



Article

Dual-Band Antenna at 28 and 38 GHz Using Internal Stubs and Slot Perturbations

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Abstract: A double-stub matching technique is used to design a dual-band monopole antenna at 28 and 38 GHz. The transmission line stubs represent the matching elements. The first matching network comprises series capacitive and inductive stubs, causing impedance matching at the 28 GHz band with a wide bandwidth. On the other hand, the second matching network has two shunt inductive stubs, generating resonance at 38 GHz. A Smith chart is utilized to predict the stub lengths. While incorporating their dimensions physically, some of the stub lengths are fine-tuned. The proposed antenna is compact with a profile of $0.75\lambda_1 \times 0.66\lambda_1$ (where λ_1 is the free-space wavelength at 28 GHz). The measured bandwidths are 27–28.75 GHz and 36.20–42.43 GHz. Although the physical series capacitance of the first matching network is a slot in the ground plane, the antenna is able to achieve a good gain of 7 dBi in both bands. The proposed antenna has a compact design, good bandwidth and gain, making it a candidate for 5G wireless applications.

Keywords: 5G; double-stub matching; impedance matching; matching network; millimeter wave antenna; Smith chart



Citation: Bhadravathi Ghouse, P.S.; Kumar, P.; Mane, P.R.; Pathan, S.; Ali, T.; Boullogeorgos, A.-A.A.; Anguera, J. Dual-Band Antenna at 28 and 38 GHz Using Internal Stubs and Slot Perturbations. *Technologies* **2024**, *12*, 84. <https://doi.org/10.3390/technologies12060084>

Academic Editor: Simone Genovesi

Received: 24 April 2024

Revised: 25 May 2024

Accepted: 3 June 2024

Published: 6 June 2024



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1. Introduction

Spectrum congestion in wireless applications is addressed by choosing higher carrier frequency or spectral efficiency (modulation) techniques. From the perspective of higher carriers, the millimeter wave (mmWave) spectrum (30–300 GHz) is trending in almost all communication fields to fulfill the demand for broadband support with low latency. At such bands, the antenna profile decreases, consequently accommodating multiple antennas through which spatial multiplexing can also be achieved. However, some of these bands, on the other hand, experience increased atmospheric attenuation and path loss. The path loss at mmWave is extremely high, being on the order of 100 dB for a distance of 50 m. Also, path loss depends on the surrounding environmental objects [1,2]. The mmWave frequency spectrum (24–100 GHz) was brought under the fifth generation (5G) category by the World Radiocommunication Conference (WRC) in 2019. The applications utilizing this spectrum expect a minimum bandwidth support of 400 to 800 MHz with a data rate of 1–20 Gbps through various modulation techniques [3,4]. In this spectrum, the 28 and 38 GHz bands are licensed for mobile and satellite applications as they are more suitable for long-range

communication. Thus, the antenna must have good sensitivity and selectivity to work efficiently in these bands.

Various antenna structures designed for mmWave operations exist, such as pillbox-distributed [5], substrate-integrated-waveguide [6], magnetoelectric dipole [7], ridge-waveguide [8], and leaky-wave antennas [9]. These antenna structures are complex, and they are challenging to embed in compact handheld wireless devices due to large antenna profiles. The solution is a planar microstrip antenna that is easy to embed and cost-efficient.

Therefore, our article focuses on designing and developing a planar antenna to cover 28 and 38 GHz bands with adequate bandwidth. The major limitation of a planar or microstrip antenna is its narrow bandwidth. The bandwidth has a direct relationship with the substrate thickness (t) and an inverse relationship with the square root of the relative permittivity ($\sqrt{\epsilon_r}$) [10]. It can be improved by introducing slots in the radiator [11], adding parasitic elements near the radiator [12,13], metamaterial structures [14], and a defected ground [15]. Any modification to the radiator through slots, slits, or the addition of parasitic elements perturbs the current flow, causing good impedance matching at the adjacent frequencies to that of the resonance band, thus enhancing the bandwidth. For example, in [12], a monopole antenna with circular patch termination generates multiple resonances with narrow bandwidth. Adding stub and parasitic elements near the radiator caused good impedance matching at the adjacent frequencies. As a result, a bandwidth ranging from 24 GHz to 40 GHz occurred. However, such wideband antennas are susceptible to noise interference in specific applications when the spectrum is shared. The solution is to have an antenna with high selectivity that serves at both the 28 and 38 GHz frequencies.

Impedance matching at these high frequencies is crucial to maximizing antenna radiation efficiency. Matching networks, such as a π -structure with tunable components [16], capacitance matrix representation in series/parallel [17], and L-type networks [18], are used in high-frequency RF applications. Stub loading is another impedance-matching technique that can be used in antennas to generate multi-band resonances [19].

In one of the recent designs in [20], the conventional circular patch is modified to generate dual resonances. The dual resonances are obtained by loading a pair of stubs and a pair of slots into the crescent-shaped radiator. However, the stubs and slots are arbitrary and are tuned from the parametric method. In [21], an inverted U-shaped slot is introduced in the conventional patch, causing dual-band resonance at 28 and 38 GHz.

Another design in [22] has quad slits (in the corner) and slots in the rectangular patch, achieving dual-resonance. However, the bandwidth at the first band (28 GHz) is narrow, and the antenna profile is large at $1.95\lambda_0 \times 2.47\lambda_0$ (at 28 GHz). Likewise, in [23] as well, a combination of slits and slots are curved on the radiator to obtain dual resonance. This structure also results in a narrow bandwidth. Since the ground plane in these designs is not disturbed, the radiation is mostly in the broadside, with good gain; however, it suffers from a narrow bandwidth.

The above limitation of narrow bandwidth is addressed in [24] by adopting a partial ground technique. The partial ground at the bottom and stub loading on the radiator generates a dual-band resonance with decent bandwidth. Similar to the previous design, [25] also incorporates a partial ground for bandwidth enhancement. The radiator combines a T-shaped monopole with an arc structure, leading to dual-band generation. In [26], a combined technique of a partial ground and slotted radiator is employed to achieve dual-band resonances. A similar concept is applied in [27] with a change in the structural design. In [28], a combination of a partial ground and a double-stub matching technique is used to achieve dual-band resonance. However, these resonances are separated by a large frequency gap.

An oval-shaped structure naturally has dual-resonance [29]. A defected ground can be used to tune the desired resonance. Although the above designs improved bandwidth through the partial ground technique, the antenna gain is reduced.

Therefore, from such a literature study, it is evident that numerous planar antennas have incorporated arbitrary stubs and slots through parametric analysis to generate dual-

resonance. However, a proof of concept of these stubs and slots behavior is unclear in the literature. Thus, here, we emphasize the characterization of the stubs and slots behavior with the matching network, through which dual-resonances at 28 and 38 GHz are achieved.

This article demonstrates the implications of the double-stub matching technique in antenna design by incorporating dual matching networks. The theoretical interpretation of the matching network is presented by the transmission line concept and the aid of the Smith chart. The embedded slots and stubs for matching are theoretically analyzed, providing dimensions in close agreement with the physical dimensions of the proposed antenna structure. However, these analytical results are further fine-tuned to achieve the desired performance. The realized antenna is compact, with a dimension of $0.75\lambda_1 \times 0.66\lambda_1$ (where λ_1 is wavelength at 28 GHz), achieving dualband at 27–28.75 GHz and 36.20–42.43 GHz. The antenna has maximum radiation on the broadside, with a realized gain of 7.2 dBi and a 90/95% total efficiency in the respective bands.

2. Design Methodology

A monopole antenna is etched on the Rogers 5880 substrate, which has a thickness of 0.254 mm. The monopole's circumference length is close to one wavelength (ML1) at 28 GHz but curved to reduce the overall antenna size, as shown in Figure 1. At the feed point, the antenna impedance is $2.1 - 4.21j$ and $12 + 23.97j \Omega$ at 28 and 38 GHz, respectively. Thus, it is fully mismatched with the 50Ω feeding line. A double-stub matching is proposed based on the transmission line theory.

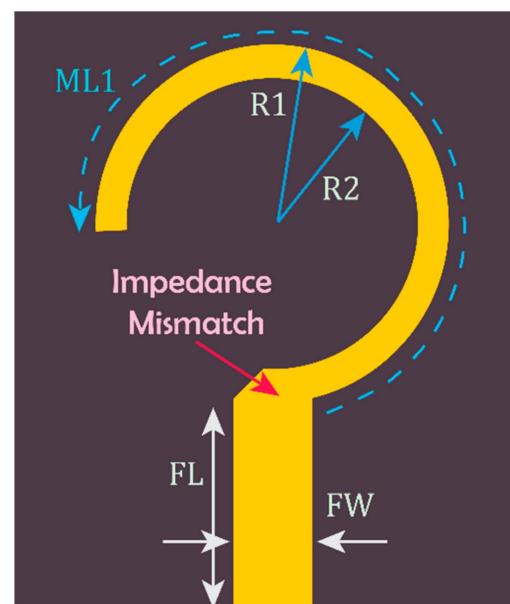


Figure 1. Bent monopole antenna. The dimensions in mm are as follows: FL = 2.9, FW = 1, R1 = 2.4, R2 = 2.

2.1. Matching Network (MN1) at 28 GHz (f_1)

At 28 GHz (λ_1), the designed matching network (MN1) utilizes a series of capacitive and inductive stubs to cancel the load reactance and match to the 50Ω impedance port, as displayed in Figure 2. The computation of these distributed elements through transmission line theory is calculated using a Smith chart [30]. A Smith chart is a convenient tool that efficiently computes the matching network elements. The double-stub matching network method using the Smith chart is explained in detail in [31].

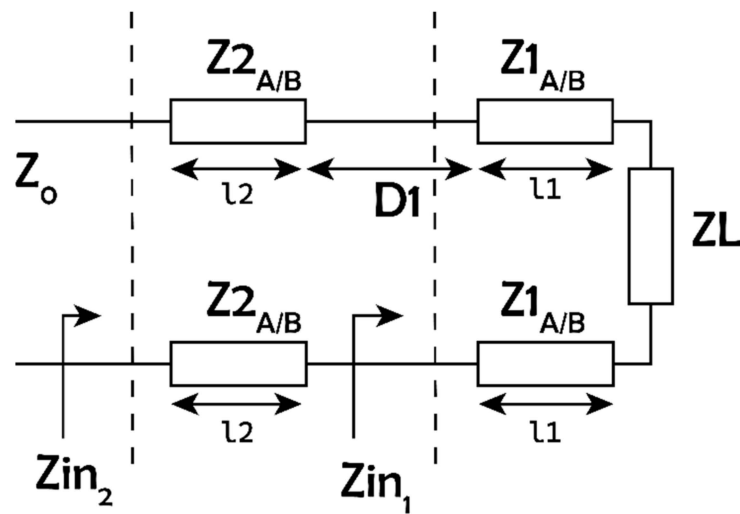


Figure 2. Equivalent circuit of distributed elements implementing double-stub matching technique at 28 GHz.

For analysis purposes, the length of the monopole antenna is calculated by considering the circumference of the outer ring (R1). Thus, the length of the monopole antenna (ML1) is given by $\frac{3}{4}(2 \times \pi \times R1)$, which is approximately equal to λ_1 . The reason for choosing a monopole length equal to λ_1 because the monopole, in our case, behaves as a transmission line for which the internal stubs are added at the distances of D1 and D2. The first stub of MN1 is the series capacitance, with its impedance $Z1_{A/B}$, directly connected to the load ZL. The second stub is a series inductance, with its impedance $Z2_{A/B}$, connected at a distance $D1 = \frac{1}{4}(2 \times \pi \times R1) = 0.35\lambda_1$ from the capacitive stub.

