Circuital characteristics and radiation properties of an UWB electric-magnetic planar antenna for Ku-band applications

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[1] A planar, directive antenna with large fractional bandwidth is introduced in this paper. A detailed discussion on the proposed antenna topology and its architecture is reported. The proposed element is a combination of a patch and a loop radiator. A proper combination of the electric field radiator (patch) with a magnetic field radiator (loop around the patch) is exploited for expanding the operational bandwidth. A parametric study is presented to investigate the effect of the antenna geometrical parameters on its performance. A general and computationally efficient procedure for extracting the antenna equivalent circuit is described and used to achieve a meaningful circuit theory-based insight into the characteristics of the radiating structure. The theoretical and experimental results are compared, and it is demonstrated that the element features over 100% fractional bandwidth, good impedance matching, and unidirectional and stable radiation patterns.


1. Introduction

[2] For a wide scope of applications including surveillance, security, through-wall or rubble imaging, and medical diagnosis, ultrawideband (UWB) radar is needed. A wide bandwidth provides a very fine range resolution for this type of radars. The wide operational bandwidths of the antennas are crucial for the total system performance; however, there are only a few radiators which are able to provide a fractional bandwidth close to 100%.

[3] In the last 10 years, many antenna elements have been designed which are suitable for UWB or wideband applications [Schantz, 2005; Qing and Chen, 2009; Yarovoy and Pugliese, 2006]. However, in the majority of the cases, the antennas are not suitable for radar applications due to their bidirectional radiation pattern. Furthermore, for dense arrays, the elements should be small enough to fit within the allowable element spacing to avoid grating lobes. A nondispersive behavior of the antenna is also required for optimal wave reception and transmission. The most widely used radiators in UWB arrays are Vivaldi-like antennas [Schaubert et al., 2003; Qing et al., 2006]. However, Vivaldi-based arrays are characterized by a large volume (3D array). For systems implemented on a mobile platform, such as through-wall imaging radar and radar systems for an unmanned aerial vehicle, planar antenna arrays are preferred to minimize the total weight. The study presented in this paper is aimed at developing a novel class of UWB electric-magnetic planar antennas which satisfies the above mentioned requirements. It is experimentally demonstrated that the considered radiating elements feature a good impedance matching from 8 to 24 GHz and, hence, are attractive candidates for short-range UWB imaging, satellite communications, and radar applications in X/Ku band. The original antenna concept has been introduced in Tran et al. [2009, 2011], and its applicability as a polarimetric array element has been addressed in Haider et al. [2010]. In this paper, emphasis is put on the enhancement of the impedance bandwidth and size reduction of the element, as well as the relevant experimental verification. Furthermore, a detailed analysis of the antenna performance as a function of the main geometrical parameters is presented in combination with a suitable equivalent circuit model.

[4] The paper is arranged in the following way. Section 2 describes the topology and the geometry of the proposed antenna. The parametric analyses are presented in section 3. In section 4, the antenna equivalent circuit is discussed. The numerical and experimental analysis of the antenna performance in time and frequency domains is detailed in section 5. Finally, section 6 contains the concluding remarks.

2. Antenna Structure

[5] For ultrawideband radiation, it is important to combine proper excitation of both electric and magnetic fields [Schantz, 2005, p. 247; Kwon et al., 2008; Farr et al., 1999]. The goal of this research is to develop a novel antenna based on a combination of a patch and a loop. The primary requirements for an array element are unidirectional radiation pattern and compact element size to avoid grating lobes. To design an UWB antenna with unidirectional pattern, we propose a topology where a patch antenna is etched within the allowable element spacing to avoid grating lobes. To design an UWB antenna with unidirectional pattern, the patch antenna is encircled by a loop as shown in Figure 1. The antenna is etched on a commercially available Taconic high-frequency
material (TLY-5) with a relative permittivity $\varepsilon_r = 2.2$. The antenna is modeled numerically by using the commercially available simulation tool CST Microwave Studio, based on the finite integration technique (FIT).

[6] The inner patch contributes more to the radiation in the higher part of the operating band where the patch length becomes quarter to half of the guided wavelength. On the other hand, the perimeter of the loop is roughly one wavelength around the center frequency. Due to the presence of the stubs, the actual loop perimeter will increase, and hence, the frequency band will expand toward the lower frequencies.

