

# Loop Antennas for Use On/Off Ground Planes

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**Abstract**— Many applications benefit from the ability of an RFID tag to operate both on and off a conducting ground plane. This paper presents an electrically small loop antenna at 433 MHz that passively maintains its free-space tune and match when located a certain distance away from a large conducting ground plane. The design achieves this using a single radiation mechanism (that of a loop) in both environments without the use of a ground plane or EBG/AMC structure. An equivalent circuit model is developed that explains the dual-environment behavior and shows that the geometry balances inductive and capacitive parasitics introduced by the ground plane such that the free-space loop reactance, and thus resonant frequency, does not change. A design equation for balancing the inductive and capacitive parasitic effects is derived. Finally, experimental data showing the design eliminates ground plane detuning in practice is presented. The design is suitable for active, “hard” RFID tag applications.

**Index Terms**—Antennas, Dual-environment, Loop antenna, Metal tag, On/Off metal, Platform-tolerant, Radiofrequency identification (RFID)

## I. INTRODUCTION

Many reported UHF RFID antenna designs are suitable for use on a conducting ground plane [1], and many RFID applications further benefit from the ability to operate *both* near to as well as away from a ground plane. This feature is known by various names in the literature: “platform tolerant,” “dual-environment,” “on/off metal,” etc. Many such designs have been reported, for example [2]-[32] and [41]-[44]. Designs based on dipoles ([2]-[11]) and PIFAs ([12]-[21]) are most common. Patch-based designs are described in [22]-[32]. Some designs, such as [4], [29], [44] operate as one type of radiator in free-space and as another type when near ground.

Most works surveyed [2]-[44] have a height about 1% of the operating wavelength  $\lambda_0$ ; only five have heights greater than  $0.03\lambda_0$ . Low-profile designs may not address the need for “hard” RFID tags that are used to track large items like shipping containers, railroad cars, and industrial machinery. Due to the large size of these items, longer read range—and thus higher antenna performance—is required. Accordingly, commercial hard RFID tags [45]-[49] have higher profile, often about

$0.03\lambda_0$ , with more robust packaging and higher cost than small, light and flexible “soft” tags used for tracking small items.

Compared to other antenna types, loops are relatively uncommon in the far-field UHF RFID literature for either on-metal only [38]-[40] or dual-environment applications [41]-[44]. However, a popular design comprising two PIFAs whose radiating edges face each other [16]-[21] (platform-tolerant) and [33]-[35] (not platform-tolerant) is similar to a loop antenna in some regards, as are “folded patch” designs [22] and [23], “high impedance unit” cell [36] and “looped bowtie” [37] designs. Because electrically small loop antennas interact with the magnetic field, they are a natural choice for use against a conducting ground plane—where tangential magnetic fields are doubled [41]. Unfortunately, detuning of loop antennas near conducting ground planes is known to be problematic for dual-environment applications [50]. Here we present a simple loop antenna that balances inductive and capacitive parasitics introduced by a ground plane such that the free-space tune and match change little when the antenna is located a fixed distance away from a ground plane. In practical hard tags (e.g., [45]-[49]), this distance is already naturally fixed by a dielectric radome.

This work contributes to the RFID literature by 1) giving a simple antenna geometry that performs well in both free-space and near a conducting plane, 2) developing a simple equivalent circuit model that explains the observed dual-environment behavior and 3) deriving a general design equation from the equivalent circuit that allows others to easily scale the concept to their own applications.

The design presented in this paper is about  $0.03\lambda_0$  in height—similar to [45]-[49]—and is suitable for 50  $\Omega$  active RFID “hard” tag applications in the 433 MHz ISM band. However, the technique presented here may be easily adapted to 915 MHz passive RFID “soft” tags. For example, because loop antennas are *inductive*, it is easy to adjust the matching network for conjugate match to passive RFID chips with capacitive input impedances—less matching capacitance is simply used than in the 50  $\Omega$  case [42]. Moreover, the loop antenna may be etched on a flexible inlay that is folded around a foam support as in [15], [22], [23] and [40]. Finally, the discrete matching capacitors in this work may be easily integrated into a two-layer etched inlay as in [42]. This work is similar to [41] in some regards, however, this work presents an equivalent circuit model, along with a design equation, to

