

Active Matching Circuits for Enhanced Antennas

EE3L1: Bachelor Graduation Project

Michaël Blaakman

Klaas de Kroon

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by

Michaël Blaakman
Klaas de Kroon

Instructor: Dr. D. Cavallo
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Abstract

Wide-band phased-array antennas are an essential technology for modern communication, radar, and radio astronomy systems because they enable operation across broad frequency ranges while maintaining performance over large scan angles. Connected array antennas provide inherent wide-band behavior, but the introduction of a backing reflector to ensure unidirectional radiation creates frequency-dependent reactive effects that limit the operational bandwidth. This project investigates the use of active matching circuits that create a negative inductance to compensate for the inductive behavior introduced by the reflector, thereby improving antenna impedance matching over a broad frequency range. A connected slot array antenna was modeled and analyzed using electromagnetic simulations. Equivalent circuit models were developed to represent the antenna and reflector system, after which three active matching topologies were studied, based on operational amplifiers, transistors, and tunneling diodes. The circuit parameters were optimized to minimize reflections and improve matching performance across the operating band and for different scanning angles. The results demonstrate that negative inductance compensation can significantly enhance the impedance bandwidth of reflector-backed connected arrays compared to the unmatched configuration. The transistor and tunneling diode circuits delivered satisfactory matching performance under broadside operation, with the tunneling diode showing slightly better behavior at higher scan angles. The operational amplifier delivered poor results, showing that its high frequency performance is restricted by its limited gain-bandwidth product. The study concludes that active non-Foster matching is a promising technique for extending the bandwidth of connected array antennas, although practical challenges related to stability, power consumption and biasing active components must be addressed before implementation in real-world wide band antenna systems.

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1

Introduction

Antenna systems are essential in wireless technologies, enabling the transmission and reception of signals for both communication and radar applications. Wide-band operation is a critical requirement of modern antenna systems to support multi-functionality, spectral agility, and high data throughput. Antennas capable of operating over broad frequency ranges enable the integration of multiple services, such as communication, sensing, and surveillance, within a single system, reducing complexity, volume, and cost. Furthermore, wide-band antennas enhance resolution, robustness to interference, and adaptability to dynamic electromagnetic environments, making them essential for next-generation wireless, radar, and space applications.

1.1. Wide-band Phased Array Antennas

A phased array is an antenna system made up of many individual radiating elements whose signals are carefully controlled in phase (and sometimes amplitude) so that the combined emission forms a directed beam. By adjusting the relative timing of the signals across the elements, the main beam of the antenna can be steered electronically without physically moving the structure. This allows rapid changes in direction, precise control over beam shape, and the ability to form multiple beams in more advanced systems. Phased arrays are a core technology in modern sensing and communication systems because they provide fast, flexible, and highly controllable radiation patterns.

Wide-band operation becomes important when a phased array must work effectively over a large range of frequencies rather than a narrow band. This capability is essential in modern environments where signals are not confined to a single frequency and where systems need to operate flexibly across multiple bands. Wide-band phased arrays are used in a variety of high-performance applications where adaptability and precision are critical. In radar systems, including modern active electronically scanned arrays, they enable high-resolution imaging, rapid target tracking, and improved resistance to interference across multiple operating modes. In wireless communications, such as advanced 5G and emerging 6G networks, they support high data rates and dynamic beam steering across different frequency bands to maintain strong, efficient connections with moving users. They are also widely used in satellite communications, remote sensing, and electronic warfare systems, where the ability to operate reliably across broad frequency ranges allows robust performance in complex electromagnetic environments.

1.2. State of the Art of Wideband Antenna Arrays

The literature has introduced a variety of wideband phased-array designs, such as tightly coupled arrays [1], the Planar Ultrawideband Modular Antenna array [2], [3], Vivaldi or flared-notch configurations [4], [5], [6], [7], and connected arrays [8], [9]. Each of these approaches involves different compromises across key performance parameters for wideband phased arrays, including bandwidth, beam steering range, radiation efficiency, complexity, and physical thickness. Among these architectures, connected arrays [8] offer a good tradeoff between the different figures of merit. In this project, this connected array antenna concept will be studied and used as the baseline reference design for the study.

Besides the antenna architecture itself, the matching network also plays an important role in achieving wideband behavior. The use of non-Foster elements to match antennas has been demonstrated before. For example, a negative impedance converter realized with GaAs FETs was used to almost double the impedance bandwidth of a microstrip patch antenna [10]. Also, a non-Foster metasurface was used to match a conductor-backed substrate over a frequency decade in an absorber design [11]. These results show that active negative-reactance elements can extend the matching bandwidth beyond the limits of passive networks.

1.3. Connected Arrays and Their Limitations

A connected array is a planar antenna architecture where the radiating elements are not discrete dipoles or slots, but instead are intentionally electrically interconnected, so the array behaves as a continuous radiating aperture rather than a collection of independent resonators. The antenna concept is described in Fig. 1.1: in arrays composed of resonant dipoles or slots, the current distribution on the dipoles or the electric field distribution across the slots is inherently sinusoidal and strongly frequency dependent. As a result, each element radiates efficiently only near its resonance, typically around the half-wavelength ($\lambda/2$) condition, which makes such arrays fundamentally narrowband.

Connected arrays address this limitation by electrically linking adjacent dipoles or slots, forming continuous elongated conductors or apertures that are periodically excited along their length. Instead of isolated resonant elements, the structure behaves as a set of long, connected dipoles or slots with distributed feeding points. Because the elements are no longer individually truncated and resonant, the current and field distributions are much less sensitive to frequency variations. This leads to a more stable behavior over a broad frequency range, resulting in inherently wideband performance.

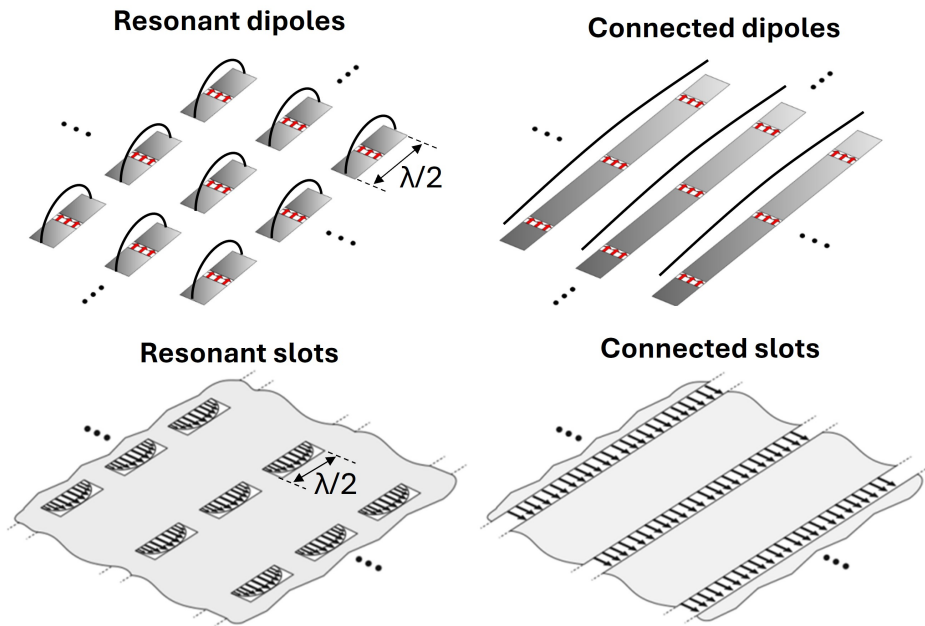


Figure 1.1: Array of resonant dipoles and slot antennas, compared with connected arrays of dipoles and slots

Although connected arrays exhibit frequency-independent behavior when radiating in free space, their efficiency is limited to approximately 50%, since the radiated power is equally split between upward and downward directions. In most practical applications, antennas are required to radiate predominantly in a single direction, making this bidirectional radiation undesirable. A common approach to overcome this efficiency loss is to introduce a backing reflector behind the array, as illustrated in Fig. 1.2. This configuration redirects the downward radiation toward the upper half-space, effectively improving the radiation efficiency by suppressing backward radiation. However, the presence of the reflector introduces a resonant effect associated with the spacing between the array and the reflector (h_d or h_s in the figure), which in turn limits the operational bandwidth.

The presence of the reflector introduces an inductive effect, particularly at lower frequencies, which degrades the impedance matching of the antenna. To compensate for this, a series tuning capacitor can be introduced in the antenna feed network to counteract the unwanted inductive reactance. However, this compensation is inherently frequency-dependent and therefore cannot be maintained over a wide bandwidth. One possible approach to achieve broadband compensation of the inductive effect introduced by the backing reflector is to employ a negative impedance converter whose effective inductance is equal in magnitude and opposite in sign to the reflector's equivalent inductance.

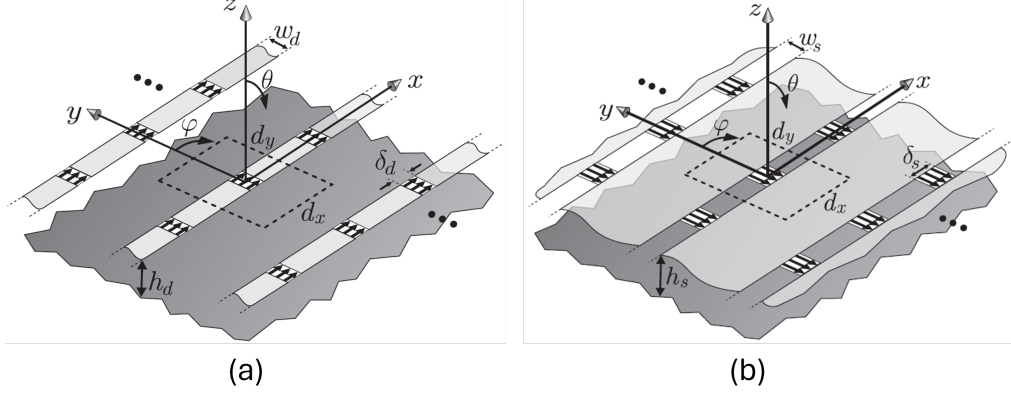


Figure 1.2: Connected arrays of (a) dipoles and (b) slots, in the presence of a metal backing reflector. Characteristic geometrical parameters and reference coordinate system are also defined.

1.4. Goal of the Project

The goal of this project is to investigate the use of negative impedance converters (NICs) for the wide band compensation of the reactive effect introduced by a backing reflector in connected antenna arrays. In particular, different implementations of negative inductance are explored, including circuits based on operational amplifiers (op-amps), MOSFET transistors, and tunneling diodes. The study initially considers idealized component models, which are combined with the connected array unit cell to evaluate the fundamental compensation principle and its impact on impedance behavior. Subsequently, more realistic equivalent circuit models are introduced in order to account for key non-idealities of the negative inductance implementations, enabling a more accurate assessment of their performance in practical wide band array applications.

1.5. Practical Limitations of Negative Impedance Converters

Negative inductance is not a physical passive component but can be realized in practice through active circuit techniques that synthesize an effective negative reactance. Practical implementations of negative inductance circuits are limited by several key factors. First, they are inherently active, meaning they require external power and introduce noise, which can degrade system sensitivity, especially in RF applications. Second, their performance is limited by parasitic effects. At RF and microwave frequencies, unintended parasitic capacitances and inductances from device packaging, interconnects, and biasing networks can significantly distort the intended negative impedance behavior. This effect has been studied for integrated NIC implementations, where the shunt and series parasitics of the transistors limit the frequency band over which the circuit linearly synthesizes the desired negative reactance, typically restricting it to a few GHz [12]. Third, they are prone to stability issues, as the same feedback that creates negative reactance can also lead to unwanted oscillations if not carefully controlled. Finally, these circuits often suffer from nonlinearity and dynamic range limitations, since active devices can saturate or behave nonlinearly at higher signal levels. In this work, several key non-idealities will be considered, with particular emphasis on the finite gain and bandwidth of the amplifier, as well as its associated parasitic effects.

1.6. About This Report

This report was written with the help of the supervisor Dr. D. Cavallo. For this thesis the following generative AI was used; Claude and Chat-GPT. They were used for creating LaTeX tables and syntax, generating images, and performing spell checks. For the parameter optimization code, GenAI was consulted on how to load data into Python.

The remainder of this report is organized as follows. First, the requirements, objectives, and success measures of the project are defined. Next, the methods used throughout the project are described. The results are then presented and discussed. Finally, conclusions are drawn and recommendations for future work are provided.

2

Program of Requirements

Now that the project has been introduced and the problem is defined, it is time to take a look at the objectives and requirements that need to be reached to complete this project. To establish these, first a look is taken at who the stakeholders of the project are, and what their needs are. Based on these needs, the objectives of the project are defined, followed by the success measures used to evaluate whether the objectives have been achieved. Finally, a complete overview of all the functions and requirements of our antenna system is provided.

2.1. Stakeholders

This project has the following stakeholders.

- Daniele Cavallo: The supervisor of the project. Intends to use our design for his work on connected array antennas.
- Research institutes and universities: Research groups working on antennas and electromagnetics are interested in non-Foster matching techniques for future wireless systems, connected arrays and ultra-wideband antennas.
- Customers
 - Communication: Companies developing wireless and satellite communication systems, such as Nokia, Ericsson, SES, Viasat and Starlink, require broadband antenna systems for 5G, 6G and satellite communication networks.
 - Defence: Defence companies and government organizations, such as Thales, Lockheed Martin, Raytheon and Northrop Grumman, use broadband phased-array antennas in radar, surveillance and UAV systems.
 - Radio Astronomy: Research organizations such as ASTRON, the Square Kilometre Array Observatory (SKAO) and the National Radio Astronomy Observatory (NRAO) require wideband antennas with low noise and high sensitivity to observe weak astronomical signals over large frequency ranges.

2.2. Stakeholder Needs

The stakeholders have different needs and expectations for the project. Communication companies require broadband antennas with compact size and low implementation complexity so that they can be used in wireless communication systems such as 5G and satellite communication. Defence applications require reliable wideband antennas that can operate in dynamically changing electromagnetic environments and that can support radar and surveillance systems. Radio astronomy applications require antenna systems with very low noise contributions and high sensitivity in order to detect extremely weak signals over large frequency ranges.

Research institutes and universities require accurate simulation models and realistic active matching circuits to further investigate non-Foster matching techniques and connected array antennas. The supervisor is interested in evaluating whether non-Foster matching circuits can practically improve the bandwidth of connected array antennas while remaining physically realizable.