With these conditions, the matching network is computed at 28 GHz: First, the load impedance ZL is normalized to the port impedance ($Z_0 = 50 \Omega$), resulting in $0.042 - 0.082j$, marked as ZL^1 in Figure 3. As the $Z1_{A/B}$ (Stub-1) is at the load, the $1 + xj$ circle is shifted towards the load by a distance of D1. Here, the Stub-1 is added, changing the impedance reactance. Further, moving from ZL^1 on a circle of constant resistance, the line intersects with the two points of the shifted unit circle, which are marked as Z1 and $Z1^1$. The impedance at these points are:

$$Z1 = 0.042 - 1j \quad (1a)$$

$$Z1^1 = 0.042 - 0.482j \quad (1b)$$

When comparing the above impedance with that of the ZL^1 impedance, a change is observed in the reactance. Therefore, the location of Stub-1 is computed by subtracting the impedance of ZL^1 from Z1 and $Z1^1$. This leads to two solutions: $Z1_A$ and $Z1_B$.

$$Z1_A = Z1 - ZL = -0.918j \quad (2a)$$

$$Z1_B = Z1^1 - ZL = -0.4j \quad (2b)$$

The respective admittances are:

$$Y1_A = \frac{1}{Z1_A} = 1.089j = bj \quad (3a)$$

$$Y1_B = \frac{1}{Z1_B} = 2.5j = bj \quad (3b)$$

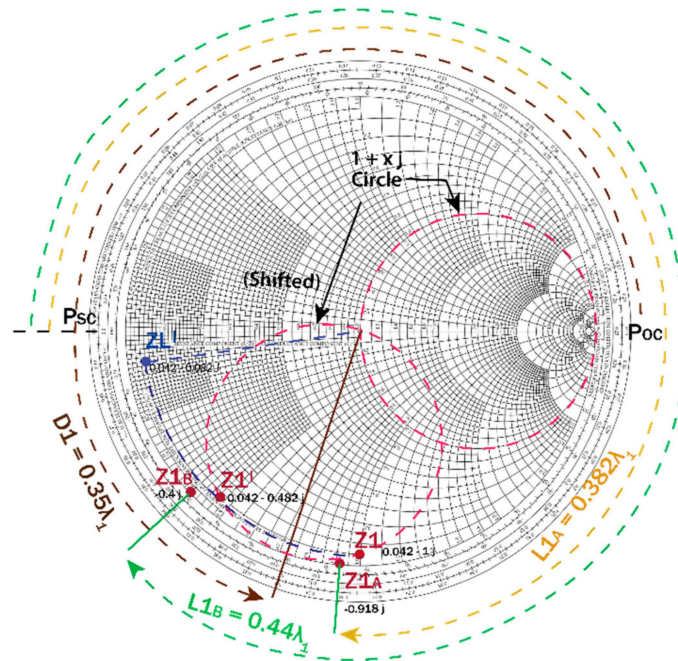


Figure 3. Stub-1 length prediction at 28 GHz using a Smith chart.

The Stub-1 length is calculated from the short-circuit impedance (P_{SC}) point to $Z1_A$ and $Z1_B$, which results in two stub lengths.

$$L1_A = 0.382 \lambda_1 \text{ at } Z1_A \quad (4a)$$

$$L1_B = 0.44 \lambda_1 \text{ at } Z1_B \quad (4b)$$

The equivalent capacitance for these stub lengths is computed using Equation (5):

$$C = \frac{b}{2 \times \pi \times f_1 \times Z_0} \quad (5a)$$

$$C1_A = 0.123 \text{ pF} \quad (5b)$$

$$C1_B = 0.284 \text{ pF} \quad (5c)$$

We choose the solution of $Z1_B$ of length $L1_B$. In our design, Stub-1 is achieved by etching a circular slot in the ground plane with a radius of G_s , as shown in Figure 4a,b. The circumference length of the slots is matched with the length of $L1_B$. As a distributed element, the stub impedance does not change when the length changes by multiples of half of the wavelength (λ_1).

As a result, the circumference of Stub-1 is tuned such that its length is twice the $L1_B$. The capacitance value of the designed slot is calculated from Equation (6) as defined in [32]. The slot capacitance is thrice the capacitance of the theoretical value.

$$C_c = \frac{\sqrt{\epsilon_{\text{reff}}} L1_{B(\text{Physical})}}{v_0 Z_0} \cong 3(C1_B) \quad (6)$$

where ϵ_{reff} is the relative permittivity of the substrate and v_0 is the wave velocity in free space.

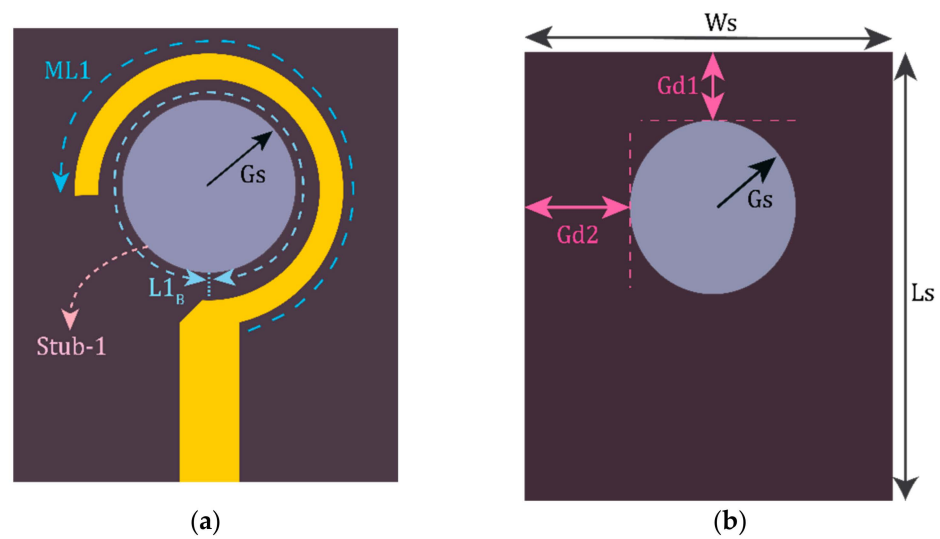


Figure 4. Monopole antenna with Stub-1 as a slot in the ground plane. (a) Top view, and (b) Bottom view. Dimensions in mm are: $G_s = 1.6$, $G_{d1} = 1.1$, $G_{d2} = 1.9$, $W_s = 7$, $L_s = 8$.

With Stub-1, the impedance is slightly improved, as shown in Figure 5. However, the actual matching occurs when Stub-2 is added.

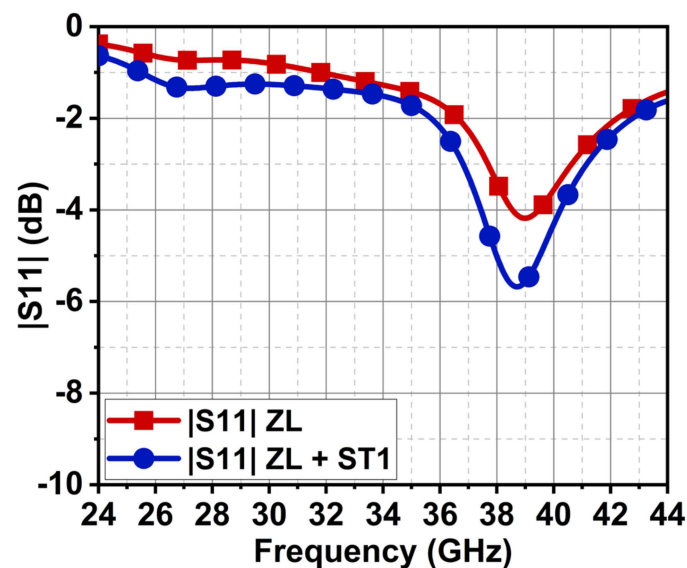


Figure 5. Comparison of the reflection coefficient of load Z_L , with and without Stub-1.

For the Stub-2 ($Z_{2A/B}$) calculation, the Z_1 and Z_1^1 are considered as new loads. First, considering the Z_1 as the load with its impedance of $0.042 - 1j$, a reflection coefficient circle (Γ_1) is drawn. Moving from Z_1 towards the generator on the Smith chart by a distance D_1 , it reaches $Z_2 = 1 + 10j$, as displayed in Figure 6a. To cancel this reactance, the Stub-2 reactance must be $-10j = xj$. Therefore, a new point is marked at $-10j$ as Z_{2A} . The Stub-2 length L_{2A} , for the new load Z_1 is calculated from the short-circuit impedance (P_{SC}) towards the generator till Z_{2A} . It resulted in $L_{2A} = 0.276\lambda_1$. Another value of the Stub-2 length is obtained by following a similar process by considering Z_1^1 as a new load, as shown in Figure 6b. Moving a distance of D_1 from Z_1^1 towards the generator reaches point Z_2^1 , having an impedance of $1 - 4.85j$. Thus, to cancel this capacitance, a series reactance

of $+4.85 j(Z_{2B})$ should be added. Thus, a length of $L_{2B} = 0.218\lambda_1$ from P_{SC} is needed. Further, from these theoretical values, the physical length of Stub-2 is calculated as:

$$L_{2A} = 0.276\lambda_1 \quad (7a)$$

$$L_{2B} = 0.218\lambda_1 \quad (7b)$$

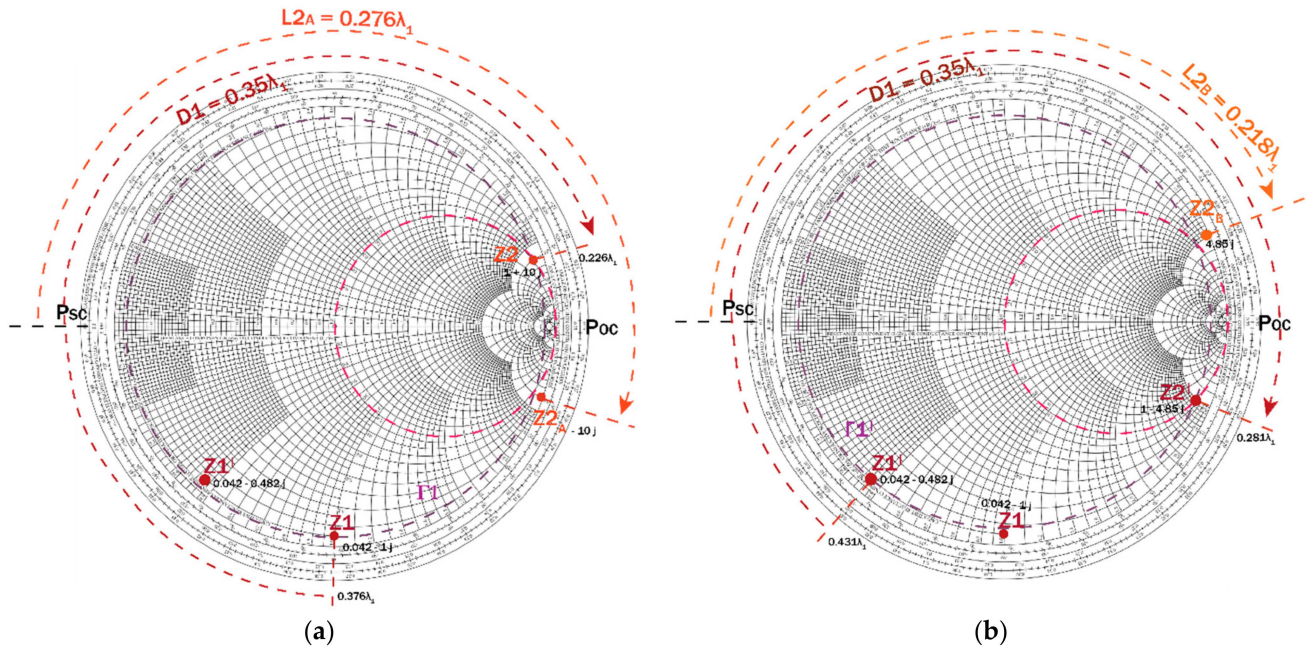


Figure 6. Stub-2 length of MN1 at 28 GHz using the Smith chart with (a) Z_1 as a load and (b) Z_1^1 as a load.

The inductance is calculated with Equation (8) as:

$$L = \frac{xZ_0}{2\pi f_{1/2}} \quad (8a)$$

$$L_{2AL} = 2.84 \text{ nH} \quad (8b)$$

$$L_{2BL} = 1.37 \text{ nH} \quad (8c)$$

The above analysis suggests that the Stub-2 length is either $L_{2A} = 2.995 \text{ mm}$ or $L_{2B} = 2.33 \text{ mm}$. Thus, the short stub is chosen and further tuned to achieve resonance at 28 GHz. Stub-2 is positioned at a distance of D_1 with a tuned length of L_{2B} . The antenna with an impedance-matching network at 28 GHz is shown in Figure 7. Figure 8 illustrates the improvement of the reflection coefficient due to the matching network at 28 GHz. The results indicate that the matching network has performed good impedance matching between the load impedance Z_L and port impedance Z_0 at 28 GHz.

