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Design, Fabrication and Testing of a Multifrequency Microstrip RFID Tag Antenna on Si

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Abstract: A configurable design of a microstrip square spiral RFID tag antenna, for a wide range of microwave frequencies in the S- and C-band, is presented. The design is parameterized in dimensions, and hence changing the design frequency (or frequencies) is easy, by changing only an initial value for the spiral geometry. A tag specimen was fabricated using a Cu electroplating technique according to the design for frequencies of interest in the areas of 2.4 and 5.8 GHz. The substrate material is 320 μm high-resistivity Si and the bridge dielectric is 15 μm polyimide PI2525. The steps of the microfabrication process involve metallic structure pattern transfer techniques with optical UV lithography procedures. The reflection coefficient and antenna gain of the specimen were measured inside an anechoic enclosure using a vector network analyzer (VNA) and a TEM horn test antenna over a frequency range of up to 6 GHz. Simulated and measured results, exhibiting reasonable agreement, are presented and discussed.

Keywords: RFID; antenna; silicon



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1. Introduction

RFID (Radio Frequency Identification) systems are used in a variety of fields, including logistics, item-level inventory tracking, access control, race timing, tool tracing, and materials management [1,2], with new applications currently appearing, e.g., health-related ones [3]. RFID tags may be classified into active, semi-passive, and passive ones. An active RFID tag uses a power supply and a transmitter for communication; a semi-passive RFID tag has a power source but no transmitter. A passive RFID tag functions using the electromagnetic wave power delivered by the reader, so it is the cheapest option. The typical frequency bands for RFID applications include 125/135 KHz (LF), 13.56 MHz (HF), 433 MHz, and 840–960 MHz (UHF), as well as 2.45 and 5.8 GHz in the microwave region [4].

Several studies [5,6] have explored Si-based designs for OCAs (On-Chip Antennas) with a view to improving gain and efficiency and reducing parasitic interconnections. Such antennas generally tend to have a narrow bandwidth (less than 10%) and a limited radiation efficiency and range; however, what distinguishes them is their small size, allowing integration into a variety of electronic devices [7]. Recent developments in the implementation of high-performance on-chip antennas for millimeter-wave and terahertz-integrated circuit applications are discussed in [8,9]. On-chip antennas show a significant potential to accommodate applications in fields such as 5G, Internet of Things (IoT), autonomous vehicles, high-data-rate point-to-point communications, GPS, Wi-Fi, WLAN, Bluetooth, and wireless sensors. In recent work (see [9] and references therein), the use of novel substrate layers (such as silicon, graphene, polyimide, and GaAs) to facilitate IC integration has been

studied, as well as a new excitation mechanism that appears to suppress the propagation of surface waves and reduce substrate loss.

The rapid advances in IC technologies in recent decades involve various aspects, such as increased complexity, reduced size, higher frequency regions, etc. In this trend, the use of on-chip integrated antennas through, e.g., complementary metal oxide semiconductor (CMOS) technology, is becoming an increasingly attractive option. The Antennas on Si substrate are of recent interest [10–13], offering advantages such as compact construction and size reduction compared to a common material like FR-4; hence, they are promising candidates for integration into current and future microwave and mm wave ICs. On the other hand, such RFID antennas have received relatively limited interest in the literature so far.

In this work, we focus on the use of Si and Polyimide for the stack-up section. Our motivation lies in developing novel RFID tag antenna structures on Si, in anticipation of a significant growth in applications for this type of antenna in the near future. An RFID design on a Si-based structure is studied, with a view to improving the impedance matching, bandwidth, and gain performance. The main motivation is to achieve a configurable design of an RFID tag antenna on Si. With the parameterized approach, one can easily change the total area of the design, and consequently the frequency region, by changing only one dimension, and subsequently modifying it as specific needs arise. The coupling between the design lines can be controlled by varying the distance between the lines. We focus on the design of an RFID antenna at 2.4 GHz and 5.8 GHz. These frequencies have been chosen because of the substantial current demand for multi-frequency components in a variety of applications (Wi-Fi, Bluetooth, IoT, Automotive, Smart Home, etc.) that may be expected to further increase in the near future.

Table 1 presents a comparison between the proposed antenna and some recent studies using FR-4, RT–Duroid 5880, jeans, and SiPh as substrates. The present design, based on a Si substrate and polyimide PI2525, appears capable of achieving a combination of relatively high gain and dual-band behavior.

Table 1. Comparison of present results with previous work.

Reference	Frequency (GHz)	Gain (dBi)	Material
[6]	20–25	1.29	SiPh
[14]	5–8.5	3.31	FR-4
[15]	2.4–5.8	2–3.5	FR-4
[16]	0.915–2.45	2.2	FR-4
[17]	2.45	2.29	FR-4
[18]	2.4–5.8	2.47–1.98	Jeans
[19]	2.4–5.8	3.6–3.3	RT–Duroid 5880
[20]	5.8	3.62	FR-4
[21]	5.8	2.84	FR-4
[22]	2.4–5.8	1–3.7	FR-4
Proposed work	2.4–5.8	3.47–4.18 ¹	Si (High resistivity)

¹ Simulated values.

2. Design Procedure and Parametric Study

To provide some background information on the physical working mechanism of the antenna, a pair of equal-width microstrip lines (1, 2) is shown in Figure 1a, with s the spacing between them, ϵ_r the dielectric constant and h the thickness of the substrate material; a typical equivalent circuit is presented in Figure 1b. The characteristic impedance of each line is determined by the self-inductances L_{11} , L_{22} , and the self-capacitances C_{11} , C_{22} . The electromagnetic coupling between the lines, depending on spacing, is described by the mutual inductance L_{12} and capacitance C_{12} .