2.3. Project Objectives

To address these stakeholder needs, the project aims to:

- Investigate the use of active non-Foster matching circuits to enhance antenna bandwidth.
- Study how negative reactance elements such as negative inductances and capacitances can compensate for the frequency-dependent reactance of antennas.
- Simulate antennas and matching circuits with the commercial electromagnetic solver CST to analyze impedance behaviour and bandwidth improvement.
- Investigate additional input resistance of the active matching circuit.
- Design a matching circuit that remains practically implementable using realistic high-frequency components.

2.4. Success Measures

To determine the success of the project the following criteria has to be successfully completed:

- A demonstration of improved antenna bandwidth compared to the unmatched antenna.
- Achieving an active reflection coefficient (ARC) of: $\Gamma_{\text{active,dB}} < -10$ dB over the target frequency range.
- Successful CST simulations of the antenna and matching circuit.
- Stable synthesized negative reactance over the simulated frequency range.
- Physically realizable component values and compact circuit implementation.
- Delivery of a literature study, simulation results, final report and final presentation.

2.5. List of Functions and Requirements

The final product should provide the following functions, along with their corresponding requirements:

- Simulate the antenna in CST
 - The CST model should include the radiating element, the connections between the slots and the back reflector
 - The simulation should operate over the 0.1 to 1 GHz and 1 to 10 GHz frequency range.
- Determine the antenna input impedance
 - The antenna impedance has to be extracted over both of the simulated frequency ranges.
- Design a non-Foster matching circuit
 - The circuit should synthesize a negative reactance.
 - The circuit should use physically realizable component values.
 - The circuit should use components capable of operating at GHz frequencies.
 - The matching network should improve the impedance bandwidth of the antenna.
- Analyze ARC and bandwidth
 - The broadside ($\theta = 0^\circ$) ARC should satisfy:

$$\Gamma_{\text{active,dB}} < -10 \text{ dB}$$

across the target bandwidth, where θ is the scan elevation angle.

- The bandwidth improvement should be quantified using CST simulations.

- Evaluate antenna performance for different scan angles
 - The antenna should be simulated at scan angles of

$$\phi = 0^\circ, \theta = 20^\circ, 40^\circ, 60^\circ.$$

$$\phi = 90^\circ, \theta = 20^\circ, 40^\circ, 60^\circ$$
 where θ is the elevation angle and ϕ the azimuth angle.
 - The ARC should satisfy

$$\Gamma_{\text{active,dB}} < -6 \text{ dB}$$
 for the maximum scan angle $\theta = 60^\circ$.
 - The effect of scan angle on the matching network performance should be evaluated.
- Evaluate circuit stability
 - The synthesized negative reactance should remain stable over the simulated frequency range.
- Ensure practical circuit realizability
 - The circuit should use commercially available components.
 - The circuit should use realistic inductance and capacitance values.
- Optimize the matching network parameters
 - The matching network parameters should be optimized to maximize bandwidth.
 - The matching network should minimize reflection losses.
- Visualize and document simulation results
 - The project should include plots of impedance, Γ_{active} and bandwidth.
 - The simulation results should be clearly documented and discussed.

3

Methods

In this section, the methodology of the project is discussed. First, the simulation method to estimate the active input impedance of the connected array antenna is described. Next, the non-Foster matching circuits are investigated and their parameters are chosen. Finally, the simulation procedure is discussed.

3.1. Connected Slot Unit Cell Simulation

The antenna considered in this project is a connected slot array antenna. This structure consists of periodically interconnected slot elements that enable continuous current flow between adjacent radiating apertures. As a result, the array supports a uniform current distribution and can operate over a wide frequency band [9]. However, as discussed in Chapter 1, the introduction of a backing reflector to enforce unidirectional radiation represents the primary limitation on the achievable operational bandwidth in practical implementations.

A schematic of the antenna element with a backing reflector is shown in Fig. 3.1. The structure is simulated using CST Studio Suite [13], a full-wave numerical electromagnetic solver. Periodic boundary conditions are applied to the unit cell, which effectively assumes an infinite periodic repetition of the structure in both the x and y directions.

The slot is excited using a discrete port (blue rectangle in Fig. 3.1). This is a commonly used idealized feeding model in full-wave electromagnetic simulations and consists of a constant voltage (or current) source over a rectangular region, effectively enforcing a specified voltage difference between two edges.

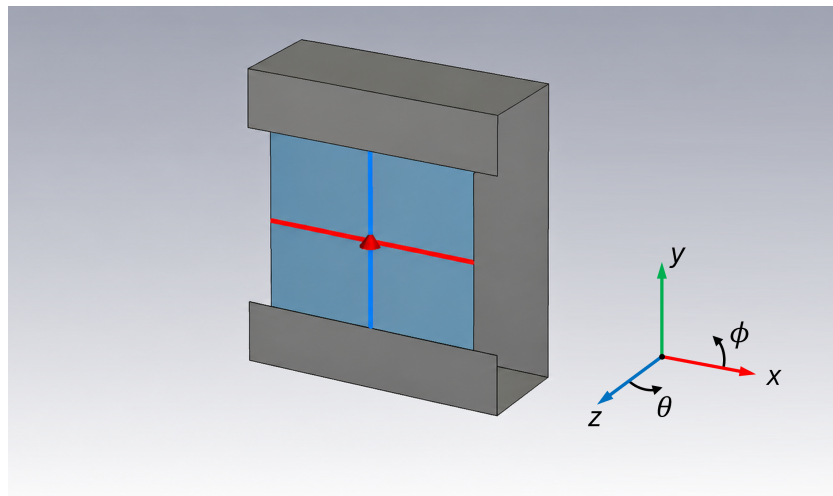


Figure 3.1: Schematic of the connected slot array unit cell with backing reflector.

In the model, both the radiating elements and the backing reflector are represented as infinitely thin perfect electric conductor (PEC) surfaces. When the scan angle is varied, electromagnetic energy can become partially trapped in the cavity formed between the slot layer and the backing reflector, leading to a reduction in radiation efficiency. To prevent this effect, the antenna layer and the backing reflector are electrically connected at two sides of the unit cell using additional PEC walls, ensuring proper field termination and improved power radiation into free space.

The dimensions of the antenna are dependent on the center wavelength of the operating frequency band. For this project we look at frequency ranges of 0.1-1 GHz and 1-10 GHz, the dimensions of the antenna are given in table 3.1.

Table 3.1: Antenna Dimensions for Different Frequency Ranges. Geometrical Parameters are defined in Figure 1.2.

Parameter	0.1–1 GHz	1–10 GHz
Period along x (d_x)	135 mm	13.5 mm
Period along y (d_y)	135 mm	13.5 mm
Antenna height (h_s)	50 mm	5 mm
Slot width (w_s)	80 mm	8 mm
Feed length (δ_s)	110 mm	11 mm

3.2. Equivalent Circuit of Connected Slot with Reflector

To better understand the strategy adopted in this study, it is convenient to introduce a simplified equivalent circuit of the connected slot unit cell. This is shown in Figure 3.2a.

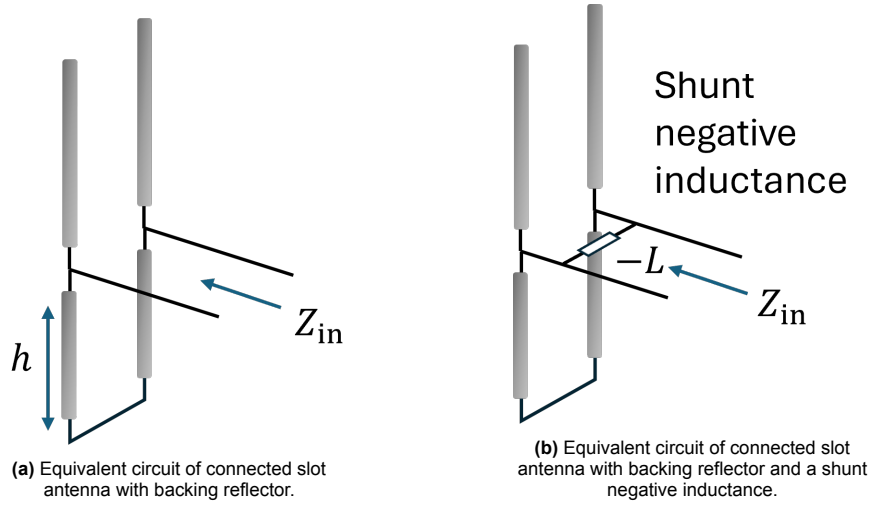


Figure 3.2: Approximate first-order equivalent circuits to represent the active input impedance of connected slot arrays.

The radiation into free space is represented as an infinite long transmission line, while the backing reflector can be modeled as a short-circuited transmission-line stub. The impedance equivalent of the backing reflector is given by

$$Z_{br} = jZ_{fs} \tan(kh) \quad (3.1)$$

where Z_{fs} is the free-space characteristic impedance, k is the propagation constant and h is the distance between the antenna slot plane and the reflector [14]. For electrically short distances, $kh \ll 1$, the tangent function can be approximated as linear, which yields

$$Z_{br} \approx jZ_{fs}kh = j\omega \frac{Z_{fs}h}{c} \quad (3.2)$$

where ω is the angular frequency and c the speed of light. This means the reflector approximately acts as an inductive reactance, where

$$L = \frac{Z_{fs}h}{c} \quad (3.3)$$

To cancel out the reflector's impedance, a negative equivalent inductance needs to be obtained and placed in shunt with the backing reflector, as seen in Figure 3.2b. This will be achieved using a negative impedance converter (NIC). Substituting the characteristic impedance of free-space (377Ω), the distance between the antenna and the reflector plates (5 mm and 50 mm for the 1-10 GHz and 0.1-1 GHz bands, respectively) and the speed of light ($c \approx 3 \cdot 10^8 \text{m/s}$) the inductances become

$$L_{eq,1} = 6.3\text{nH}, \quad L_{eq,2} = 0.63\text{nH}$$

where $L_{eq,1}$ assumes that the antenna operates in the 0.1-1 GHz range and $L_{eq,2}$ in the 1-10 GHz range.

3.3. Parameter Optimization

All three negative impedance converters (NICs) have extensive equivalent impedance formulas, because of the added parasitic component introduced in their equivalent model. To obtain the best result, approximations do no longer suffice. With the extracted impedance results of the connected slot antenna with backing reflector, it is possible to calculate the ARC Γ_{active} by the following formula;

$$\Gamma_{\text{active}}(f) = \frac{Z_{\text{model}}(f) - Z_{\text{ref}}}{Z_{\text{model}}(f) + Z_{\text{ref}}} \quad (3.4)$$

where $Z_{\text{model}}(f)$ is the input impedance of the total circuit, which depends on the frequency f , and Z_{ref} is the reference normalization impedance [14]. Using a Python script, design variables that can be controlled were swept such that, for both $\phi = 0^\circ$ and $\phi = 90^\circ$, the ARC satisfies $\Gamma_{\text{active,dB}} \leq -10, \text{dB}$ for $\theta = 0^\circ$ and $\Gamma_{\text{active,dB}} \leq -6, \text{dB}$ for $\theta = 60^\circ$. A full overview of the configuration used by the code is given in Figure 3.3. The NICs are modeled using equivalent circuit representations, whose parameters are selected based on realistic values for components realizable at the target frequency ranges.

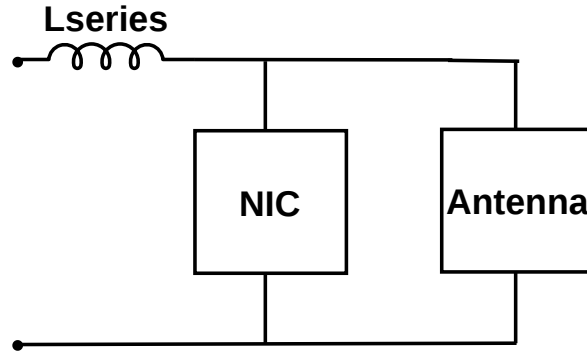


Figure 3.3: Schematic of the configuration of the inductor, NIC and antenna.

As can be observed in the figure, an inductor is placed in series with a parallel combination of the NIC and the antenna with backing reflector. This inductance compensates for the earlier small-argument approximation of the reflector impedance, used in Equation 3.2. While this approximation is accurate at lower frequencies, the deviation between $\tan(kh)$ and its linear approximation increases as frequency increases [14]. The added series inductance partially compensates for this deviation, resulting in a better impedance match over the upper portion of the operating bandwidth. The code calculates $\Gamma_{\text{active}}(f)$ for every swept parameter. All results will be penalized if they do not meet the requirements. For results that do meet the requirements, those with the largest margin below the threshold will be favored.

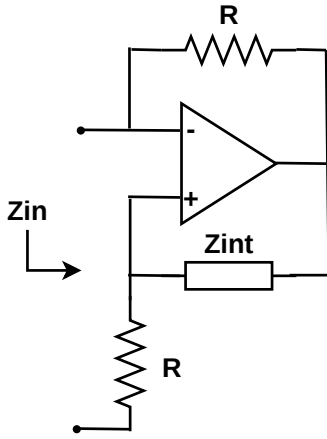
The code can be found in Appendix A. The parameter sweep was performed differently for each NIC circuit; this is further explained in section 3.4.

3.4. Negative Impedance Converter (NICs)

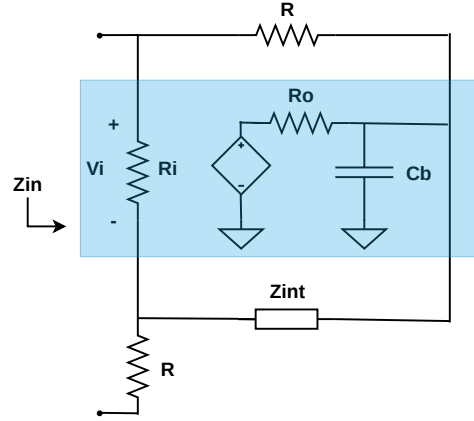
A negative impedance converter (NIC) is used to cancel the inductance of the circuit. For this project, three different NICs are studied: one using an op-amp, one using an NMOS transistor, and one using a tunneling diode. These different circuits offer different trade-offs in bandwidth and performance.