2.2. Matching Network (MN2) at 38 GHz (f_2)

After adding the first matching network (MN1), its impedance is Z_{in2} when looked toward the load, as depicted in Figure 2. Ideally, it must have been purely resistance of 50Ω . However, after tuning, it resulted in $50 - 5.5 j \Omega$ at 28 GHz. Whereas at 38 GHz, it is $26.73 + 32.67 j \Omega$, as shown in Figure 9. The next task is to design the second matching network (MN2) at 38 GHz. To begin with, again, a double-stub matching technique is

adopted. Nevertheless, in this case, the proposed matching network (MN2) is a pair of stubs with a shunt inductance, as shown with an equivalent transmission line circuit in Figure 10. The first shunt inductance is connected at the load (Z_{in2}). The second shunt inductance is at the distance of $D2 = 0.95\lambda_2$ (where λ_2 is the wavelength at 38 GHz) from the Z_{in2} towards the generator. In this case, as the impedance is connected in a shunt, working with admittance is more convenient mathematically. Therefore, a normalized load admittance (Y_{in2}^l) is used as:

$$Y_{in2}^l = \frac{1}{Z_{in2}} = 0.76 - 0.94j \quad (9)$$

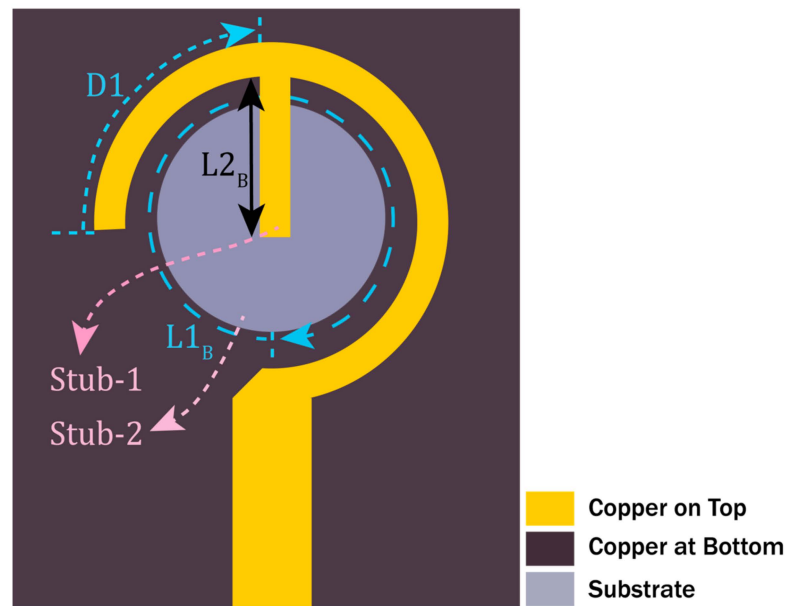


Figure 7. Antenna with the matching network for 28 GHz. Stub-1 length $L2_B$ is tuned to 2.25 mm with a thickness of 0.4 mm.

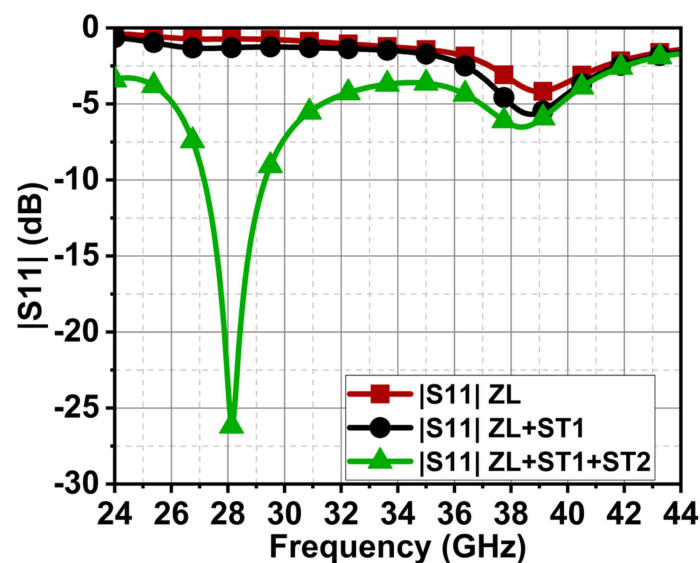


Figure 8. Reflection coefficient due to the first matching network (MN1).

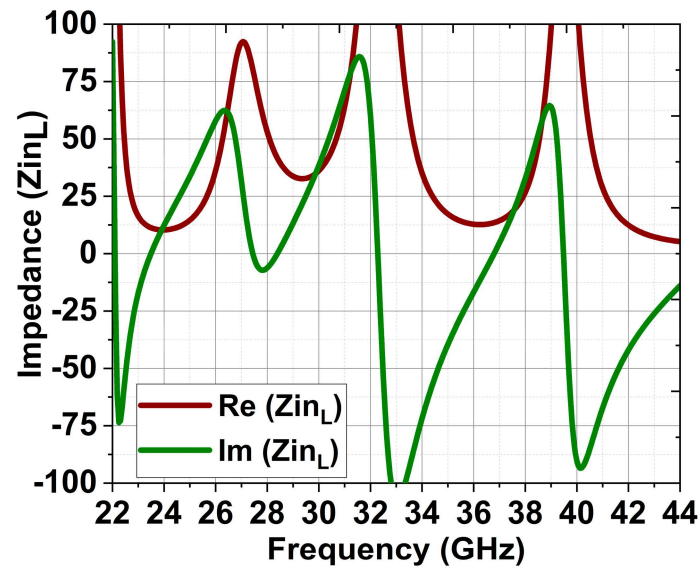


Figure 9. Impedance Z_{inL} after adding the first matching network (MN1).

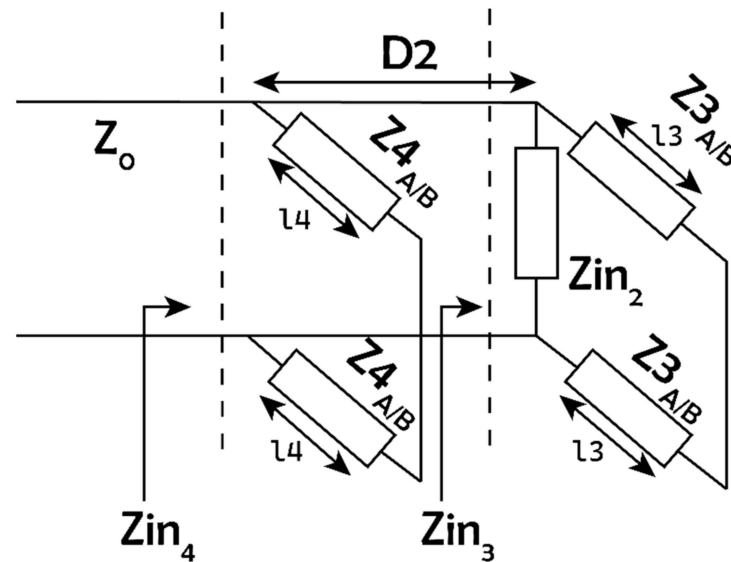


Figure 10. Equivalent circuit of distributed elements of the second matching network (MN2) applying the double-stub matching technique at 38 GHz.

The Stub-3 ($Z_{3A/B}$) analysis requires first reaching the unit circle $1 + xj$. This is achieved by a shift in distance of $D2 = 0.95\lambda_2$, towards the generator, or $0.45\lambda_2$ on the Smith chart. Moving from load admittance (Y_{in2}^l) along the conductance path, the change is observed only in the susceptance. Further, it intersects with two points of the shifted unit circle, which are marked as $Y3$ and $Y3^l$. The admittance at these points are $0.76 - 0.36j$ and $0.76 - 8j$, as demonstrated in Figure 11. With the two intersection points, two solutions for Stub-3 are possible by subtracting the load admittance from $Y3$ and $Y3^l$.

$$Y_{3A} = Y3 - Y_{in2} = 0.58j \quad (10a)$$

$$Y_{3B} = Y3^l - Y_{in2} = -7.06j \quad (10b)$$

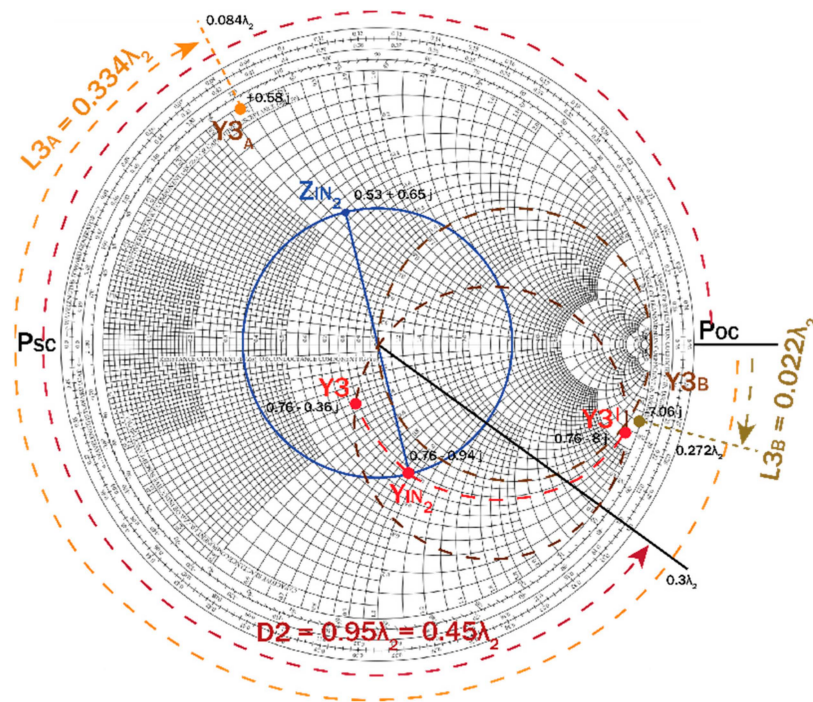


Figure 11. Stub-3 admittance of MN2 at 38 GHz using Smith chart.

The respective impedances are:

$$Z_{3A} = \frac{1}{Y_{3A}} = -1.72 j \quad (11a)$$

$$Z_{3B} = \frac{1}{Y_{3B}} = +0.14 j \quad (11b)$$

The first length of Stub-3 due to Y_{3A} from the shorted admittance (P_{OC}) moving towards the generator is $L_{3A} = 0.334\lambda_2$. The second possible Stub-3 length due to Y_{3B} is $L_{3B} = 0.022\lambda_2$. The physical lengths for Stub-3 are:

$$L_{3A} = 0.334\lambda_2 \quad (12a)$$

$$L_{3B} = 0.022\lambda_2 \quad (12b)$$

The respective inductance values of L_{3A} and L_{3B} are calculated using Equation (8), considering Stub-3 reactances, that is Z_{3A} and Z_{3B} , resulting in:

$$L_{3AL} = 3.6 \text{ nH} \quad (13a)$$

$$L_{3BL} = 0.29 \text{ nH} \quad (13b)$$

L_{3A} is quite long and overlaps with L_{1B} . On the other hand, the L_{3B} is too small, which may cause fabrication errors. Thus, L_{3A} is reduced by $\lambda_2/2$, as shown in Figure 12.

As the double-stub matching technique is chosen for the second matching network (MN2), with only Stub-3, the impedance matching cannot be achieved at 38 GHz. However, a Stub-3 of MN2 generated multiple resonances at 34 and 40 GHz without disturbing the resonance at 28 GHz, as shown in Figure 13. Though multiple resonances are generated, the objective of the design is strictly to achieve dual resonance at 28 and 38 GHz. Therefore, computational analysis for the fourth stub (Stub-4) continues.