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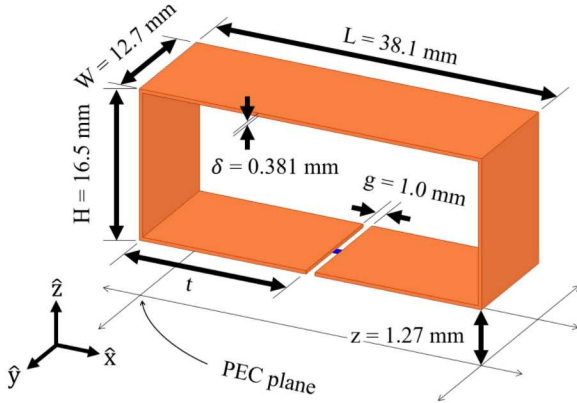


Figure 1. Copper ( $\sigma=58\text{MS/m}$ ) loop antenna above an infinite metal ground plane at  $z=0$ ; the loop is fed at the  $g=1.0$  mm gap.  $t$  is set in Sections II & III. An L-section matching network is used to tune and match the loop for a free-space ( $z=\infty$ ) environment (see Fig. 2).

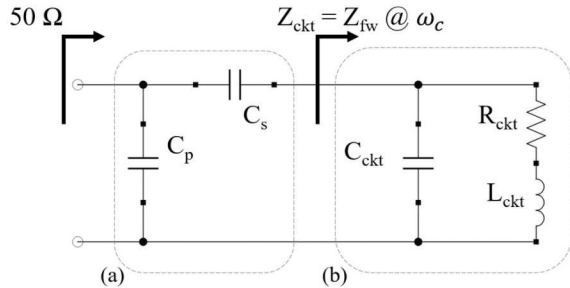


Figure 2. (a) L-section matching network [53], and (b) equivalent circuit of the loop in free-space [54].

eliminate detuning on/off metal as well as experimental data showing that the dual-environment antenna works in practice.

## II. LOOP ANTENNA DETUNING NEAR A GROUND PLANE

The severity and nature of the detuning problem for small loop antennas used in both free-space and near ground planes is demonstrated via full-wave calculation of the Fig. 1 geometry with  $t=18.55$  mm using Ansys HFSS [51]. Note that due to basic image theory considerations, the orientation shown is the most profitable for a loop antenna near a ground plane [52]. At 433 MHz, this loop has electrical size  $ka \sim 0.2$  (where  $k$  is the wavenumber and  $a$  is the radius of the smallest sphere enclosing the antenna). Accordingly, the antenna has a high radiation  $Q$  and narrow impedance bandwidth.

The resulting full-wave reactance,  $X_{fw}$ , is shown in Fig. 3(a). The free-space loop is matched to  $50 \Omega$  at 433 MHz with an L-section matching network [53] consisting of  $C_s=4.19$  pF series and  $C_p=149$  pF shunt capacitors, as shown in Fig. 2(a); the resulting reflection coefficient is shown in Fig. 3(b). When this same loop geometry is simulated  $z=1.27$  mm away from a ground plane and this same matching network is applied, the loop reactance shifts 2% to 424 MHz. In this state, the mismatch loss at 433 MHz is greater than 20 dB. However, near 424 MHz the return loss is 10 dB, resulting in only 0.5 dB mismatch loss; if this resonance shift can be eliminated, good performance can be achieved in both environments.

