenna oscillator. The estimated radiated output power is calculated as

$$P_{\text{out}} = G_a \eta_{\text{rad}} |V_{\text{out}}|^2 \quad (4.1)$$

where G_a is the antenna conductance at the oscillation frequency, η_{rad} is the radiation efficiency obtained from the HFSS simulation, and V_{out} is the oscillation voltage amplitude extracted from the harmonic balance simulation. Using an antenna conductance of approximately $G_a \approx 0.003$ S and a radiation efficiency of about 68%, the maximum radiated output power is estimated to be approximately 86 μ W at around 650 GHz.

4.2 Complete Oscillator Structure Including Parasitics

4.2.1 HFSS Analysis of the Complete Oscillator Structure

After the simplified antenna-only model was analysed, the complete oscillator structure including parasitic elements was simulated in HFSS. The idealized complete model described in Section 3.4 was first used, where the airbox was placed directly adjacent to the substrate sidewalls and bottom surface to reduce substrate-related reflections.

The slot dimensions obtained from the simplified antenna-only model, $w = 4$ μ m and $l = 57$ μ m, were first applied to the complete oscillator structure. The extracted admittance is shown in Fig. 4.6.

At 650 GHz, the extracted admittance of the complete oscillator structure is approximately

$$Y_{a,\text{complete}} \approx 0.0032 - j0.0123 \text{ S}. \quad (4.2)$$

For comparison, the antenna-only model with the same slot dimensions gave

$$Y_a \approx 0.0033 - j0.0123 \text{ S}. \quad (4.3)$$

The difference between Y_{circ} and Y_a is small, indicating that the added parasitic structures do not significantly shift the admittance at the target frequency. Therefore, the slot dimensions obtained from the simplified antenna-only model remain a suitable starting point for the complete oscillator analysis.

To further optimize the complete oscillator structure, a sweep of slot width and slot length was performed. The selected geometries are summarized in Table 4.2. Since increasing either the slot width or the slot length reduces the magnitude of the susceptance, the sweep was chosen such that the susceptance around 650 GHz

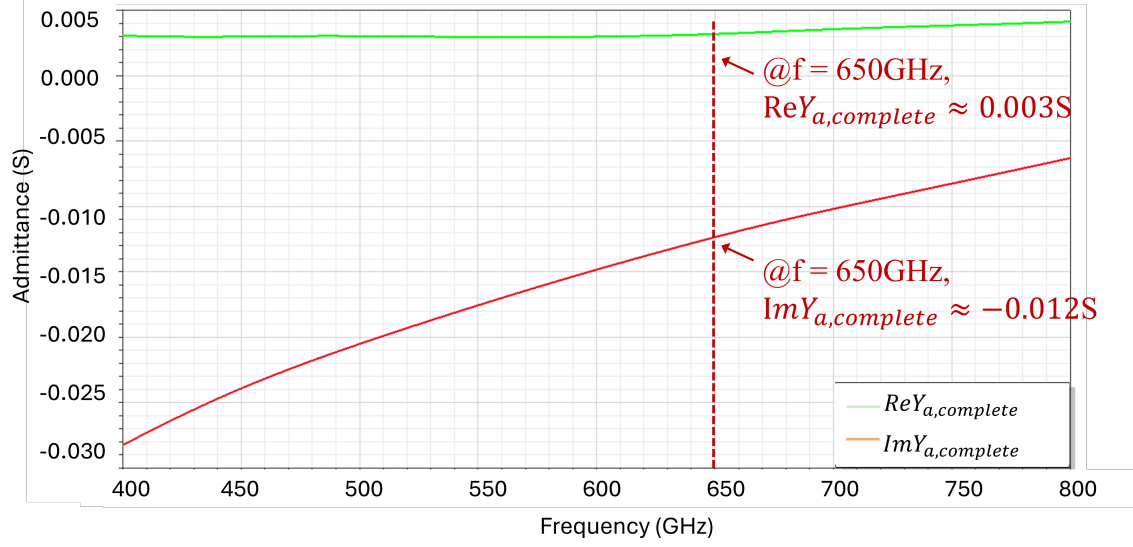


Figure 4.6: Extracted admittance of the complete oscillator structure with $w = 4 \mu\text{m}$ and $l = 57 \mu\text{m}$.

remained close to the target value of approximately -0.012 S . Starting from the initial geometry of $w = 4 \mu\text{m}$ and $l = 57 \mu\text{m}$, two sweep directions were therefore considered: decreasing the slot width while increasing the slot length, and increasing the slot width while decreasing the slot length, , as presented in Table 4.2

Table 4.1: Slot geometries used in the HFSS sweep of the complete oscillator structure.

Slot width w [μm]	Slot length l [μm]
1	70
2	65
3	61
4	57
5	54
6	51
7	48
8	45
9	42
10	39

The imaginary part of the extracted admittance for the swept geometries is shown in Fig. 4.7. All selected geometries provide a susceptance close to -0.012 S near 650 GHz . This confirms that the sweep was able to keep the resonance condition approximately fixed while changing the slot width and length.

The corresponding real part of the extracted admittance is shown in Fig. 4.8. Although the susceptance remains close to the target value for all cases, the conductance changes significantly with geometry. In general, shorter slot lengths lead to smaller $G_{a,\text{complete}}$. For example, the geometry with $w = 1 \mu\text{m}$ and $l = 70 \mu\text{m}$

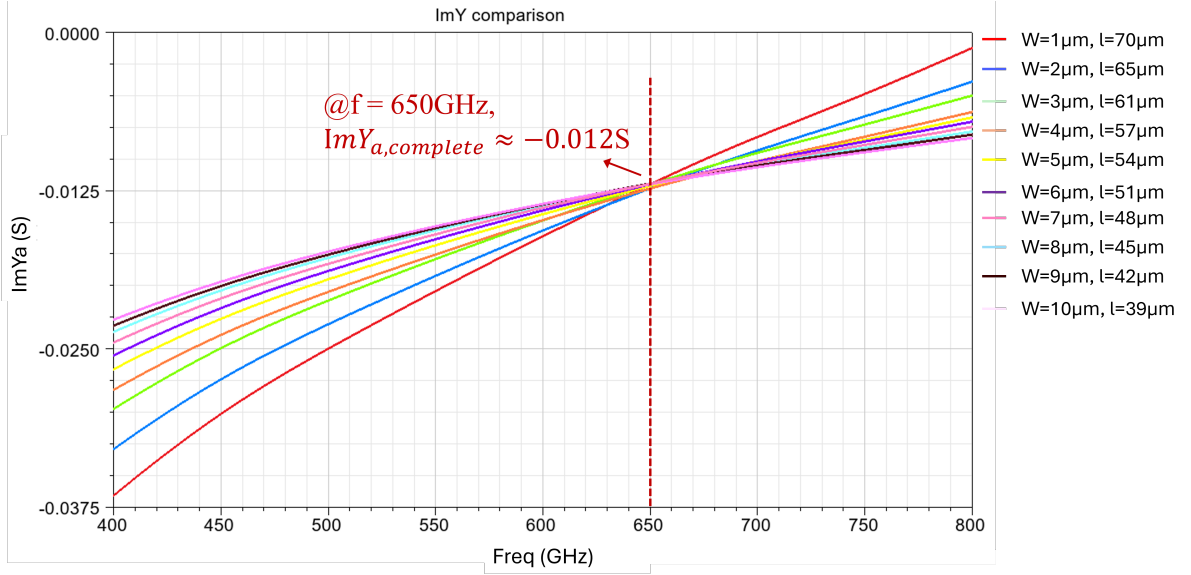


Figure 4.7: Comparison of the imaginary part of the extracted admittance for different slot geometries in the complete oscillator structure. The geometries were selected to keep $\text{Im}(Y_{a,\text{complete}})$ close to -0.012 S near 650 GHz.

