

Near-field planar antenna for microwave excitation of paramagnetic quantum emitters

Stefano Lagomarsino^{1,*} and Mario Agio^{1,2,†}¹*National Institute of Optics of the National Research Council (CNR-INO), 50019, Sesto Fiorentino, Italy*²*Laboratory of Nano-Optics, University of Siegen, 57072, Siegen, Germany*

(Received 23 January 2026; revised 8 June 2026; accepted 23 July 2026; published 14 September 2026)

The concept of a microwave planar antenna for the excitation of dielectric paramagnetic media is proposed. The shape and size of the electrodes are designed for the generation of a uniform magnetic field parallel to the surfaces of a single- or multilayer dielectric, with a specific resonance frequency and bandwidth. One or both electrodes can be semireflective to allow the optical accessibility of the paramagnetic media and, possibly, to enhance the collection efficiency by beaming the optical emission of quantum emitters. This device is particularly useful in optically detected magnetic resonance due to the large bandwidth of the microwave excitation and favorable conditions for optical excitation and detection. Applications are devised especially in vector magnetometry and microwave spectral analysis because of the paramagnetic quantum properties of emitters such as the nitrogen-vacancy center in diamond.

DOI: [10.1103/rhmf-tf2r](https://doi.org/10.1103/rhmf-tf2r)

I. INTRODUCTION

Microwave excitation of quantum transitions in atoms, molecules, or defects in solids is useful in a variety of applications, such as manipulation of qubits, hyperfine transition spectroscopy, and optically detected magnetic resonance (ODMR). This last technique is useful both in materials science and in microsensing and nanosensing of several physical parameters, but primarily in vector magnetometry [1–2] and detection or spectroscopy of rf frequencies [3–4]. ODMR exploits the transitions induced by microwave (MW) radiation on paramagnetic systems subject to external magnetic fields. In the MW excitation of paramagnetic systems, the key problems include achieving uniform and sufficiently strong MW magnetic fields, optimizing bandwidth, and managing power consumption and heat dissipation, all while maintaining optical accessibility. In magnetometry, for instance, a uniform field distribution is crucial for sensitivity, which depends on the number of active color centers. In MW spectroscopy, an external field gradient must extend over a wide area, on which a uniform and intense exciting magnetic field guarantees a wide spectroscopic bandwidth. In these

applications, the MW magnetic field strikes a balance: strong enough to ensure good contrast without causing power broadening. Another crucial question is also compactness, which is significant to enable optical integration and device miniaturization.

Several kinds of geometries have been adopted to design antennas addressing these problems. For example, straight microwires [5] allow the generation of quite intense MW magnetic fields (max approximately 80 G at 1 W power, 50 Ω impedance, for a 10 μm -diameter wire), but they are strongly non-uniform. Ω -shaped wires [6,7] gain in uniformity ($B_z = 10\%$ in a 100 μm -wide central area of a 300 μm -diameter loop coil), but with loss of intensity (approximately 8 G at the center of the coil with 1 W power, 50 Ω impedance). More sophisticated architectures exploit the resonances of planar or coplanar structures fabricated either on the printed circuit board, where the sample is fixed on [8–12], or on the sample itself [13–15], by suitable deposition of metal layers. In this case, the quality factor of the antenna limits the bandpass to less than GHz and, consequently, the range of MW measurements, but it is possible to obtain higher, more uniform fields in a larger volume (order of 1 mm^3).

In general, planar solutions enhance field uniformity, but at the cost of limited bandwidth, while wire-based configurations provide a broader bandwidth, but with reduced field uniformity.

The device described in this article is designed to match wide bandwidth, field strength, and uniformity, maintaining full optical accessibility. This is allowed by a planar architecture given by two thin parallel metal electrodes,

*Contact author: stefano.lagomarsino@ino.cnr.it†Contact author: mario.agio@cnr.it

Published by the American Physical Society under the terms of the [Creative Commons Attribution 4.0 International](https://creativecommons.org/licenses/by/4.0/) license. Further distribution of this work must maintain attribution to the author(s) and the published article's title, journal citation, and DOI.

whose currents generate a nearly uniform magnetic field oscillating with a resonant frequency and bandwidth that can be tuned by convenient design of the shape, size, and thickness of the electrodes. The paramagnetic medium, e.g., a dielectric doped with paramagnetic centers, is placed between the two electrodes like a spacer. The semi-transparency of metallic thin electrodes, possibly joined to their beaming properties when included in a Fabry-Perot-like geometry [16], allows full optical accessibility of the dielectric layer between the antenna electrodes. Specifically, here we report solutions for the generation of a nearly uniform, 1–10 G strong MW magnetic field in the frequency range of at least 2 decades in the interval of 100 MHz–100 GHz, in a thin layer compared to the MW wavelength, e.g., 0.1–4 mm wide and 0.1–100 μm thick, permitting the optical observation of the entire volume interested in the magnetic field.

The planar antenna concept has been designed in two steps. First, a one-dimensional (1D) model based on the telegrapher's equation for an open line with a tapered profile has been applied. Tapered lines are usually employed to match lines with different impedance characteristics [17–20]. Here we are rather interested in how the taper can model the impedance, the bandpass, and the magnetic field profile in an open line acting as a near-field antenna. First, we have developed the analytical theory for a single exponential, open taper. Then, with a continuous succession of exponentially tapered profiles, an analytical solution has been found, and the impedance of the antenna, along with the magnetic field distribution along the line as a function of a one-dimensional position, has been calculated as a function of frequency. As a last step, a more rigorous full-3D simulation of the behavior of the antenna has been performed with COMSOL Multiphysics®, and its behavior in terms of bandpass and distribution of the magnetic field has been compared with the analytical model.