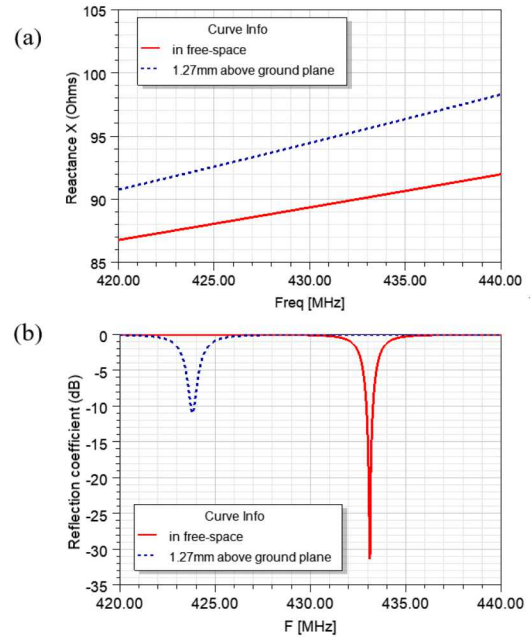


Figure 3. (a) Full-wave reactance,  $X_{fw}$ , of the Fig. 1 loop (i.e., with no matching network) with  $t=18.55$  mm for free-space and  $z=1.27$  mm above a ground plane. (b) Resulting reflection coefficient when the antenna is matched with the same network of Fig. 2(a). The resonance shifts only 2%, but causes 20 dB mismatch loss at 433 MHz.

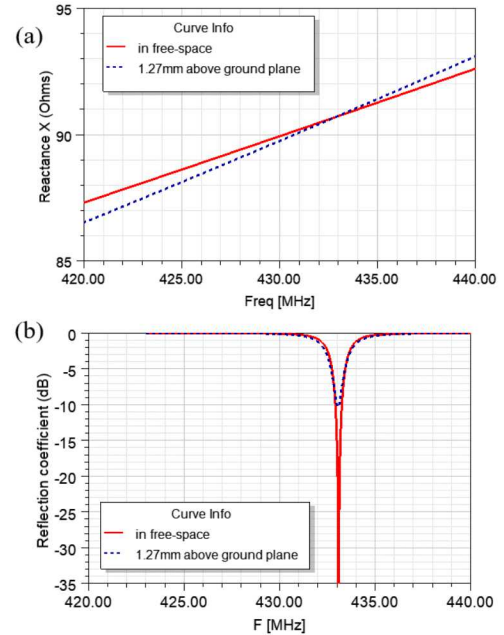


Figure 4. (a) Full-wave reactance,  $X_{fw}$ , of the Fig. 1 loop (i.e., with no matching network) with  $t=7.2$  mm for free-space and  $z=1.27$  mm above a ground plane. The reactance in both environments is equal at 433 MHz. (b) Resulting reflection coefficient when the antenna is matched with the same network of Fig. 2(a), showing negligible detuning between the two environments.

## III. A LOOP ANTENNA WITH STABLE TUNING

The Fig. 1 loop is altered such that  $t=7.2$  mm. The full-wave loop reactance,  $X_{fw}$ , is calculated for the loop in free-space and  $z=1.27$  mm away from the ground plane. As seen in Fig.

4(a),  $X_{fw} = 90.7 \Omega$  at 433 MHz for *both* environments. The free-space antenna may be matched to  $50 \Omega$  at 433 MHz with  $C_s = 4.17$  pF series and  $C_p = 148$  pF shunt capacitors as shown in Fig. 2(a). Because the reactance does not change between the two environments, the resonant frequency shifts negligibly, as shown in Fig. 4(b). There is a modest change in the full-wave loop resistance,  $R_{fw}$ , between the environments (about  $0.122 \Omega$  in free-space and  $0.230 \Omega$  when  $z = 1.27$  mm away from the ground plane). However, the mismatch loss remains minimal in practice (only about 0.5 dB) despite this.

The full-wave calculated radiation efficiency of the loop is 57% in free-space and 76% when  $z = 1.27$  mm away from the ground plane; matching network losses may be considered separately. Because the loop circumference is only 16% of the wavelength, the current distribution is substantially uniform. Consequently, the radiation pattern approximates a toroid in free-space and a half-toroid when the loop is located close to a large ground plane; currents and patterns are shown in Fig. 5.