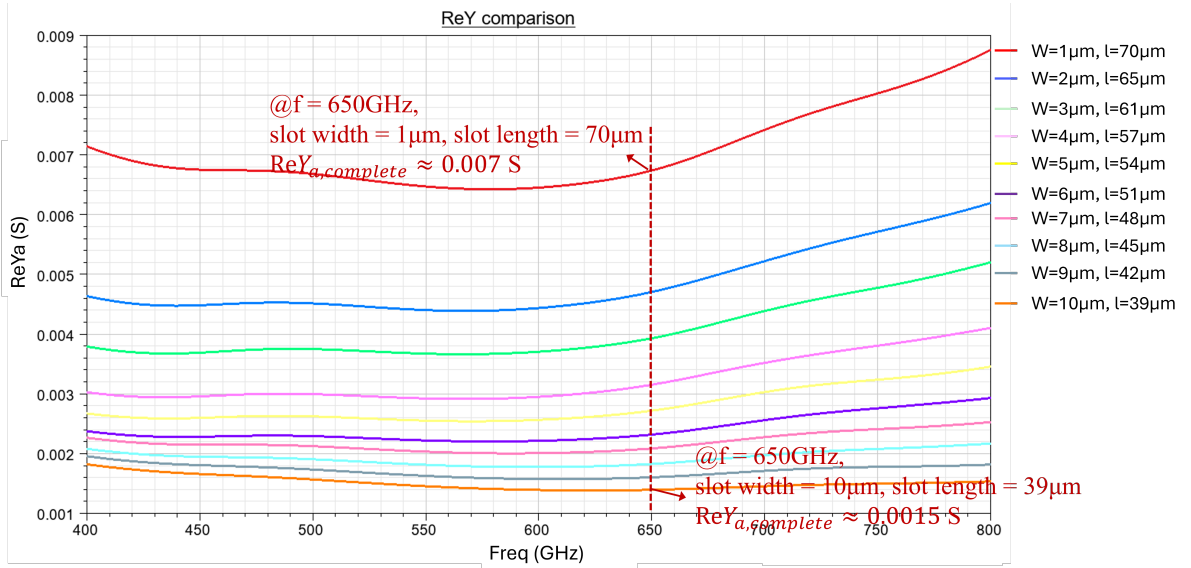


Figure 4.8: Comparison of the real part of the extracted admittance for different slot geometries in the complete oscillator structure. Shorter slot lengths generally lead to smaller conductance around 650 GHz.

gives a conductance of approximately 0.007 S at 650 GHz, while the geometry with $w = 10 \mu\text{m}$ and $l = 39 \mu\text{m}$ gives a conductance of approximately 0.0015 S. A smaller conductance reduces the loss that must be compensated by the RTD negative differential conductance and therefore increases the available bias range for oscillation. However, a smaller conductance does not necessarily correspond to the maximum output power, since the radiated power also depends on the radiation conductance, radiation efficiency, and oscillation voltage amplitude.

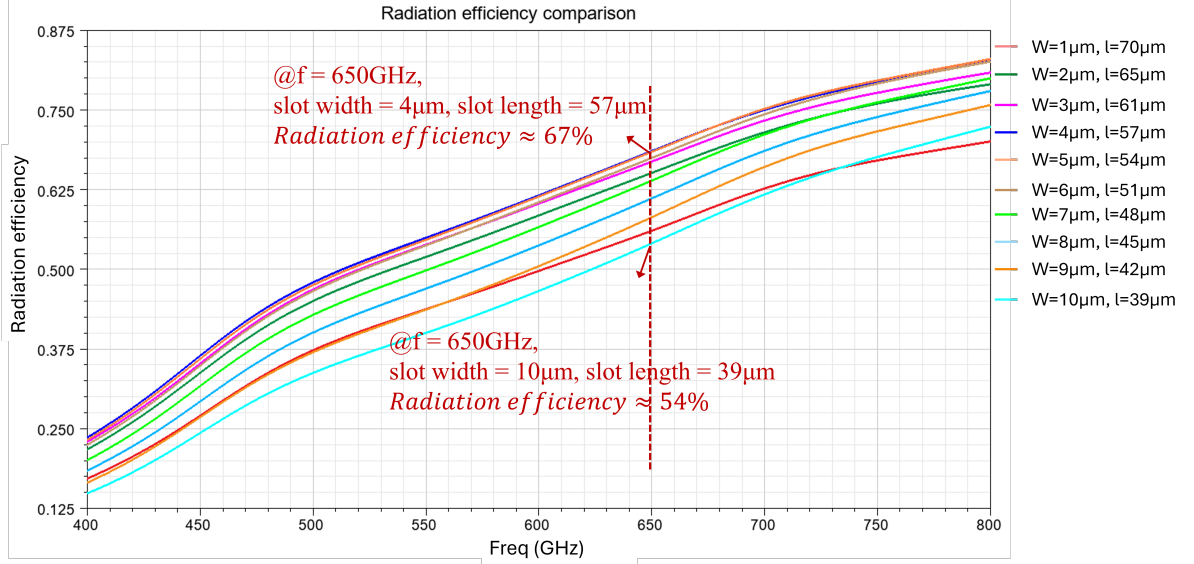


Figure 4.9: Radiation efficiency comparison for different slot geometries in the complete oscillator structure. The radiation efficiency changes with slot geometry and must be considered together with the extracted admittance.

The radiation efficiency for the same geometries is shown in Fig. 4.9. The radiation efficiency also varies with geometry. Around 650 GHz, the initial geometry with $w = 4 \mu\text{m}$ and $l = 57 \mu\text{m}$ has a radiation efficiency of approximately 67%, while the geometry with $w = 10 \mu\text{m}$ and $l = 39 \mu\text{m}$ has a lower radiation efficiency of approximately 54%. This indicates that reducing the conductance by using shorter slot lengths can also reduce the radiation efficiency.

Therefore, the optimum geometry cannot be determined from $G_{a,\text{complete}}$ or radiation efficiency alone. Instead, the output power must be evaluated by combining the extracted admittance, the antenna admittance, the radiation efficiency, and the oscillation voltage obtained from ADS harmonic balance simulations. The output power is calculated using

$$P_{\text{out}} = \frac{1}{2} G_{a,\text{complete}} \eta_{\text{rad}} |V_{\text{out}}|^2. \quad (4.4)$$

The detailed ADS analysis and output power comparison for the swept geometries are presented in the following section.