3.4.1. Op-Amp Circuit

The first NIC, which will be studied, is an op-amp circuit [15]. The ideal circuit is shown in figure 3.4a. For an accurate study of this circuit an equivalent model of the op-amp is used, given in [16]. However, op-amps are not able to operate at high frequencies due to the limited gain-bandwidth product; to account for this, an equivalent circuit of the op-amp is used with a capacitor to mimic a low-pass filter at the output. The drop-off will be identical to the behavior of the closed loop op-amp gain at high frequencies. Despite the op-amp's limited bandwidth, it was chosen for its simple topology, making it suitable as a proof of concept.



(a) The NIC using the op-amp, here R and Z_{int} are the chosen to match the antenna impedance



(b) Equivalent circuit of the op-amp NIC. Here R_o is the output resistance, R_i the input resistance, C_b output capacitor and V_i the voltage over the op-inputs.

Figure 3.4: Negative inductance matching network and its equivalent circuit.

The equivalent input impedance of this circuit can be written as

$$Z_{in} = R_1 \frac{A_v R_1 R_i Z_{int} - R_1 R_i R_o + R_1 R_i Z_{int} + R_1 R_o Z_{int} + 2R_1 Z_{int} + R_i R_o Z_{int} + 2R_i Z_{int}}{-A_v R_i Z_{int}^2 - R_1^2 R_o + R_1^2 Z_{int} - R_1 R_i R_o + R_1 R_i Z_{int} + 2R_1 Z_{int}^2 + R_i R_o Z_{int} + 2R_i Z_{int}} \quad (3.5)$$

However, making the assumption that $A_v \gg 1$, this simplifies to

$$Z_{in} \approx -\frac{R^2}{Z_{int}} \quad (3.6)$$

Then Z_{int} is defined as a capacitor

$$Z_{int} = \frac{1}{j\omega C} \quad (3.7)$$

This makes the total input impedance equal to

$$Z_{in} \approx -j\omega C R^2 \quad (3.8)$$

where C and R are chosen based on the reflector impedance derived in Section 3.1.

Determining Parameters

For the equivalent op-amp model, the parameters are the open-loop gain A_{v0} , gain-bandwidth product (GBP), the bandwidth limiting capacitance C_b , input resistance R_i , and the output resistance R_o . Except for the values of R_i , R_o , and C_b , all parameters could be read directly from the datasheet [17]. These values are listed in Table 3.2. The input resistance R_i is directly stated in the OPA847 datasheet as a differential input impedance of $2.7\text{ k}\Omega \parallel 2.0\text{ pF}$ [17]. The resistive part gives $R_i = 2.7\text{ k}\Omega$. The parallel capacitance of 2.0 pF is not included in the equivalent circuit model. At the lower band edge ($f = 0.1\text{ GHz}$) its impedance is approximately $800\text{ }\Omega$, which is non-negligible relative to R_i , and at $f = 1\text{ GHz}$ it drops to $80\text{ }\Omega$, where it dominates the input impedance. Neglecting this capacitance therefore introduces a modeling error that increases with frequency and represents a known limitation of the equivalent circuit at the upper end of the band. The output resistance R_o is not directly stated in the OPA847 datasheet. Only the closed-loop output impedance is given: $R_{o,CL} = 0.003\text{ }\Omega$ at $G = +20$ and $f < 100\text{ kHz}$ [17]. The open-loop output resistance is recovered using the standard feedback relation [18]:

$$R_{o,CL} = \frac{R_o}{1 + A_{v0}\beta}, \quad (3.9)$$

where $\beta = 1/G = 1/20$ is the feedback factor at the specified gain. Rearranging and substituting $A_{v0} = 10^{98/20} = 79\,433$:

$$R_o = R_{o,CL} \cdot (1 + A_{v0}\beta) = 0.003 \times \left(1 + 79\,433 \times \frac{1}{20}\right) \approx 12\text{ }\Omega. \quad (3.10)$$

In a real op-amp, the open-loop gain rolls off with frequency according to the single-pole model [18]:

$$A_v(f) = \frac{A_{v0}}{1 + j A_{v0} f / \text{GBP}}, \quad (3.11)$$

where A_{v0} is the DC open-loop gain and GBW is the gain-bandwidth product. At frequencies well above the dominant pole, the magnitude of the open-loop gain reduces to [18]:

$$|A_v(f)| \approx \frac{\text{GBP}}{f}. \quad (3.12)$$

The condition $A_v \gg 1$ therefore degrades as the operating frequency approaches the GBW. The minimum GBW required for the approximation to remain valid at the upper band edge $f_{\max}$ is thus:

$$\text{GBP} \gg f_{\max}. \quad (3.13)$$

The exact minimum GBW at which NIC performance becomes unacceptable is determined in simulation by sweeping the GBW while evaluating the deviation of Z_{in} . In the equivalent circuit this is implemented by varying C_b , since:

$$C_b = \frac{A_{v0}}{2\pi R_o \cdot \text{GBP}}, \quad (3.14)$$

with A_{v0} and R_o held fixed at the values in Table 3.2. The results of this sweep are presented below.

Parameter Optimization

The parameters chosen above cannot be changed due to their values being physically determined by the component. However, the $Z_{\text{int}} = \frac{1}{j\omega C}$ and R mentioned in Equation 3.5 can be altered. Using the parameter optimization configuration mentioned in Figure 3.3, the L_{series} is also swept. For the both frequency ranges, the result of the optimization code and the other variables are given in Table 3.2.

Table 3.2: Fitted Circuit Parameters for Different Frequency Ranges.

Parameter	0.1–1 GHz	1–10 GHz
R (Ω)	100	100
A_{v0}	7.94×10^4	7.94×10^4
R_i (Ω)	2.7×10^3	2.7×10^3
R_o (Ω)	12	12
GBP (Hz)	3.9×10^9 *	3.9×10^9 *
C_p (F)	6.8979×10^{-11}	6.5918×10^{-12}
L_{series} (H)	0.1×10^{-9} **	0.1×10^{-9} **
Z_0 (Ω)	200***	90***

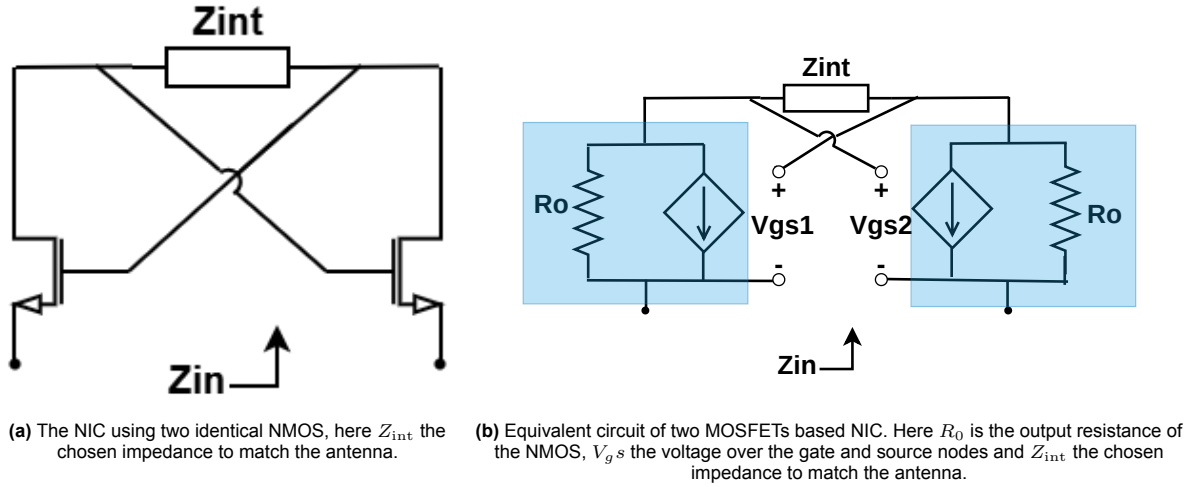
*However, as mentioned above, for high frequency ranges there exists no op-amp (yet) that is able to provide a useful gain over these frequency ranges. This effect can be seen in the results section 4.3.

** For significantly lower L_{series} , the code produced slightly better results. However, an assumption for a lower bound of 0.1nH for an inductor was used.

*** The parameter optimization script gave for both frequency ranges an optimized value of $Z_0 = 35\Omega$. However, the code looks at minimizing the total distance from the thresholds. While working with the current normalization frequencies, makes for a limited frequency bands still being able to be meet the requirements.

3.4.2. NMOS Circuit

The next NIC is made using NMOS transistors [19]. A circuit with an NMOS transistor was chosen for its high-frequency capability and compatibility with integrated circuit technologies but due to it consisting of a latch, it is prone to instability [15]. The circuit and its equivalent circuit are shown in figure 3.5b, in the equivalent circuit, the gate capacitances are neglected.

**Figure 3.5:** Negative inductance matching network and its equivalent circuit using two identical MOSFETs.

This circuit has an equivalent input impedance of

$$Z_{in} = \frac{r_o (-Z_{int} r_o g_m^2 + 2g_m r_o + 1)}{(g_m r_o)^2 + 2g_m r_o + 1} \quad (3.15)$$

where Z_{int} is the impedance of the chosen element, r_o is the output resistance of the transistor and g_m the transconductance of the transistors. To simplify this, factor the denominator to get

$$(g_m r_o)^2 + 2g_m r_o + 1 = (g_m r_o + 1)^2$$

Substituting into Equation 3.15 gives

$$Z_{\text{in}} = \frac{-r_o^2 g_m^2 Z_{\text{int}} + 2g_m r_o^2 + r_o}{(g_m r_o + 1)^2}$$

Now, assuming $g_m r_o \gg 1$, meaning that the transistor intrinsic gain is large, this expression becomes

$$(g_m r_o + 1)^2 \approx (g_m r_o)^2$$

This gives

$$Z_{\text{in}} \approx \frac{-r_o^2 g_m^2 Z_{\text{int}} + 2g_m r_o^2 + r_o}{(g_m r_o)^2}$$

Separate the terms and simplify to obtain

$$Z_{\text{in}} \approx -Z_{\text{int}} + \frac{2}{g_m} + \frac{1}{g_m^2 r_o} \quad (3.16)$$

Now, applying the assumption ($g_m r_o \gg 1$), gives

$$\frac{1}{g_m^2 r_o} \ll \frac{2}{g_m}$$

so the last term may be neglected, and a final impedance is obtained as

$$Z_{\text{in}} \approx -Z_{\text{int}} + \frac{2}{g_m} \quad (3.17)$$

Here Z_{int} is chosen to be

$$Z_{\text{int}} = \frac{2}{g_m} + j\omega L \quad (3.18)$$

such that the resulting input impedance becomes

$$Z_{\text{in}} \approx -j\omega L \quad (3.19)$$

where L is selected to match the inductance calculated in section 3.1.

Determining Parameters

The two parameters of the NMOS equivalent circuit are the transconductance g_m and the output resistance r_o . The fundamental constraint for component selection is that the transistor must remain an active, low-noise amplifying device up to the highest frequency of interest. This is characterized by the transistor cutoff frequency f_T , defined as the frequency at which the small-signal current gain falls to unity. To ensure that the transistor provides adequate current gain over the target band, the cutoff frequency f_T must be well above the operating frequency. Following [18], the transistor gain is approximately 0 dB at $f_T \geq 3 \cdot f_{\text{max}}$. This gives the frequency constraint on device selection:

$$f_T \geq 3 \cdot f_{\text{max}} \quad (3.20)$$

A commercially available component is the NE3508M04 [20]. The associated datasheet does not specify f_T directly. Instead, the operating frequency range of the device is used as a direct indicator: the NE3508M04 is characterized at 2 GHz with 14 dB of associated gain, which confirms that the device provides substantial gain well into the lower portion of the 1–10 GHz band. For a GaAs HJ-FET of this gate length (0.25 μm), f_T is established in the range 60–80 GHz from parameter extraction studies on this device class [21], satisfying $f_T \geq 30$ GHz with significant margin for both the 1–10 GHz and 0.1–1 GHz band. Stated in its datasheet [20], it has a typical transconductance value of $g_m = 100\text{mS}$. However, the output resistance r_o is not directly stated. Using the same parameter extraction studies on this device class [21]: a 0.25 μm GaAs pHEMT, the intrinsic gain $g_m r_o$ is typically in the range 15–40 [21]. Using $g_m = 100\text{mS}$ and a conservative $g_m r_o = 20$:

$$r_o = \frac{g_m r_o}{g_m} = \frac{20}{100\text{mS}} = 200\ \Omega, \quad (3.21)$$

Parameter Optimization

The only parameter that can be changed in this NIC is the Z_{int} . However, as can already be observed from the ideal expression for Z_{in} in Equation 3.18, the NIC inherently adds a real part to the circuit. To be able to get the best optimization, the following was swept; $Z_{\text{int}} = R + j\omega L$. The result can be seen Table 3.3.

Table 3.3: Fitted circuit parameters for different frequency ranges for NMOS circuit

Parameter	0.1–1 GHz	1–10 GHz
r_o (Ω)	200	200
L_p (H)	6.611×10^{-8}	7.4141×10^{-9}
g_m (S)	0.1	0.1
R (Ω)	13.158	12.947
L_{series} (H)	8.421×10^{-9}	5.631×10^{-10}
Z_0 (Ω)	260	280

3.4.3. Tunneling Diode Circuit

The final circuit uses a tunneling diode, this consists of a heavily doped p–n junction diode with a very thin depletion region that allows electrons to tunnel through the barrier, enabling it to exhibit a negative resistance. By combining a negative resistance with an inductance, a negative impedance is obtained. The tunneling diode was chosen for its potential to operate at high frequencies [22]. The schematic of the NIC circuit is given in the figure below.