II. DESIGN OF THE ANTENNA

In principle, an open rectangular planar waveguide of suitable length, fed at one of the ends, can resonate at whatever frequency, generating a transverse nearly uniform magnetic field in the central region. Nevertheless, the frequencies normally involved in ODMR of nitrogen-vacancy (NV) centers in diamond (around 3 GHz) would require a length of about 2 cm, which is quite impractical: commercial diamond monocrystals do not exceed a few millimeters in size. The resonance of a millimeter-sized waveguide around the gigahertz frequency range requires a structure resembling an LRC circuit, whose scheme is presented in Fig. 1: a first part with a low cross-section w , acting mostly as an inductor with inductance per unit length $L = \mu_0 \frac{d}{w}$ (d is the thickness of the dielectric layer); a second rectangular part with a resistive character, generating the nearly uniform magnetic field between

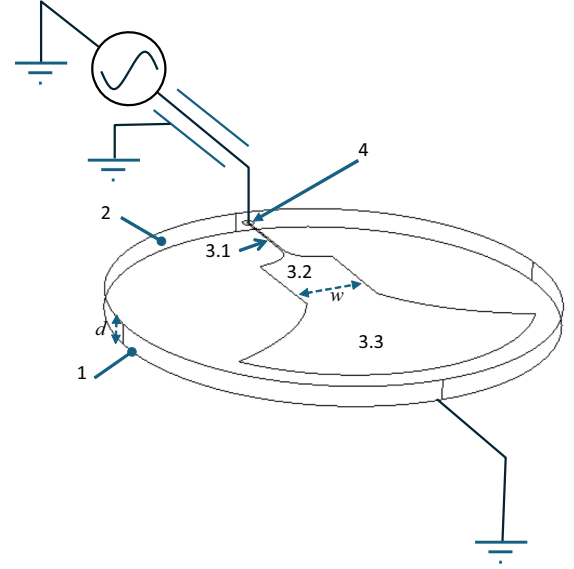


FIG. 1. Scheme of the near-field planar antenna: 1. Ground electrode. 2. Dielectric layer (containing the paramagnetic centers). 3.1. Inductive part of the waveguide. 3.2. Resistive section generating the nearly uniform magnetic field. 3.3 Capacitive part. w is the width of the waveguide, depending on the position along the guide, d is the interelectrode distance.

the electrodes, with a resistance per unit length given by $R = \frac{\rho}{w} \frac{th_{\text{ground}} + th_{\text{phase}}}{th_{\text{ground}} \cdot th_{\text{phase}}}$ (ρ is the resistivity and $th_{\text{ground,phase}}$ the thicknesses of the ground and phase electrodes, respectively); and a third large-area part with a capacitive role, whose capacitance per unit length is $C = \epsilon_r \cdot \epsilon_0 \frac{w}{d}$ (ϵ_0 and μ_0 have the usual meanings).

Of course, each of the three parts shares inductive, resistive, and capacitive characteristics to be properly considered for the calculation of the currents at each point of the waveguide as a function of frequency. Nevertheless, a rough calculation based on concentrated constants can be made, whose results can be corrected and refined with more rigorous simulations. For instance, assuming the lengths of the inductive, resistive, and capacitive sections are 0.4–0.5 mm and 0.75 mm, respectively; their width w is 0.01, 0.4, and 1.7 mm, respectively; the thicknesses of the gold electrodes are 15 and 100 nm, respectively; and the diamond thickness is $d = 5 \mu\text{m}$, we have a real impedance of about 50Ω (the impedance has been calculated by taking also into account the resistivity of the inductive section) at a resonant frequency of around 3 GHz (the zero-magnetic-field transition frequency between the spin sublevels of the ground state is 2.87 GHz), and a 3-dB-bandpass as high as 30 GHz. At resonance, a 200-mA alternating current (1 W of dissipated power) gives a maximum magnetic field in the central section of about 6.3 G. Thus, we have a resonant structure less than 2 mm long (instead of 20 mm for a rectangular waveguide), with a much wider bandpass.

A more rigorous calculation of the parameters of the antenna will be made in two steps:

- A) Adopting a linear-waveguide approximation, we will solve analytically the equations for a succession of exponentially tapered sections. For instance, for the structure shown in Fig. 1, we will have at the beginning a nearly constant-section inductive part, followed by a rapidly increasing matching section to the following resistive part, whose slightly decreasing exponential profile compensates for the decreasing current to generate a uniform magnetic induction intensity, followed by an exponentially increasing-width part with a mostly capacitive role.
- B) After having designed the antenna with the desired impedance, resonant frequency, and bandpass with step 1, a full-3D COMSOL Multiphysics® simulation will be performed for a further refinement of the results.

Step A, being semianalytical in nature, is much faster in finding the correct working point of the system, but step B is necessary because when the rate of growth of w approaches unity, the current is not constant across the section of the guide, and the 1D model fails.

III. STEP A: 1D-MODEL OF THE WAVEGUIDE

The Telegrapher's equation for a tapered, lossy line along the x direction is

$$\frac{\partial^2 I}{\partial x^2} - \frac{1}{v^2} \frac{\partial^2 I}{\partial t^2} - \frac{1}{\Lambda} \frac{\partial I}{\partial x} - \frac{1}{D} \frac{\partial I}{\partial t} = 0. \quad (1)$$