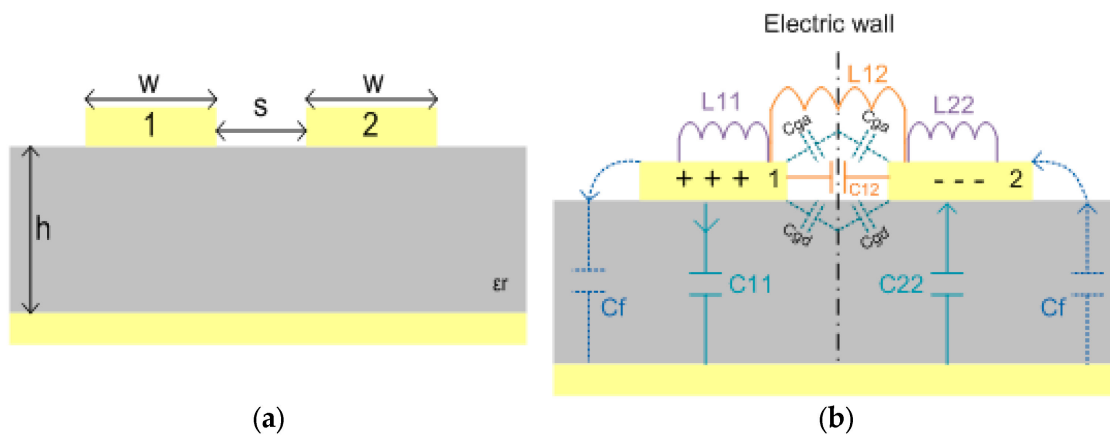


Figure 1. (a) Coupled microstrip lines and (b) equivalent circuit at odd mode and the arrows indicate the fringe capacitance.

According to the stimulation (symmetric/antisymmetric), the line operates in different modes (even/odd), each with a different effective dielectric constant [23,24]. A typical view of the corresponding electric and magnetic fields is depicted in Figure 2. In the odd mode, capacitances of adjacency [23,24] appear due to the edge phenomenon (C_{ga} , C_{gd}), and C_f is the fringe capacitance [24]. The mutual coupling between two lines is primarily attributed to the fields present at the air–dielectric interface and depends on their relative position, as in the case of slot or wire antennas; hence, the spacing (s) is of critical importance in this regard.

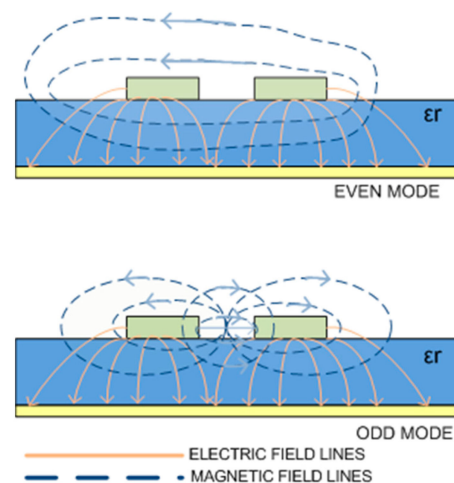


Figure 2. Electric and magnetic fields in coupled lines.

An outline of the design process is shown in Figure 3. According to the rationale discussed in the Introduction, we based the design of the tag antenna on a Si substrate. A rectangular spiral shape was chosen for space saving and compactness. In short, we optimized the process by varying a specific parameter at a time and running the simulation; the procedure was repeated for all design parameters, arriving at a (partially) optimized value set. Then, the whole process was reiterated. At each iteration of the parametric search, the S_{11} and gain parameters, as well as the overall behavior of the radiation pattern in the 2.4 GHz and 5.8 GHz frequencies, were checked.

We started from an initial design with $b_{sub} = 0.32$ mm, copper plate thickness $b_{cop} = 3$ μm , $b_{sp} = 0.018$ mm and $b_{dielectric} = 3$ μm . The total area of the antenna component was 70×78.5 mm². The spacing (s) and width (x) of the spiral lines were 0.9 mm and 2.7 mm, respectively.

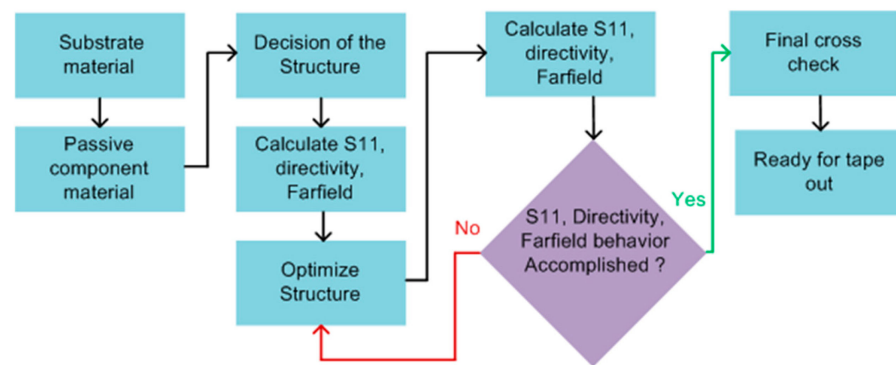


Figure 3. Design methodology flowchart.

A first search was performed by varying x for resonances at 2.4 GHz and 5.8 GHz. Figure 4 shows the results of a series of simulations for S_{11} , where $s = 0.85$ mm and x is varied from 2.3 mm to 2.7 mm. The value of $x = 2.5$ mm yielded the best results.

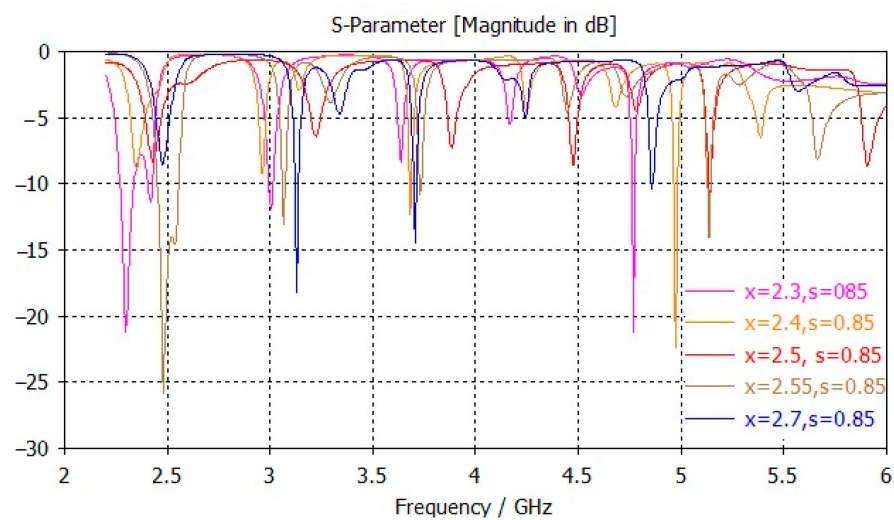


Figure 4. Example of parametric simulation for x .

Further on, a search varying s from 0.8 mm to 0.95 mm was performed, as shown in Figure 5 for the simulated values of S_{11} . The value of 0.85 was found to achieve good tuning close to 5.8 GHz.