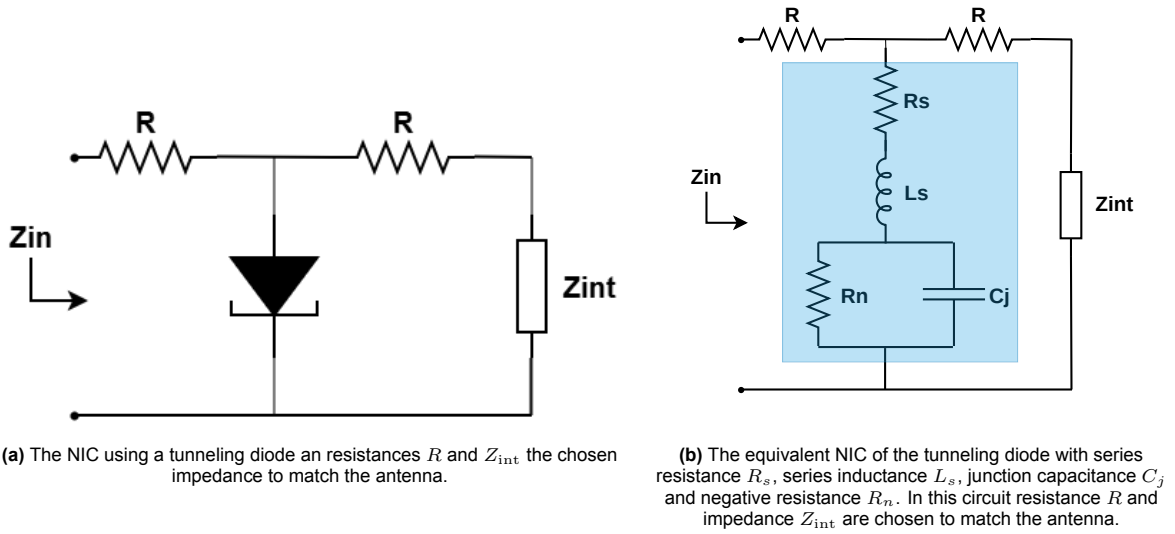


Figure 3.6: Negative inductance matching network and its equivalent circuit using tunneling diodes.

This circuit has an input impedance of

$$Z_{\text{in}} = R + \left[\frac{1}{R_s + sL_s + \frac{R_n \left(\frac{1}{sC_j} \right)}{R_n + \frac{1}{sC_j}}} + \frac{1}{R + \frac{1}{sC_p}} \right]^{-1} \quad (3.22)$$

In an ideal case [15], the equivalent input impedance of the circuit can be assumed as

$$Z_{\text{in}} \approx -\frac{R^2}{Z_{\text{int}}} \quad (3.23)$$

where Z_{int} was chosen to be

$$Z_{\text{int}} = j\omega \frac{L}{R^2} \quad (3.24)$$

making the equivalent input impedance

$$Z_{\text{in}} \approx -j\omega L \quad (3.25)$$

where again L was chosen to be the inductance calculated in section 3.1.

Determining Parameters

The four parameters of the tunneling diode equivalent circuit are the series resistance R_s , series inductance L_s , junction capacitance C_j , and negative resistance R_n . Their values are constrained by two frequency conditions derived from the equivalent circuit [23]. The resistive cutoff frequency f_r , at which the real part of the diode impedance passes through zero, is given by

$$f_r = \frac{1}{2\pi|R_n|C_j} \sqrt{\frac{|R_n|}{R_s} - 1}, \quad (3.26)$$

and the reactive cutoff frequency f_x , at which the imaginary part passes through zero, is given by

$$f_x = \frac{1}{2\pi} \sqrt{\frac{1}{L_s C_j} - \frac{1}{(|R_n|C_j)^2}}. \quad (3.27)$$

For useful NDR operation, both f_r and f_x must exceed the upper band edge f_{max} [23].

The negative resistance is determined by the slope of the I-V characteristic in the NDR region and is determined by the peak-to-valley current ratio and junction doping of the specific device [23]. A measured value of $|R_n| = 120 \Omega$ and $R_s = 3.5 \Omega$ are taken from a GaAs resonant tunneling diode characterized at microwave frequencies [24], and is used throughout the following derivations. With R_n and R_s fixed, rearranging (3.26) gives the maximum permissible C_j as a function of f_{max} :

$$C_{j,\text{max}} \leq \frac{1}{2\pi f_{\text{max}}|R_n|} \sqrt{\frac{|R_n|}{R_s} - 1}. \quad (3.28)$$

For the 0.1–1 GHz band ($f_{\text{max}} = 1 \text{ GHz}$):

$$C_{j,\text{max}} \leq \frac{1}{2\pi \times 10^9 \times 120} \sqrt{\frac{120}{3.5} - 1} \approx 7.5 \text{ pF}. \quad (3.29)$$

For the 1–10 GHz band ($f_{\text{max}} = 10 \text{ GHz}$):

$$C_{j,\text{max}} \leq \frac{1}{2\pi \times 10^{10} \times 120} \sqrt{\frac{120}{3.5} - 1} \approx 0.75 \text{ pF}. \quad (3.30)$$

The measured device of [24] has $C_j = 0.7 \text{ pF}$, which satisfies both bounds.

Rearranging (3.27) and requiring $f_x \geq f_{\text{max}}$ gives the maximum permissible L_s :

$$L_s \leq \frac{1}{C_j \left[(2\pi f_{\text{max}})^2 + \frac{1}{(|R_n|C_j)^2} \right]}. \quad (3.31)$$

For the 0.1–1 GHz band with $C_j = 0.7 \text{ pF}$:

$$L_s \leq \frac{1}{0.7 \times 10^{-12} \left[(2\pi \times 10^9)^2 + \frac{1}{(120 \times 0.7 \times 10^{-12})^2} \right]} \approx 51 \text{ nH}. \quad (3.32)$$

For the 1–10 GHz band:

$$L_s \leq \frac{1}{0.7 \times 10^{-12} \left[(2\pi \times 10^{10})^2 + \frac{1}{(120 \times 0.7 \times 10^{-12})^2} \right]} \approx 0.35 \text{ nH}. \quad (3.33)$$

The measured device of [24] reports $L_s = 3.5 \text{ nH}$ for a packaged device, consistent with the 0.1–1 GHz bound. A lower series inductance could be achieved by removing the package and using the raw semiconductor chip. There is no measurement done without the packaging in [24], so the series inductance for the 1–10GHz band is assumed to be $L_s = 0.1 \text{ nH}$.

Parameter Optimization

For the optimization of the tunneling diode circuit, only the $Z_{\text{int}} = \frac{1}{j\omega L}$ and the R can be swept. This results in the values stated in Table 3.4.

Table 3.4: Fitted circuit parameters for different frequency ranges

Parameter	0.1–1 GHz	1–10 GHz
R_s (Ω)	3.5	3.5
L_s (H)	1.0×10^{-10}	1.0×10^{-10}
R_n (Ω)	-120	-120
C_j (F)	7.0×10^{-13}	7.0×10^{-13}
C_p (F)	3.5789×10^{-12}	0.1×10^{-12} *
R (Ω)	123.68	100
L_{series} (H)	1.205×10^{-8}	0.1×10^{-9} *
Z_0 (Ω)	130	95

* Similarly for the op-amp NIC the code produced slightly better results for significantly lower L_{series} . However, an assumption for a lower bound of 0.1nH for an inductor was used.

3.5. Simulation Procedure

All simulations were performed using CST Studio Suite. The simulation procedure consisted of the following steps:

- Simulate the connected slot array antenna without a reflector.
- Add the reflector and determine the influence on the antenna input impedance.
- Extract the antenna impedance over the operating frequency range.
- Design the three NIC topologies to match the reflector impedance.
- Optimize the circuit parameters to get the best results for the given thresholds.
- Determine the ARC Γ_{active} and the resulting bandwidth improvement.
- Evaluate the results of each topology.
- Repeat the simulations for elevation angles of 0° , 20° , 40° and 60° and also azimuth angle of 0° and 90° .

The performance of each matching network was evaluated based on the achieved ARC and input impedance over the operational frequency range.

Results and Discussion

The simulation results are presented and discussed in this chapter. The analysis begins with the antenna operating in free space, i.e., without a reflector, followed by the antenna with a backing reflector. For both configurations, the ARC and input impedance are evaluated under broadside conditions, meaning that no scan angle is applied. Next, the performance of the reflector-backed antenna combined with the matching NIC is investigated. For these results, the ARC and input impedance are examined for broadside and at different scan angles to evaluate the effectiveness of the matching network under varying operating conditions. In the final section, the insertion losses of the circuits are presented.

4.1. Free-Space Antenna

The antenna is simulated without reflector and under broadside conditions. The schematic of the antenna in CST is given in figure 4.1 and the simulation results are displayed in figure 4.2. The ARC of the antenna remains under -10 dB for the entire frequency range, meaning only a small part of the power is being reflected back. The input impedance is slightly decreasing over the frequency range, but shows no oscillations, meaning it is relatively steady over the frequency band.

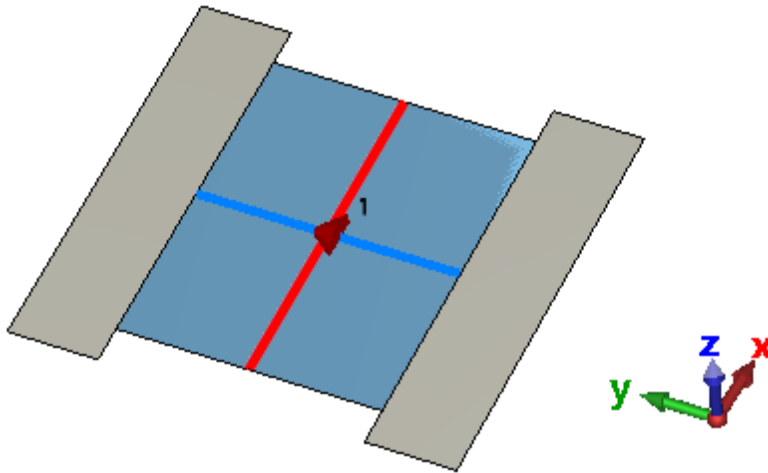


Figure 4.1: Schematic of the connected slot array unit cell in free-space.

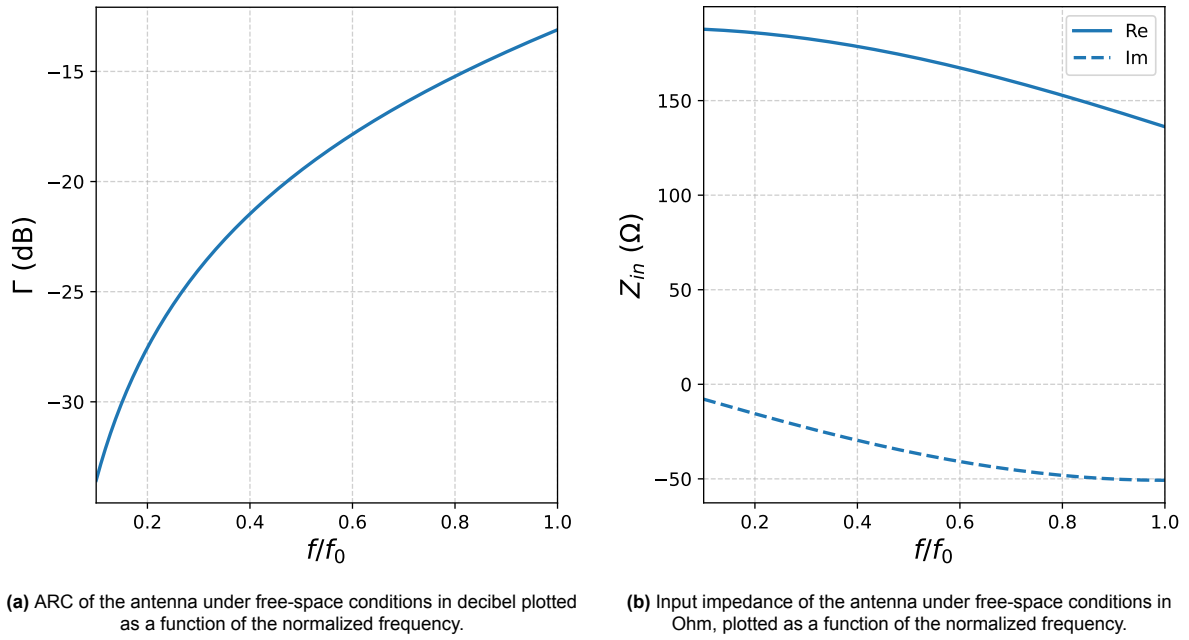


Figure 4.2

4.2. Antenna with Reflector

The simulations of the antenna with a backing reflector are shown in figures 4.3 and 4.4, the schematic of the antenna was given in figure 3.1. The ARC exceeds the -10 dB level until $0.6f_0$, meaning too much of the power is being reflected back. In the impedance graph, the imaginary part seems to have a peak at $0.6f_0$; this is likely caused by a resonance due to the distance from the ground plane, as shown in equation 3.1. The resonance is clearly visible by the frequency behavior of the impedance in figure 4.4. The results agree with the results found in the literature [9].

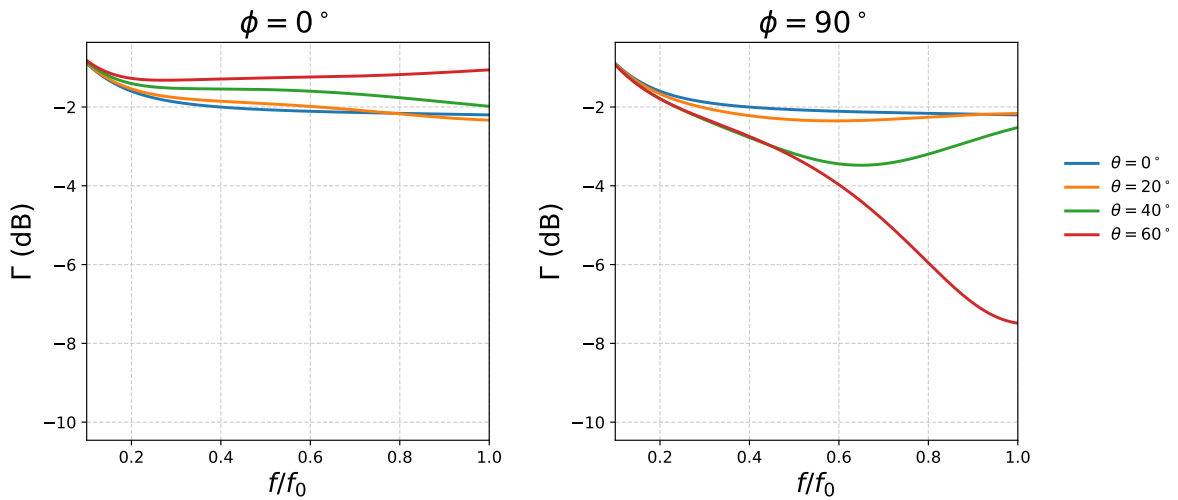


Figure 4.3: ARC of the antenna with reflector in decibel plotted as a function of the normalized frequency for varying incidence angles θ , plotted as a function of normalized frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

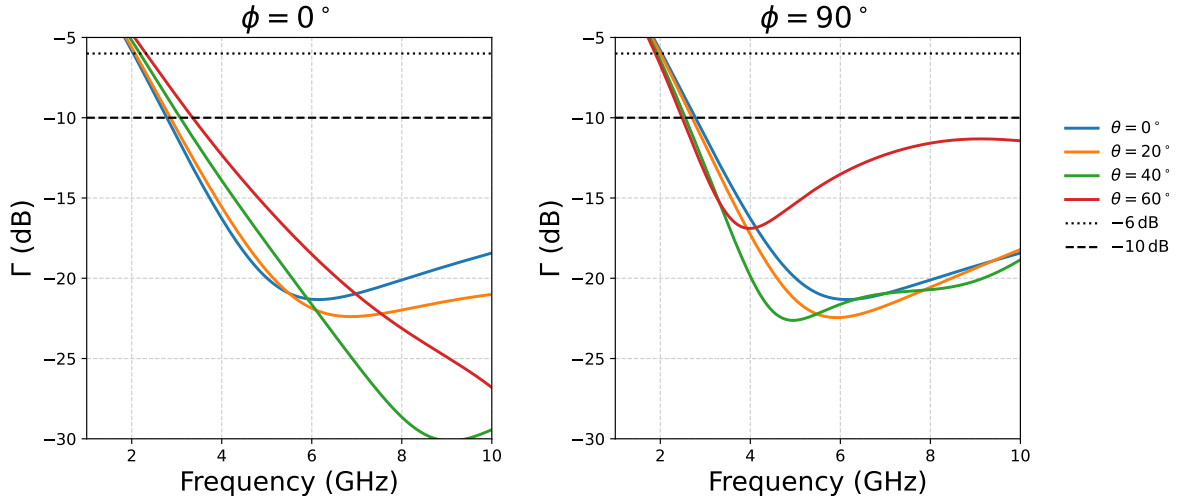


Figure 4.6: ARC of the op-amp NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 90\Omega$.