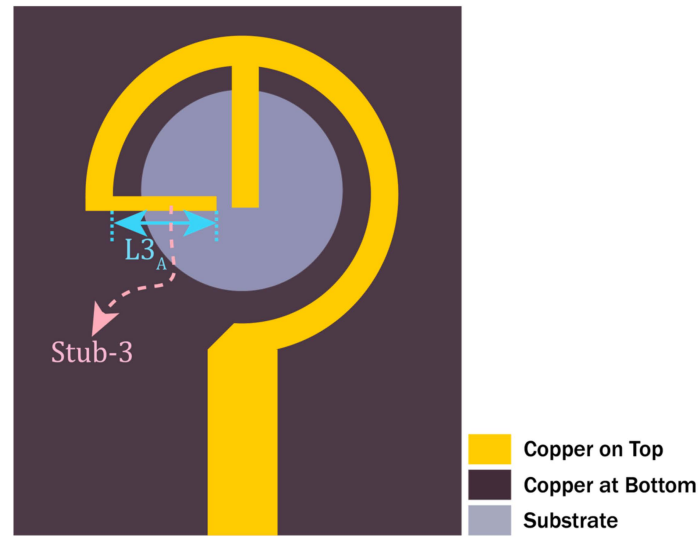


Figure 12. Antenna after adding Stub-3 to improve matching at 38 GHz. The actual length of L_{3A} is half of the calculated value, which is further tuned to 1.6 mm with a width of 0.2 mm.

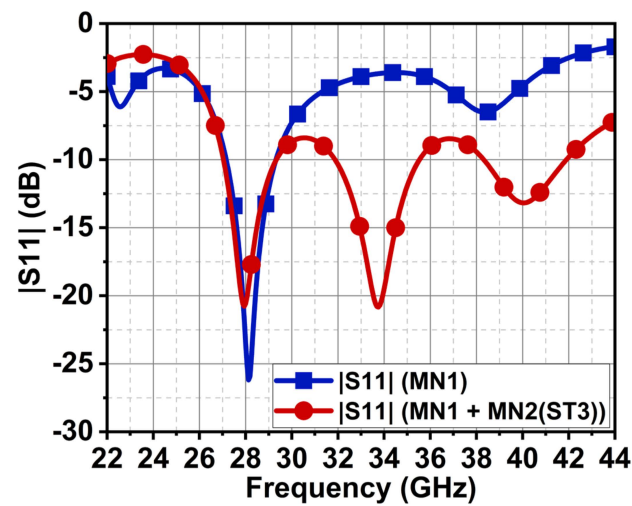


Figure 13. Reflection coefficient after incorporating Stub-3 of MN2 along with MN1.

The Y_3 and $Y_3^{\dagger}$ admittances are now considered as new individual loads, providing two possible solutions. First, considering the Y_3 admittance, a distance of $0.45\lambda_2$ is moved along its reflection coefficient circle Γ_2 in Figure 14, which intersects at point Y_4 . The Y_4 has an admittance of $1 - 5.2j$. Stub-4 must have a susceptance of $+5.2j$ to cancel the susceptance of Y_4 . Therefore, a point Y_{4A} is marked at $+5.2j$, and the Stub-4 (L_{4A}) length is estimated from the short-circuit admittance (P_{oc}) to Y_{4A} towards the generator. This resulted in the first solution for Stub-4 with L_{4A} of $0.45\lambda_2$.

Further, considering $Y_3^{\dagger}$ as the load, a similar procedure is followed by drawing the reflection coefficient circle ($\Gamma_2^{\dagger}$) and moving a distance of $0.45\lambda_2$ towards the generator in Figure 15. It intersects at $Y_4^{\dagger}$, resulting in an admittance of $1 + 6.8j$. The second possible Stub-4 susceptance can be $-6.8j$ at Y_{4B} . The second possible Stub-4 length (L_{4B}) from P_{oc} to Y_{4B} is $0.023\lambda_2$. Thus, the physical lengths of L_{4A} and L_{4B} are:

$$L_{4A} = 0.47 \lambda_2 \quad (14a)$$

$$L_{4B} = 0.023 \lambda_2 \quad (14b)$$

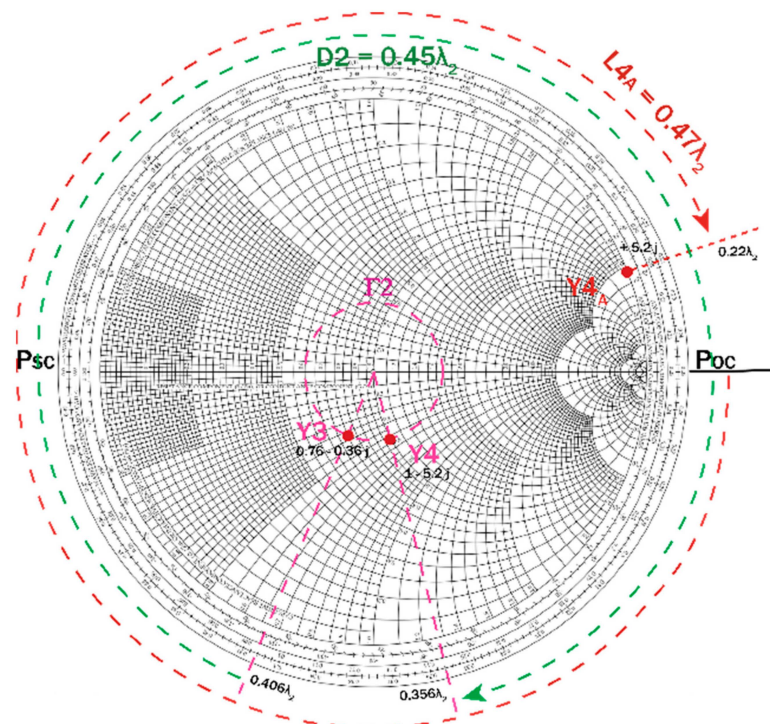


Figure 14. First possible solution for Stub-4 of MN2 at 38 GHz using the Smith chart.

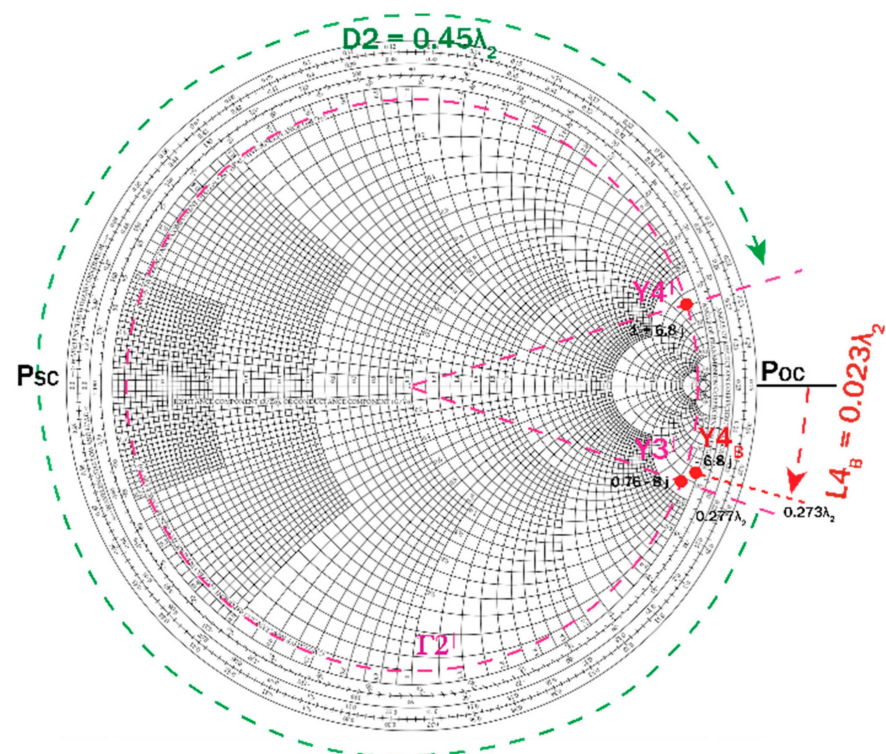


Figure 15. Second possible solution for Stub-4 in MN2 at 38 GHz using the Smith chart.

The Stub-4 impedances are:

$$Z_{4A} = \frac{1}{Y_{4A}} = -0.19j \quad (15a)$$

$$Z_{4B} = \frac{1}{Y_{4B}} = 0.14 j \quad (15b)$$

The inductance of L_{4A} and L_{4B} using Equation (8) are:

$$L_{4AL} = 0.39 \text{ nH} \quad (16a)$$

$$L_{4BL} = 0.29 \text{ nH} \quad (16b)$$

From the above physical dimensions of L_{4A} and L_{4B} , choosing L_{4A} with a reduced dimension by a half-wavelength for the design is appropriate. It is implemented at a distance of $D2$, as displayed in Figure 16. With the combination of two matching networks, that is MN1 and MN2, a dual-resonance is generated at 28 and 38 GHz with a decent bandwidth ranging from 27–28.75 GHz and 36.20–42.43 GHz, as displayed in Figure 17.

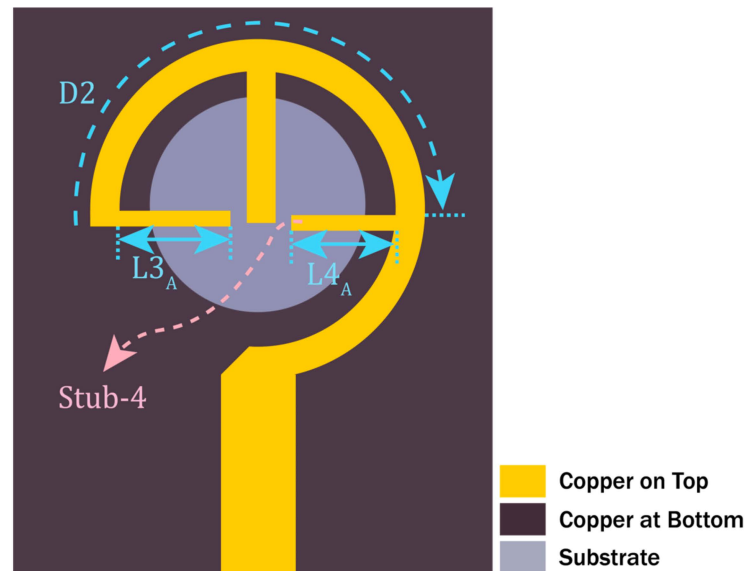


Figure 16. Antenna with Stub-4 (L_{4A}) at a distance of $D2$. The incorporated L_{4A} is approximately half of the calculated value; after tuning it, the final length is 1.5 mm.

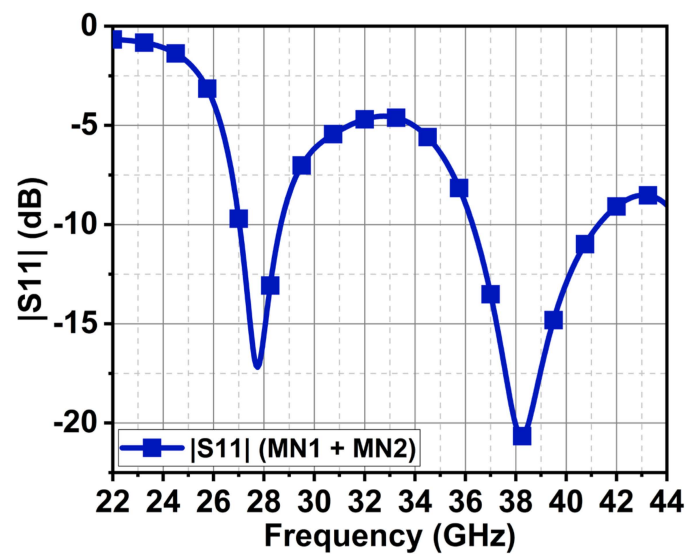


Figure 17. Simulated reflection coefficient results of the proposed antenna with both the matching networks.

3. Parametric Analysis

In this section, the variation in the stub dimensions of L_{1B} , L_{2B} , L_{3A} and L_{4A} are studied to establish its relationship with the impact on antenna impedance. L_{1B} is the series capacitance of the MN1. Its dimension is varied by changing the radius of the slot in the ground plane. On the other hand, L_{2B} is the series inductance of MN1, and L_{3A} and L_{4A} are the shunt inductances of MN2, which are varied in length to study the effect on the antenna impedance.

3.1. Series Capacitance L_{1B} of MN1

From the theoretical analysis, the Stub-1 length L_{1B} resulted in 4.71 mm. It is introduced in the design as a slot in the ground plane. The circumference of the slot is matched by tuning it with twice the calculated value of L_{1B} . Therefore, the tuned value of L_{1B} is 10.05 mm, whereas in terms of radius (G_s), it is 1.6 mm. The change in L_{1B} and its corresponding effect is observed in the reflection coefficient $|S_{11}|$ curve and its respective bandwidth. The change in the length of L_{1B} is also related to the respective slot radius in Table 1.