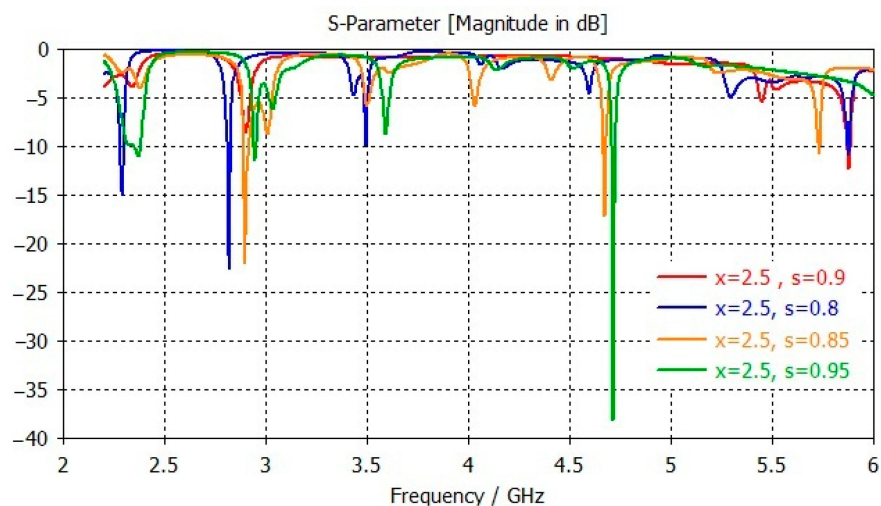


Figure 5. Example of parametric simulation for s .

A subsequent search was carried out for x_0, y_0 , which related to the overall dimension of the antenna, as shown in Figure 6, with x_0 values ranging from 16 mm to 18 mm and y_0 values ranging from 19 mm to 20 mm. For $x_0 = 17.5$ mm, $y_0 = 20$ mm, good behavior is achieved at 2.4 GHz and 5.8 GHz, resulting in the reduced total dimensions of the antenna (an additional benefit).

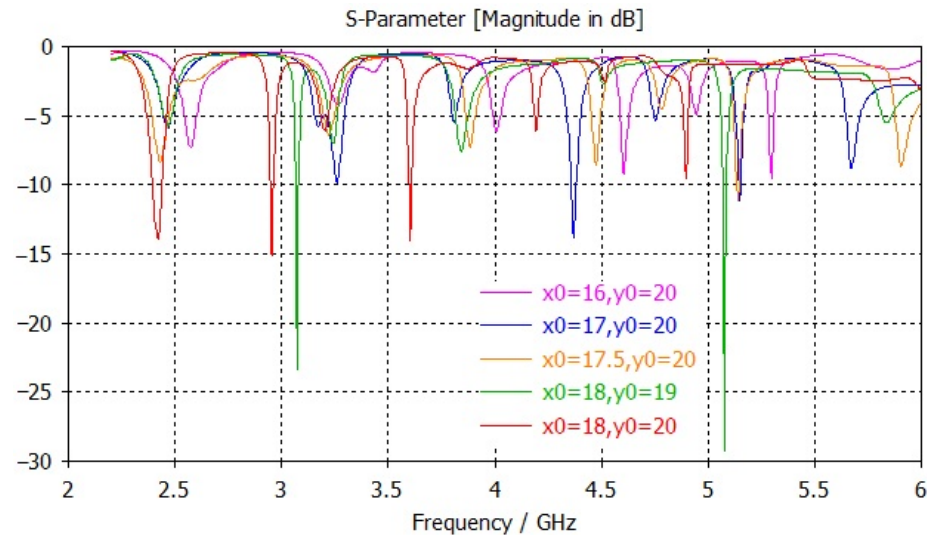


Figure 6. Example of parametric simulation for x_0, y_0 .

In summary, the most important parameters appear to be x (width) and s (spacing). As mentioned above, these values are responsible for the electromagnetic coupling between the lines and have a special role in the design. The x_0, y_0 values are also important, significantly affecting the behavior of the antenna. More specifically, the modification of x_0, y_0 , i.e., the reduction in the overall dimension, results in better tuning at 5.8 GHz, as well as an improved radiation pattern in both bands at 2.4 GHz and 5.8 GHz. Reducing the value of the x -dimension also improves, to some extent, the tuning at 2.4 GHz. Figure 7 shows the initial (a) and final (b) designs after optimization. Table 2 shows the final parameter values in detail.

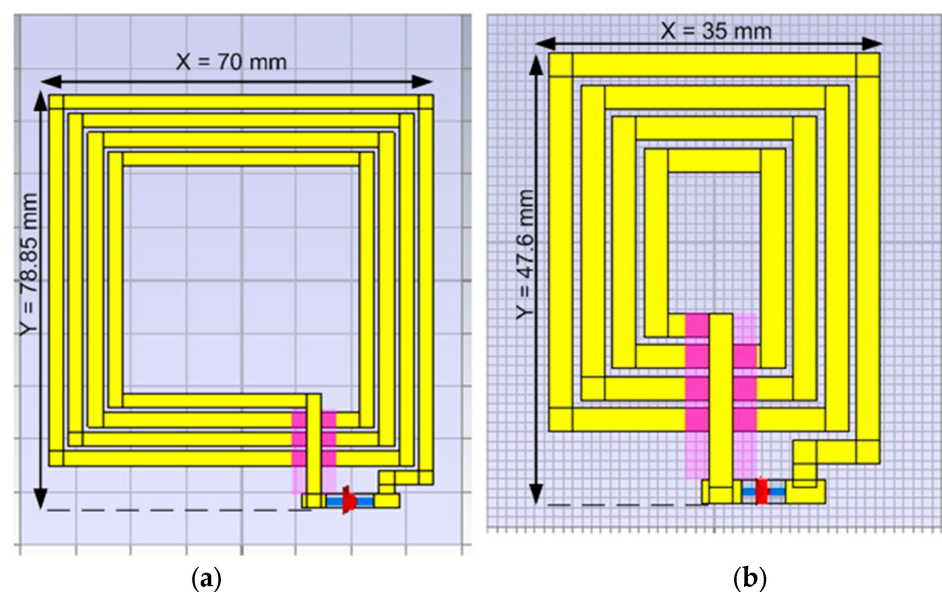


Figure 7. Evolution of RFID tag antenna: (a) initial design and (b) proposed design. Purple denotes Polymide and yellow the copper.