However, even for the 3-10 GHz range, this is not a successful result. At these frequencies, far beyond its gain-bandwidth product of $GBP = 3.9\text{GHz}$, the NIC does not produce a negative inductor anymore. This is shown in Figure 4.7b, with the equivalent circuit of a low gain op-amp in Figure 4.7a.

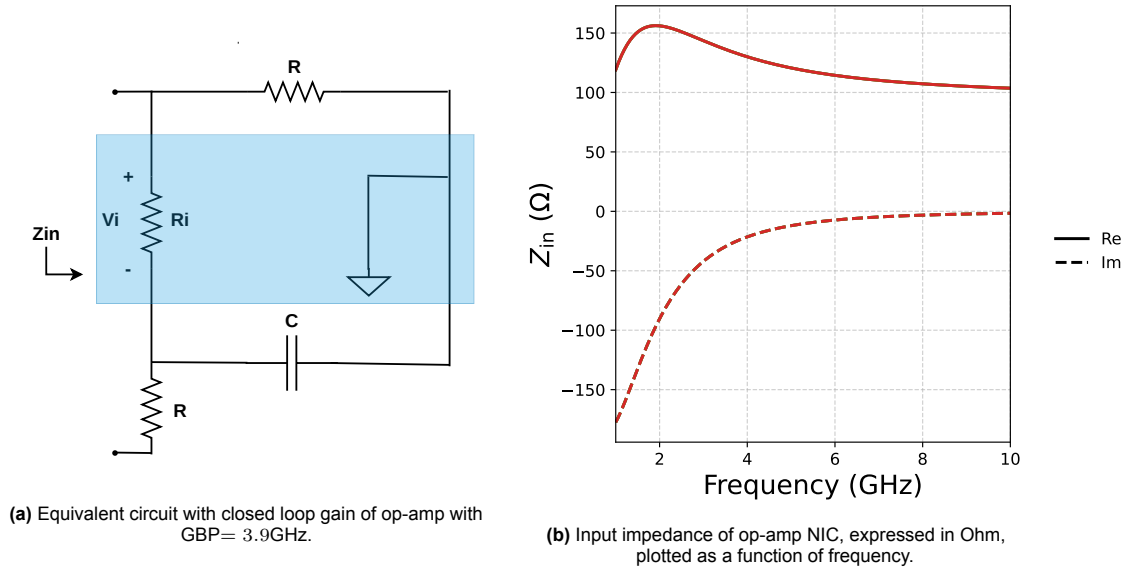


Figure 4.7: Equivalent circuit and input impedance of the op-amp NIC

It can be observed that not only does it not act as a negative inductance, the NIC now also adds a significant amount of resistance to the total setup. This can also be observed when taking $A_v \rightarrow 0$ and $R_i \rightarrow \infty$ (because this is significantly larger than the other resistances in the circuit) in Equation 3.5. Now, after some derivation, $Z_{in} \approx R$. With the free-space characteristic impedance being $Z_{fs} = 377\Omega$, approximately 70% of the input power is going to the op-amp NIC, because the two are in parallel, as shown in Figure 3.3.

The input impedance results obtained with the op-amp NIC are shown in Figure 4.8. Because the ARC and the input impedance are linked through Equation 3.4, the more it resembles the input impedance of the antenna without backing reflector, the better. However, the figures show in the low frequency ranges

deviation from this more flat goal. Hence, this where there is no matching in section 4.3. This behavior is caused by the limited gain-bandwidth product of the operational amplifier. Using Equation 3.12, at a frequency of 1GHz and the GBP mentioned in Table 3.2, resulting in a closed loop gain $|A_v| = 3.9$. This is clearly insufficient gain for the NIC to get an appropriate result.

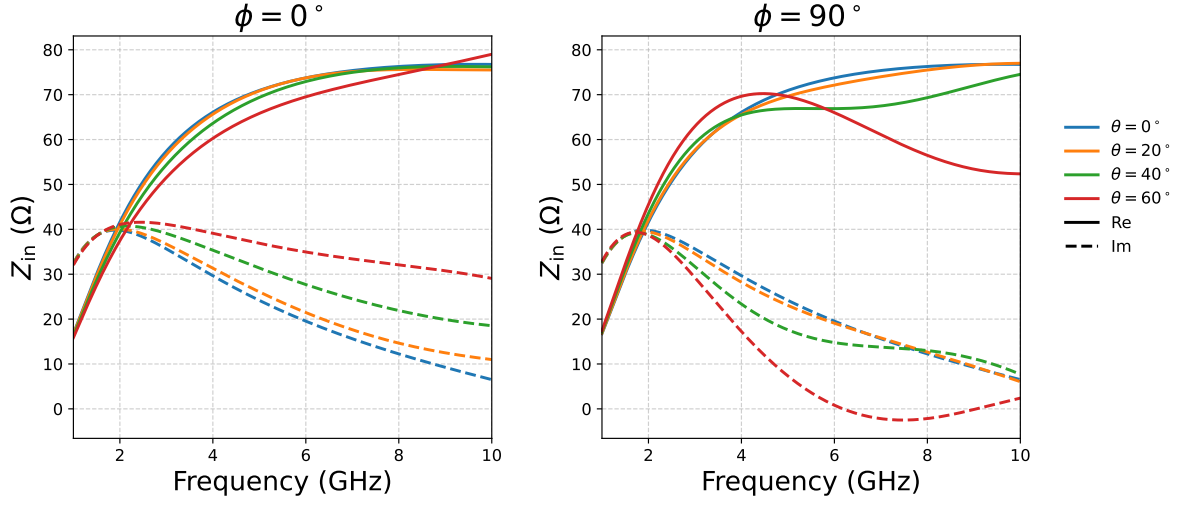


Figure 4.8: Input impedance of the op-amp NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

To get the lowest needed GBP for a result that meets the thresholds, a parameter sweep was performed. For the 1–10 GHz range, the lowest needed GBP ≈ 308 GHz. This is far beyond the current limits of high-frequency operational amplifiers.

4.3.2. 0.1-1 GHz

The results for the op-amp NIC in the 0.1 – 1 GHz range are given in figures 4.9 and 4.10. These results seem worse than those of the 1–10 GHz frequency range. Showing that the op-amp NIC does not meet the requirements to properly match the impedance caused by the reflector at this frequency range either. Using Equation 3.12 again, but now at a frequency of 0.1GHz with the same gain-bandwidth product of GBP = 3.9GHz, gives $|A_v| = 39$. This does not provide enough gain for the NIC to produce a negative impedance.

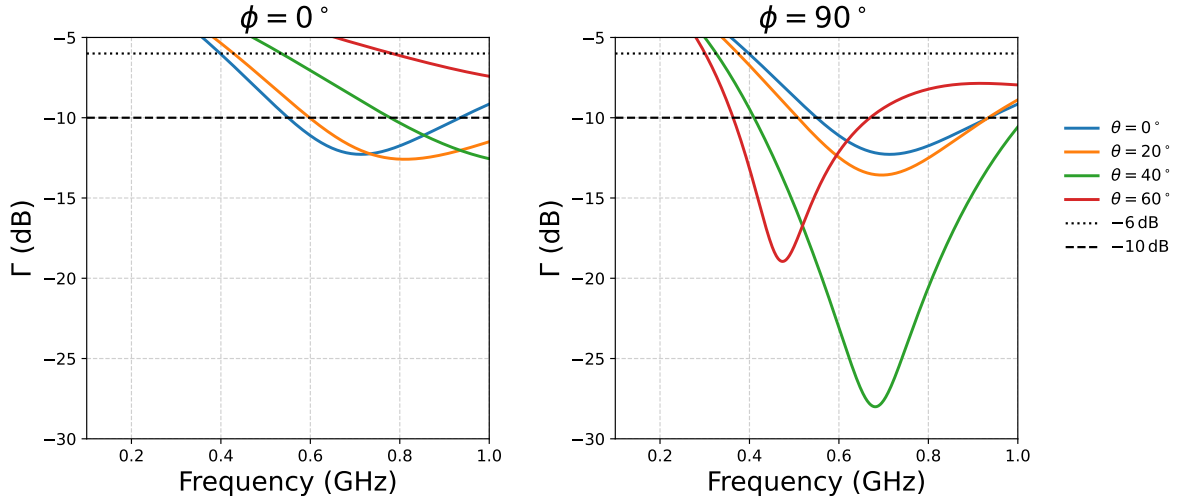


Figure 4.9: ARC of the op-amp NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 200\Omega$.

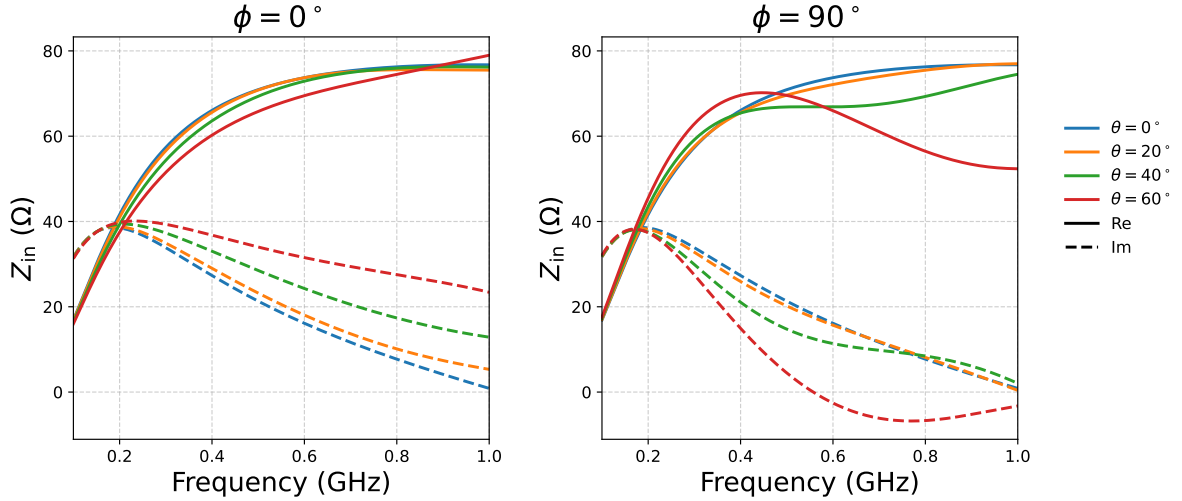


Figure 4.10: Input impedance of the op-amp NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

As for the 1–10 GHz frequency range, a parameter sweep was performed to determine the minimum required gain-bandwidth product. For the 0.1–1 GHz range, the minimum required GBP was found to be approximately 30 GHz. This reduction by roughly one order of magnitude compared to the 308 GHz requirement in the 1–10 GHz range is consistent with the corresponding decade decrease in operating frequency.

4.4. NMOS NIC

Next, the NMOS NIC was connected to the antenna. The simulation results are presented and discussed in this section. First, the results for the 1–10 GHz frequency range are shown, followed by those for the 0.1–1 GHz range. For each case, the active reflection coefficient and input impedance are evaluated for $\phi = 0^\circ$ and $\phi = 90^\circ$, and for the scan angles $\theta = 0^\circ$, 20° , 40° , and 60° .

4.4.1. 1-10 GHz

The ARC results obtained in the 1–10 GHz frequency range using the NMOS-based NIC are shown in Figure 4.11. For $\phi = 0^\circ$, the reflection coefficient remains below the -10 dB threshold for scan angles up to $\theta = 40^\circ$ across the entire frequency range. For $\theta = 60^\circ$, it barely crosses the -6 dB threshold near $f = 9$ GHz, although it remains close to this limit. This indicates that the matching performance is satisfactory for all scanning angles.

For $\phi = 90^\circ$, similar results are observed. The ARC remains below -10 dB for $\theta = 0^\circ, 20^\circ, 40^\circ$. For $\theta = 60^\circ$, the ARC, like in the $\phi = 0^\circ$ case, is close to the threshold of -6 dB, but remains in sufficient range. Consequently, the NMOS-based NIC satisfies the matching requirements for all scan-angle combinations.