Table 1. Relation of Stub-1 length (L_{1B}) to the radius of slot G_s in mm.

Stub-1 Length (L_{1B})	Slot Radius (G_s)
6.28	1
7.54	1.2
8.79	1.4
10.05	1.6
11.3	1.8
12.56	2

From Figure 18a,b, it can be noticed that the change in the radius of the slot (G_s) or in length L_{1B} has a direct impact on the drift in resonance frequencies at both bands. It is because the change in slot dimension (G_s) influences all other stubs. As a result, a change in the resonance is observed. For a G_s of 1 mm or an L_{1B} of 6.28 mm, both resonances are shifted to close by at 31 and 33 GHz. Also, a third resonance at 40.5 GHz is observed; however, it has a high impedance mismatch.

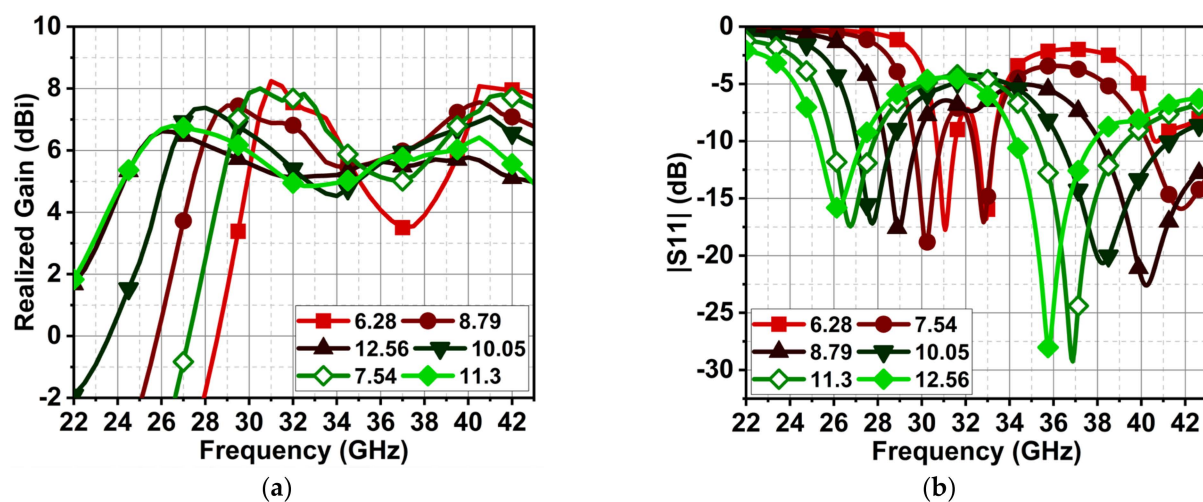


Figure 18. Variation in the $|S_{11}|$ in (a) and realized gain in (b) due to variations in Stub-1 length L_{1B} .

A similar response with the realized gain is also seen in Figure 18b. As the radius of G_s is increased, the resonances drift apart, and it can be seen that for a G_s of 2 mm, the resonances are at 26 and 36 GHz.

3.2. Series Inductance L_{2B} of MN1

The change in length of Stub-2 (L_{2B}) of MN1 has only a significant impact on the first resonance, as shown in Figure 19a,b. For an L_{2B} of 0.85 to 1.65 mm, it has no effect on impedance matching in MN1. As a result, the first resonance is abstained. However, from 2.05 to 2.85 mm, the impedance matching at the first band shows improvement. Also, with the increase in the Stub-2 length, a shift in resonance at a lower frequency is observed. The bandwidth remains the same for 2.05 to 2.85 mm. However, the resonance can be tuned with an increase or decrease between these dimensions. Figure 19b indicates that the antenna behaves as a stop-band filter before the first resonance, consequently radiating nothing. As a result, a negative gain is observed. It has a realized gain of 7 dBi at the first and second resonances.

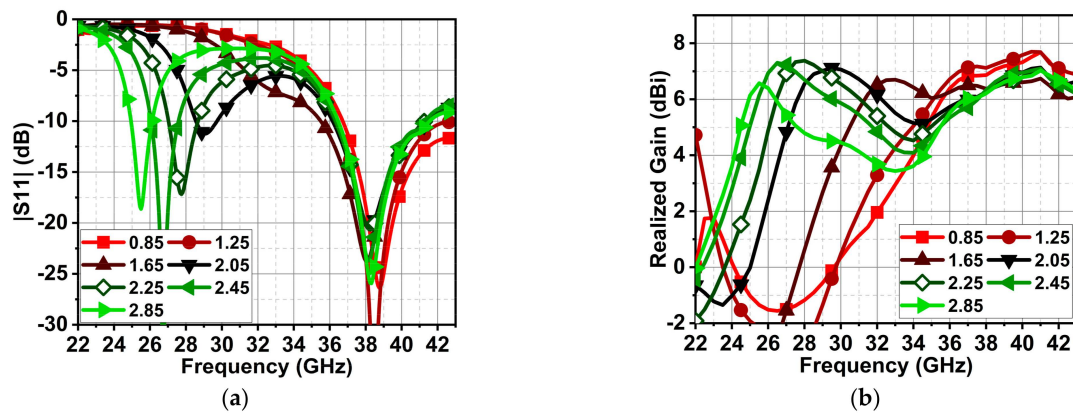


Figure 19. Variation in the $|S_{11}|$ in (a) and realized gain in (b) due to variation in Stub-2 length L_{2B} .

3.3. Shunt Inductance L_{3A} of MN2

The Stub-3 (L_{3A}) is varied from 0.6 to 1.8 mm in steps of 0.2 mm. From Figure 20a, it is evident that the variation in Stub-3 length directly impacts the second resonance impedance with almost negligible drift. For 0.6 and 0.8 mm, it has resulted in high impedance matching of 50 and 30 dB with a resonance at 37.7 GHz. For 1 mm and above, decent impedance matching is achieved at 20 dB with a resonance at 38 GHz. At 1.8 mm, impedance further improved; however, a drift in the resonance is observed at 39 GHz. Figure 20b illustrates the realized gain plot for Stub-3 variations. As indicated, the gain at 28 GHz remains the same at 7 dBi, whereas at 38 GHz, a slight change occurs due to antenna impedance changes.

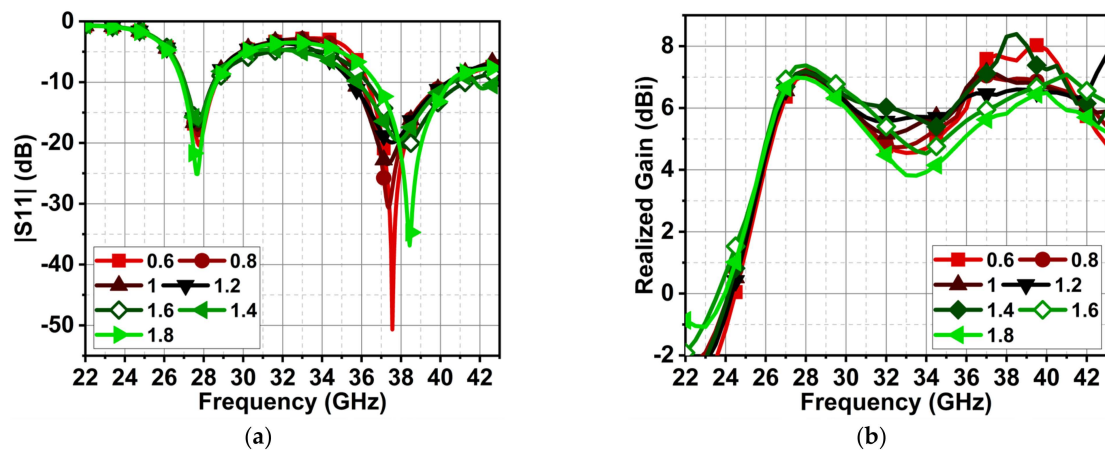


Figure 20. Variation in the $|S_{11}|$ in (a) and realized gain in (b) due to variations in Stub-3 length L_{3A} .

3.4. Shunt Inductance L_{A_4} of MN2

In this case, the L_{A_4} is varied from 0.4 to 1.6 mm. For 0.4 to 0.8 mm, the second resonance occurred at 33.5 GHz, as shown in Figure 21a. Further, from 1 to 1.6 mm, the impedance matching is shifted to 38 GHz, increasing from 15 to 32.5 dB. The corresponding shift in the realized gain from 33 GHz to 38 GHz is also observed in Figure 21b.

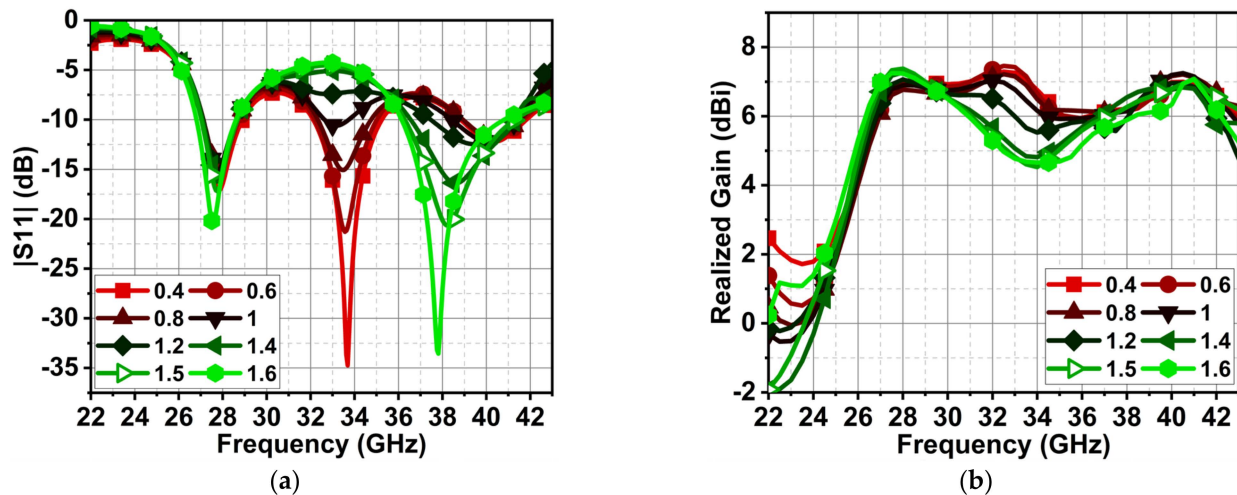


Figure 21. Variation in the $|S_{11}|$ in (a) and realized gain in (b) due to variations in Stub-4 length L_{A_4} .

4. Results and Discussion

4.1. |S-Parameter| and Radiation Pattern

The proposed antenna with a dual-matching network has an antenna profile of $0.75\lambda_1 \times 0.66\lambda_1$. The focus of the article is to present the concept of dual-band generation with the aid of internal stubs and slots embedded in a monopole antenna. The proposed antenna is highly compact and can be connected directly to an RF circuitry, as demonstrated in [33], instead of a standalone antenna, thus avoiding the requirement of SMA connectors. Further, in the future, the proposed antenna may be expanded to a large array structure for beamforming applications; in such cases, it can operate as a standalone antenna for which an SMA connector interface is used. However, to validate the simulation results of the proposed antenna, the design is prototype fabricated and uses a commercially available 2.92 mm end-launch SMA connector from Johnson (operating range up to 40 GHz) to interface [31]. An alternative method, such as direct coaxial wire feeding for a compact implantable antenna, as in [34,35], may not be effective at mmWave frequencies; consequently, 2.92 mm SMAs are used.

Before performing the prototype antenna measurement, the effect of the SMA connector on the proposed antenna is simulated, as shown in Figure 22. The 2.95 mm SMA connector has a normalized impedance of 50Ω , as a result, it has a negligible impact on the antenna impedance and reflection coefficient. However, the inner core of the SMA connector adds an additional length to the feed line, causing a slight drift of the resonances to the lower frequency, as observed in Figure 23. As the dimension of the SMA connector is comparable with the proposed antenna, the connector edges cover most of the antenna area; consequently, a slight distortion in the radiation pattern is observed in Figure 22.