#### IV. EQUIVALENT CIRCUIT

A simple circuit model explains the observed phenomenon and yields a design equation for preventing frequency shift with other loop geometries at an arbitrary frequency below the loop self-resonance. The free-space loop can be represented by the parallel-resonant equivalent circuit of Fig. 2(b) [54]. Circuit element values can be selected such that the equivalent circuit impedance  $Z_{ckt}$  matches the full-wave model impedance  $Z_{fw} = R_{fw} + j\omega_c L_{fw}$  at the operating frequency  $\omega_c$  as follows. First, the self-resonant frequency  $\omega_0$  of the loop in free-space is calculated via the full-wave model. Next, the full-wave model is solved at the operation frequency  $\omega_c$  yielding  $Z_{fw}$ . The equivalent circuit resistance  $R_{ckt}$  and inductance  $L_{ckt}$  are then chosen using the following approximations derived from basic circuit analysis of the Fig. 2(b) equivalent circuit:

$$R_{ckt} \sim R_{fw} (1 - \omega_c^2 / \omega_0^2)^2, \quad (1)$$

$$L_{ckt} \sim L_{fw} (1 - \omega_c^2 / \omega_0^2). \quad (2)$$

Finally,  $C_{ckt}$  is determined according to:

$$C_{ckt} = 1 / (\omega_0^2 L_{ckt}). \quad (3)$$

For the Fig. 1 geometry with  $t = 7.2$  mm, full-wave simulation indicates the free-space loop resonant frequency  $\omega_0 = 2\pi 1.2$  GHz and  $Z_{fw} = 0.122 + j90.7 \Omega$  at  $\omega_c = 2\pi 433$  MHz. By (1), (2) and (3),  $R_{ckt} = 0.092 \Omega$ ,  $L_{ckt} = 29.0$  nH and  $C_{ckt} = 0.606$  pF.

When the free-space loop is brought near a ground plane, the primary perturbation is magnetic flux coupling between the loop and its image in the ground plane, shown in Fig. 6 as mutual inductance  $M$ . The capacitance  $C_1$  in Fig. 6 represents the parasitic capacitance between the feed node and its ground-plane image; with respect to the feed,  $C_1$  appears in series with the larger capacitance,  $C_2$ , of the remaining loop bottom face and thus the effective value  $C_{br} = (C_1 C_2) / (C_1 + C_2)$  is approximately  $C_1$  when  $t \ll L$ . This implies that  $C_{br}$  is

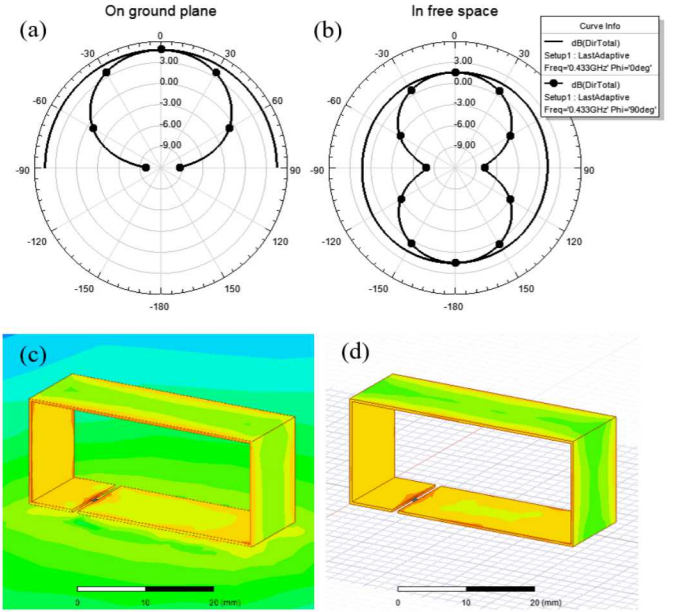


Figure 5. Directivity (a) and (b) and current magnitude (c) and (d) for the antenna of Fig. 1 with  $t = 7.2$  mm near a large ground plane (a) and (c) and in free space (b) and (d). The patterns approximate an ideal toroid in free space and a half-toroid when near the ground plane. The predominant vector direction of the loop current is in the plane of the loop; the color scale is +40 dBA/m to -40 dBA/m.