4.2.2 ADS Analysis of the Complete Oscillator Structure

Based on the HFSS optimization presented in the previous section, the geometry with a slot width of $6\ \mu\text{m}$ and a slot length of $51\ \mu\text{m}$ was selected for detailed circuit-level analysis in ADS. This geometry provides a favorable trade-off between conductance, susceptance, and radiation efficiency, and was found to give the highest estimated output power among the investigated designs.

4.2.2.1 S-parameter Analysis

The extracted admittance of the complete oscillator was imported into ADS and combined with the RTD admittance model. A small-signal S-parameter analysis was performed to evaluate the oscillation condition under different RTD bias voltages. Figure 4.10 shows the admittance comparison and the corresponding Smith chart at $V_{\text{bias}} = 2.01\ \text{V}$. The conductance condition is just satisfied, while the susceptance curves intersect close to $646\ \text{GHz}$. Therefore, this bias voltage can be regarded as the lower boundary of the oscillation region. The Smith chart provides an alternative view of the same result. At $650\ \text{GHz}$, the extracted oscillator admittance is approximately $Y_{\text{circ}} \approx 0.0031 - j0.012\ \text{S}$, while the RTD admittance is $Y_{\text{RTD}} \approx -0.003 + j0.012\ \text{S}$. The two admittances are nearly conjugate matched, indicating that the susceptances cancel each other and the conductance condition is approximately fulfilled. This agrees well with the admittance comparison shown in Fig. 4.10(a).

At $V_{\text{bias}} = 2.10\ \text{V}$, shown in Fig. 4.11(a), the RTD provides the largest negative differential conductance. The susceptance intersection remains close to $646\ \text{GHz}$, indicating that the oscillation frequency is still primarily determined by the balance between the complete-oscillator susceptance and the RTD capacitance. Figure 4.11(b) shows the corresponding Smith chart representation. Around $646\ \text{GHz}$, the imaginary parts of the complete-oscillator admittance and RTD admittance remain balanced, while the magnitude of the RTD negative conductance is significantly larger than the oscillator conductance. This bias point therefore provides the largest conductance margin and represents the most favourable condition for oscillation.

As the bias voltage increases further, the magnitude of the RTD negative conductance decreases again. Figure 4.12 shows the condition close to the upper boundary of the oscillation region. Although the susceptance cancellation condition is still satisfied near $650\ \text{GHz}$, the conductance margin becomes smaller, eventually limiting the oscillation range.

Based on the S-parameter analysis, the complete oscillator is expected to oscillate within the approximate bias range of $2.01\ \text{V}$ to $2.23\ \text{V}$. Compared with the simplified antenna-only model, which oscillates from approximately $2.02\ \text{V}$ to $2.20\ \text{V}$, the complete oscillator exhibits a slightly wider oscillation region. This indicates that the additional parasitic structures only have a minor influence on the oscillation condition and do not significantly alter the admittance matching around the target frequency.

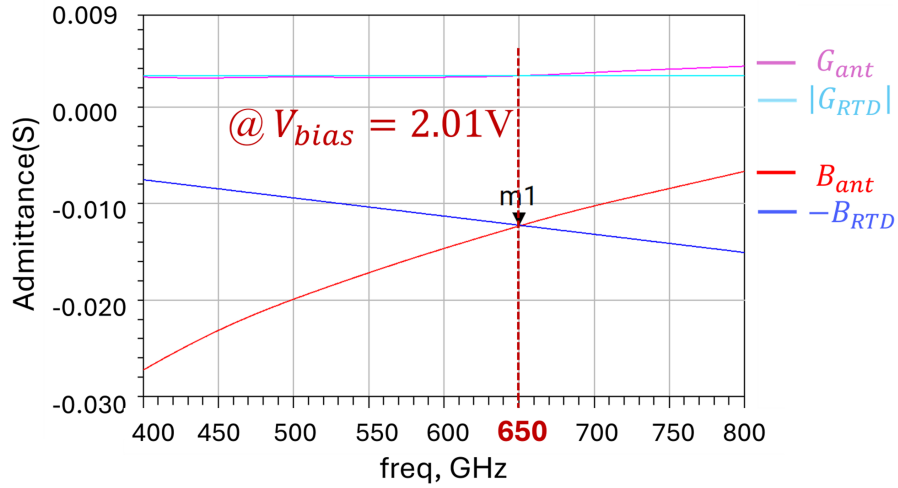
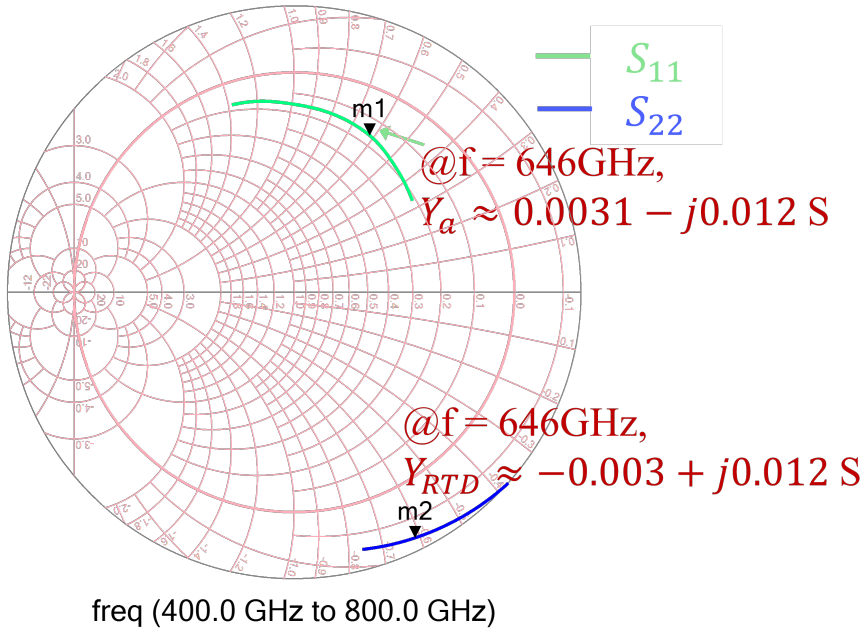
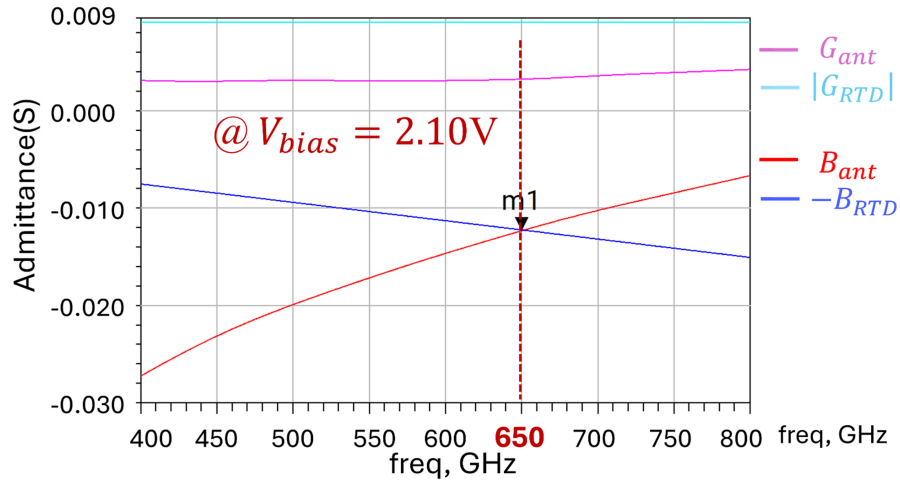
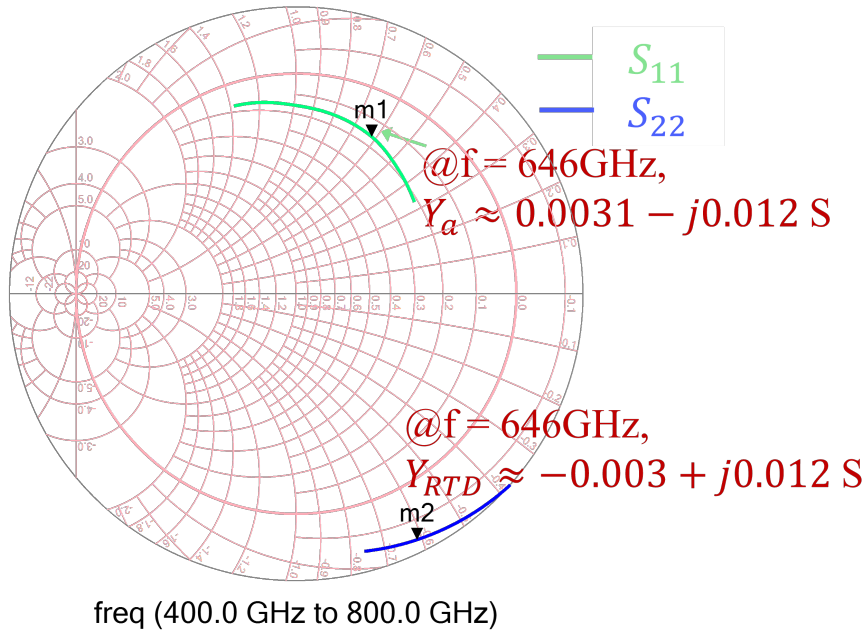