The Figure 24 shows the fabricated antenna. The reflection coefficient of the antenna is measured with an Anritsu (S820E) two-port VNA and radiation pattern in an anechoic chamber, as demonstrated in Figure 25. The simulated results indicate that the first matching network comprising a series capacitive and inductive stub generates a resonance at 28 GHz with a bandwidth ranging from 27–28.68 GHz.

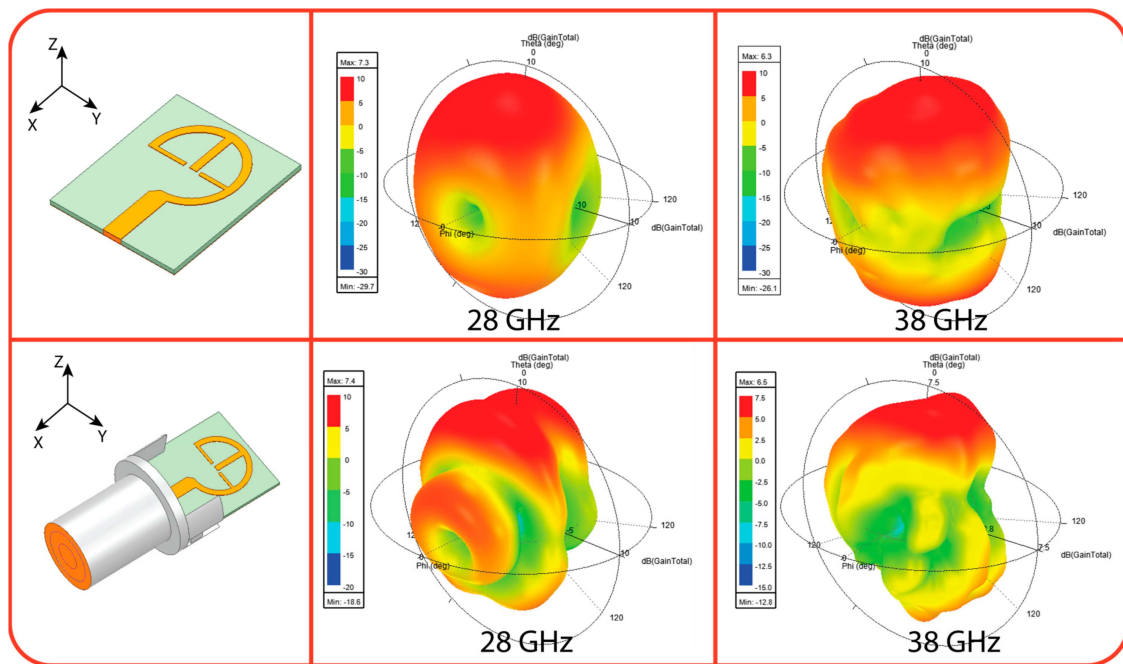


Figure 22. Proposed MIMO antenna with and without SMA connector interface and its respective radiation pattern at 28 and 38 GHz.

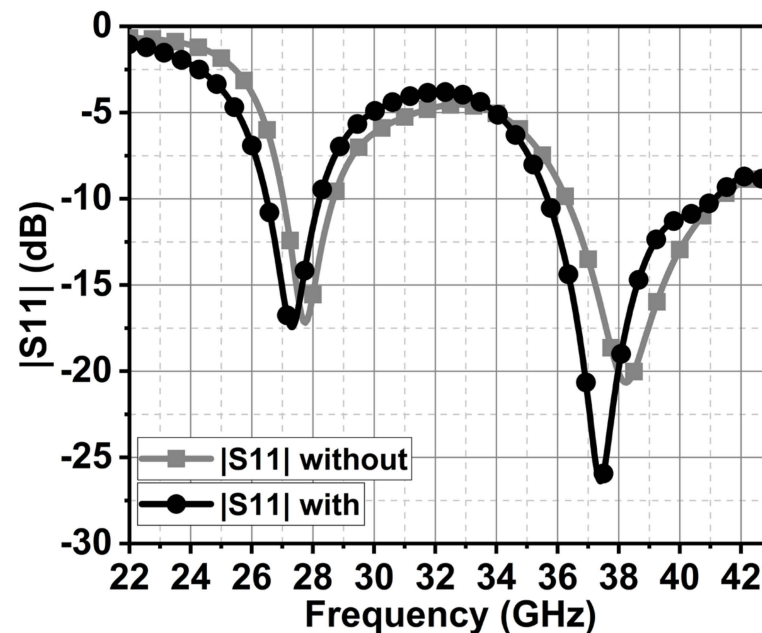


Figure 23. Simulated reflection coefficient of the proposed antenna with and without SMA connector.

Validating these results with measured results indicates the bandwidth occurred in the 26.7–28.9 GHz range. On the other hand, the second matching network with two shunt inductances generates resonance at 38 GHz with bandwidth ranging from 36.25–41.31 GHz. Due to the limitation of the frequency range of the VNA and the connector, the second band can only measure up to 40 GHz. As a result, the measured bandwidth for $|S_{11}| > 10$ dB begins at 36.1 GHz and continues up to 40 GHz. However, observing the trace in Figure 26, it can be estimated that the upper cutoff frequency could reach up to 41.5 GHz. Therefore, the second resonance bandwidth could be 5.4 GHz. The measured results are in good agreement with the simulation.

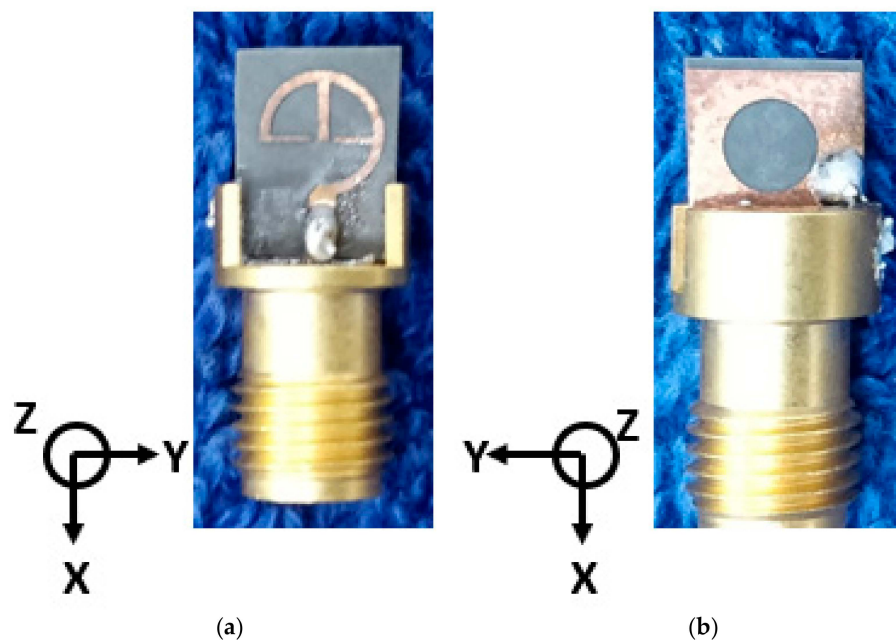


Figure 24. Prototype fabricated antenna with dual impedance matching networks.

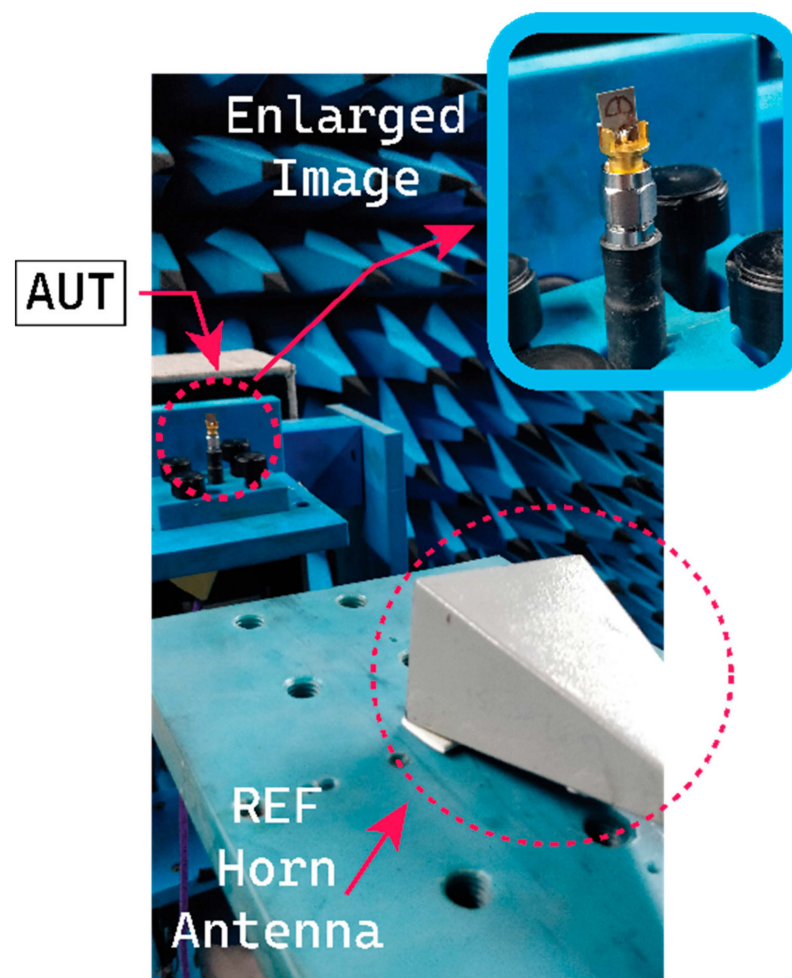


Figure 25. Measurement setup of gain and radiation pattern of proposed antenna in an anechoic chamber.

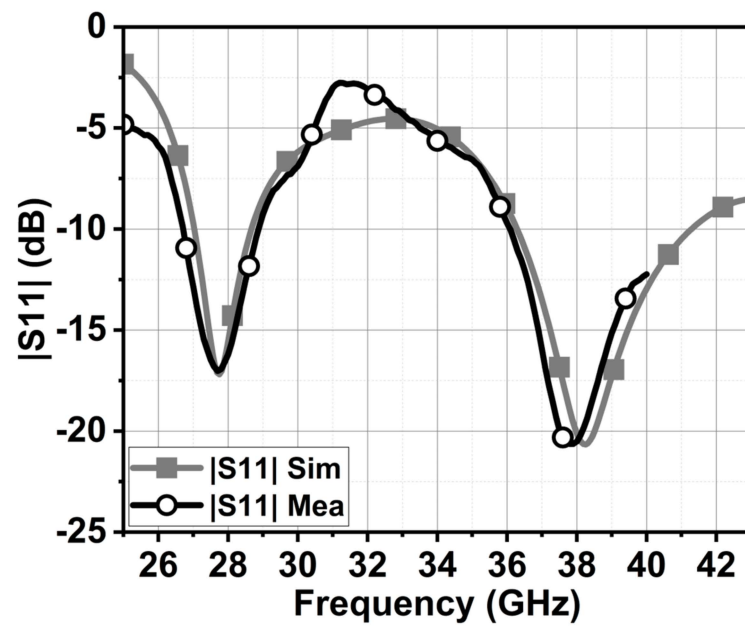


Figure 26. Simulated and measured reflection coefficient curve of the proposed antenna.

The series capacitance of the first matching is a slot in the ground plane. Due to this, a portion of radiation escapes through back radiation. Nevertheless, maximum radiation is achieved in the broadside with a half-power-beamwidth (HPBW) of 60° in the XZ plane at 28 GHz, as displayed in Figure 27a. However, in the YZ plane, it results in omnidirectional radiation, as shown in Figure 27b. In the case of 38 GHz, the XZ plane has omnidirectional radiation, whereas the YZ plane has bidirectional radiation with HPBW in the broadside at 62° , as shown in Figure 28a,b.