[7] To concurrently feed the radiating patch and the loop, a grounded coplanar waveguide (CPWG) is the most appropriate choice. Furthermore, it is well known that the CPW structure has many advantages for wideband applications, such as constant effective permittivity, low radiation and conductor losses, less dispersion, simple structure, and very wide operational bandwidth.

[8] To further expand the impedance bandwidth, both the patch and the loop geometry have been optimized. First, two side notches are added to the radiation patch to extend the electrical length which implies that the current will be forced to follow a longer path. Four stubs are attached to the upper side of the ring. The length of the outer stubs is important for the center frequency. By introducing these outer stubs, we are reducing the loop area. The inner stubs are important for the lower resonance. These stubs are producing additional path length for the current, and the operational band shifts toward the lower frequencies.

3. Parametric Analyses

[9] A parametric analysis has been carried out by evaluating the effect of the variation of geometrical parameters on the element performance. In this section, the impact of the key geometrical parameters of the antenna is analyzed and discussed thoroughly. In doing so, only one parameter value has been changed at a time while keeping the others unaltered.

[10] In Figure 2, the effect of the patch width ($P_4$) on the impedance bandwidth is illustrated. The patch width has a clear balancing effect between the lower and higher frequency bands. When $P_4$ is reduced to less than 6 mm, a strong resonance around 18 GHz is observed. On the other hand, a patch width larger than 6.6 mm results in a strong resonance around 11 GHz. This parametric analysis shows prospective of the proposed structure as a frequency-reconfigurable antenna element. For a wide operational band, it is essential to maintain balance between the resonances, and hence, $P_4 = 6.25$ mm has been chosen. The patch length ($P_3$) has been varied from 3 to 4 mm, and the effect is shown in Figure 3. When $P_3$ is larger than 4 mm, it has been found out that the return loss level in the lower frequency band increases. On the other hand, when the length is decreased, the antenna shows two clear separate resonances around 10 GHz and 18 GHz.

[11] In Figure 4, the effect of the outer stub length ($P_{15}$) is illustrated. This study shows that $P_{15}$ has a negligible impact on the total bandwidth. On the other hand, it plays an important role for the antenna matching condition in the center.
The impact of the curve radius $P_{13}$ is analyzed in Figure 5. The obtained numerical results clearly show that a smooth transition from the input CPWG to the radiating patch, namely, a larger value of $P_{13}$, results in an enhanced return loss response in the frequency spectrum of interest. The purpose of the impedance transformer is to provide a good matching between the $50\,\Omega$ coax cable and the antenna input impedance over a large frequency band. Figure 6 shows that the transformer width ($P_9$) has the most effect on the center frequency. This study shows that maximizing the transformer width will provide good balance in the return loss.

The influence of other parameters, e.g., patch inner stub length and notch length, has been studied and fine tuned to achieve a wide bandwidth (VSWR lower than 2 between 8 GHz and 24 GHz) and linear phase response. In Figure 7, a flowchart of the parametric analyses procedure has been outlined, and the selected values of different geometrical parameters are listed in Table 1. The performance of this radiator is further discussed in section 5.

**4. Antenna Equivalent Circuit**

Typically, electromagnetic field solvers and measurement systems, such as network analyzers, generate scattering parameter representations of microwave components and antennas. However, electronic circuit simulators, such as SPICE [Nagel and Pederson, 1973], can natively handle conventional circuits consisting of lumped frequency-independent elements, whose behavior can be conveniently modeled by differential algebraic equations which are solved using suitable implicit integration methods in combination with sparse matrix techniques. In this context, a general and computationally efficient procedure for extracting an equivalent circuit from a given $S$-parameter representation is highly desirable [Timmins and Wu, 2000].