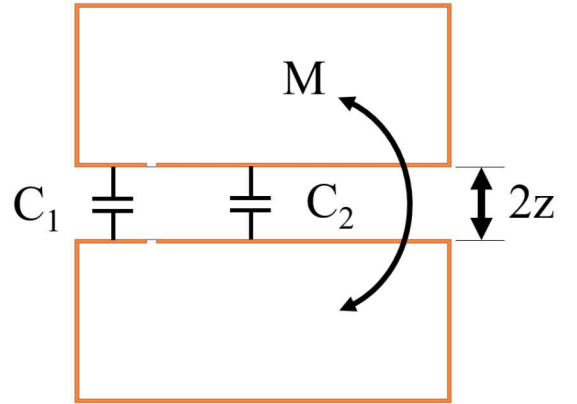


Figure 6. Loop antenna with  $t = 7.2$  mm (top) and its ground-plane image (bottom) separated by a distance  $2z$ . The mutual inductance  $M$  between the loops and two parasitic capacitances are indicated. Because  $C_1$  is smaller than  $C_2$  and the two appear in series, the total series capacitance  $C_{br}$  is close to  $C_1$ .

maximum when the feed gap is centered along the loop length  $L$ ; due to symmetry, either increasing or decreasing  $t$  from the centered condition will reduce  $C_{br}$  from the maximum value. Without the “bridging capacitance”  $C_{br}$ , there is no frequency below self-resonance where the reactances of the free-space and ground-plane equivalent circuits are equal.

With  $t = 7.2$  mm and  $z = 1.27$  mm, a simple estimate of  $C_{br} \sim C_1$  using the parallel plate capacitor formula is 0.325 pF. The mutual inductance  $M$  is calculated from the full-wave two-port s-parameters of the Fig. 5 geometry at 10 MHz as 3.42 nH. The transformer coefficient  $k$  is thus  $M/L_{ckt} = 0.118$ .

The equivalent circuit of the loop near a ground plane *without* a matching circuit is shown in Fig. 7(a). Because image theory

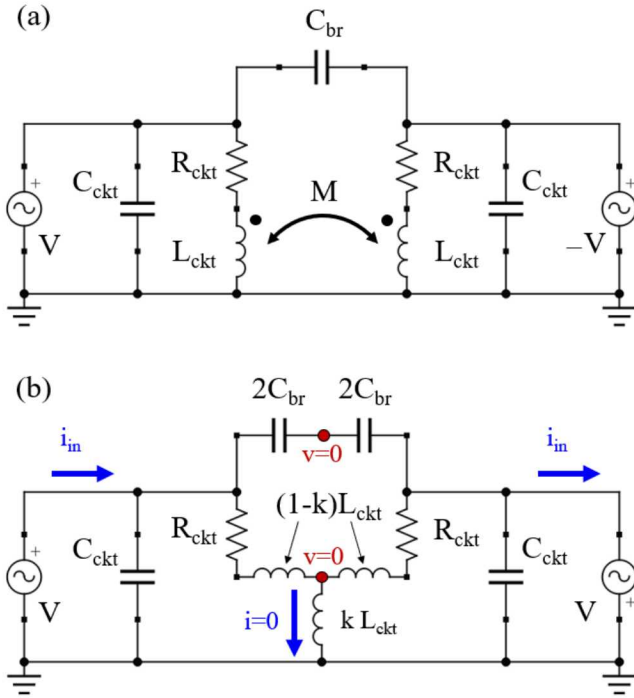


Figure 7. (a) Equivalent circuit of the loop near a ground plane, and (b) modified equivalent circuit. Image theory dictates an antiphase source image and thus odd (anti-symmetric) mode operation, resulting in zero shunt inductor current and zero potential at the nodes indicated; thus, these nodes may be grounded. No matching circuit is shown in either (a) or (b).