In Eq. (1), $v = 1/\sqrt{LC}$ is the velocity of the electromagnetic radiation in the dielectric media, $D = 1/(RC)$ is a diffusion constant, and $\Lambda = C/(\frac{\partial C}{\partial x})$ is a length scale depending essentially on the rate of growth of w with respect to the coordinate x . If Λ is constant in a given interval, w has an exponential dependence on x :

$$w(x) = w_0 \exp\left(\frac{x}{\Lambda}\right), \quad (2)$$

and Eq. (1) can be solved analytically, in the frequency domain, as a sum of two damped traveling waves propagating in opposite directions:

$$I(x, t) = I_+ e^{\left(\frac{1}{2\Lambda} - l\right)x} e^{j(\omega t - kx)} + I_- e^{\left(\frac{1}{2\Lambda} + l\right)x} e^{j(\omega t + kx)}, \quad (3)$$

with k and l equaling

$$k = \sqrt{\frac{\sqrt{\left(\frac{\omega^2}{v^2} - \frac{1}{4\Lambda^2}\right)^2 + \left(\frac{\omega}{D}\right)^2} + \left|\frac{\omega^2}{v^2} - \frac{1}{4\Lambda^2}\right|}{2}},$$

$$l = \sqrt{\frac{\sqrt{\left(\frac{\omega^2}{v^2} - \frac{1}{4\Lambda^2}\right)^2 + \left(\frac{\omega}{D}\right)^2} - \left|\frac{\omega^2}{v^2} - \frac{1}{4\Lambda^2}\right|}{2}}. \quad (4)$$

The behavior of an antenna made with a single exponential profile of Eq. (2) is completely defined by the boundary conditions $I(0, t) = I_0 \cdot e^{j\omega t}$ and $I(L, t) = 0$ (open line). The magnetic field along the line, perpendicular to the x axis, is given by $B(x, t) = \frac{\mu_0}{w(x)} I(x, t)$, and the impedance of the antenna is $Z = \frac{V(0, t)}{I(0, t)} = \frac{-1}{j\omega C} \frac{\partial I}{\partial x} / I|_{x=0}$, since the electric potential V varies along the line according to $V(x, t) = \frac{-1}{j\omega C} \frac{\partial I}{\partial x}$. As an example, the impedance of a single-exponentially tapered antenna of length L , length scale, initial width w_0 , takes the form $Z = Z_R + jZ_I$, with

$$Z_R = \frac{d}{2\omega\epsilon_0\epsilon_r w_0 \Lambda} \cdot \frac{\beta \sinh\left(\alpha \frac{L}{\Lambda}\right) - \alpha \sin\left(\beta \frac{L}{\Lambda}\right)}{\cosh\left(\alpha \frac{L}{\Lambda}\right) - \cos\left(\beta \frac{L}{\Lambda}\right)}, \quad (5)$$

$$Z_I = \frac{d}{2\omega\epsilon_0\epsilon_r w_0 \Lambda} \cdot \left[1 - \frac{\alpha \sinh\left(\alpha \frac{L}{\Lambda}\right) + \beta \sin\left(\beta \frac{L}{\Lambda}\right)}{\cosh\left(\alpha \frac{L}{\Lambda}\right) - \cos\left(\beta \frac{L}{\Lambda}\right)} \right], \quad (6)$$

and

$$j\alpha - \beta = \sqrt{\left(\frac{4\Lambda^2\omega^2}{v^2} - 1\right) - 4j \frac{\epsilon_0\epsilon_r\Lambda^2}{\sigma \cdot d \cdot th}}. \quad (7)$$

Following Eqs. (5)–(7) it is possible to fix the parameters for a device operating within given geometrical and physical constraints. Figure 2 plots, for different values of the equivalent electrode thickness, the length of an antenna

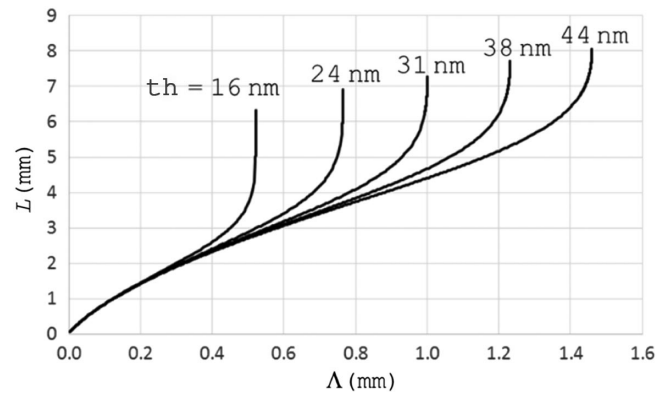


FIG. 2. For different values of the equivalent thickness of the gold electrodes, the length of an exponentially tapered antenna resonating at 2.87 GHz is shown as a function of the length scale Λ .

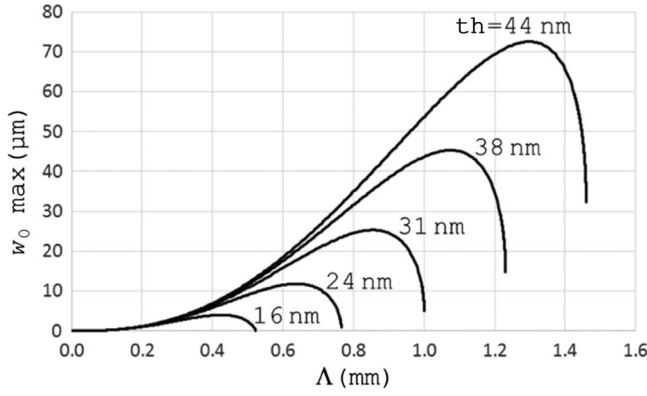


FIG. 3. For different values of the equivalent thickness of the gold electrodes, the maximum initial width w_0 permitting $\frac{dw}{dx} < 1$ of an exponentially tapered antenna resonating at 2.87 GHz is shown as a function of the length scale Λ .

whose impedance is purely real at 2.87 GHz as a function of the length scale Λ .

In Fig. 3, for the same values of the equivalent thickness, the maximum value of the initial width w_0 is shown that fulfills the condition $\frac{dw}{dx} < 1$, for an antenna resonating at 2.87 GHz. In Fig. 4, for the values of L and w_0 given by Figs. 2 and 3, the impedance of the device is plotted against the scale Λ .