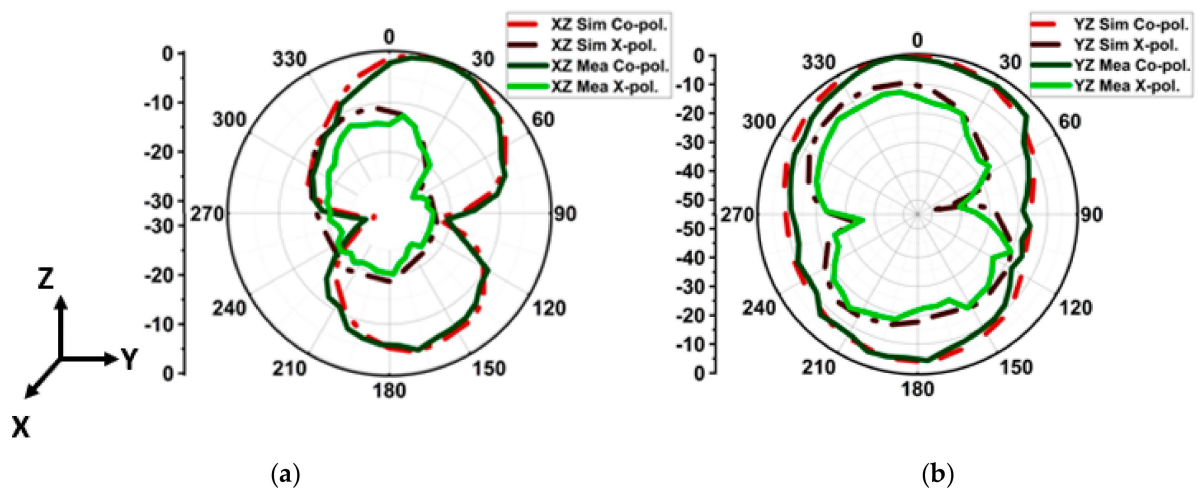


Figure 27. Normalized simulated and measured radiation pattern at 28 GHz in (a) XZ plane and (b) YZ plane.

Figures 27b and 28b indicate high cross-polarization in the YZ plane at 28 and 38 GHz. The reason is analyzed by comprehending the Poynting vector plot at both frequencies. Figure 29a,b shows the movement of the electric field in the radiator is circular (indicated with a black arrow). Also, in the ground plane around the slot, it moves circularly (marked with a red arrow). However, in Stub-3 and Stub-4 of MN2, the E-field is along the horizontal direction, which creates the high cross-polarization. Though the E-field is circular in the radiator and ground, the orthogonal E-field magnitudes are not the same to cause circular polarization. As a result, the proposed antenna has linear polarization. The antenna

with maximum broadside radiation achieved a realized gain of 7.2 dBi at 28 and 38 GHz. Also, the total efficiency of the antenna at these bands is 90% and 95%, respectively, as demonstrated in Figure 30.

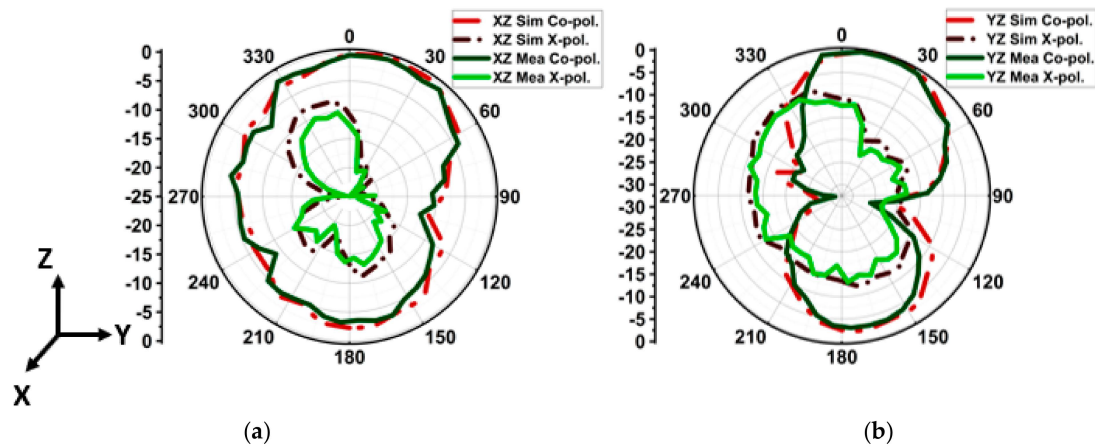


Figure 28. Normalized simulated and measured radiation pattern at 38 GHz in (a) XZ plane and (b) YZ plane.

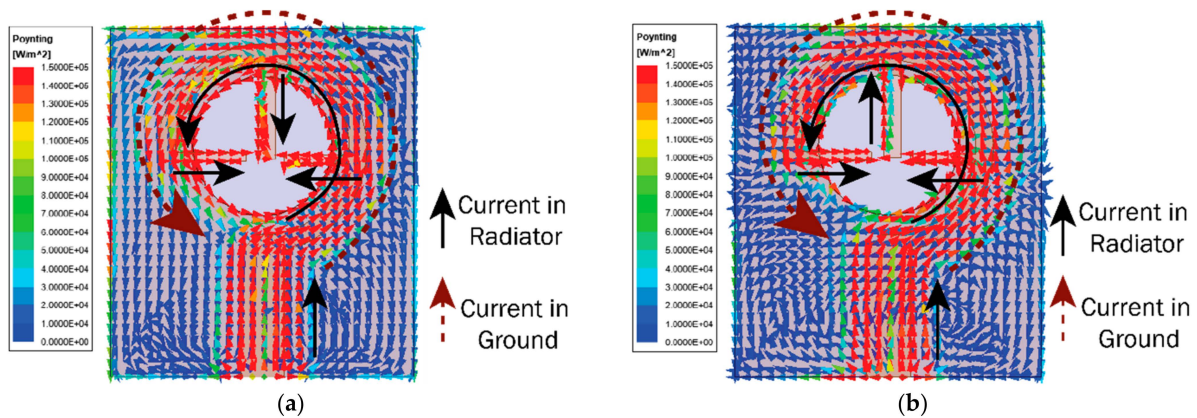


Figure 29. Representation of the electric field of an antenna using Poynting vector at (a) 28 GHz and (b) 38 GHz.

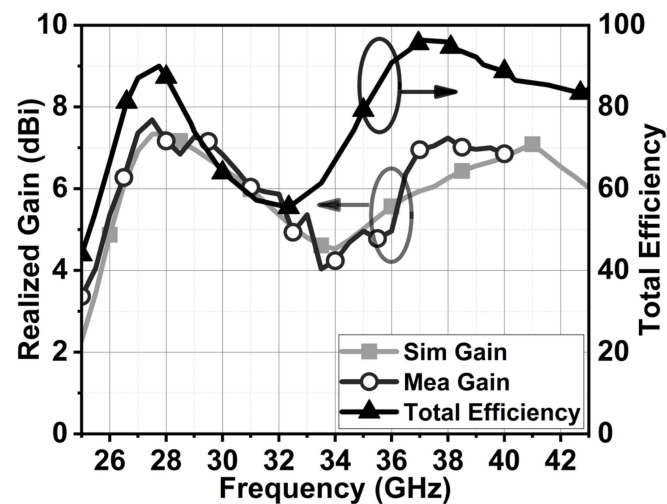


Figure 30. Realized gain and total efficiency of the proposed antenna.

4.2. Comparative Analysis

The proposed antenna with dual matching networks is compared with the existing state-of-the-art designs in Table 2. The major benefit of the proposed antenna over other designs in Table 2 is its simplicity in design. With the aid of internal matching stubs and slots, the monopole antenna could achieve decent bandwidth and gain. The proposed antenna is compact compared to other designs except with [36]; however, the antenna in [36] results in a very narrow bandwidth. Due to its compact nature, the proposed antenna could be interfaced directly with the RF circuit for applications like implantable health monitoring devices, mobile phones, smart watches, or any handheld wireless devices. The proposed antenna delivers decent bandwidth and gain compared to its other competitive designs. In all the existing designs, stub loading begins with an arbitrary dimension and is then fine-tuned with a parametric analysis. However, in our case, we presented a proper mathematical model using the double-stub technique to obtain the dimensions of the stubs for impedance matching in the antenna, thus authenticating the approach and design.

Table 2. Comparison of the proposed antenna with the existing state-of-the-art designs.

Ref.	Dimensions in mm	Dimensions in λ	Resonance	Bandwidth (GHz) at $ S_{11} > 10$ dB	Realized Gain (dBi)	Impedance Matching Technique
[37]	29×70	$2.52\lambda \times 6.08\lambda$	26	24–28.5	12 *	Stub-loaded + linear Array
[36]	2.28×3.06	$0.25\lambda \times 0.34\lambda$	33.5/60.8	33–34.2/60–60.2	6.6/7.1	Arbitrary slot-loading
[38]	15×13	$1.5\lambda \times 1.3\lambda$	30/39	27.5–32/37–41.5	7/7.5 **	Arbitrary (parametric) slot + stub loading
[39]	14×11	$1.31\lambda \times 1.03\lambda$	28/37	27–28.7/35–40	6.8/6.5	Parasitic element + L-stub
[40]	14.5×13	$1.35\lambda \times 1.22\lambda$	28/38	22–30/36.8–40	5.8 **	Arbitrary slots + stub
[41]	18×10	$1.5\lambda \times 0.84\lambda$	25/29	23–25.5/28.4–32	7	$\lambda/4$ transformer + SRRs
[42]	8.5×8	$0.8\lambda \times 0.75\lambda$	28	27.4–28.8	6.1	$\lambda/4$ transformer
[43]	9.2×5	$0.85\lambda \times 0.46\lambda$	27.5/38	25.2–29.4/36–40	6	Arbitrary slots + stub
Prop.	8×7	$0.75\lambda \times 0.66\lambda$	28/38	26.7–28.9/36.1–41.5	7.2	Double-stub matching with dual-matching network

* series of four-stub element array, ** Simulated gain.

5. Conclusions

The article carefully presented the double-stub matching technique in the antenna design, in which two such matching networks achieved dual resonance. The feed line and the curved monopole antenna had an impedance of $2.1 - 5.21j$ and $12 + 23.97j$ at 28 and 38 GHz due to the impedance mismatch. The first matching network, consisting of a series capacitance and inductance, matched the impedance at 28 GHz, resulting in $50 - 5.5j$. With the aid of the double-matching technique and a Smith chart, the computed Stub-1 length, L_{1B} , was $0.44\lambda_1$, which was later fine-tuned to twice the computed length in the physical design. Likewise, the Stub-2 length, L_{2B} , incorporated was $0.218\lambda_1$. The approximated capacitance and inductance values of these stubs were 0.852 pF and 1.37 nH. After accomplishing the impedance matching at the first resonance, the second matching network was employed with two-shunt inductance. The length of these stubs was L_{3A} and L_{4A} with half-wavelength at λ_2 . Notably, these matching networks are effective for the desired band and cause no interference or impedance mismatching to other bands. Thus, the proposed antenna achieved a decent bandwidth and gain at the respective frequency. The results attained put the antenna at the forefront of performance for 5G communication in the future.

Author Contributions: Conceptualization, P.S.B.G. and T.A.; methodology, P.S.B.G. and P.R.M.; software, P.S.B.G., P.K. and S.P.; validation, A.-A.A.B., J.A. and T.A.; formal analysis, P.S.B.G. and T.A.;

writing—original draft preparation, P.S.B.G.; writing—review and editing, P.R.M., A.-A.A.B., J.A. and T.A.; supervision, T.A. All authors have read and agreed to the published version of the manuscript.

Funding: This research received no external funding.

Institutional Review Board Statement: Not applicable.

Informed Consent Statement: Not applicable.

Data Availability Statement: Data is provided in the manuscript.

Conflicts of Interest: The authors declare no conflicts of interest.