The procedure used to extract the equivalent circuit is based on a heuristic modification of Foster’s network synthesis technique [Guillemin, 1965], useful to account for ohmic and radiation losses occurring in the antenna structure. To this end, suitable resistive elements are introduced in the equivalent network (see Figure 8) used to model the input

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**Table 1. Geometrical Dimension of the X-Band Antenna Element**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Dimension (mm)</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_1$</td>
<td>10</td>
<td>Antenna length</td>
</tr>
<tr>
<td>$P_2$</td>
<td>10</td>
<td>Antenna width</td>
</tr>
<tr>
<td>$P_3$</td>
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<td>Patch length</td>
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<td>$P_4$</td>
<td>6.25</td>
<td>Patch width</td>
</tr>
<tr>
<td>$P_5$</td>
<td>1.94</td>
<td>Signal width of CPWG</td>
</tr>
<tr>
<td>$P_6$</td>
<td>0.102</td>
<td>Gap width of CPWG</td>
</tr>
<tr>
<td>$P_7$</td>
<td>N/A</td>
<td>Impedance transformer length</td>
</tr>
<tr>
<td>$P_8$</td>
<td>3</td>
<td>Length of the feeding section</td>
</tr>
<tr>
<td>$P_9$</td>
<td>1.94</td>
<td>Transformer width</td>
</tr>
<tr>
<td>$P_{10}$</td>
<td>0.99</td>
<td>Notch length</td>
</tr>
<tr>
<td>$P_{11}$</td>
<td>0.17</td>
<td>Notch width</td>
</tr>
<tr>
<td>$P_{12}$</td>
<td>1.03</td>
<td>Slot curve radius</td>
</tr>
<tr>
<td>$P_{13}$</td>
<td>0.80</td>
<td>CPWG curve radius</td>
</tr>
<tr>
<td>$P_{14}$</td>
<td>0.72</td>
<td>Patch curve radius</td>
</tr>
<tr>
<td>$P_{15}$</td>
<td>0.672</td>
<td>Outer stub length</td>
</tr>
<tr>
<td>$P_{16}$</td>
<td>1.52</td>
<td>Inner stub length</td>
</tr>
<tr>
<td>$P_{17}$</td>
<td>1.06</td>
<td>Inner stub separation</td>
</tr>
<tr>
<td>$P_{18}$</td>
<td>1.49</td>
<td>Inner stub width</td>
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<tr>
<td>$P_{19}$</td>
<td>1</td>
<td>Ring width</td>
</tr>
<tr>
<td>$H$</td>
<td>5.52</td>
<td>Height of the substrate</td>
</tr>
</tbody>
</table>

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**Figure 5.** The effect of the transition curve radius ($P_{13}$) on the input reflection coefficient of the antenna.

**Figure 6.** The effect of the transformer width ($P_9$) on the input reflection coefficient of the antenna.

**Figure 7.** Parametric analyses flowchart.

**Figure 8.** The proposed Foster-like network topology for the antenna equivalent circuit.
impedance $Z_{\text{FIT}}(f)$ of the device as computed by the adopted FIT-based electromagnetic field solver. As a consequence, the antenna frequency behavior is described by the following series expansion [Caratelli et al., 2006]:

$$Z_{\text{in}} = j2\pi f L_0 + \frac{R_0}{1 + j2\pi f \tau_0} + \sum_{n=1}^{N} \frac{R_n}{1 + jQ_n \left( \frac{f_n}{f} - \frac{1}{2} \right)}.$$  (1)

where $\tau_0 = R_0 C_0$, $R_0$ and $C_0$ being the quasi-static input resistance and capacitance, respectively. In (1), $L_0$ is the inductance accounting for the effect of the higher-order modes and the feeding structure, whereas $N$ denotes the number of resonant modes needed to properly synthesize the input impedance of the antenna. In particular, each mode is assumed to be characterized by natural frequency $f_n$, quality factor $Q_n$, and damping resistance $R_n$.

A simple analysis of the equivalent network topology shown in Figure 8 readily allows evaluation of the quasi-static resistance $R_0$ and capacitance $C_0$ as follows:

$$R_0 = \lim_{f \to 0} \text{Re}\{Z_{\text{FIT}}^{\text{in}}(f)\},$$  (2)

$$C_0 = \lim_{f \to 0} \frac{\text{Im}\{Y_{\text{FIT}}^{\text{in}}(f)\}}{2nf} = \frac{1}{2\pi} \lim_{f \to 0} \frac{\partial \text{Im}\{Y_{\text{FIT}}^{\text{in}}(f)\}}{\partial f},$$  (3)