Figure 5 shows the map of the magnetic field generated along the median plane of the membrane by a single-exponentially tapered antenna with a gold thickness of 24 nm and a 50Ω real impedance at 2.87 GHz. The nonuniformity of the magnetic field along the antenna profile is apparent, making this device suitable in ODMR only if a gradient in the magnetic field is not detrimental. To obtain uniform magnetic fields on relatively large areas (typically, in magnetic field measurement), the taper must include a section with nearly uniform width.

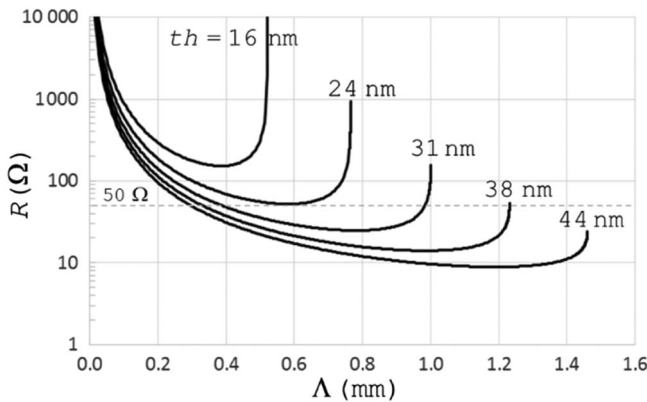


FIG. 4. For each value of the length scale Λ , the length L is calculated giving a real impedance, along with the value of w_0 of the antenna giving $\frac{dw}{dx} < 1$, and the corresponding value of the resistance R . The plot is drawn for several values of the equivalent thickness $th = \frac{th_{\text{phase}} \times th_{\text{ground}}}{th_{\text{phase}} + th_{\text{ground}}}$ of the gold electrodes.

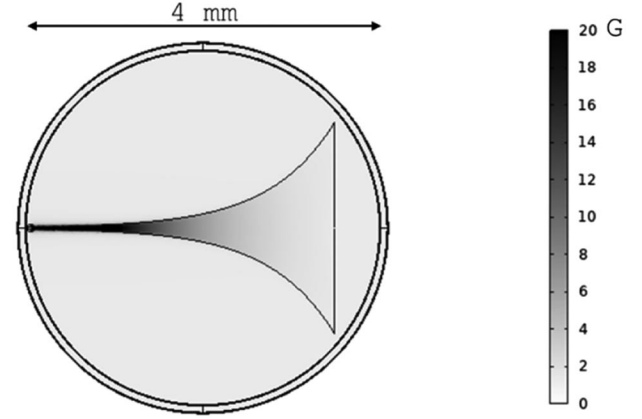


FIG. 5. Map of the magnetic field intensity along an antenna designed to have a 50Ω impedance at 2.87 GHz, dissipating 1 W of electric power, with gold electrodes having an equivalent thickness of 24 nm. The initial width of the profile is $w_0 = 10 \mu\text{m}$.

Suppose the profile of a tapered line is defined as in Eq. (2) on each of the elements of the connected union of disjoint intervals (Fig. 6 gives a relevant example with four intervals). Thus, the current $I(x, t)$ is expressed, on each interval, according to Eq. (3), with different values of k , l , and $I_{\pm}$ for each domain. The values of $I_{\pm}$ have to satisfy the boundary conditions between adjacent intervals, which are the continuity of both the current $I(x, t)$ and the potential $V(x, t) = \frac{-1}{j\omega C} \frac{\partial I}{\partial x}$, along with $I(0, t) = I_0 \cdot e^{j\omega t}$ and $I(L, t) = 0$ (open line). Analytical expressions for the fields and the impedance are attainable, though not in as closed and simple a form as Eqs. (5)–(7), and it is possible to design the geometry of an antenna of given characteristics of resonant frequency, impedance, and generated magnetic field.

IV. STEP B: THREE DIMENSIONAL COMSOL SIMULATIONS

The validity of the Telegrapher's equation, Eq. (1), requires a constant current density along the transverse cross section of the waveguide. This approximation is justified if the rate of growth of the profile cross section $\frac{dw}{dx}$ is much less than unity along all the guide length, that is, for a taper described by Eq. (2), if $w(x) \ll \Lambda$. If it is not so, the flux lines of the current tend to form large angles with the x axis, and a more rigorous full 3D calculation must be made. For this study, we have performed the calculation of the electric and the magnetic fields in the space between and around the electrodes by means of COMSOL Multiphysics® software. We have simulated the diamond membrane between the electrodes as an insulating layer with a dielectric constant $\epsilon_r = 5.7$ independent of frequency, surrounded by vacuum, modeling the gold electrodes deposited on the membrane by means of two-dimensional transition boundary conditions, specifying the conductivity ($4.1 \times 10^7 \text{ S/m}$), and a

thickness equal to the physical thickness of the electrodes (considering that the skin depth of gold at the frequencies of interest is much larger than the physical thickness).

The phase electrode has a shape described by a continuous succession of profiles as in Eq. (2), but with the “Continuous first derivative” option to attenuate convergence problems at the corners. For the ground electrode, two solutions have been simulated: the first with an electrode identical to the phase electrode and facing it. The second with an infinitely extended conductive layer. The first solution presents constructive complications due to the necessity of mask alignment at the micrometer level on the two sides of the membrane. Field uniformity and response at low frequencies, on the other hand, tend to be significantly different in the two cases due to the poorer confinement effect of the second solution, especially at low frequencies. The feed is applied by means of a small cylinder (50- μm diameter) crossing the volume of the diamond layer, in which an alternating current with fixed amplitude and frequency is forced to pass. The corresponding voltage is calculated by integrating the electric field between the faces of the cylinder, and the current is normalized to maintain a constant amplitude for the feed voltage at every frequency.

For each frequency, the corresponding antenna impedance and the value of the magnetic field at every point of the dielectric media have been calculated.