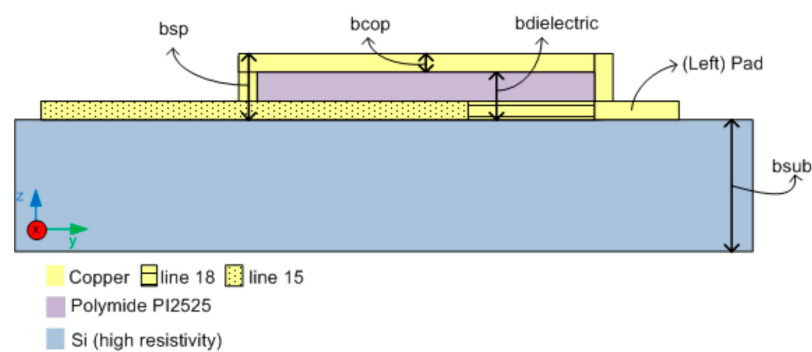
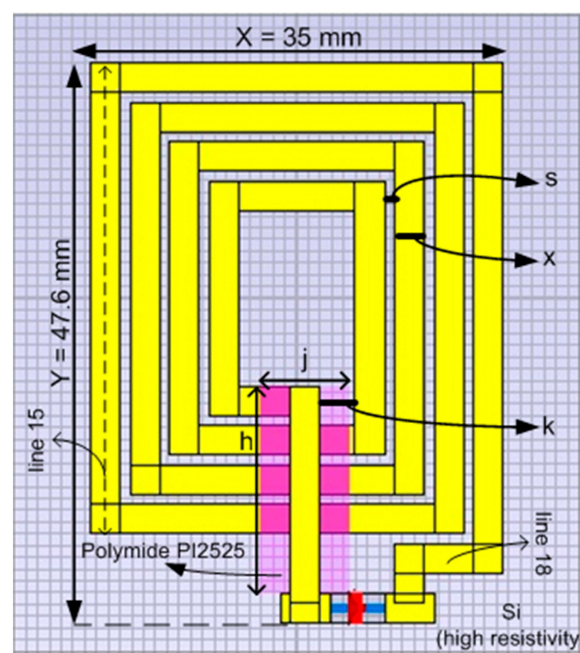
Table 2. Parameters of initial and final designs (proposed here).

Name	Initial Design ¹	Final Design ¹
b_{cop}	0.003	0.003
$b_{\text{dielectric}}$	0.003	0.003
b_{sp}	0.018	0.018
b_{sub}	0.320	0.320
k	7	3
x	0.9	0.85
s	2.7	2.5
x_0	35	17.5
y_0	35	20
Total area (mm ²)	70×78.5	35×47.6

¹ mm.

3. The Antenna Design

Figures 8 and 9 depict the substrate structure (cross-section) and the corresponding antenna layout (top view) of the proposed RFID tag antenna design. A set of values for the design dimensions were selected (Table 2) and used for the simulation and fabrication of the antenna as follows.

**Figure 8.** Substrate with Si, Polyimide, and Cu.**Figure 9.** Antenna design viewed from top layout view.

The substrate material for the design is Si (high resistivity) and Polyimide PI2525 is used for the bridge dielectric. The dielectric constant ϵ_r for high-resistivity Si is 11.68 and that for PI2525 is 2.68. In Figure 8, b_{sub} is 0.320 mm, b_{cop} is 3 μm , b_{sp} is 0.018 mm, $b_{dielectric}$ is 3 μm , and k (the correction factor, see below) is 3 mm. The total area of the antenna component is $35 \times 47.6 \text{ mm}^2$. The spacing (s) and width (x) of the spiral lines, as shown in Figure 9, are 0.85 mm and 2.5 mm, respectively, and h and j are 17.6 mm and 7.5 mm, respectively.

The antenna operation is based on electromagnetic coupling, as described previously. A decrease in the gap between subsequent lines corresponds to a stronger electromagnetic coupling.

For the parameterized design, the following formulas have been used:

$$x_0 + N(i+1)x + M(i-1)s = x_i \quad (1)$$

$$y_0 + N(j+1)x + M(j-1)s = y_j \quad (2)$$

where N is the number of turns and

$$N(i+1) = N + i, \text{ where } i = 0:3 \text{ with step } 1,$$

$$M(i-1) = M - i, \text{ where } i = 1:3 \text{ with step } 1,$$

$$N(j+1) = N + j, \text{ where } j = 0:3 \text{ with step } 1,$$

$$M(j-1) = M - j, \text{ where } j = 1:3 \text{ with step } 1,$$

$$\text{for } x_i, i = 0:3 \text{ with step } 1 \text{ and for } y_j, j = 0:3 \text{ with step } 1.$$

Equations (1) and (2) are the basic conditions for creating the spiral antenna shown in Figure 9. N corresponds to the number of turns, and M corresponds to the number of spaces between the metal paths (In Figure 9, $M = 3$). We note that Equation (1) refers to the x dimension and (2) to the y dimension, with $x_0 = X/2$, $y_0 = Y/2$, respectively, where X and Y are the maximum dimensions. We use (1) and (2) for each metal path. For the first inner metal path, we apply a correction parameter k ($k \geq 3$) when $x_0 y_0 \leq 19$.

$$y_0 = N(j+1)x + M(j+1)s - k \quad (3)$$

Equations (1)–(3) determine the design and lead to the final design of a parameterized antenna through a series of simulations. The substrate thickness b_{sub} along the z axis is also taken into account in the numerical simulations.

The inductance is calculated by the following equations (see, e.g., [24] pp. 95–96):

$$L \text{ (nH)} = 0.03937 \frac{a^2 N^2}{8a + 11c} K_g \quad (4)$$

$$a = \frac{x_0 + x_i}{4} = \frac{x_0 + y_0 - Nx - (N-1)s - k}{4} \quad (5a)$$

$$c = \frac{x_0 - x_i}{2} = \frac{x_0 - y_0 + Nx + (N-1)s + k}{2} \quad (5b)$$

$$K_g = 0.057 - 0.145 \ln(x/b_{sub}) \quad (6)$$

We use these equations, especially (5a) and (5b), for each one of the 19 metal paths. As already mentioned, the values of x (width) and s (space) for the lines, as shown in Figure 9, have been specifically selected for optimization of the S_{11} parameter at the frequencies of interest.