 (a) Admittance comparison at $V_{\text{bias}} = 2.01$ V.

 (b) Smith chart at $V_{\text{bias}} = 2.01$ V.

Figure 4.10: Oscillation-condition verification for the complete oscillator at $V_{\text{bias}} = 2.01$ V. The admittance comparison and Smith chart consistently indicate that the oscillation condition is reached close to 646 GHz.



(a) Admittance comparison at $V_{bias} = 2.10$ V.



(b) Smith chart at $V_{bias} = 2.10$ V.

Figure 4.11: Oscillation-condition verification for the complete oscillator at $V_{bias} = 2.10$ V.

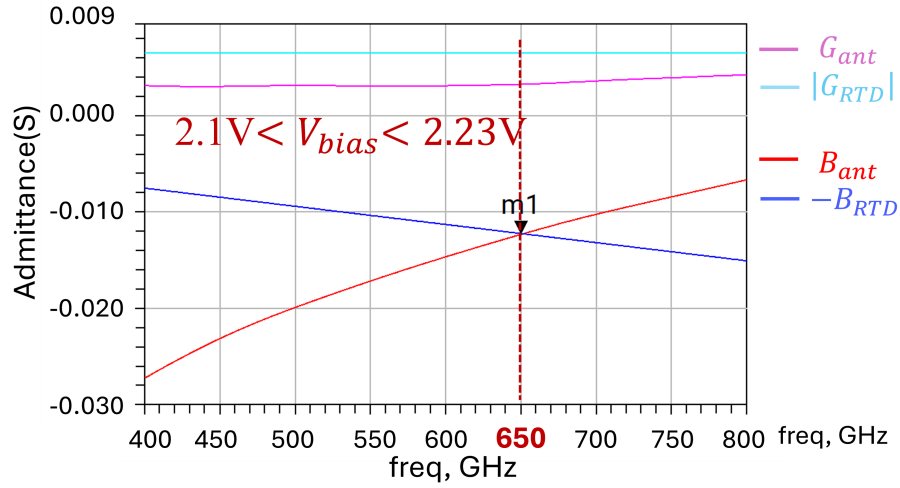
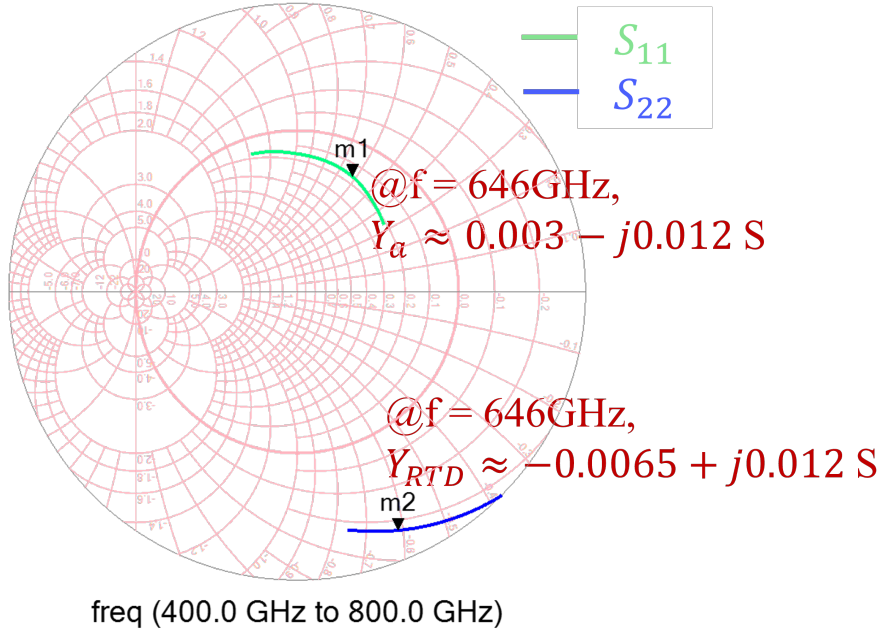

 (a) Admittance comparison at $V_{\text{bias}} = 2.15$ V.

 (b) Smith chart at $V_{\text{bias}} = 2.15$ V.

Figure 4.12: Oscillation-condition verification for the complete oscillator at $V_{\text{bias}} = 2.15$ V.

4.2.2.2 Harmonic Balance Analysis

After identifying the oscillation region from the S-parameter analysis, harmonic balance (HB) simulations were performed in ADS. The RTD bias voltage was swept from approximately 2.01 V to 2.23 V in order to investigate the dependence of the oscillation frequency and oscillation amplitude on the bias voltage. The resulting HB analysis is shown in Fig. ??.