V. A CASE STUDY

We proceed to design an antenna with a resonant frequency of about 3 GHz, with a bandpass of at least 30 GHz, generating a magnetic field in a central region exceeding 5 G, with good uniformity. A succession of exponentially tapered profiles will be studied. Here we consider the profile given in Fig. 6, whose shape has been fixed in a way that the inductance, the resistance, and the capacitance, calculated by integrating the distributed constants along the antenna, determine (with the simple LRC model) the desired characteristics of frequency and bandpass, with a thickness of the diamond membrane of 5 μm , and ground- and phase-electrode thicknesses of 15 and 100 nm, respectively.

The values of the concentrated constants are $R = 50 \Omega$, $L = 0.28 \text{ nH}$, and $C = 0.11 \text{ pF}$. Since at low frequencies the current tends to decrease along the guide, a bottleneck has been designed in the central region to make it constant as possible with respect to the superficial current density (and the magnetic field). The geometrical parameters defining sections AB, BC, CD, and DEF are specified in Table I.

We expect that, at sufficiently low frequencies, the complex impedance of the antenna resembles that of an *RLC* circuit. Figure 7 reports the real and the imaginary parts of the antenna impedance in the hundreds-of-megahertz to hundreds-of-gigahertz range, calculated by means of the 1D model deduced by the Telegraphers’ equation. Two different

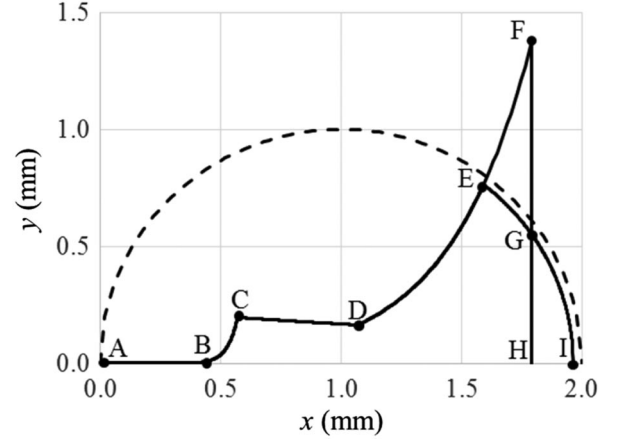


FIG. 6. Plot of the function describing half of the profile of the antenna (the other half is symmetric with respect to the x axis). The 1D simulation is performed considering the curve ABCDEFGH; the 3D COMSOL Multiphysics[®] simulation considers the profile ABCDEGI so as to fit the shape of a circular diamond membrane of 2 mm in diameter (see the dashed line). The two profiles, which are quite different in the terminal part, are fixed in a way that the area EFG is equivalent to GIH so that the capacitance of the terminal parts is nominally the same. The parameters describing the curve are reported in Table I.

regimes are clearly evidenced: an *RLC*-like regime, up to about 20 GHz, with a nearly constant real part R and an imaginary part approximated by $(\omega L - \frac{1}{\omega C})$. A stationary-wave regime then appears, clearly setting at about 30 GHz, when the wavelength approaches twice the length of the antenna.

As a matter of fact, instead of diverging at high frequencies, the impedance of the device oscillates, also reaching lower values than those typical of the first resonance frequency. It is worth exploring the possibility that the bandpass of the antenna, with respect to the magnetic field produced in the central part, is larger than that predicted by the low-frequency model.

TABLE I. Parameters defining the geometrical shape of the antenna. The profile given by ABCDEFGH in Fig. 6 represents a succession of exponentially tapered guides modeled with the 1D model; the profile ABCDEGI is terminated by the circular section EGI, whose radius is 1 mm. This last profile (used in the 3D-COMSOL Multiphysics[®] simulations) fits within a diamond membrane of 2 mm in diameter and has the same surface as the previous one.

Section	x_{initial} (mm)	w_{initial} (mm)	Λ (mm)
AB	0	0.01	4
BC	0.4	0.011	0.042
CD	0.55	0.393	−2.8
DEF	1.05 (final 1.8)	0.329 (final 1.5)	0.34
EGI	1.6 (final 2.0)	(arc of circle radius 1 mm)	

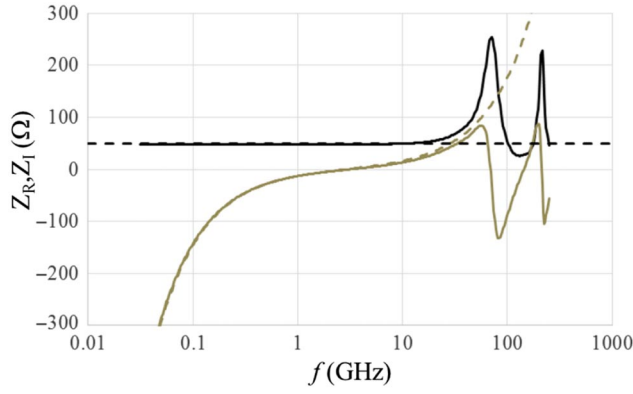


FIG. 7. Plot of the real (black curve) and the imaginary part (gray curve) of the antenna impedance as a function of frequency. Dashed lines: the values calculated with the RLC approximation, with $R = 50$, $L = 0.28$ nH and $C = 0.11$ pF.

The physical quantities that are important here are the strength and the uniformity of the magnetic field in the region limited by the boundary CD in Fig. 6, assuming that the antenna is excited with a constant alternating potential that, at the resonant frequency, produces a fixed power dissipation of 1 W.

Figure 8 reports the amplitudes of the oscillating magnetic field along the planar antenna (on a logarithmic scale) calculated for selected frequencies, both in the RLC -like and stationary-wave regimes. It is apparent that at very high frequencies the strength of the magnetic field (which is greater than 6 G in the central part at the resonance) can be higher than that at low frequencies.