4. Simulation Results

Figure 10 shows the results of the numerical simulation for the reflection coefficient $|S_{11}|$ at the antenna driving point, plotted versus the frequency. Some dips (quasi-resonant frequencies) with $|S_{11}|$ about 7 dB or better are observed in the region between 2.2 and 6 GHz. The dip frequencies, besides 2.4 and 5.9 GHz, are approximately 3.2 GHz, 3.9 GHz, 4.4 GHz, 4.7 GHz, and 5.2 GHz. The dips' position and depth depend on several design

parameters, including s (spacing), x (width) and b_{cop} (thickness of copper). Some simulation results for the frequencies of 2.4 GHz and 5.8 GHz are summarized in Table 3, including the antenna's maximum directivity (3.47 dBi and 4.18 dBi, respectively). Detailed far-field results and polar patterns are shown in Figure 11a,b. For both frequencies of interest, the antenna radiation pattern is quite broad. At 2.4 GHz (Figure 11a), it comprises two almost symmetric wide lobes around the z -axis in the $\varphi = 90^\circ, 270^\circ$ plane. At 5.8 GHz (Figure 11b), a frontal lobe and a somewhat smaller backward lobe are observed around the x -axis.

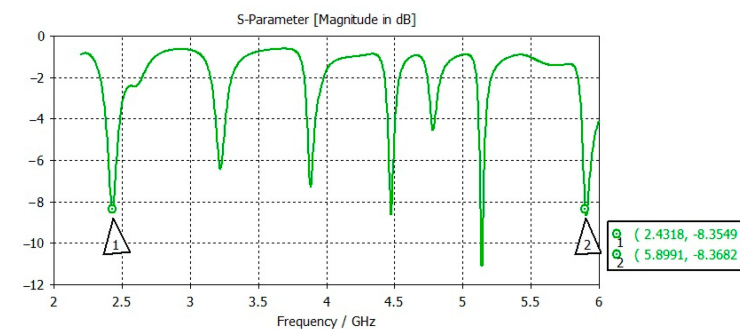


Figure 10. Simulated $|S_{11}|$ values.

Table 3. Simulation results for 2.4 GHz and 5.8 GHz.

Frequency (GHz)	$ S_{11} $ (dB)	Directivity (Max) (dBi)	Directivity at $\theta=0^\circ$ (dBi)
2.4	−8.1	3.47	2.6
5.8	−2	4.18	0.4

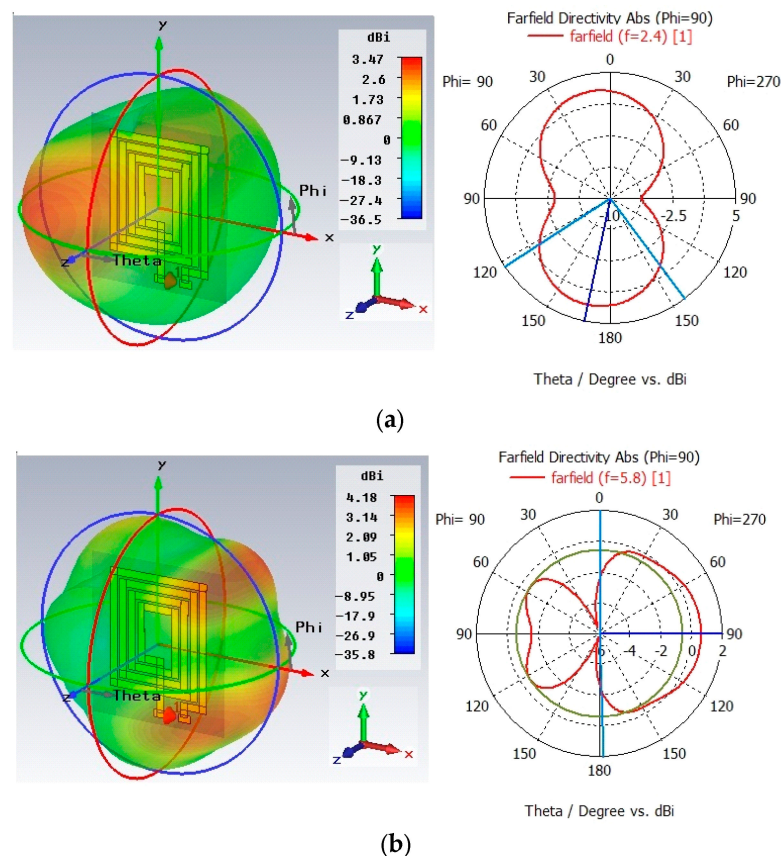


Figure 11. (a) Simulated far-field polar pattern at 2.4 GHz. (b) Simulated far-field and polar pattern at 5.8 GHz.

Snapshots of simulated surface currents at 2.4 GHz and 5.8 GHz are shown in Figure 12a,b, respectively. Also, the coupling effect is visualized in the additional detail view of Figure 12a. In the case of 2.4 GHz (Figure 12a), a maximum surface current value is observed along the two outer lines of the spiral, while on the inner lines, the surface current is up to approximately half the maximum value. A smooth overall current distribution is observed. As shown in Figure 12b, the surface currents for 5.8 GHz are smaller due to a higher operating frequency.