where $Y_{\text{FIT}}^{\text{in}}(f) = 1/Z_{\text{FIT}}^{\text{in}}(f)$ denotes the input admittance, and $Af$ is the frequency step adopted to perform the discrete Fourier’s transform of the computed time domain antenna response. Furthermore, the resonant frequencies of the RLC subnetworks can be calculated, at a first approximation, by determining the values $v_n^{(0)}$ for which the real part of $Z_{\text{FIT}}^{\text{in}}(f)$ exhibits local maxima. Thereby, the initial estimates of $R_n$ and $Q_n$ are found to be as follows:

$$R_n^{(0)} = \text{Re}\{Z_{\text{FIT}}^{\text{in}}(v_n^{(0)})\},$$  (4)

$$Q_n^{(0)} = -\frac{v_n^{(0)} \partial \text{Im}\{Z_{\text{FIT}}^{\text{in}}(v_n^{(0)})\}}{2R_n^{(0)}},$$

$$= \frac{\text{Im}\{Z_{\text{FIT}}^{\text{in}}(v_n^{(0)} + Af)\} - \text{Im}\{Z_{\text{FIT}}^{\text{in}}(v_n^{(0)} - Af)\}}{4R_n^{(0)} Af},$$  (5)

Figure 9. Frequency domain behavior of the input reflection coefficient relevant to the considered ultrawideband radiator as a function of the width $P4$ of the metal patch: antenna response (a) computed by the finite integration technique and (b) synthesized by a Foster-like equivalent network of order $N=9$.

Figure 10. Quasi-static resistance, inductance, and capacitance relevant to the Foster-like equivalent circuit of order $N=9$ used to synthesize the computed input impedance of the antenna as a function of the patch width $P4$. 

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Figure 11. Natural frequencies of the RLC bipoles relevant to the Foster-like equivalent circuit of order $N = 9$ used to synthesize the computed input impedance of the antenna as a function of the patch width $P_4$.

Figure 12. Quality factors of the RLC bipoles relevant to the Foster-like equivalent circuit of order $N = 9$ used to synthesize the computed input impedance of the antenna as a function of the patch width $P_4$. 
respectively, with the values of $L_n^{(0)}$ and $C_n^{(0)}$ following from their mutual relationships, that is,

$$ L_n^{(0)} = \left(\frac{R_n^{(0)}}{Q_n^{(0)}}\right)^2 , \quad C_n^{(0)} = \frac{R_n^{(0)}}{2\pi v_n^{(0)} Q_n^{(0)}}. \quad (6) $$

[17] Upon doing so, it is straightforward to show that the quasi-static inductance can be estimated as follows:

$$ L_0^{(0)} = \lim_{f \to 0^+} \frac{\text{Im}\{Z_{in}^{(f)}(f)\}}{2\pi f} - \sum_{n=1}^{N} L_n^{(0)} $$

$$ \simeq \frac{\text{Im}\{Z_{in}^{(f)}\Delta f(f)\}}{2\pi \Delta f} - \sum_{n=1}^{N} L_n^{(0)}. \quad (7) $$

[18] The parameter values so derived can be used as initial guess within an iterative nonlinear fitting procedure, based on the gradient method [Fletcher, 1980], and aimed to minimize, in the frequency band $B$ of interest, the relative mean square error:

$$ e^{(i)}(R_n^{(i)}, L_n^{(i)}, C_n^{(i)}) = \sqrt{\frac{\int_B \left| \Gamma_{in}^{(f)}(f) - \Gamma_{in}^{(f,R_n^{(i)},C_n^{(i)})}\right|^2 df}{\int_B \left| \Gamma_{in}^{(f)}(f)\right|^2 df}}. \quad (8) $$

with $\Gamma_{in} = (Z_{in} - Z_0)/(Z_{in} + Z_0)$ denoting the input reflection coefficient, with respect to the reference impedance $Z_0 = 50 \Omega$, relevant to the equivalent network shown in Figure 8, expressed as a function of the circuit parameters $R_n^{(i)}, L_n^{(i)}, C_n^{(i)}$ ($n = 0, 1, \ldots, N$) at the $i$th iteration. In this way, the synthesis of the Foster-like circuit model that mimics the computed structure response can be numerically accomplished in a simple and straightforward manner, thus providing a useful tool for the design and optimization of the antenna front-end.