In fact, if the length of the antenna is longer than about one-quarter of the wavelength, the feeding current excites modes having a maximum inside the planar antenna, but in this case the maximum can be narrow, at the expense of uniformity. Consequently, it is convenient to consider the actual performance of the planar antenna in terms of both

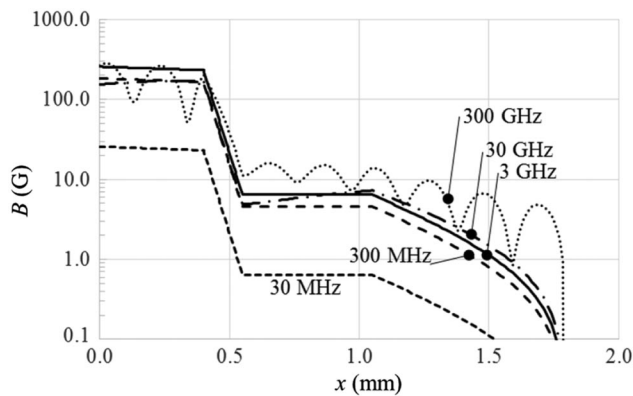


FIG. 8. Plot of the magnetic field as a function of the position along the planar antenna, applying the potential difference that produces, at the resonant frequency (3 GHz), the power dissipation of 1 W.

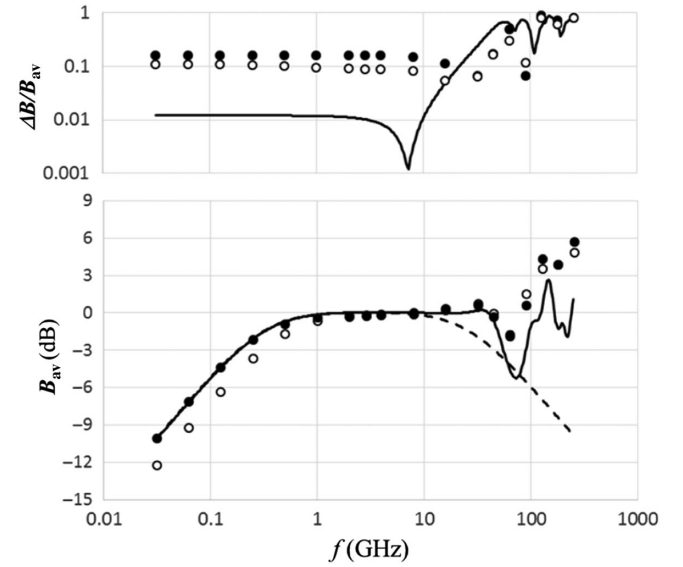


FIG. 9. Performance of the case-study planar antenna in terms of the average magnetic induction field in the central region (lower panel) and the uniformity (upper panel). Solid lines: analytical 1D model. Black dots: 3D COMSOL Multiphysics[®] simulation with the two electrodes identical to each other. White dots: 3D COMSOL Multiphysics[®] simulation with an infinitely extended ground electrode. The identical-electrode antenna exhibits better bandpass, but slightly poorer field uniformity.

the average magnetic field, B_{av} , and the ratio between the maximum oscillation and the average value, $\frac{\Delta B}{B_{av}}$.

These two quantities are reported in Fig. 9 as a function of frequency. At the level of the 1D simulations (solid lines), the low-frequency regime (up to two octaves over the resonance) is consistent with the RLC model, whereas the uniformity is assured below the level of a few percent. At higher frequencies, the performance in terms of average magnetic field remains unaltered, well behind the limit of the RLC model, but the uniformity progressively degrades.

Further insight into the performance of the planar antenna is given by the 3D simulations. COMSOL Multiphysics[®] simulations have been applied at a profile described by the line ABCDEGI in Fig. 6 rather than ABCDEFGH, in a way to fit the actual size of a 2-mm-diameter diamond membrane. The area of the EGH surface has been designed to be equal to that of HGI, thereby leaving the capacitance of the last section unaltered.

Figure 10 reports in gray tones the map of the magnetic field intensity at the median plane between the two electrodes at two specific frequencies. The scale is saturated above 10 G to show the behavior, especially in the central region.

In Fig. 9, the average magnetic field and the amplitude of the relative oscillations are also reported, calculated by means of the 3D simulations.

The identical-electrode case is more exactly approximated by the 1D analytical model and follows its

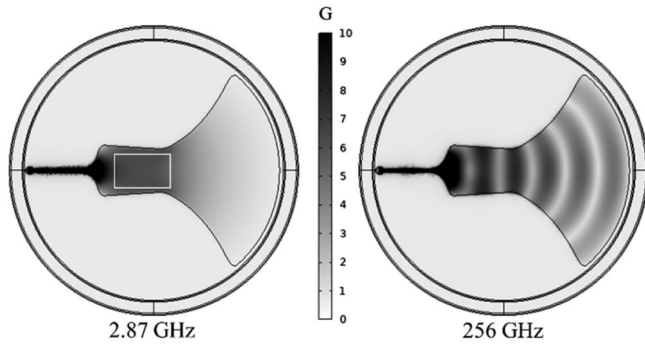


FIG. 10. Map of the magnetic-field amplitude on the median plane of the planar antenna for two different values of the frequency: (left panel) 2.87 GHz, at resonance, corresponding to the energy difference between the spin sublevels of the ground states of the NV color center of diamond, which is in the *RLC*-like range; (right panel) 256 GHz, deep in the stationary-wave range. The gray-tone mapping saturates at 10 G, so the map of the thin initial inductive section is not accurate.

prediction very well from low frequencies and up to about four octaves over the resonance, well beyond the validity of the *RLC* model, which is limited to a couple of octaves (dashed line in Fig. 9). On the other hand, the extended-ground electrode case shows a low-frequency cut because of the worst confinement of the magnetic field at low frequencies. Nevertheless, the actual uniformity of the magnetic field is slightly better for the planar antenna with the extended-ground electrode, as seen from the upper panel of Fig. 9.