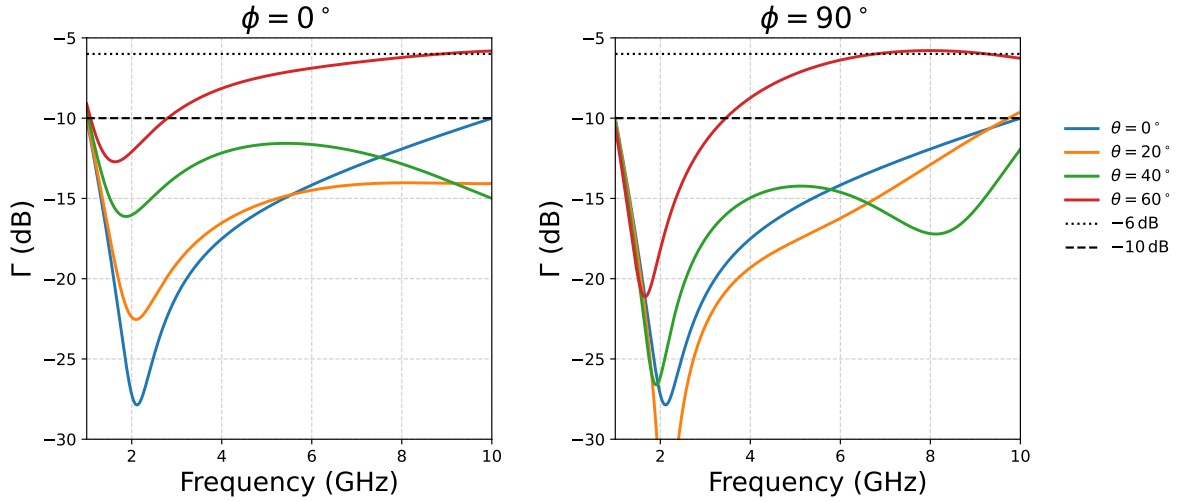


Figure 4.11: ARC of the NMOS NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 280\Omega$.

The input impedance results obtained with the NMOS NIC are shown in Figure 4.12. For both $\phi = 0^\circ, 90^\circ$, the real and imaginary part of the impedance remain stable over the frequency band, resembling the free-space results, shown in Figure 4.2b. This indicates a sufficient results, which can be seen in Figure 4.11. For both cases, the $\theta = 60^\circ$ seems unstable and does not closely resemble the free-space impedance. This is confirmed by the inferior matching.

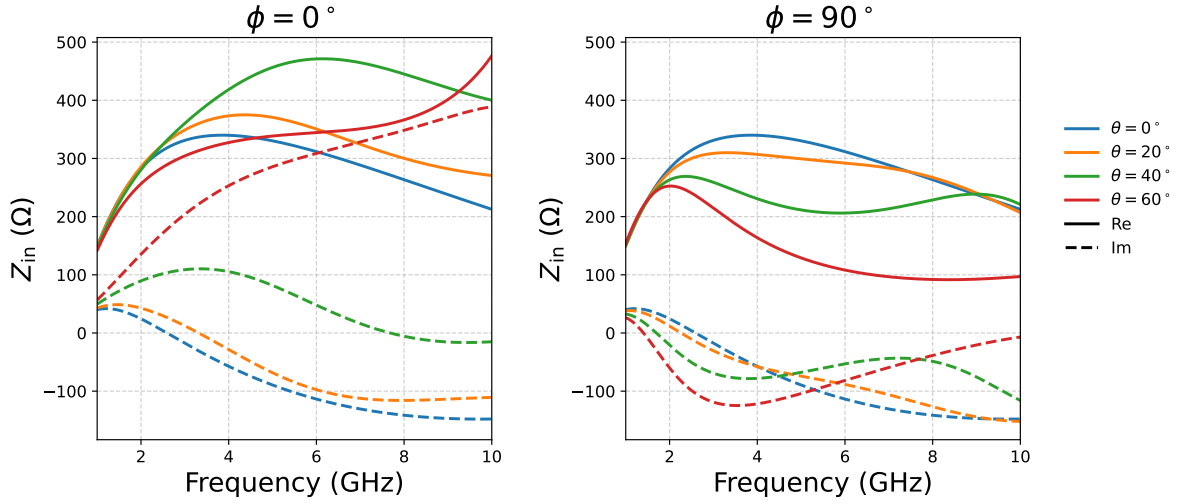


Figure 4.12: Input impedance of the NMOS NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

4.4.2. 0.1-1 GHz

The ARC results of the NMOS NIC in the 0.1–1 GHz range are shown in figure 4.13. The results are similar to the 1–10 GHz case, shown in Figure 4.11. Again for both ϕ cases $\theta = 0^\circ$, 20° , and 40° remain below the -10 dB requirement across the entire frequency range. For $\theta = 60^\circ$, the reflection coefficient exceeds the -10 dB limit, but remains mostly under its threshold of -6 dB, meaning a successful result for this frequency range.

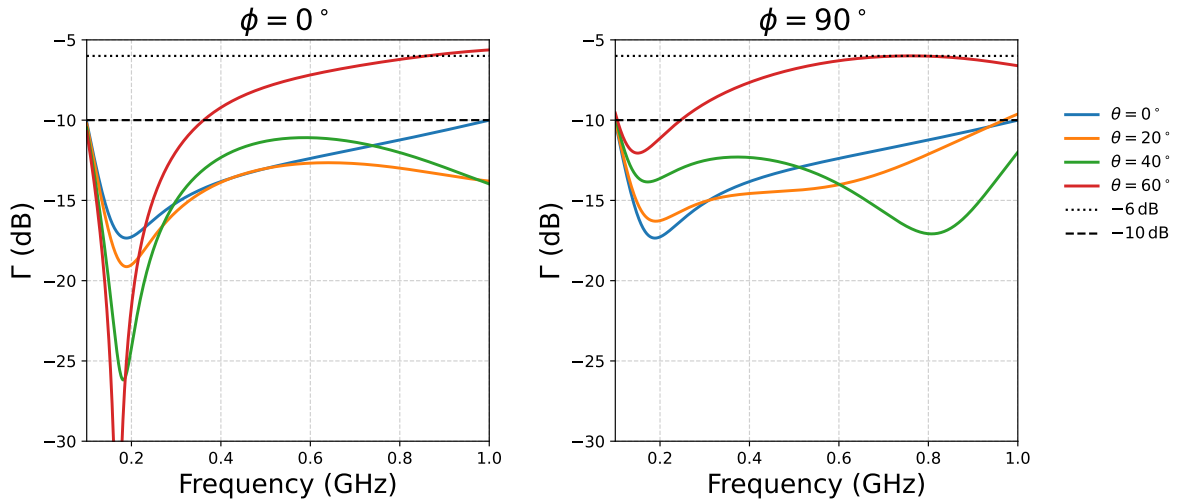


Figure 4.13: ARC of the NMOS NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 260\Omega$.

The input impedance results obtained with the NMOS NIC are shown in Figure 4.12. These results are again similar to the 1–10 GHz range. Because the impedance and the ARC are linked by Equation 3.4, this is expected when the ARC for both frequency ranges are similar too. Like before, the more it matches the shape of the free-space ARC results, shown in Figure 4.2a, the lower the reflection. Again, for $\phi = 0^\circ$ and $\theta = 60^\circ$, a big increase in the imaginary part can be observed, which is likely caused by the series inductance. However, lowering this inductance for better matching on this angle will simultaneously decrease the ARC of the other angles, indicating a trade-off. Like before, the inferior matching of the $\theta = 60^\circ$ at the $\phi = 90^\circ$ is because of the matching to the designated Z_0 mentioned in

Table 3.3. Because of the stable impedance results at these angles, it is possible to get a lower ARC by using a different Z_0 . However, this will result in inferior matching for the other angles, again indicating a trade-off.

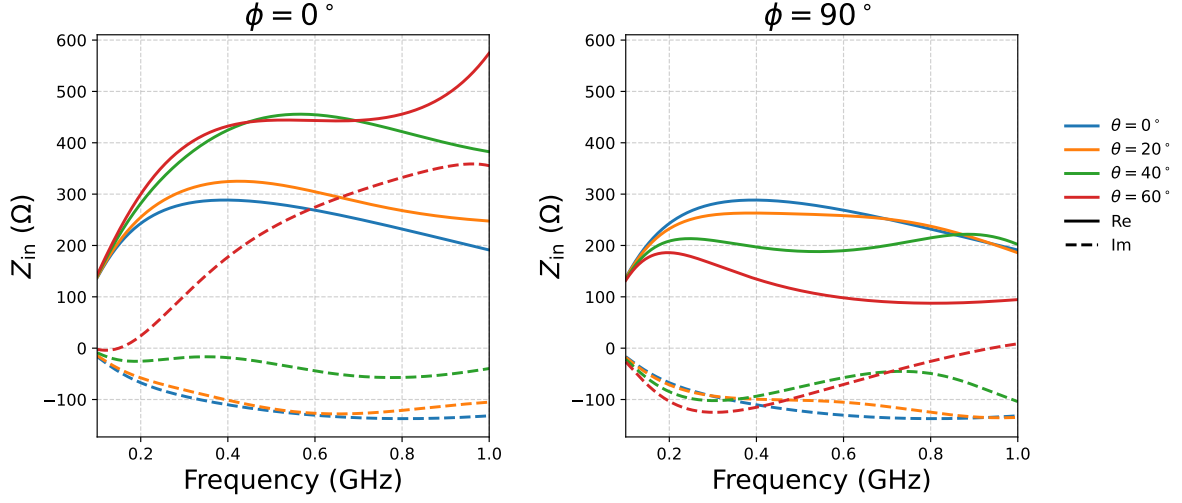


Figure 4.14: Input impedance of the NMOS NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

4.5. Tunneling Diode NIC

The tunneling diode NIC was the final circuit analyzed in this project. The final simulation results are presented and discussed in this section. First, the results for the 1-10 GHz frequency range are shown, followed by those for the 0.1–1 GHz range. For each case, the active reflection coefficient and input impedance are evaluated for $\phi = 0^\circ$ and $\phi = 90^\circ$, and for the scan angles $\theta = 0^\circ$, 20° , 40° , and 60° .

4.5.1. 1-10 GHz

The ARC results obtained using the tunnel-diode-based NIC are shown in figure 4.15. For $\phi = 0^\circ$ and $\phi = 90^\circ$, all angles θ start slightly above the -10 dB limit, but drop down around $f = 1.2$ GHz and stay below the limit. This indicates that the matching performance is good for all scan angles. This slight mismatch is because of the C_j being close to the $C_{j,max}$ mentioned in Equation 3.30. Lowering C_j slightly by 10% to 0.63 pF gives a result where for all frequencies it is below the designated threshold. Consequently, the tunnel-diode-based NIC satisfies the matching requirements for all scan-angle combinations on the 1 – 10 GHz range.

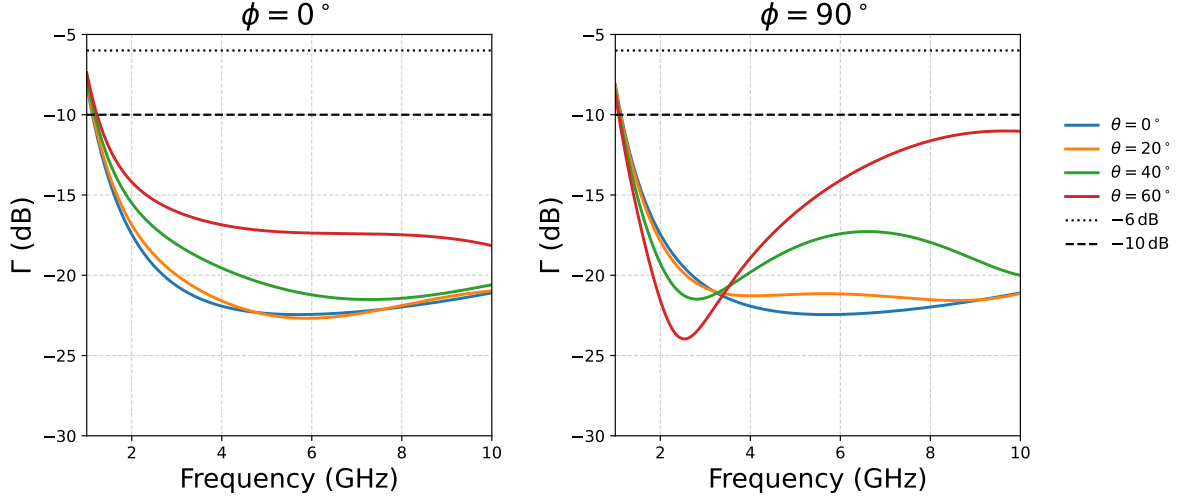


Figure 4.15: ARC of the tunneling diode NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 95\Omega$.

The input impedance results obtained using the tunnel-diode-based NIC are shown in figure 4.16. For both $\phi = 0^\circ, 90^\circ$, the impedance parameters resemble the free-space impedance results, shown in Figure 4.2b. Indicating sufficient matching over the frequency band. Like mentioned before, only at the frequencies close to 1GHz is there a deviation with the free-space impedance. This can be seen in ??, where the difference can be explained by a slightly too large C_j .

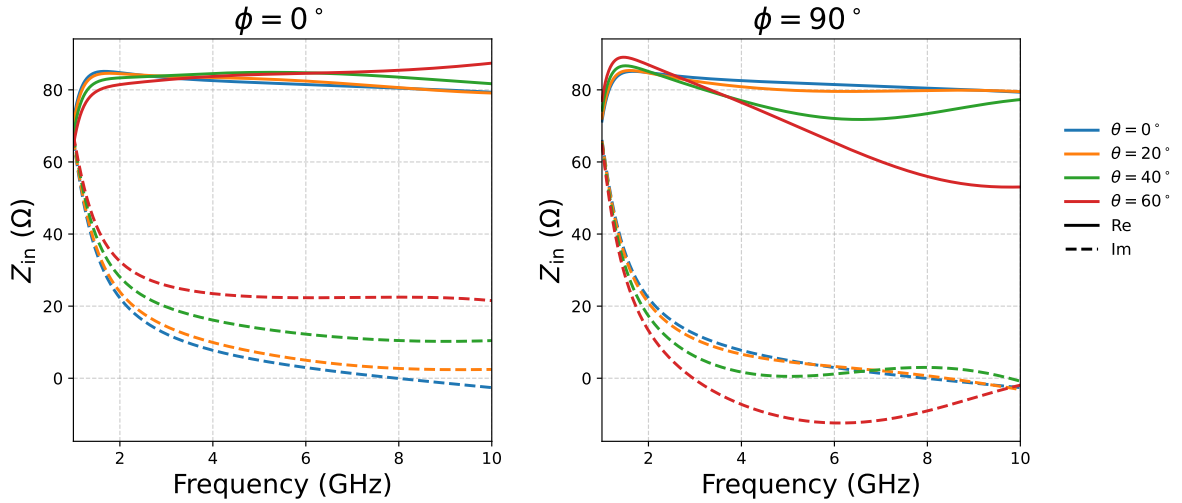


Figure 4.16: Input impedance of the tunneling diode NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of normalized frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

4.5.2. 0.1-1 GHz

The active reflection coefficient results of the tunneling diode NIC are shown in Figure 4.17. For $\phi = 0^\circ$ and $\phi = 90^\circ$, all scan angles comfortably satisfy their respective thresholds across the entire frequency range. This is due to $C_j \ll C_{j,max}$ for this frequency range, meaning it is well within the bound mentioned in Equation 3.28.