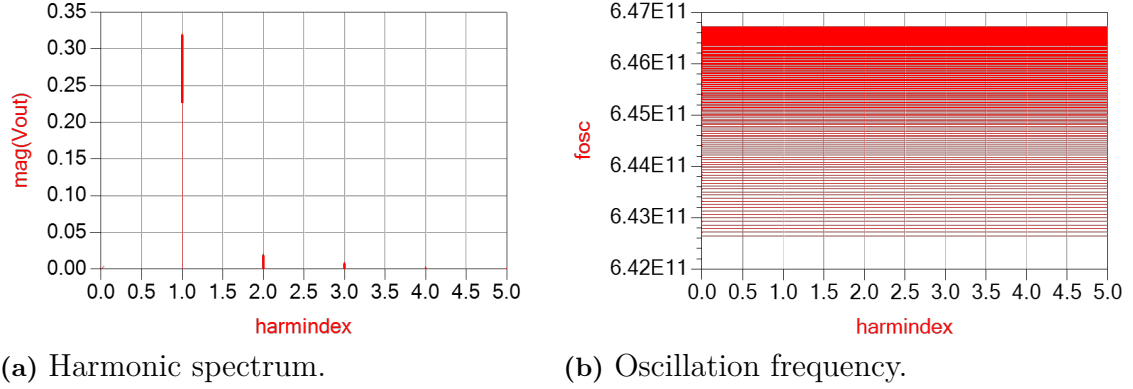


Figure 4.13: Harmonic balance results of the complete oscillator: (a) harmonic spectrum and (b) Oscillation frequency over the bias range from 2.01 V to 2.23 V.

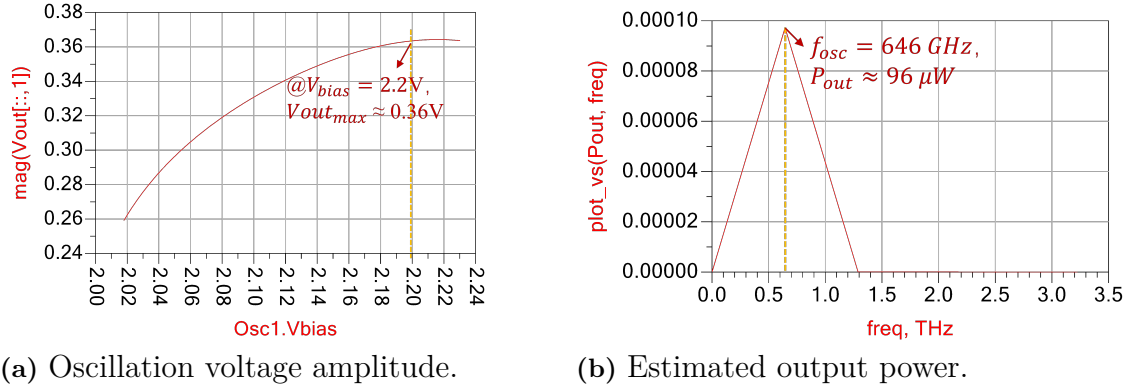


Figure 4.14: Bias-dependent oscillation amplitude and estimated output power: (a) oscillation voltage amplitude and (b) estimated output power.

Figure 4.13a shows the harmonic spectrum of the oscillation voltage, while Fig. 4.13b presents the oscillation frequency obtained at different RTD bias voltages. Similar to the simplified model, the oscillation frequency remains relatively stable around 646 GHz. However, a slight frequency shift can be observed as the bias voltage changes.

Figure 4.14a shows the oscillation voltage amplitude as a function of bias voltage. The oscillation amplitude increases as the RTD negative differential conductance grows, reaches a maximum near $V_{\text{bias}} \approx 2.2$ V, and then decreases as the available negative conductance becomes smaller toward the end of the NDC region. Using

the oscillation voltage amplitude obtained from the harmonic balance simulation together with the admittance and radiation efficiency extracted from HFSS, the output power is estimated as shown in Fig. 4.14b, the radiated output power is approximately $P_{\text{out,max}} \approx 96 \mu\text{W}$.

4.2.2.3 Geometry Comparison and Output Power Optimization

The complete oscillator analysis described above was repeated for all geometries listed in Table 4.2. For each geometry, the optimum RTD bias voltage was first identified through harmonic balance simulations, after which the oscillation frequency, oscillation voltage amplitude, and output power were extracted.

Table 4.2: Performance comparison of the investigated complete oscillator geometries.

Slot Width (μm)	Slot Length (μm)	G_a (S)	η_{rad} (-)	V_{out} (V)	P_{out} (μW)
1	70	0.0070	0.559	0.11	25
2	65	0.0050	0.654	0.18	56
3	61	0.0040	0.664	0.23	69
4	57	0.0033	0.678	0.28	85
5	54	0.0028	0.674	0.32	95
6	51	0.0023	0.669	0.36	96
7	48	0.0020	0.630	0.39	93
8	45	0.0017	0.599	0.42	85
9	42	0.0014	0.569	0.46	84
10	39	0.0012	0.528	0.50	77

Table 4.2 shows that the oscillation voltage amplitude increases monotonically as the slot width increases from 1 μm to 10 μm while the slot length decreases from 70 μm to 39 μm . Over this range, the extracted oscillation amplitude increases from approximately 0.11 V to 0.50 V. However, the output power does not follow the same monotonic trend.

As shown in Tab. 4.2, both G_a and η_{rad} gradually decrease as the slot width increases and the slot length decreases. Consequently, although the oscillation voltage continues to increase, the effective radiation conductance

$$G_r = G_a \eta_{\text{rad}} \quad (4.5)$$

becomes progressively smaller. Therefore, the increase in oscillation amplitude is eventually outweighed by the reduction in radiated power conversion efficiency.

As a result, the estimated output power first increases and then decreases across the investigated geometries. The maximum output power is obtained for the geometry with a slot width of 6 μm and a slot length of 51 μm , where a favorable balance between oscillation amplitude, antenna conductance, and radiation efficiency is achieved. This geometry delivers an estimated output power of approximately 96 μW and is therefore selected as the optimum design of the complete RTD-slot antenna oscillator.

4.2.2.4 Wideband Verification and Transient Analysis

To verify that no additional oscillation modes exist outside the frequency range investigated previously, the complete antenna structure with a slot width of $6\ \mu\text{m}$ and a slot length of $51\ \mu\text{m}$ was simulated in HFSS over a wide frequency range from 1 GHz to 2 THz, and then implemented in transient analysis.

The resulting real and imaginary parts of the antenna admittance are shown in Fig. 4.15 and Fig. 4.16, respectively.