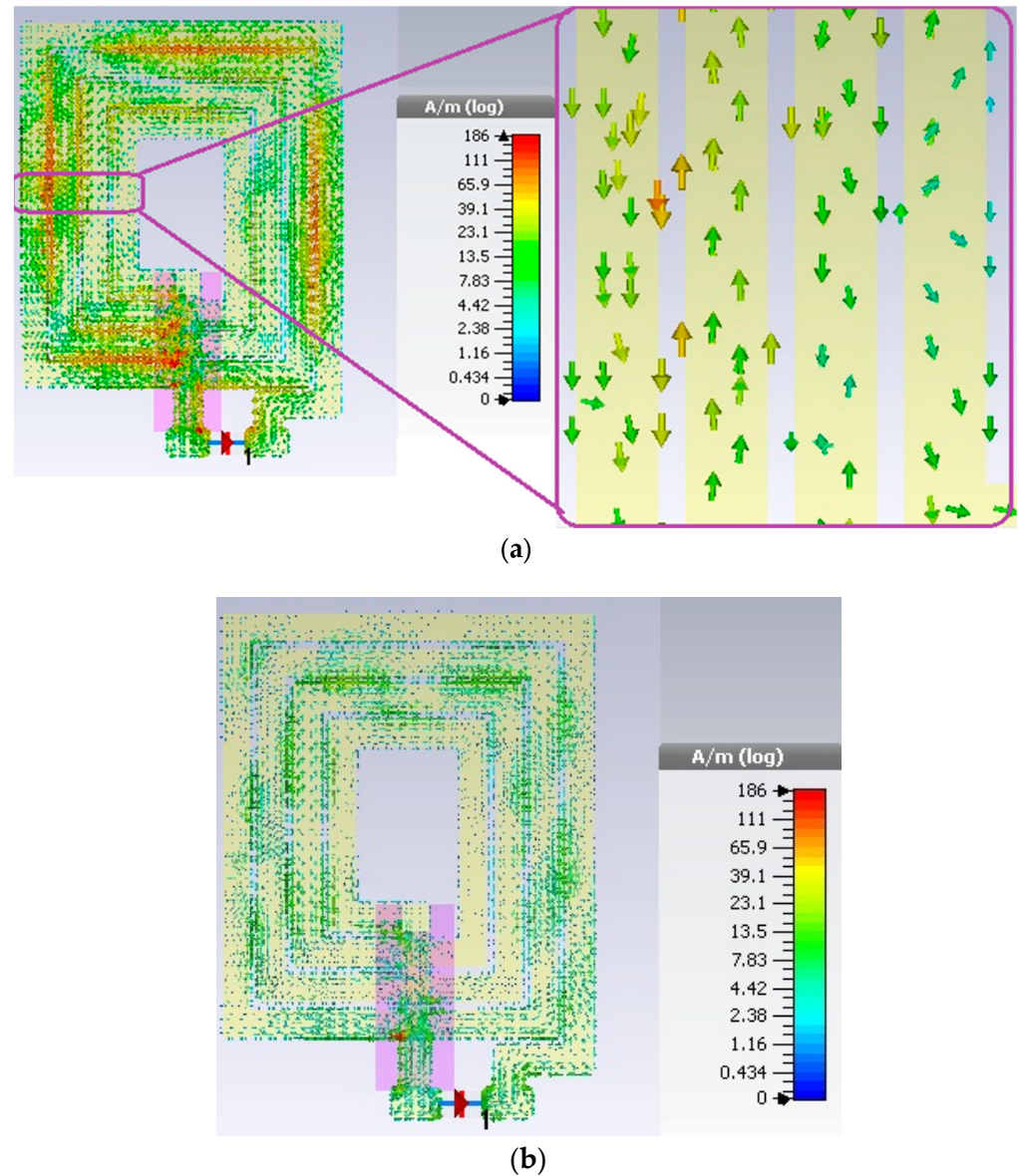


Figure 12. (a) Simulated surface current for 2.4 GHz. (b) Simulated surface current for 5.8 GHz.

5. Fabrication and Testing

5.1. Fabrication

An RFID tag antenna specimen was fabricated at the microelectronics laboratory of the Foundation for Research and Technology–Hellas and tested at the telecommunications laboratory of Hellenic Naval Academy. The microfabrication process involves metallic structure pattern transfer techniques, including optical UV lithography procedures (for the pattern transfer of CAD layout designed patterns), metal depositions using electroplating, and spin-coated stay-on dielectrics.

The procedure begins with the standard degreasing of a Silicon substrate (76 mm/3 inches in diameter), which is a Float-Zone (FZ) Silicon with a wafer thickness of 380 μm , a crystallographic orientation of (100) and, most importantly, a high resistivity of 10,000–1,000,000 $\Omega\cdot\text{cm}$ (like a pure dielectric substrate). The degreasing step is a standard organic solvent cleaning procedure involving the ultrasonic-agitated dipping of the Si substrate in the following order: Acetone, Isopropanol, and finally deionized H_2O . The substrate is dried using nitrogen. Following this step, the AZ9260 photoresist is spin-coated with a thickness of 0.006 mm onto the substrate and thermally stabilized using a hot plate at 110 $^\circ\text{C}$ for 2 min. The transfer pattern is obtained by exposing the layout mask of the bottom inductor spiral using i-line UV lithography with the mask aligner Suss MA6, and developing the exposed photoresist with the use of the AZ826 MIF developer.

The bottom inductor spiral electrode of the RFID is metalized with a 3 μm thick copper (Cu) electroplating using the previous lithography technique. As a dielectric “bridge” support for the bridge interconnect that is intended to connect the center of the spiral coil to the outside electrode, the spin-on polyimide dielectric PI2525 is used. Directly above the polyimide layer, another photoresist layer is spin-coated. The previous optical lithographic technique is once more used to define and transfer the pattern of the dielectric bridge support with the co-development of both the photoresist and the underlying polyimide layer. After the selective resist stripping of the photoresist mask, the polyimide is stabilized by a thermal treatment curing procedure on a programmable hot plate with ramped heating from room temperature up to 350 $^\circ\text{C}$ under an ambient nitrogen atmosphere.

Finally, similar to the first lithographic patterning and Cu electrodeposition process, the bridge Cu interconnect, with a thickness of 5 μm , is created in order to electrically connect the inner core part of the bottom spiral inductor with the outer electrode. The fabricated RFID is presented in Figure 13a,b.

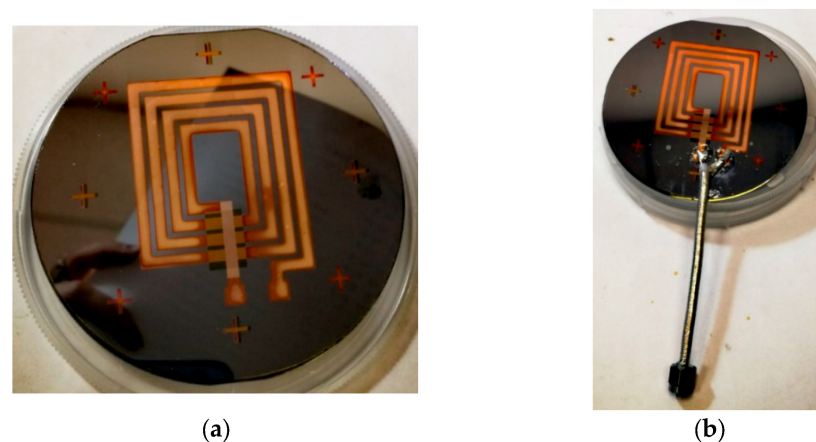


Figure 13. (a) RFID tag specimen fabricated on Si wafer. (b) RFID tag with attached SMA cable.