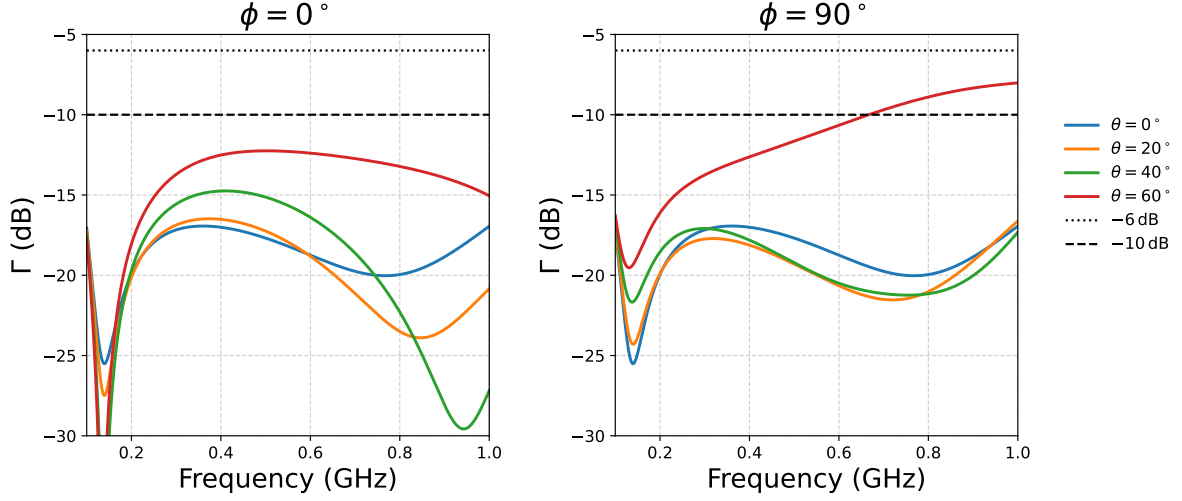


Figure 4.17: ARC of the tunneling diode NIC, expressed in decibel, for varying incidence angles θ , plotted as a function of frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right), normalized to $Z_0 = 130\Omega$.

The input impedance results obtained with the tunneling diode NIC are shown in Figure 4.18. For $\phi = 0^\circ$, the real part of the impedance remains positive and relatively smooth, increasing with scan angle. The imaginary part remains small compared to the real part and varies very little with frequency. This indicates effective compensation of the reflector reactance, resembling that of the free-space impedance in Figure 4.2b.

For $\phi = 90^\circ$, the impedance exhibits a stronger dependence on scan angle for $\theta = 60^\circ$. The real part decreases significantly, while the imaginary part changes sign at higher frequencies. This behavior corresponds with the degradation observed in the reflection coefficient results, because the two are linked through Equation 3.4. Again, like in section 4.4, decreasing the series inductance in Figure 3.3 gives a better matching for the $\phi = 90^\circ$ and $\theta = 60^\circ$ angle. However, this raises the ARC for the other angles, pointing to the before mentioned trade-off.

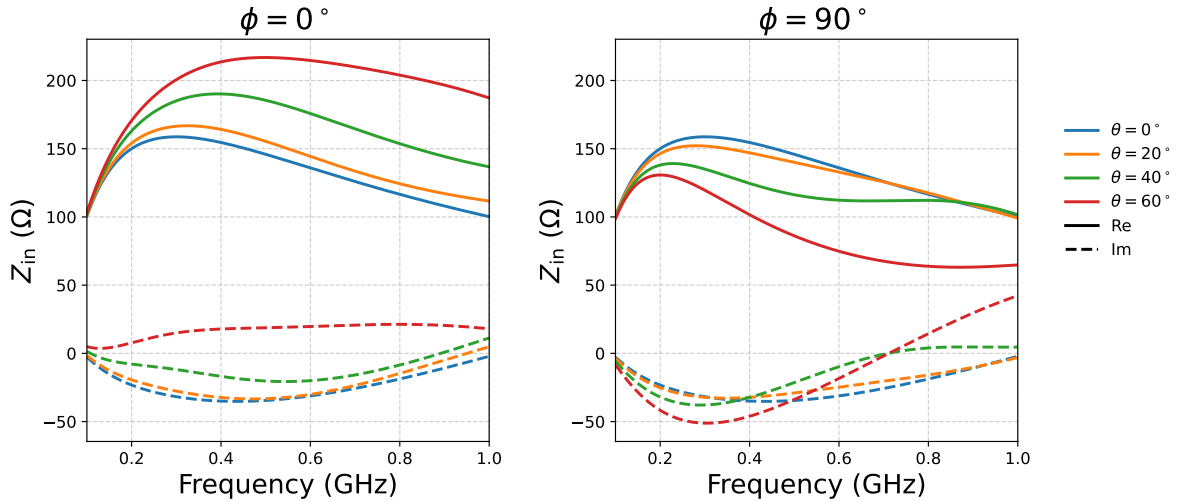


Figure 4.18: Input impedance of the tunneling diode NIC, expressed in Ohm, for varying incidence angles θ , plotted as a function of normalized frequency for $\phi = 0^\circ$ (left) and $\phi = 90^\circ$ (right).

4.6. Estimated Insertion Losses

Because the matching circuits also contain resistive components, the insertion losses of the NIC implementations are also considered. An evaluation performed with CST indicates that both the NMOS and tunneling diode NICs exhibit an insertion loss over the considered frequency range of $S_{21} \approx -0.25$ dB. This corresponds to a power transmission efficiency of 94%, meaning only a small fraction of the signal power is dissipated in the matching network. These results indicate that the improvements of the matching circuit do not significantly affect the antenna efficiency.

5

Conclusion

This chapter summarizes the main findings of the project and evaluates the extent to which the objectives and requirements have been successfully achieved. The performance of the investigated non-Foster matching circuits is assessed based on the simulation results, and the possibility of real-life applications is discussed. Finally, the recommendations for future work on this topic are given, followed by some final thoughts on the project.

5.1. Summary of findings

This project investigated the use of active non-Foster matching circuits to improve the bandwidth performance of connected array antennas with a backing reflector. The main objective was to determine whether negative impedance converter can effectively compensate the frequency-dependent inductive behavior introduced by the reflector, and thereby enable wide-band impedance matching beyond the limitations of conventional passive matching techniques. The results show that non-Foster matching can significantly improve the impedance behavior of the antenna system over a wide frequency range. The NMOS and tunneling diode implementations achieve good matching performance over the investigated bandwidth, while the op-amp implementation shows limited effectiveness due to its insufficient gain-bandwidth product at GHz frequencies. The results of the NMOS and tunneling diode confirm that the fundamental compensation principle of negative inductance works and is effective in enhancing antenna bandwidth. An evaluation of the insertion losses indicates a loss of only -0.25 dB for the NMOS and tunneling diode implementations, demonstrating that the bandwidth enhancement can be achieved without significant dissipation losses.

It was also found that increasing the scan angle has a negative effect on the matching performance. As the scan angle increases, the active reflection coefficient generally increases, indicating a reduction in matching quality. Although the matching performance decreases with increasing scan angle, both the NMOS and tunneling diode implementations continue to satisfy the scan-angle requirements for all cases.

Among the investigated topologies, the tunneling diode implementation provided the most consistent matching performance across the different scan angles. The NMOS implementation achieved comparable results, but with slightly lower margins from the specified thresholds. In contrast, the op-amp implementation was unable to provide adequate compensation due to the limited gain-bandwidth product of practical operational amplifiers at GHz frequencies.

5.2. Requirements and Objectives

The project objectives and requirements were mostly met. The main goal of demonstrating bandwidth enhancement using non-Foster matching in simulation was successfully achieved. The antenna system with active matching consistently outperformed the unmatched case in terms of impedance bandwidth. Both the NMOS and tunneling diode topologies demonstrated good wide-band matching performance over the operating frequency ranges and scanning angles. The results demonstrate that non-Foster

matching can be implemented successfully in practical circuit topologies, but the performance remains dependent on physical limitations of the components. From an application perspective, the results indicate that non-Foster matching circuits are promising for improving wide-band phased-array performance in communication, radar, and sensing systems.

5.3. Future Work

The next iteration of this project should focus on improving stability in the designs, including the effects of noise and nonlinearity, and investigating appropriate biasing strategies for all active components and quantifying the associated power consumption. Once these points are addressed, a next step could be looking into the real-life fabrication of these NICs, and then testing if their performance matches the theory.

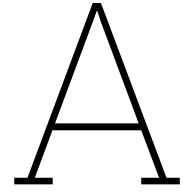
5.4. Final Remarks

Overall, this project demonstrates that active non-Foster matching is a viable approach for extending the bandwidth of connected array antennas. This project also provided deeper insight into wide-band antenna behavior, impedance matching techniques and the creation of negative-reactance networks. The knowledge and experience gained from this work forms a valuable foundation for future research on active matching networks and wide-band phased array antennas.

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Parameter Optimization Python Code

```
1 import itertools
2 import re
3 from pathlib import Path
4 import matplotlib.pyplot as plt
5 import numpy as np
6
7 CA_ZIN_FILE = Path("Antenna_data/CA_Zin.txt")
8
9 ZO = 90
10
11 FREQ_MODE = "1-10GHz"
12 # FREQ_MODE = "0.1-1GHz"
13
14 THRESHOLD_dB = {0: -10.0, 60: -6.0}
15
16 # Data loading
17 def parse_CA_Zin_file(filepath):
18     data = {}
19     current_key = None
20     freqs, reals, imags = [], [], []
21
22     with open(filepath, "r") as f:
23         for line in f:
24             line = line.strip()
25             if not line:
26                 continue
27             if line.startswith("#"):
28                 phi_m = re.search(r"phi=(\d+)", line)
29                 theta_m = re.search(r"theta=(\d+)", line)
30                 if phi_m and theta_m:
31                     if current_key is not None and freqs:
32                         data[current_key] = (
33                             np.array(freqs),
34                             np.array(reals) + 1j * np.array(imags),
35                         )
36                         freqs, reals, imags = [], [], []
37                         current_key = (int(phi_m.group(1)), int(theta_m.group(1)))
38                     elif current_key is not None:
39                         parts = line.split()
40                         if len(parts) == 3:
41                             try:
42                                 freqs.append(float(parts[0]))
43                                 reals.append(float(parts[1]))
44                                 imags.append(float(parts[2]))
45                             except ValueError:
46                                 pass
47
48     if current_key is not None and freqs:
49         data[current_key] = (np.array(freqs), np.array(reals) + 1j * np.array(imags),
```

```

50     )
51
52     return data
53
54
55 def split_by_phi(all_data):
56     br_phi0 = {theta: d for (phi, theta), d in all_data.items() if phi == 0}
57     br_phi90 = {theta: d for (phi, theta), d in all_data.items() if phi == 90}
58     return br_phi0, br_phi90
59
60
61 def make_sim_freq(f_data_GHz):
62     f_sim_GHz = f_data_GHz / 10.0 if FREQ_MODE == "0.1-1GHz" else f_data_GHz
63     return f_sim_GHz, f_sim_GHz * 1e9
64
65 # Circuit models
66 def Zin_opamp_GBP(f_Hz, R1, Av, Ri, Ro, Cp, GBP):
67     s = 1j * 2.0 * np.pi * f_Hz
68     fp = GBP / Av
69     A_s = Av / (1 + 1j * f_Hz / fp)
70     Zint = 1.0 / (s * Cp)
71     ZC = 1e20
72     num = R1 * (A_s*R1*Ri*ZC*Zint - R1*Ri*Ro*ZC + R1*Ri*Ro*Zint + R1*Ri*ZC*Zint + R1*Ro*ZC*
73             Zint + R1*Ro*Zint**2 + R1*ZC*Zint**2 + Ri*Ro*ZC*Zint + Ri*Ro*Zint**2 + Ri*ZC*Zint**2)
74     den = (-A_s*Ri*ZC*Zint**2 - R1**2*Ro*ZC + R1**2*Ro*Zint + R1**2*ZC*Zint - R1*Ri*Ro*ZC +
75             R1*Ri*Ro*Zint + R1*Ri*ZC*Zint + 2*R1*Ro*Zint**2 + 2*R1*ZC*Zint**2 + Ri*Ro*ZC*Zint +
76             Ri*Ro*Zint**2 + Ri*ZC*Zint**2)
77     return num / den
78
79 def Zin_nmos(f_Hz, gm, ro, Lp, R):
80     s = 1j * 2.0 * np.pi * f_Hz
81     Z_int = s * Lp + R
82     num = ro * (-Z_int * ro * gm**2 + 2 * gm * ro + 1)
83     den = (gm * ro)**2 + 2 * gm * ro + 1
84     return num / den
85
86 def Zin_tunnel_diode(f_Hz, Cj, Rn, Rs, Ls, R, Cp):
87     s = 1j * 2.0 * np.pi * f_Hz
88     ZCj = 1.0 / (s * Cj)
89     Z_par = (Rn * ZCj) / (Rn + ZCj) # Rn Cj
90     ZTD = Rs + s * Ls + Z_par
91     Zint = 1.0 / (s * Cp)
92     Z_right = R + Zint
93     return R + 1.0 / (1.0 / ZTD + 1.0 / Z_right)
94
95 # Map names to functions
96 ZIN_models = {
97     "opamp_GBP": Zin_opamp_GBP,
98     "nmos": Zin_nmos,
99     "tunnel_diode": Zin_tunnel_diode,
100 }
101
102 def Zin(f_Hz, **params): # call the active circuit model selected
103     return ZIN_models[CIRCUIT](f_Hz, **params)
104
105 # Series inductor
106 def Z_ind(f_Hz, Lseries):
107     return 1j * 2.0 * np.pi * f_Hz * Lseries
108
109 # System impedance and S11
110 def compute_Z_model(Z_in, Z_br, Z_ind_val=0.0):
111     # return Z_in
112     return Z_ind_val + Z_in * Z_br / (Z_in + Z_br) # antenna case: Z_ind + Zin || Z_br
113
114 def compute_S11(Z, Z0=Z0): # S11 = (Z - Z0) / (Z + Z0)
115     return (Z - Z0) / (Z + Z0)
116
117 # Sweep over phi=0 and phi=90
118 def sweep_parameters(f_sim_Hz, br_per_angle_phi0, br_per_angle_phi90, sweep_params,
119                     fixed_params, sweep_zind_params=None, fixed_zind_params=None, angle_weights=None,
120                     sweep_Z0=None):