In Fig. 11, a map of the magnetic field intensity at the resonant frequency of 2.87 GHz (*y* component) is also shown along the vertical symmetry plane (*xz* in the figure) and along a plane parallel to the *yz* plane, 150 μm apart from the center of the circular membrane. It is apparent that the field is highly uniform along the *z* direction, while the variation along the *xy* plane is within the limits

specified by Fig. 9. As to the comparison between the 1D analytical model and the full-3D simulation, it is apparent from Fig. 10 that in the region where $w(x)$ is comparable with or greater than Λ , the isolines of the magnetic field are not actually vertical as assumed in the 1D model. Especially at the beginning of the central part limited by CD (see Fig. 6), this alters the distribution of the field, so the average field B_{av} and the oscillation ΔB have been calculated on a more limited region, evidenced by a white rectangle in the figure.

This deformation also affects the resistance of the device. Since the currents tend to concentrate along the axis of the system, the impedance is also higher. To correct this effect, a thickness of 100 nm instead of 15 nm has been attributed to an initial section of the inductive part AB so as to restore the 50 Ω impedance at resonance. From the constructive point of view, this expedient does not introduce too much complication; it is sufficient to prolong the metal deposition on the first part of the guide by interposing a screen in front of the second section. The length of this initial section is 210 μm for the identical-electrode case and 150 μm for the extended-ground-electrode case.

From Fig. 10, it is also apparent the effect of the stationary-wave regime at high frequencies. The geometric resonance of this structure makes the current along the various parts of the device much higher than that characteristic of the low-frequency regime, but each maximum is quite narrow, and the uniformity does not extend to the entire central section of the device.

VI. CONCLUSION

A concept for a MW near-field antenna for the stimulation of paramagnetic quantum systems has been presented, allowing optical addressing of the region between the electrodes and good uniformity and strength of the oscillating magnetic field over a wide MW spectral range.

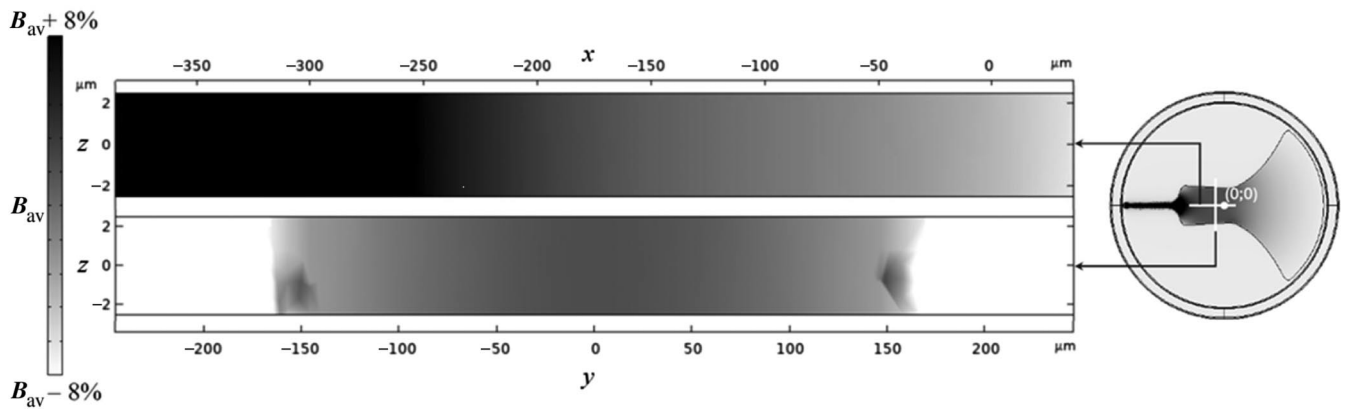


FIG. 11. Map of the magnetic field intensity variations (component *y*) along the *xz* plane (upper panel) between the metal electrodes and along the *yz* plane (lower panel) at the frequency of 2.87 GHz, as calculated by COMSOL Multiphysics® simulation. The *z* scale is tenfold expanded for convenience. The grayscale is chosen to highlight around 20% of global variation. Excluding a restricted area near the edges of the electrodes, where boundary effects occur, the field variation along *z* is negligible.

The case study presented here is particularly useful for NV centers in diamond, permitting on the footprint of commercial samples a bandpass exceeding 2 decades in the 100 MHz–100 GHz interval. Both the bandpass and magnetic field intensity are significantly greater than those reached with a MW antenna external to the sample, at the same level of power consumption, permitting a superior degree of integration. The fabrication of a device of the kind depicted in this article depends on the availability of diamond membranes several millimeters large and a few micrometers thick, which are readily available for polycrystalline materials (see Ref. [21] and related references) but are more challenging for single crystals [22]. However, principles and methods have been presented allowing the application of this concept also to different ranges of intensity and frequency and other materials, such as silicon carbide, available also in large, thin samples, varying the size and shape of the planar antenna and the dielectric thickness.

ACKNOWLEDGMENTS

This work received funding from the European Defence Fund (EDF) under grant agreement 101103417—project ADEQUADE. Funded by the European Union. Views and opinions expressed are, however, those of the author(s) only and do not necessarily reflect those of the European Union or the European Commission. Neither the European Union nor the granting authority can be held responsible for them.

Patent Pending 102025000018688.

DATA AVAILABILITY

The data that support the findings of this article are not publicly available. The data are available from the authors upon reasonable request.