References

1. Ullah, U.; Kamboh, U.R.; Hossain, F.; Danish, M. Outdoor-to-Indoor and Indoor-to-Indoor Propagation Path Loss Modeling Using Smart 3D Ray Tracing Algorithm at 28 GHz mmWave. *Arab. J. Sci. Eng.* **2020**, *45*, 10223–10232. [\[CrossRef\]](#)
2. Kohli, M.; Adhikari, A.; Avci, G.; Brent, S.; Dash, A.; Moser, J.; Hossain, S.; Kadota, I.; Garland, C.; Mukherjee, S.; et al. Outdoor-to-Indoor 28 GHz Wireless Measurements in Manhattan: Path Loss, Environmental Effects, and 90% Coverage. *IEEE/ACM Trans. Netw.* **2024**, *preprint*, 1–16. [\[CrossRef\]](#)
3. Ahamed, M.M.; Faruque, S. 5G Network Coverage Planning and Analysis of the Deployment Challenges. *Sensors* **2021**, *21*, 6608. [\[CrossRef\]](#)
4. Singh, C.; Sharma, C.; Tripathi, S.; Sharma, M.; Agrawal, A. A Comprehensive Survey on Millimeter Wave Antennas at 30/60/120 GHz: Design, Challenges and Applications. *Wirel. Pers. Commun.* **2023**, *133*, 1547–1584. [\[CrossRef\]](#)
5. Sun, G.-H.; Wong, H. A Planar Millimeter-Wave Antenna Array with a Pillbox-Distributed Network. *IEEE Trans. Antennas Propag.* **2020**, *68*, 3664–3672. [\[CrossRef\]](#)
6. Sun, Q.; Ban, Y.-L.; Lian, J.-W.; Liu, Y.; Nie, Z. Millimeter-Wave Multibeam Antenna Based on Folded C-Type SIW. *IEEE Trans. Antennas Propag.* **2020**, *68*, 3465–3476. [\[CrossRef\]](#)
7. Borhani-Kakhki, M.; Dadgarpour, A.; Antoniadis, M.A.; Sebak, A.R.; Denidni, T.A. Magnetolectric Dipole Antennas Loaded with Meta-Lens for 5G MIMO Pattern Diversity Applications. *IEEE Trans. Antennas Propag.* **2022**, *70*, 7112–7117. [\[CrossRef\]](#)
8. Dadgarpour, A.; Bayat-Makou, N.; Antoniadis, M.A.; Kishk, A.A.; Sebak, A. A Dual-Polarized Magnetolectric Dipole Array Based on Printed Ridge Gap Waveguide with Dual-Polarized Split-Ring Resonator Lens. *IEEE Trans. Antennas Propag.* **2020**, *68*, 3578–3585. [\[CrossRef\]](#)
9. Alibakhshikenari, M.; Virdee, B.S.; See, C.H.; Abd-Alhameed, R.A.; Falcone, F.; Limiti, E. High-Isolation Leaky-Wave Array Antenna Based on CRLH-Metamaterial Implemented on SIW with $\pm 30^\circ$ Frequency Beam-Scanning Capability at Millimetre-Waves. *Electronics* **2019**, *8*, 642. [\[CrossRef\]](#)
10. Lee, K.F.; Tong, K.F. Microstrip Patch Antennas—Basic Characteristics and Some Recent Advances. *Proc. IEEE* **2012**, *100*, 2169–2180. [\[CrossRef\]](#)
11. Liu, P.; Zhu, X.-W.; Zhang, Y.; Wang, X.; Yang, C.; Jiang, Z.H. Patch Antenna Loaded with Paired Shorting Pins and H-Shaped Slot for 28/38 GHz Dual-Band MIMO Applications. *IEEE Access* **2020**, *8*, 23705–23712. [\[CrossRef\]](#)
12. Ramanujam, P.; Arumugam, C.; Venkatesan, R.; Ponnusamy, M. Design of Compact Patch Antenna with Enhanced Gain and Bandwidth for 5G Mm-wave Applications. *IET Microw. Antennas Propag.* **2020**, *14*, 1455–1461. [\[CrossRef\]](#)
13. Farahat, A.E.; Hussein, K.F.A. Dual-Band (28/38 GHz) Wideband MIMO Antenna for 5G Mobile Applications. *IEEE Access* **2022**, *10*, 32213–32223. [\[CrossRef\]](#)
14. Han, K.; Yan, Y.; Yan, Z.; Wang, C. Low-Profile Millimeter-Wave Metasurface-Based Antenna with Enhanced Bandwidth. *Micromachines* **2023**, *14*, 1403. [\[CrossRef\]](#)
15. Kumar, A.; Kumar, A.; Kumar, A. Defected Ground Structure Based High Gain, Wideband and High Diversity Performance Quad-Element MIMO Antenna Array for 5G Millimeter-Wave Communication. *Prog. Electromagn. Res.* **2023**, *101*, 1–16. [\[CrossRef\]](#)
16. Hoarau, C.; Corrao, N.; Arnould, J.-D.; Ferrari, P.; Xavier, P. Complete Design and Measurement Methodology for a Tunable RF Impedance-Matching Network. *IEEE Trans. Microw. Theory Techn.* **2008**, *56*, 2620–2627. [\[CrossRef\]](#)
17. Lim, Y.; Tang, H.; Lim, S.; Park, J. An Adaptive Impedance-Matching Network Based on a Novel Capacitor Matrix for Wireless Power Transfer. *IEEE Trans. Power Electron.* **2014**, *29*, 4403–4413. [\[CrossRef\]](#)
18. Kim, M.G.; Yoon, S.; Kim, H.H.; Shung, K.K. Impedance Matching Network for High Frequency Ultrasonic Transducer for Cellular Applications. *Ultrasonics* **2016**, *65*, 258–267. [\[CrossRef\]](#)
19. Zhang, Y.H.; Spiegel, R.J.; Fan, Y.; Joines, W.T.; Liu, Q.H.; Da Xu, K. Design of a Stub-Loaded Ring-Resonator Slot for Antenna Applications. *IEEE Trans. Antennas Propag.* **2015**, *63*, 517–524. [\[CrossRef\]](#)
20. Cuneray, K.; Akcam, N.; Okan, T.; Arican, G.O. 28/38 GHz Dual-Band MIMO Antenna with Wideband and High Gain Properties for 5G Applications. *AEU-Int. J. Electron. Commun.* **2023**, *162*, 154553. [\[CrossRef\]](#)
21. Islam, S.; Zada, M.; Yoo, H. Low-Pass Filter Based Integrated 5G Smartphone Antenna for Sub-6-GHz and Mm-Wave Bands. *IEEE Trans. Antennas Propag.* **2021**, *69*, 5424–5436. [\[CrossRef\]](#)
22. Alnemr, F.; Ahmed, M.F.; Shaalan, A.A. A Compact 28/38 GHz MIMO Circularly Polarized Antenna for 5 G Applications. *J. Infrared Millim. Terahertz Waves* **2021**, *42*, 338–355. [\[CrossRef\]](#)

23. Chaudhary, S.; Kansal, A. Compact High Gain 28, 38 GHz Antenna for 5G Communication. *Int. J. Electron.* **2023**, *110*, 1028–1048. [\[CrossRef\]](#)
24. Sabek, A.R.; Ali, W.A.E.; Ibrahim, A.A. Minimally Coupled Two-Element MIMO Antenna with Dual Band (28/38 GHz) for 5G Wireless Communications. *J. Infrared Millim. Terahertz Waves* **2022**, *43*, 335–348. [\[CrossRef\]](#)
25. Kamal, M.M.; Yang, S.; Kiani, S.H.; Sehrai, D.A.; Alibakhshikenari, M.; Abdullah, M.; Falcone, F.; Limiti, E.; Munir, M. A Novel Hook-Shaped Antenna Operating at 28 GHz for Future 5G Mmwave Applications. *Electronics* **2021**, *10*, 673. [\[CrossRef\]](#)
26. Esmail, B.A.F.; Koziel, S. Design and Optimization of Metamaterial-Based Dual-Band 28/38 GHz 5G MIMO Antenna with Modified Ground for Isolation and Bandwidth Improvement. *Antennas Wirel. Propag. Lett.* **2023**, *22*, 1069–1073. [\[CrossRef\]](#)
27. Kiouach, F.; Aghoutane, B.; Das, S.; Islam, T.; El Ghzaoui, M.; Madhav, B. A Dual Operating (27/38 GHz) High Performance 2×4 MIMO Antenna Array for 5G New Radio Applications. *Phys. Scr.* **2023**, *98*, 115504. [\[CrossRef\]](#)
28. Islam, S.; Zada, M.; Yoo, H. Highly Compact Integrated Sub-6 GHz and Millimeter-Wave Band Antenna Array for 5G Smartphone Communications. *IEEE Trans. Antennas Propag.* **2022**, *70*, 11629–11638. [\[CrossRef\]](#)
29. Khabba, A.; Amadid, J.; Ibnyach, S.; Zeroual, A. Pretty-Small Four-Port Dual-Wideband 28/38 GHz MIMO Antenna with Robust Isolation and High Diversity Performance for Millimeter-Wave 5G Wireless Systems. *Analog. Integr. Circ. Signal Process* **2022**, *112*, 83–102. [\[CrossRef\]](#)
30. Chan, K.-C.; Harter, A. Impedance matching and the Smith chart-The fundamentals. *RF Design* **2000**, *23*, 52–66.
31. Ida, N. *Engineering Electromagnetics*; Springer International Publishing: Cham, Switzerland, 2015; ISBN 978-3-319-07805-2.
32. Caloz, C.; Okabe, H.; Iwai, T.; Itoh, T. A Simple and Accurate Model for Microstrip Structures with Slotted Ground Plane. *IEEE Microw. Wireless Compon. Lett.* **2004**, *14*, 127–129. [\[CrossRef\]](#)
33. Lyu, H.; Wang, J.; La, J.-H.; Chung, J.M.; Babakhani, A. An Energy-Efficient Wirelessly Powered Millimeter-Scale Neurostimulator Implant Based on Systematic Codesign of an Inductive Loop Antenna and a Custom Rectifier. *IEEE Trans. Biomed. Circuits Syst.* **2018**, *12*, 1131–1143. [\[CrossRef\]](#) [\[PubMed\]](#)
34. Li, H.; Wang, B.; Guo, L.; Xiong, J. Efficient and Wideband Implantable Antenna Based on Magnetic Structures. *IEEE Trans. Antennas Propag.* **2019**, *67*, 7242–7251. [\[CrossRef\]](#)
35. Alrawashdeh, R.S.; Huang, Y.; Kod, M.; Abu Bakar Sajak, A. A Broadband Flexible Implantable Loop Antenna with Complementary Split Ring Resonators. *Antennas Wirel. Propag. Lett.* **2015**, *14*, 1506–1509. [\[CrossRef\]](#)
36. Farooq, U.; Rather, G.M. A Miniaturised Ka/V Dual Band Millimeter Wave Antenna for 5G Body Centric Network Applications. *Alex. Eng. J.* **2022**, *61*, 8089–8096. [\[CrossRef\]](#)
37. Guo, Y.Q.; Pan, Y.M.; Zheng, S.Y.; Lu, K. A Singly-Fed Dual-Band Microstrip Antenna for Microwave and Millimeter-Wave Applications in 5G Wireless Communication. *IEEE Trans. Veh. Technol.* **2021**, *70*, 5419–5430. [\[CrossRef\]](#)
38. Hussain, M.; Awan, W.A.; Ali, E.M.; Alzaidi, M.S.; Alsharef, M.; Elkamouchi, D.H.; Alzahrani, A.; Fathy Abo Sree, M. Isolation Improvement of Parasitic Element-Loaded Dual-Band MIMO Antenna for Mm-Wave Applications. *Micromachines* **2022**, *13*, 1918. [\[CrossRef\]](#) [\[PubMed\]](#)
39. Sadeghi-Marasht, S.; Sharawi, M.S.; Zhu, A. Dual-Band Circularly Polarized Antenna Array for 5G Millimeter-Wave Applications. *IEEE Open J. Antennas Propag.* **2022**, *3*, 314–323. [\[CrossRef\]](#)
40. Ali Esmail, B.; Koziel, S. High Isolation Metamaterial-Based Dual-Band MIMO Antenna for 5G Millimeter-Wave Applications. *AEU-Int. J. Electron. Commun.* **2023**, *158*, 154470. [\[CrossRef\]](#)
41. Aboualalaa, M.; Mansour, I. Dual-Band End-Fire Four-Element MIMO Antenna Array Using Split-Ring Structure for Mm-Wave 5G Applications. *IEEE Access* **2023**, *11*, 57383–57390. [\[CrossRef\]](#)
42. Ghosh, S.; Baghel, G.S.; Swati, M.V. A Low-Profile, High Gain, Dual-Port, Planar Array Antenna for Mm-Wave Powered 5G IoT Systems. *AEU-Int. J. Electron. Commun.* **2022**, *155*, 154354. [\[CrossRef\]](#)
43. Khan, D.; Ahmad, A.; Choi, D.-Y. Dual-Band 5G MIMO Antenna with Enhanced Coupling Reduction Using Metamaterials. *Sci. Rep.* **2024**, *14*, 96. [\[CrossRef\]](#) [\[PubMed\]](#)

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