5.2. Test Setup

The reflection coefficient and gain of the fabricated RFID antenna were measured with a PicoVNA 106 Vector Network Analyzer (VNA) over the instrument’s frequency range of up to 6 GHz. To this end, an SMA connector was attached (as shown in Figure 13b) to the input pads of the RFID antenna via a small piece (10 cm approx.) of 50 ohm RG400 coaxial cable. The SMA input port thus obtained was connected to the VNA port 1 test lead to measure the reflection coefficient $|S_{11}|$. The transmission coefficient $|S_{21}|$ was subsequently measured by connecting a wideband dual-ridged TEM horn antenna (Vector Telecom Part No VT10180DRHA10SK) to the second VNA port test lead (Figure 14a). The horn antenna was rated, according to specifications, at a maximum VSWR of 2.5 and an average gain of 10 dB over the 1–18 GHz frequency range. Calibration of the VNA was carried out by the standard SOLT method at the test leads, using its accompanying mechanical calibration kit (furnished by the manufacturer). The measurement of $|S_{21}|$ was repeated for two identical horn antennas, connected to the same VNA test leads (Figure 14b). In all cases, the antennas

were placed inside an anechoic enclosure with dimensions of about $150 \times 90 \times 60$ cm. For mechanical support, expanded polystyrene was used to minimize the multiple scattering effects inside the enclosure.

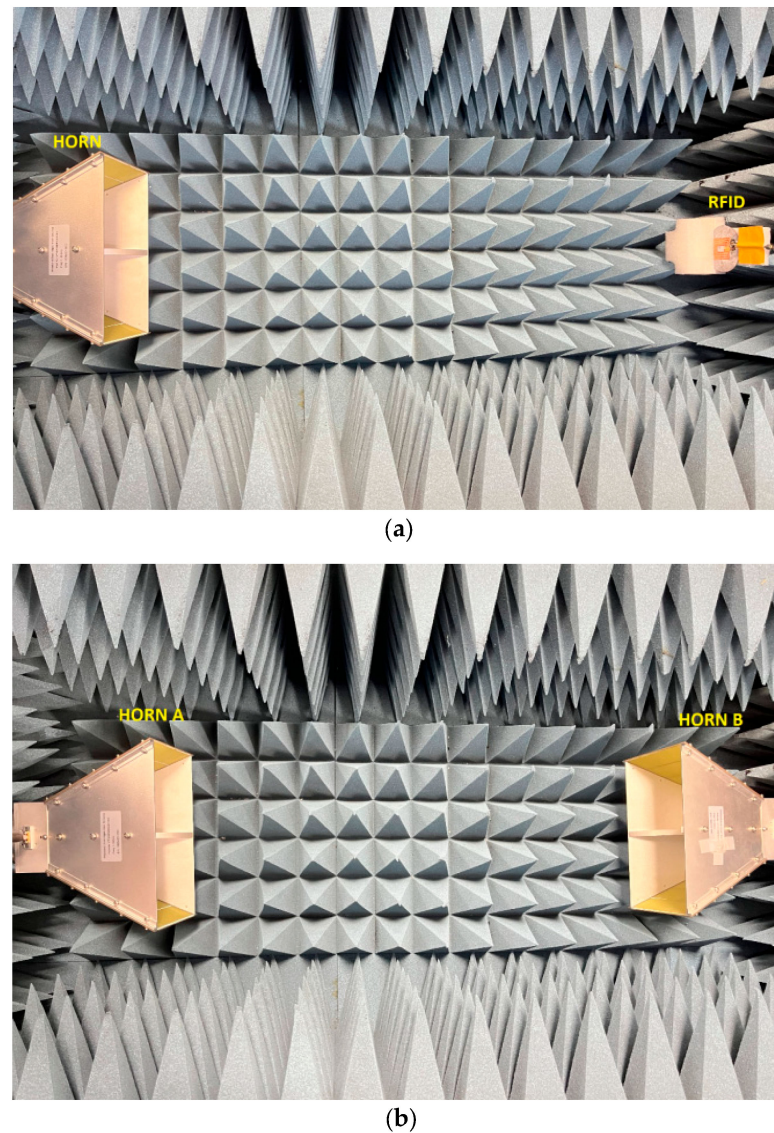


Figure 14. (a) Transmission measurement configuration: horn and RFID tag specimen. (b) Transmission measurement configuration: two identical horns.

Using the well-known Friis transmission equation (see, e.g., [25] pp. 86–88), and taking into account the antenna input mismatch losses, we obtained the following for the gain G_H of the horn antennas and subsequently for the gain G_R of the RFID antenna:

$$|S_{21,H}|^2 = \left(\frac{\lambda}{4\pi R_2} \right)^2 G_H'^2 \Rightarrow G_H' = \left(\frac{4\pi R_2}{\lambda} \right) |S_{21,H}| \quad (7)$$

$$|S_{21,R}|^2 = \left(\frac{\lambda}{4\pi R_1} \right)^2 G_H' G_R' \Rightarrow G_R' = \frac{(4\pi R_1)^2 |S_{21,R}|^2}{\lambda^2 G_H'} \quad (8)$$

where

$$G_H' = (1 - |S_{11,H}|^2) G_H \quad G_R' = (1 - |S_{11,R}|^2) G_R \quad (9)$$

and $|S_{21,H}|$ and $|S_{21,R}|$ are the S-matrix transmission coefficient values measured for the two configurations of Figure 14a,b. The scattering parameters are defined in the standard manner (see, e.g., [26]):

$$S_{ij} = \frac{V_i^-}{V_j^+}$$

where V_j^+ denotes the voltage wave incident on port j , and V_i^- denotes the voltage wave reflected from port i (in the absence of any other incident wave).

Obviously, (7) and (8) represent a simplified version of the well-known three-antenna method, using two identical horns. The validity of the simplification was verified by interchanging the two horns transmitting towards the RFID tag antenna and observing closely matched results for the transmission coefficient. Both G'_H and G'_R incorporate, via (9), the mismatch losses at the test lead ports connecting the horns and the RFID antenna (hence the primed notation). The effect of insertion and mismatch losses along the test leads is compensated by the calibration of the VNA at the test lead ports. Thus, the value of the gain G_R of the RFID antenna is obtained via (9). To compensate for the insertion loss of the small coaxial cable stub attached to the input pads of the RFID antenna, the insertion loss was measured for a cable of the same type and length with SMA connectors at both ends. A value of approximately 0.2 dB was found and added to the results of (9).