[19] As it can be noticed in Figure 9, only a few resonant modes ($N = 9$) are needed to accurately model the characteristics of the proposed antenna element over a very wide band. In particular, the dependence of the equivalent network parameters on the width $P^4$ of the radiating patch (see Figure 1) is shown in Figures 9–13. In this way, one can easily gain a meaningful circuit theory-based insight into the properties of the natural resonant modes which

![Figure 13](image13.png)

**Figure 13.** Damping resistances of the RLC bipoles relevant to the Foster-like equivalent circuit of order $N = 9$ used to synthesize the computed input impedance of the antenna as a function of the patch width $P^4$.

![Figure 14](image14.png)

**Figure 14.** Experimental prototype of the Ku-band element with the SMA connector.
occur in the device resulting in an ultrawideband (UWB) or rather multi-band \( P4 > 6.5 \) mm behavior of the antenna within the considered frequency range up to \( f_{\text{max}} = 30 \) GHz.

5. Performance Analysis and Experimental Verification

[20] In order to verify the theoretical results presented in the previous sections, an experimental prototype has been designed and fabricated (see Figure 14). The antenna is printed on a Taconic (TLY-5) substrate with relative permittivity \( \varepsilon_r = 2.2 \) and 17 \( \mu \)m thick copper cladding. The thickness of the substrate has been set to 5.52 mm, by bonding two 1.58 mm thick layers on both sides of a 2.36 mm thick substrate.

5.1. Impedance Bandwidth

[21] The impedance bandwidth of the proposed antenna structures has been experimentally analyzed. The Agilent E8364B network analyzer has been used for measurements. The frequency domain behavior of the input reflection coefficient in X/Ku band is shown in Figure 15, and a good agreement between the experimental and theoretical results can be noticed. In both cases, a good impedance matching (with respect to 10 dB return loss level) is predicted in the frequency range from 8 to 24 GHz (Figure 15),

![Figure 15](image1.png)

**Figure 15.** Simulated and measured input reflection coefficient of the antenna element shown in Figure 14.

![Figure 16](image2.png)

**Figure 16.** Realized gain of the antenna at 15 GHz: (a) co-polarized and (b) cross-polarized component.

![Figure 17](image3.png)

**Figure 17.** Normalized copolarized and cross-polarized patterns for the \( H \) plane (\( XOZ \) plane) at (a) 10 GHz, (b) 15 GHz, and (c) 25 GHz. The SMA connector has been included in the numerical model.
corresponding to a fractional bandwidth of 100%. The slight differences in the results are possibly caused by the additional parasitic element of the SMA (SubMiniature version A) connector, which is not explicitly included in the numerical model of the radiating structure.

5.2. Radiation Pattern

[22] A thorough investigation of the radiation properties of the antenna has been, also, carried out. In Figure 16, the three-dimensional view of the realized gain pattern at the center frequency $f = 15$ GHz is shown. The antenna element possesses a unidirectional radiation pattern with about 10 dB front-to-back ratio (FBR). The FBR can be further improved by extending the ground plane size. The half power beam width at 15 GHz is about 65° in the $E$ plane and about 90° in the $H$ plane. Furthermore, as expected from theory, Figure 16 reveals that cross-polarization level is minimum in the $E$ plane ($\varphi = 90^\circ$) and maximum in the $H$ plane ($\varphi = 0^\circ$).

![Figure 16](image)

**Figure 16.** Normalized copolarized patterns for the $E$ plane (YOZ plane) at (a) 10 GHz, (b) 15 GHz, and (c) 25 GHz.

![Figure 17](image)

**Figure 17.** Simulated $z$ component of Poynting’s vector flow at (a) 8 GHz, (b) 16 GHz, and (c) 25 GHz.
The normalized radiation patterns along the $H$ plane ($xz$ plane) and $E$ plane ($yz$ plane) of the antenna at three different frequencies are shown in Figures 17 and 18, respectively. An excellent agreement between the experimental and numerical results can be noticed. In particular, the measurements have been taken in DUCAT (Delft University Chamber for Antenna Test) by using a reference TEM horn (GZ0126ATP) as the transmit antenna. The radiation pattern of the antenna is moderately stable over the frequency range of interest with 2–6 dBi maximum realized gain level. At lower frequencies, the radiation pattern is slightly tilted in the $E$ plane due to the presence of the SMA connector as predicted by the numerical analysis (see Figure 18). Along the $E$ plane, the cross-polarization level is very small over the entire band (in practice, values around the numerical errors of simulations are observed) due to the antenna symmetry and therefore has not been plotted in Figure 18.