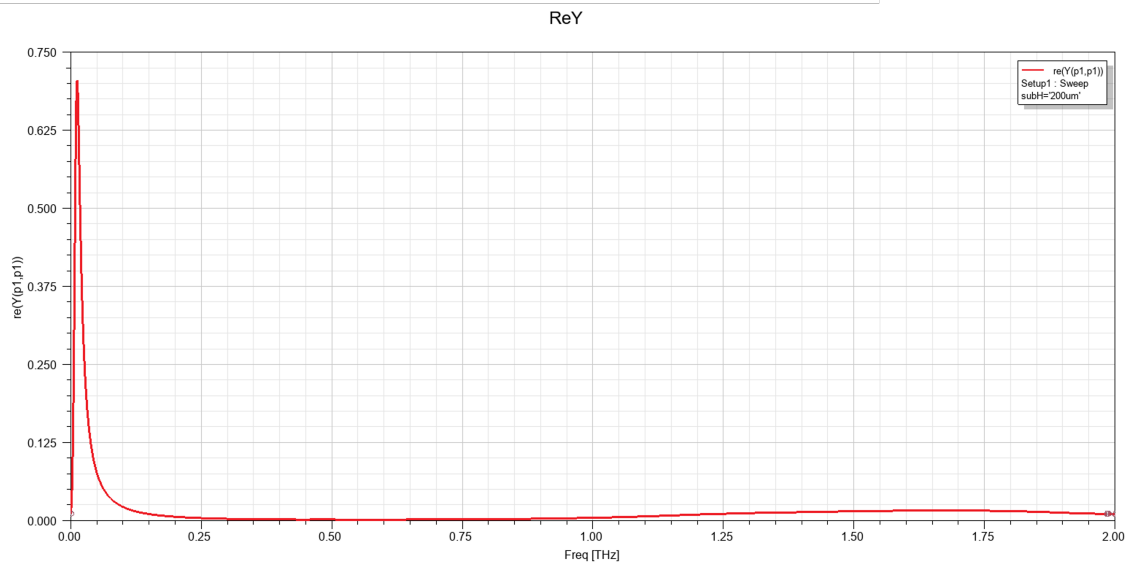


Figure 4.15: Wideband real part of the oscillator admittance from 1 GHz to 2 THz.

Figure 4.15 shows that the real part of the oscillator admittance is relatively large at very low frequencies and rapidly decreases as the frequency increases. This behavior is mainly caused by the shunt resistor introduced in the complete oscillator structure. At low frequencies, the large capacitive reactance of the RTD limits current flow through the RTD branch, causing a significant portion of the current to flow through the shunt resistor. As the frequency increases, the capacitive reactance decreases and more current is coupled into the RTD and antenna network. Consequently, the relative contribution of the shunt resistor to the overall admittance becomes smaller. And although the conductance appears close to zero over a wide frequency range, this is mainly due to the plotting scale. The actual conductance values are much smaller than those at low frequencies but still exhibit noticeable variations, particularly around the oscillation frequency near 600 GHz.

The imaginary part of the oscillator admittance is shown in Fig. 4.16. The observed frequency dependence can be understood from the extended equivalent circuit in Fig. 2.6. The reactive part of Y_{circ} includes the antenna inductance L_{ant} , the antenna-related parasitic capacitance C_{par} , and the MIM capacitor. Therefore, the reactive response is strongly influenced by the capacitive parasitic network. As the frequency increases, the inductive reactance becomes significant and the antenna inductance starts to dominate the susceptance of the passive oscillator structure. For an inductive branch, the susceptance is

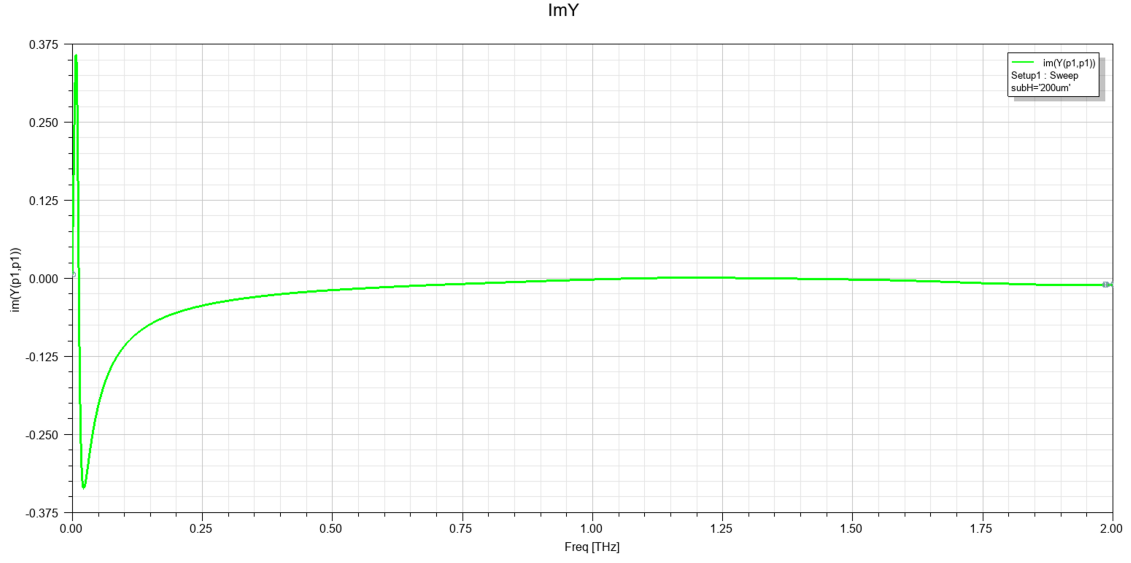


Figure 4.16: Wideband imaginary part of the oscillator admittance from 0.1 GHz to 2 THz.

$$B_L = -\frac{1}{\omega L}. \quad (4.6)$$

Therefore, the magnitude of the negative susceptance decreases as the frequency increases. This explains why $\text{Im}(Y_{\text{circ}})$ gradually approaches zero in the higher-frequency range. Around the intended oscillation frequency, the extracted susceptance remains close to the value required to compensate the RTD capacitance, while no additional strong susceptance crossing corresponding to another oscillation mode is observed within the investigated frequency range.

The wideband admittance results therefore indicate that the oscillator possesses only one physically meaningful oscillation region around 646 GHz.

To further verify the oscillation behavior, a transient simulation was performed for the complete oscillator structure. The resulting time-domain waveform and the corresponding FFT spectrum are shown in Fig. 4.17.

As shown in Fig. 4.17(a), the oscillation starts from a small perturbation and gradually grows until a steady-state amplitude is reached after approximately 200 ps. This confirms the successful start-up of the oscillator.

The steady-state waveform was then transformed into the frequency domain using a Fast Fourier Transform (FFT). The resulting spectrum is shown in Fig. 4.17(b). A dominant spectral component appears at approximately

$$f_{\text{osc}} \approx 646 \text{ GHz}, \quad (4.7)$$

which agrees well with the oscillation frequency predicted by both the S-parameter and harmonic balance analyses.

The peak power obtained from the FFT spectrum is approximately

$$P_{\text{out,max}} \approx -10.4 \text{ dBm} \approx 90 \text{ } \mu\text{W}. \quad (4.8)$$

This value is close to the output power obtained from the harmonic balance simulation, providing an independent time-domain verification of the oscillator performance.

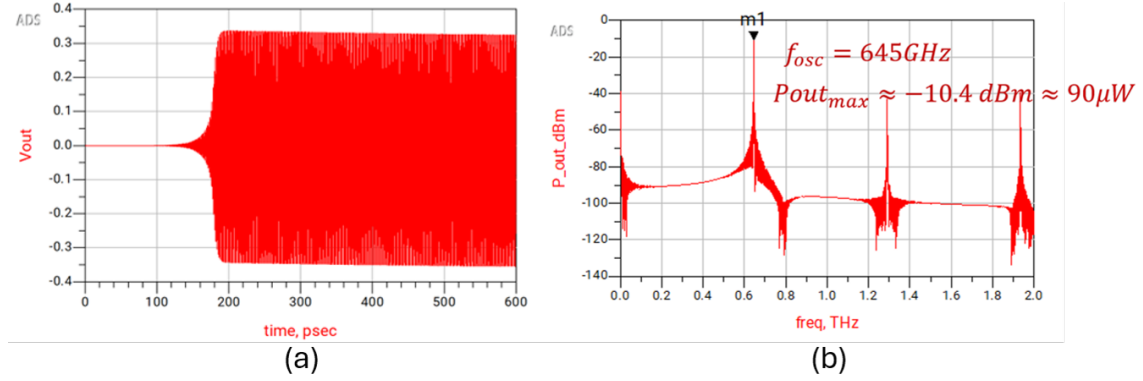


Figure 4.17: Transient response and FFT spectrum of the complete oscillator structure. (a) Time-domain oscillation waveform showing the start-up behavior and steady-state oscillation. (b) FFT spectrum of the steady-state waveform, indicating the oscillation frequency and output power.