To calculate directivity from the measured gain values for comparison with the simulated ones, the radiation efficiency of the RFID tag specimen was measured at frequencies of 2.4 and 5.8 GHz using the classic Wheeler Cap method [27–29]. The antenna under test was placed in a cylindrical box (tin can) of approximately 9.7 cm in diameter and 6 cm in height, covered with a circular aluminum lid with a small opening for the antenna feed coaxial cable. Following the suggestions of [28], care was taken to ensure good electrical contact between the box and the lid, and to place the specimen as near to the center of symmetry of the box as possible. Using the same VNA (PicoVNA 106) with test leads and calibration data as above, the complex reflection coefficient $|S_{11}|$ was measured. The measured value was subsequently transformed by phase rotation along a lossless transmission line, as suggested in [29], to achieve a good approximation of the antenna input impedance via a series equivalent circuit. Taking the real part of the input impedance calculated via $|S_{11}|$ as the loss resistance, the radiation efficiency of the tag specimen is given by the standard Wheeler Cap formula [27]:

$$\eta = \frac{\text{Re}\{Z_i\} - \text{Re}\{Z_w\}}{\text{Re}\{Z_i\}} \quad (10)$$

where Z_i and Z_w are the input impedances without and with the cap, respectively. It was verified that $\text{Re}\{Z_i\} > \text{Re}\{Z_w\}$ at both frequencies of interest, i.e., the measured values were consistent with the underlying assumption of a series equivalent circuit for the antenna under test.

5.3. Testing Results

The measured $|S_{11}|$ values for the RFID tag are shown in Figure 15, bearing a close similarity with the simulated values of Figure 10. A small shift in the dip frequencies is observed, possibly due to manufacturing discrepancies and the impact of the attached cable and connector on the antenna load. Interestingly, the measured results seem to exhibit an improved behavior (i.e., better matching) in the dip regions, which may be attributed to the additional loss caused by specimen preparation (e.g., contact and soldering losses) and/or some reactance introduced by the connector used.

The measured results for the gain G_R of the RFID antenna are shown in Figure 16. In all, they are several dB lower than the simulated directivity values, especially in the region of 5.8 GHz. The difference is probably due to significant resistive losses over the RFID antenna structure. Such losses are to be expected in printed antennas to a greater extent than wire antennas, while larger losses at higher frequencies are usually due to the skin effect. As a check, the radiation efficiency measurements (described above) at

2.4 and 5.8 GHz yield efficiency values of $\eta = 0.85$ and 0.6, respectively, corresponding to differences of 0.7 and 2.2 dB between the directivity and gain values at these frequencies.

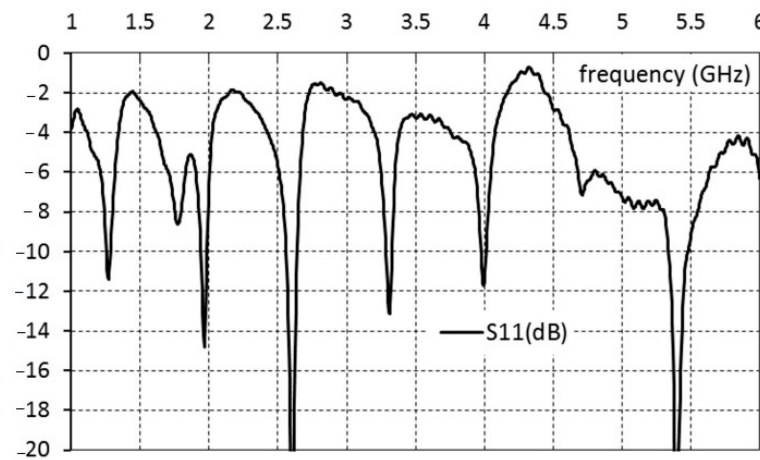


Figure 15. Measured $|S_{11}|$ values for the RFID tag antenna.

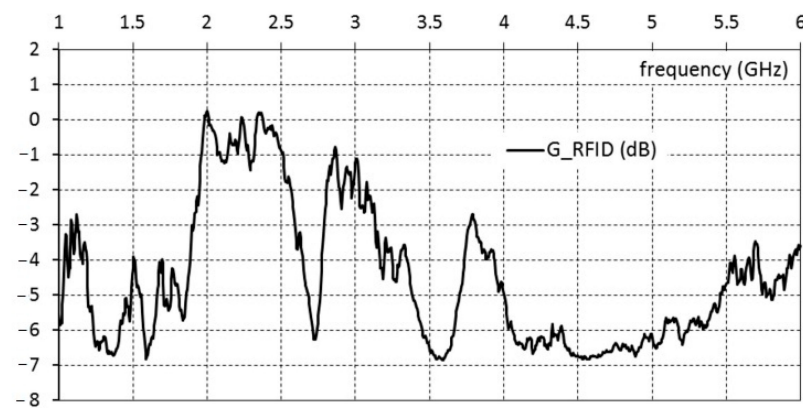


Figure 16. Measured gain G_R of the RFID antenna for frequencies of 1 to 6 GHz, with the antenna oriented toward the positive z direction ($\theta = 0^\circ$), as defined in Figure 11a,b.

6. Discussion and Conclusions

Upon comparison of the simulated and measured results for the $|S_{11}|$, as shown in Figures 10 and 15, it can be seen that the measured dip frequencies are close to the simulated ones, with the exception of the vicinity of 5.8 GHz where no dip is found, although the measured $|S_{11}|$ value remains below 4 dB. On the other hand, the dip near 2.4 GHz is displaced to the right compared with the simulated results, although the measured $|S_{11}|$ value at 2.4 GHz is about 3.6 dB. In future practical applications, a transmission line segment could be fabricated on the same substrate between the antenna pads and the input terminal, adjusting the parameters s and x (Figure 9) and possibly including matching components or stubs as appropriate to further improve reflection losses. The measured gain (Figure 16) appears to maintain considerable stability at relatively high values around both frequencies of interest (2.4 and 5.8 GHz), where a large beamwidth around the broadside direction has also been found (Figure 11a,b). As already noted, the measured gain values indicate significant losses on the antenna, which appear to be confirmed by the efficiency measurement obtained using the Wheeler cap technique. In all, the simulated and measured values were in good agreement, demonstrating the promising characteristics of the proposed RFID design on Si for a variety of applications.

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