```

```

116 sweep_zind_params = sweep_zind_params or {}
117 fixed_zind_params = fixed_zind_params or {}
118 use_zind = bool(sweep_zind_params) or bool(fixed_zind_params)
119
120 aw = angle_weights or {}
121
122 zin_names = list(sweep_params.keys())
123 zind_names = list(sweep_zind_params.keys())
124 all_grids = [np.asarray(sweep_params[k]) for k in zin_names] + \
125             [np.asarray(sweep_zind_params[k]) for k in zind_names]
126 n_zin = len(zin_names)
127 n_combos = int(np.prod([len(g) for g in all_grids])) if all_grids else 1
128
129 ZO_grid = np.atleast_1d(np.asarray(sweep_Z0, dtype=float)) \
130             if sweep_Z0 is not None else np.array([Z0])
131 total_combos = n_combos * len(ZO_grid)
132
133 print(f"Frequency_mode: {FREQ_MODE}")
134 print(f"Circuit: {CIRCUIT}")
135 print(f"Sweeping_Zin_parameter(s): {zin_names}")
136 if zind_names:
137     print(f"Sweeping_Z_ind_parameter(s): {zind_names}")
138 if len(ZO_grid) > 1:
139     print(f"Sweeping_Z0: {ZO_grid[0]:.1f} - {ZO_grid[-1]:.1f} ohm ({len(ZO_grid)} pts)")
140 if aw:
141     print(f"Angle_weights: {aw}")
142 print(f"Active_thresholds: {THRESHOLD_dB}")
143 print(f"Total_grid_points: {total_combos:,}")
144
145 # Precompute Z_br per angle (constant across the sweep)
146 Z_br_phi0 = {a: br_per_angle_phi0[a][1] for a in sorted(br_per_angle_phi0)}
147 Z_br_phi90 = {a: br_per_angle_phi90[a][1] for a in sorted(br_per_angle_phi90)}
148
149 active_phi0 = [(a, THRESHOLD_dB[a]) for a in sorted(br_per_angle_phi0) if a in
150               THRESHOLD_dB]
151 active_phi90 = [(a, THRESHOLD_dB[a]) for a in sorted(br_per_angle_phi90) if a in
152               THRESHOLD_dB]
153
154 best_zin_params = None
155 best_zind_params = None
156 best_Z0_val = float(ZO_grid[0])
157 best_cost = np.inf
158 report_step = max(1, total_combos // 10)
159 global_i = 0
160
161 for ZO_val in ZO_grid:
162     for combo in itertools.product(*all_grids):
163         zin_params = dict(fixed_params)
164         for name, val in zip(zin_names, combo[:n_zin]):
165             zin_params[name] = val
166
167         zind_params = dict(fixed_zind_params)
168         for name, val in zip(zind_names, combo[n_zin:]):
169             zind_params[name] = val
170
171         # Zin and Z_ind are angle, compute once per combo
172         Z_in = Zin(f_sim_Hz, **zin_params)
173         Z_ind_val = Z_ind(f_sim_Hz, **zind_params) if use_zind else 0.0
174
175         feasible = True
176         margin_sum = 0.0
177         violation_sum = 0.0
178
179         for dataset, active_angles in [(Z_br_phi0, active_phi0), (Z_br_phi90,
180               active_phi90)]:
181             for a, thresh in active_angles:
182                 Z_model = compute_Z_model(Z_in, dataset[a], Z_ind_val)
183                 S11 = compute_S11(Z_model, ZO_val)
184                 max_dB = float(np.max(20.0 * np.log10(np.maximum(np.abs(S11), 1e-12))))
185                 w = aw.get(a, 1.0)

```

```

184         if max_dB > thresh:
185             feasible = False
186             violation_sum += w * (max_dB - thresh)
187         else:
188             margin_sum += w * (thresh - max_dB)
189
190     # Two-tier cost: any feasible solution beats any infeasible one
191     cost = -margin_sum if feasible else 1e6 + violation_sum
192
193     if cost < best_cost:
194         best_cost = cost
195         best_zin_params = dict(zin_params)
196         best_zind_params = dict(zind_params)
197         best_Z0_val = float(Z0_val)
198
199     global_i += 1
200     if total_combos > 1000 and global_i % report_step == 0:
201         print(f"_{100*_global_i//_total_combos}%_complete...")
202
203     return best_zin_params, best_zind_params, best_Z0_val, best_cost
204
205 # Plotting
206 def plot_results(f_sim_GHz, best_zin_params, best_zind_params, br_per_angle_phi0,
207                 br_per_angle_phi90, Z0_plot=Z0):
208     f_sim_Hz = f_sim_GHz * 1e9
209     Z_in = Zin(f_sim_Hz, **best_zin_params)
210     Z_ind_val = Z_ind(f_sim_Hz, **best_zind_params) if best_zind_params else 0.0
211
212     angles = sorted(set(br_per_angle_phi0) | set(br_per_angle_phi90))
213
214     for angle in angles:
215         fig, ax = plt.subplots(figsize=(9, 4))
216
217         if angle in br_per_angle_phi0:
218             _, Z_br = br_per_angle_phi0[angle]
219             S11_dB = 20.0 * np.log10(np.maximum(
220                 np.abs(compute_S11(compute_Z_model(Z_in, Z_br, Z_ind_val), Z0_plot)), 1e-12))
221             ax.plot(f_sim_GHz, S11_dB, linewidth=2, label=f"|S11|_phi=0")
222
223         if angle in br_per_angle_phi90:
224             _, Z_br = br_per_angle_phi90[angle]
225             S11_dB = 20.0 * np.log10(np.maximum(
226                 np.abs(compute_S11(compute_Z_model(Z_in, Z_br, Z_ind_val), Z0_plot)), 1e-12))
227             ax.plot(f_sim_GHz, S11_dB, linewidth=2, linestyle="--", label=f"|S11|_phi=90")
228
229         threshold = THRESHOLD_dB.get(angle)
230         if threshold is not None:
231             ax.axhline(threshold, color="r", linestyle=":", linewidth=1.5, label=f"{threshold}
232                 _dB_target")
233
234         ax.set_xlabel("Frequency (GHz)")
235         ax.set_ylabel(f"|S11| (dB)")
236         ax.set_title(f"|S11|_theta={angle}_deg_{(FREQ_MODE)}, Z0={Z0_plot:.1f}_ohm")
237         ax.legend()
238         ax.grid(True, alpha=0.3)
239         plt.tight_layout()
240         plt.savefig(f"spar_extended_theta{angle}.png", dpi=150)
241
242     plt.show()
243
244 if __name__ == "__main__":
245     # Select circuit model:
246     CIRCUIT = "opamp_GBP"
247
248     # Load all (phi, theta) from the combined data file
249     all_data = parse_CA_Zin_file(CA_ZIN_FILE)
250     br_phi0, br_phi90 = split_by_phi(all_data)
251
252     print(f"phi=0_theta_angles_found:_{sorted(br_phi0.keys())}_deg")
253     print(f"phi=90_theta_angles_found:_{sorted(br_phi90.keys())}_deg")

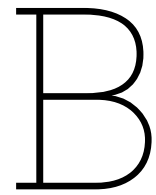
```

```

253
254 # Frequency axis from the first available block
255 f_data_GHz = next(iter(all_data.values()))[0]
256 f_sim_GHz, f_sim_Hz = make_sim_freq(f_data_GHz)
257
258 # Circuit parameter sweep
259 if CIRCUIT == "opamp_GBP":
260     sweep_params = {
261         # "Cp": np.linspace(0.1e-12, 10.0e-12, 50),
262     }
263     fixed_params = {
264         "R1": 100,
265         "Av": 79400,
266         "Ri": 2700,
267         "Ro": 12,
268         "GBP": 3.9e9,
269         "Cp": 6.5918e-12,
270     }
271
272 elif CIRCUIT == "nmos":
273     sweep_params = {
274         # "gm": np.linspace(0.02, 0.5, 100),
275     }
276     fixed_params = {
277         "ro": 200,
278         "Lp": 7.4141e-9,
279         "gm": 0.1,
280         "R": 13,
281     }
282
283 elif CIRCUIT == "tunnel_diode":
284     sweep_params = {
285         "Cp": np.linspace(3.0e-12, 4.0e-12, 20),
286         "R": np.linspace(100, 150, 20),
287     }
288     fixed_params = {
289         "Rs": 3.5,
290         "Ls": 0.1e-9,
291         "Rn": -120,
292         "Cj": 0.7e-12,
293     }
294
295 # Series inductor sweep (clear both dicts to omit)
296 sweep_zind_params = {
297     # "Lseries": np.linspace(0.1e-9, 10e-9, 50),
298 }
299 fixed_zind_params = {
300     "Lseries": 0.1e-9,
301 }
302
303 # Angle weights (0.72 = ratio of linear-scale thresholds= 10^(-6/20) / 10^(-10/20))
304 angle_weights = {0: 1.0, 60: 0.72}
305
306 # Z0 sweep
307 sweep_Z0 = np.linspace(20, 60, 50)
308 # sweep_Z0 = None
309
310 # Run sweep and report
311 best_zin_params, best_zind_params, best_Z0, best_cost = sweep_parameters(f_sim_Hz,
    br_phi0, br_phi90, sweep_params, fixed_params, sweep_zind_params=sweep_zind_params,
    fixed_zind_params=fixed_zind_params, angle_weights=angle_weights, sweep_Z0=sweep_Z0)
312 feasible = best_cost < 0
313 print("\nResult:", "FEASIBLE" if (all(thresholds_met_for_both_phi)) else "INFEASIBLE" if (thresholds_not_met))
314 print(f"Best_Z0={best_Z0:.1f} ohm")
315 print("Best_Zin parameters:")
316 for k, v in best_zin_params.items():
317     print(f"{k:8s}={v:.6g}")
318 if best_zind_params:
319     print("Best_Z_ind parameters:")
320     for k, v in best_zind_params.items():

```

```
322         print(f"_{k:8s}={v:.6g}")
323     if feasible:
324         print(f"Weighted_margin:_{-best_cost:.4g}_dB")
325     else:
326         print(f"Weighted_violation:_{best_cost-1e6:.4g}_dB")
327
328     plot_results(f_sim_GHz, best_zin_params, best_zind_params, br_phi0, br_phi90, best_Z0)
```



Ethical Implications of the Product and Engineering Process

Societal Impacts

The final product is an active non-Foster matching network used to increase the bandwidth of antenna systems. If it is successfully implemented, this technology could have several positive societal impacts. Wideband antennas can replace multiple narrowband antennas, reducing system complexity, size, weight and material usage. This may lead to more efficient communication systems, improved wireless connectivity and lower deployment costs. The main beneficiaries of this are telecommunications companies, satellite operators and scientific researchers.

However, there could also be negative impacts. Improved antenna performance could be applied in military systems such as radar, surveillance equipment, and unmanned aerial vehicles. While these applications may contribute to national security, they may also contribute to military operations and surveillance activities that raise ethical concerns. Also, increased wireless capabilities may contribute to electromagnetic spectrum usage, which could potentially increase interference and energy consumption. The main stakeholders affected by these impacts include communication providers, researchers, defense organizations, governments, and the general public.

Ethical Aspects of Development

The engineering process mostly consisted of literature research, mathematical analysis, optimization routines performed in Python and electromagnetic simulations performed in CST Studio Suite. Since no human participants, personal data, or animal testing were involved, there were no concerns regarding privacy, informed consent, or participant welfare.

Nevertheless, environmental considerations are relevant. The electromagnetic simulations require computational resources, like electrical energy or water consumption due to cooling, and indirectly contribute to environmental impacts. The scale of these simulations is small though, especially compared to large industrial projects. But engineers should still remain aware of the environmental footprint of computational work.

Another ethically relevant aspect is the responsible use of scientific literature. The project relied heavily on published research regarding non-Foster circuits, antenna theory, and active matching techniques. Proper citation and accurate representation of previous work were therefore important responsibilities throughout the project. Failure to acknowledge sources appropriately could undermine academic integrity and the reliability of engineering research.

Ethical Evaluation

The most significant ethical concern associated with the product is its potential dual-use nature. Which means that technologies that are intended for civilian use, might be used in military or surveillance systems. As engineers, it is important to recognize that these technological advancements can have both positive and negative applications. While the project itself focuses on antenna performance rather than specific real life applications, the broader implications of the technology should be acknowledged.

But a lot of aspects of the project are ethically valuable. Increasing antenna bandwidth can reduce the need for multiple specialized antennas, lowering material consumption and system complexity. Improved wireless communication capabilities can also support scientific research and global connectivity. A trade-off exists between maximizing performance and maintaining practical implementation constraints, like power consumption and circuit complexity. The project attempts to balance these factors by investigating realistic non-Foster circuit topologies rather than purely theoretical solutions.

Reflection on Design Choices

Ethical considerations had an influence on several design decisions during the project. In addition to maximizing bandwidth, attention was paid to practical realizability and component selection. This was done so that the antenna at one point can be used in real life applications. This reflects a broader engineering responsibility to consider whether a proposed solution can be safely and reliably used in practice.

Additional attention could have been given to evaluating the energy efficiency of the matching circuits. Since non-Foster networks require active components and external power sources, their power consumption may negatively affect the overall antenna performance. Taking the power consumption into account could have strengthened the ethical evaluation of the design.

Responsibility Looking Forward

If the antenna were to be developed into a real product, several measures could be taken to ensure ethical responsibility. First, the circuit should be thoroughly tested for stability and reliability under realistic operating conditions. Second, the power consumption and environmental impact should be evaluated. Third, the compliance with relevant electromagnetic compatibility and communication regulations should be ensured.

Finally, developers and organizations implementing this technology should remain aware of the potential dual-use concerns. Even though engineers cannot fully control how their technology is ultimately used, transparency regarding capabilities, limitations, and intended applications can help to promote responsible deployment. By considering technical performance alongside societal impacts, future development of non-Foster matching circuits can be conducted in a manner that is both innovative and ethically responsible.