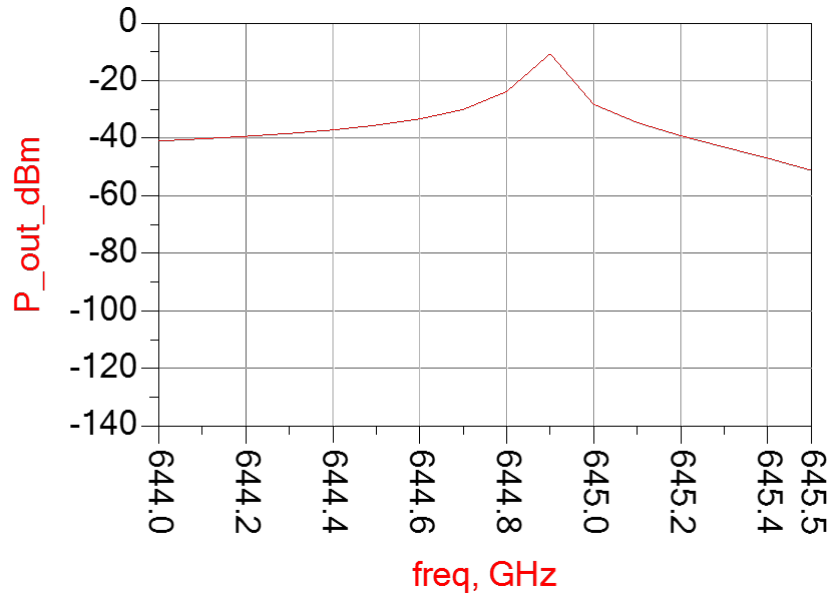


Figure 4.18: Zoomed frequency spectrum around the oscillation frequency obtained from the FFT analysis.

To further examine the spectral purity of the oscillation, the frequency spectrum around the oscillation peak is shown in Fig. 4.18. A single dominant peak is observed near 645 GHz, with no noticeable spectral broadening within the plotted frequency range. The linewidth is therefore estimated to be significantly smaller than 0.1 GHz, indicating a stable oscillation with good spectral purity.

4.2.2.5 Directivity Analysis

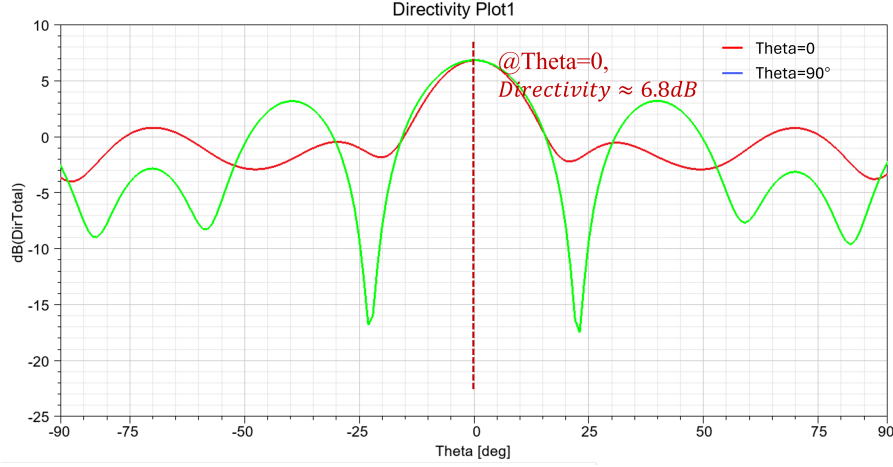


Figure 4.19: Directivity pattern of the complete oscillator antenna structure at the oscillation frequency.

To evaluate the radiation characteristics of the complete oscillator structure, the directivity pattern at the oscillation frequency was extracted from EM simulation and is shown in Fig. 4.19. To obtain reliable far-field results, the radiation boundary was placed sufficiently far from the antenna structure so that the near-field components decay before reaching the boundary. The airbox dimensions used in this work satisfy the far-field requirement at the operating frequency. The red and blue curves correspond to the radiation patterns for the two principal cuts, $\phi = 0^\circ$ and $\phi = 90^\circ$, respectively. A clear main lobe is observed around $\theta = 0^\circ$, where the peak directivity is approximately 6.8 dB. The result indicates that the slot antenna radiates preferentially in the broadside direction while maintaining a relatively symmetric radiation pattern around the main beam.

4.2.2.6 Comparison Between the Ideal and Realistic Models

To investigate the influence of substrate reflections on the extracted oscillator admittance, an additional HFSS simulation was performed using a more realistic substrate configuration. In this model, the InP substrate thickness was increased to $600 \mu\text{m}$, while an air region of $200 \mu\text{m}$ was retained above the structure. The extracted imaginary and real parts of the antenna admittance are shown in Fig. 4.20 and Fig. 4.21, respectively.

As shown in Fig. 4.20, the imaginary part of the admittance remains very similar to that obtained from the ideal model. The susceptance at the intended oscillation frequency changes only slightly, indicating that the reactive behavior of the oscillator is primarily determined by the local electromagnetic fields around the slot antenna and the RTD feed region.

According to the equivalent circuit shown in Fig. 2.6, the reactive part of the oscillator admittance mainly originates from the antenna inductance L_{ant} , the parasitic capacitance C_{par} , and the contact capacitance C_c . These elements are determined

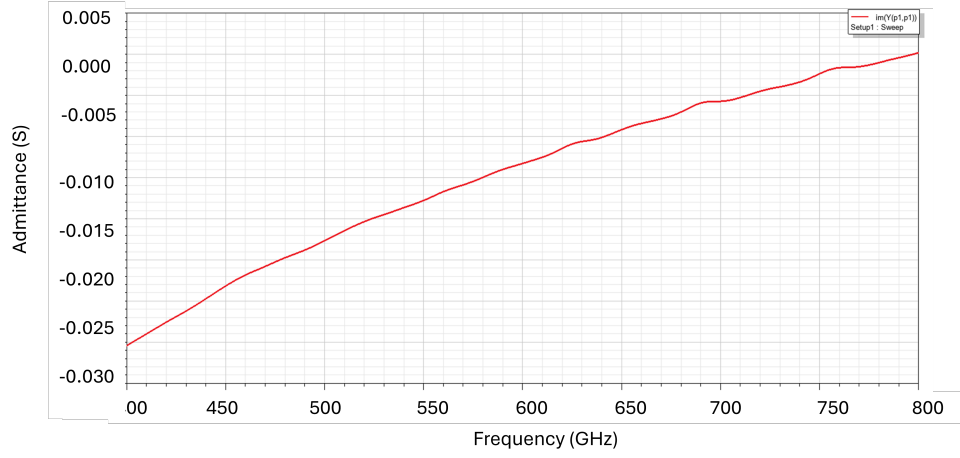


Figure 4.20: Imaginary part of the oscillator admittance extracted from the realistic model with a substrate thickness of $600\ \mu\text{m}$. The susceptance remains close to that obtained from the ideal model, indicating that the reactive behavior is mainly determined by the local slot antenna and RTD feed geometry.