5.3. Poynting’s Vector Flow

In order to investigate the near-field behavior of an antenna, usually surface current distribution, electromagnetic field strength, or Poynting’s vector flow is analyzed. Poynting’s vector indicates how the energy of the radiated wave propagates around the antenna. In Figure 19, the distribution of the $z$ component of Poynting’s vector is depicted. It is clear from this figure that Poynting’s vector flow is dominated by the loop-like radiation contribution at the lower frequencies while the inner patch contributes more in the upper part of the operational band of the antenna.

5.4. Time Domain Behavior

The performance of an UWB system is heavily dependent on the time domain behavior of the radiated and received pulse [Montoya and Smith, 1996; Caratelli and Yarovoy, 2010]. In this section, the time domain characteristics of the element are investigated. To this end, the mono-pulse fired by the pulse generator (GZ1117DN-25), shown in Figure 20, has been used in all the measurements. For numerical simulations, the same transmit pulse has been used as excitation signal. The duration of the pulse is 0.1 ns for a level of 10\% of the peak amplitude. The power spectrum density (PSD) of the pulse is shown in Figure 21. At 25 GHz, the PSD is about $-30$ dB with respect to the maximum.

In order to estimate the radiated electric field distribution, two identical elements have been used in the measurement setup. In this way, the receiving transfer function of the individual antenna, $h_2(f,\omega)$, for a fixed position can be extracted from the measured coupling coefficient $S_{21}(f,\omega)$.
between the radiating elements by applying the following relation [McLean et al., 2007; Licul and Davis, 2005; Farr and Baum, 1992]:

\[ h_R(j\omega) = \sqrt{\frac{2\pi c_0 RS_{21}(j\omega)}{j\omega}} \eta_0, \tag{9} \]

where \( R \) is the distance between the antennas, \( \beta \) is the propagation constant, and \( c_0 \) denotes, as usual, the speed of light in free space. Afterwards, the normalized incident electric field strength, \( e^{inc}(j\omega) \), can be expressed in terms of the normalized received voltage, \( V_R(j\omega) \), as follows:

\[ e^{inc}(j\omega) = -\frac{V_R(j\omega)}{h_R(j\omega)} \eta_0, \tag{9} \]

\( \eta_0 \) being the wave impedance of the medium. By doing so, the magnitude of the copolarized component of the electric field has been evaluated for different scan angles. As it can be noticed in Figure 22, a good agreement between the experimental measurements and the numerical results has been achieved. A marginal disagreement in the amplitudes of the peaks has been noticed due to tolerances in the antenna manufacturing and nonidealities of the measurement setup.

[27] It is worth noting that the proposed antenna features a reasonably good fidelity factor [Lamensdorf and Susman, 1994] varying between 0.6 and 0.8 for different scan angles. Such performance is particularly important in UWB imaging applications where a reduced signal distortion in the scanning field of view is needed in order to increase the radar detectability of the target.

6. Conclusion

[28] A novel UWB antenna topology has been proposed and utilized in the development of an antenna element with 3:1 impedance bandwidth. This investigation confirms that proper combination of electric and magnetic topologies can increase the impedance bandwidth of a planar radiator to 100% which is a substantial improvement compared to 1 to 2% fractional bandwidth of a microstrip antenna. Besides the enormous operational bandwidth, this element can provide unidirectional pattern with stable radiation properties in time and frequency domains. In addition, the proposed antenna is planar and compact in size (half of the free-space wavelength at the center frequency). Therefore, the proposed radiator is an attractive candidate for radar-based imaging sensors. In this context, in order to ease the codesign of the antenna with the relevant RF front-end in a SPICE-like simulator, a suitable frequency-independent equivalent circuit has been derived. This circuit model provides an insight into the natural resonant processes which occur in the device and eventually affect the frequency response of the antenna.

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