- [1] G. Balasubramanian, I. Y. Chan, R. Kolesov, M. Al-Hmoud, J. Tisler, C. Shin, C. Kim, A. Wojcik, P. R. Hemmer, A. Krueger, T. Hanke, A. Leitenstorfer, R. Bratschitsch, F. Jelezko, and J. Wrachtrup, Nanoscale imaging magnetometry with diamond spins under ambient conditions, *Nature (London)* **455**, 648 (2008).
- [2] P. Reuschel, M. Agio, and A. M. Flatae, Vector magnetometry based on polarimetric optically detected magnetic resonance, *Adv. Quantum Technol.* **5**, 2200077 (2022).
- [3] M. Ludovic and D. Thierry, Nitrogen-vacancy centers in diamond for instantaneous spectral analysis in the radio-frequency domain up to 18 GHz, in *2018 International Topical Meeting on Microwave Photonics (MWP), Toulouse, France* (2018), pp. 1–4.
- [4] Y. Liu, Y. Qin, Q. Wang, H. Guo, J. Li, Z. Li, H. Wen, Z. Ma, J. Tang, and J. Liu, Microwave spectrum detection at microscopic scale based on nitrogen vacancy center in diamond, *Phys. Status Solidi RRL* **17**, 2200498 (2023).
- [5] N. D. Lai, D. Zheng, F. Jelezko, F. Treussart, and J.-F. Roch, Influence of a static magnetic field on the photoluminescence of an ensemble of nitrogen-vacancy color centers in a diamond single-crystal, *Appl. Phys. Lett.* **95**, 133101 (2009).
- [6] P. L. Stanwix, L. M. Pham, J. R. Maze, D. Le Sage, T. K. Yeung, P. Cappellaro, P. R. Hemmer, A. Yacoby, M. D. Lukin, and R. L. Walsworth, Coherence of nitrogen-vacancy electronic spin ensembles in diamond, *Phys. Rev. B* **82**, 201201 (2010).
- [7] M. Chipaux, A. Tallaire, J. Achard, S. Pezzagna, J. Meijer, V. Jacques, J.-F. Roch, and T. Debuisschert, Magnetic imaging with an ensemble of nitrogen-vacancy centers in diamond, *Eur. Phys. J. D* **69**, 166 (2015).
- [8] M. Mrózek, J. Mlynarczyk, D. S. Rudnicki, and W. Gawlik, Circularly polarized microwaves for magnetic resonance study in the GHz range: Application to nitrogen-vacancy in diamonds, *Appl. Phys. Lett.* **107**, 013505 (2015).
- [9] K. Sasaki, Y. Monnai, S. Saijo, R. Fujita, H. Watanabe, J. Ishi-Hayase, K. M. Itoh, and E. Abe, Broadband, large-area microwave antenna for optically detected magnetic resonance of nitrogen-vacancy centers in diamond, *Rev. Sci. Instrum.* **87**, 053904 (2016).
- [10] K. Bayat, J. Choy, M. Farrok Baroughi, S. Meesala, and M. Loncar, Efficient, uniform, and large area microwave magnetic coupling to NV centers in diamond using double split-ring resonators, *Nano Lett.* **14**, 1208 (2014).
- [11] R. Pellicer-Guridi, K. Custers, J. Solozabal-Aldalur, A. Brodolin, J. T. Francis, M. Varga, A. Mongelos, J. Casanova, M. M. Paulides, and G. Molina-Terriza, Versatile quadrature antenna for precise control of large electron spin ensembles in diamond, *Adv. Quantum Technol.* **8**, 2400142 (2025).
- [12] O. R. Opaluch, N. Oshnik, R. Nelz, and E. Neu, Optimized planar microwave antenna for nitrogen vacancy center based sensing applications, *Nanomaterials* **11**, 2108 (2021).
- [13] G. De Lange, T. van der Sar, M. Blok, Z.-H. Wang, V. Dobrovitski, and R. Hanson, Controlling the quantum dynamics of a mesoscopic spin bath in diamond, *Sci. Rep.* **2**, 382 (2012).
- [14] F. Münzhuber, F. Bayer, V. Marković, J. Brehm, J. Kleinlein, L. W. Molenkamp, and T. Kiessling, Polarization-assisted vector magnetometry with no bias field using an ensemble of nitrogen-vacancy centers in diamond, *Phys. Rev. Appl.* **14**, 014055 (2020).
- [15] C. Zhang, H. Yuan, N. Zhang, L. X. Xu, B. Li, G. D. Cheng, Y. Wang, Q. Gui, and J. C. Fang, Dependence of high-density nitrogen-vacancy center ensemble coherence on electron irradiation doses and annealing time, *J. Phys. D* **50**, 505104 (2017).
- [16] S. Checcucci, P. Lombardi, S. Rizvi, F. Sgrignuoli, N. Gruhler, F. B. C. Dieleman, F. S. Cataliotti, W. H. P. Pernice, M. Agio, and C. Toninelli, Beaming light from a quantum emitter with a planar antenna, *Light Sci.* **6**, e16245 (2017).
- [17] R. E. Collin, The optimum tapered transmission line matching section, *Proc. IRE* **44**, 539 (1956).

- [18] R. W. Klopfenstein, A transmission line taper of improved design, *Proc. IRE* **44**, 31 (1956).
- [19] A. Toscano, L. Vegni, and F. Urbani, Exponentially tapered nonuniform transmission lines for high-pass filter design, *Microwave Opt. Technol. Lett.* **16**, 227 (1997).
- [20] F. Vega, N. Mora, F. Rachidi, N. Peña, and F. Roman, Design and simulation of a coaxial exponential transmission line for a half impulse radiating antenna, in *Proceedings of the 2011 XXXth URSI General Assembly and Scientific Symposium, Istanbul, Turkey* (2011), pp. 1–4.
- [21] A. M. Flatae, F. Sledz, H. Kambalathmana, S. Lagomarsino, H. Wang, N. Gelli, S. Sciortino, E. Wörner, C. Wild, B. Butz, and M. Agio, Single-photon emission from silicon-vacancy color centers in polycrystalline diamond membranes, *Appl. Phys. Lett.* **124**, 094001 (2024).
- [22] J. Heupel, M. Pallmann, J. Körber, D. Hunger, J. P. Reithmaier, and C. Popov, Fabrication of high-quality thin single-crystal diamond membranes with low surface roughness, *Phys. Status Solidi A* **220**, 2200465 (2023).