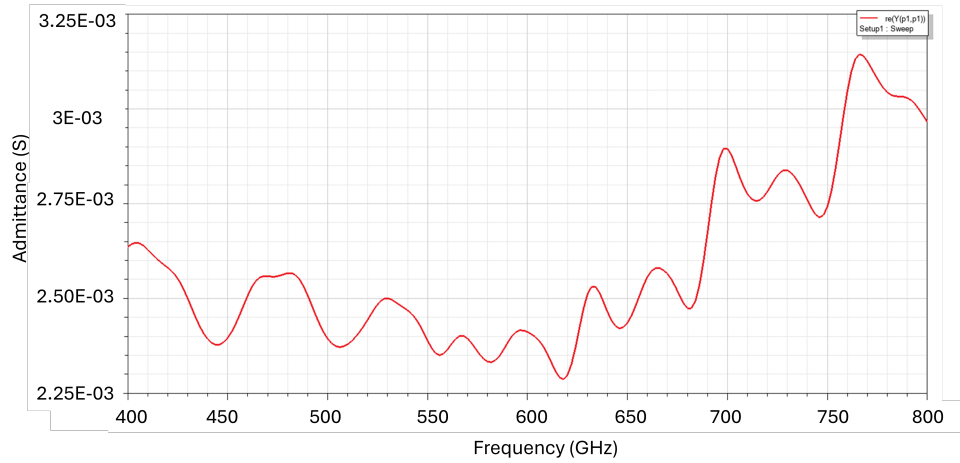


Figure 4.21: Real part of the oscillator admittance extracted from the realistic model with a substrate thickness of $600\ \mu\text{m}$. Compared with the ideal model, stronger frequency-dependent variations are observed due to substrate reflections affecting the effective radiation conductance.

by the slot dimensions, RTD feed geometry, and near-field electromagnetic distribution. Since these geometrical features remain unchanged between the two models, the extracted susceptance exhibits only minor variations.

In contrast, the real part of the admittance shows a significantly stronger dependence on the substrate configuration, as illustrated in Fig. 4.21. Compared with the ideal model, noticeable oscillations appear in the extracted conductance over the investigated frequency range.

From the equivalent circuit perspective, the real part of the oscillator admittance can be expressed as

$$G_{a,\text{complete}} = G_r + G_{\text{loss}} + G_p, \quad (4.9)$$

where G_r represents the effective radiation conductance, G_{loss} accounts for conductor and dielectric losses, and G_p denotes the additional parasitic conductance introduced by the contact structure.

Among these terms, G_r is the most sensitive to substrate reflections. When the substrate thickness is increased, electromagnetic waves radiated toward the backside of the substrate experience stronger reflections at the substrate-air interface and may propagate back toward the antenna. The reflected fields modify the effective radiated power and consequently alter the extracted radiation conductance. As a result, the conductance exhibits a stronger frequency dependence and larger fluctuations than in the ideal model. And a possible improvement is to integrate a silicon lens after the substrate[29].

Although the substrate reflections introduce noticeable variations in the extracted conductance, the oscillator is still capable of sustaining stable oscillation. A transient simulation of the realistic model predicts an output power of approximately

$$P_{\text{out}} \approx 80 \mu\text{W}, \quad (4.10)$$

which is only slightly lower than the value obtained for the ideal model. This indicates that the oscillator performance remains relatively robust even when substrate reflections are included.

5

Conclusion

In this thesis, a resonant tunneling diode (RTD) integrated slot antenna oscillator operating around 650 GHz was designed and analyzed. The objective was to develop a systematic design methodology for achieving stable terahertz oscillation while understanding how the slot antenna geometry and parasitic structures influence the oscillator performance.

A combined electromagnetic and circuit-level design approach was adopted throughout this work. The antenna geometry was first estimated using analytical methods and subsequently optimized using full-wave simulations in HFSS. The extracted admittance was then incorporated into ADS together with a nonlinear RTD model in order to evaluate the oscillation condition and predict the oscillator performance. Following the analysis of a simplified RTD-slot antenna structure, a complete oscillator model including the RTD contact geometry, MIM dielectric layer, air bridge, semiconductor contact region, and shunt resistor was developed to investigate the influence of parasitic effects.

The results demonstrate that the simplified analytical model provides a useful starting point for the antenna design, although full-wave optimization is necessary to accurately determine the admittance at frequencies near 650 GHz. For the simplified antenna-only structure, the extracted admittance satisfied the oscillation condition close to the target frequency and harmonic balance simulations predicted stable oscillation with output power on the order of several tens of microwatts.

After introducing the complete oscillator structure, the extracted oscillator admittance remained close to that obtained from the simplified model. This indicates that the local parasitic structures only slightly modify the oscillation condition and that the antenna geometry remains the dominant factor determining the operating frequency. A systematic sweep of slot width and slot length further showed that the maximum output power is obtained through a trade-off between oscillator conductance, radiation efficiency, and oscillation voltage amplitude rather than by optimizing a single parameter alone. Among the investigated geometries, the structure with a slot width of 6 μm and a slot length of 51 μm provided the highest estimated output power around 80 μW while maintaining oscillation near the target frequency.

Wideband admittance analysis and transient simulations were additionally performed to verify the oscillator behaviour. The transient response confirmed stable self-oscillation and showed excellent agreement with the oscillation frequency predicted by the harmonic balance simulations. The investigation of a more realistic substrate configuration further revealed that the imaginary part of the oscillator admittance is mainly determined by the local antenna and feed geometry, whereas the

real part is considerably more sensitive to substrate reflections and cavity effects. Overall, this work establishes a complete design methodology for RTD-integrated slot antenna oscillators and provides insight into the interaction between antenna geometry, parasitic elements, and oscillation behaviour at terahertz frequencies. The obtained results demonstrate the feasibility of compact RTD-based oscillators operating near 650 GHz through a combination of electromagnetic optimization and nonlinear circuit analysis.

Future work should focus on fabrication and experimental characterization of the proposed oscillator structure. More advanced RTD models incorporating voltage-dependent capacitance and large-signal nonlinear effects could further improve the prediction accuracy. In addition, substrate engineering, lens integration, and antenna-array implementations may be investigated as potential approaches for increasing the radiated output power and extending the operating frequency toward higher terahertz frequencies. Another possible direction is the optimization of the RTD active area. A larger RTD area can provide higher peak current and negative differential conductance, which may increase the available oscillator power, although this must be balanced against the accompanying increase in device capacitance and parasitic effects.

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