dictates the source image has  $180^\circ$  phase [52], a driven circuit simulation is required to calculate the input impedance of this equivalent circuit. The antiphase source image causes the equivalent circuit to operate in the “odd” (anti-symmetric) mode only [55]. As illustrated in Fig. 7(b), odd mode operation allows for a greatly simplified ground-plane equivalent circuit that is topologically equivalent to the free-space circuit of Fig. 4(b), however, from inspection of Fig. 7(b), we see the capacitor value is  $C_{\text{ckt}} + 2C_{\text{br}}$  and the inductor value is  $(1 - k)L_{\text{ckt}}$ . Using basic circuit theory and a high  $Q$  approximation, we find the reactance of this simplified ground-plane equivalent circuit is equal to that of the free-space equivalent circuit at the operation frequency  $\omega_c$  when:

$$\omega_c^2 \sim \frac{1}{2} \left( \frac{k}{1-k} \right) \left( \frac{1}{C_{\text{br}} L_{\text{ckt}}} \right) \quad (4)$$

This relation predicts that the equal-reactance frequency decreases as  $C_{\text{br}}$  is increased (e.g., by increasing  $t$  from 7.2 mm) and as the loop inductance is increased (e.g., by decreasing  $W$ )—as is observed in both the full-wave model as well as experimentally.

For  $\omega_c = 2\pi 433$  MHz,  $k=0.118$ , and  $L_{\text{ckt}}=29.0$  nH, (4) indicates  $C_{\text{br}}$  must be 0.312 pF, in good agreement with the parallel-plate capacitance estimate. Circuit model results are shown in Fig. 8.

From (4), we see that only certain combinations of  $k$ ,  $C_{\text{br}}$  and  $L_{\text{ckt}}$  yield equal reactance at the operating frequency. Thus, for

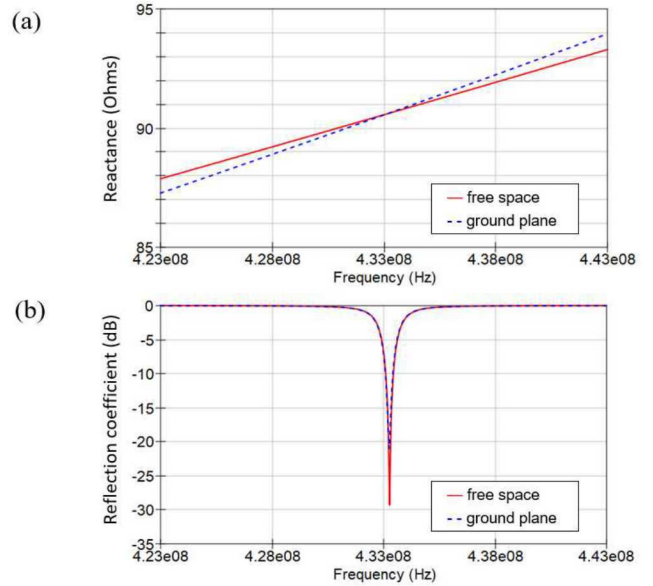


Figure 8. Circuit model results for free-space (i.e., Fig. 4(b)) and  $z=1.27$  mm above ground plane (i.e., Fig. 7(a)), with  $C_{\text{br}}$  tuned to 0.305 pF (the value predicted by (4) is 0.312 pF); (a) unmatched loop reactance and (b) 50  $\Omega$  reflection coefficient with  $C_s = 4.17$  pF and  $C_p = 148$  pF L-section matching network. The circuit models show equal-reactance at 433 MHz and resulting zero resonance shift.

a given loop design, stable tuning occurs only for a particular ground plane separation  $z$ . In practical hard RFID tags [45]–[49], this distance is already fixed by the radome. Because  $C_{\text{br}}$  is influenced by the radome dielectric, its effect should be accounted for via full-wave simulation. Despite this, the principles of the technique presented above remain unchanged.

## V. IMPLEMENTATION AND TESTING

The Fig. 1 loop geometry with  $t=7.2$  mm was implemented with copper sheet and an FR-4 printed circuit board (PCB); the PCB hosts the matching network and a U.FL coaxial connector. The loop is fixed  $z=1.27$  mm away from a 16” round metal ground plane using a nylon screw and nylon washers. The reflection coefficient was measured with a vector network analyzer (VNA) whose cable was arranged to minimally couple to the antenna loop (i.e., the loop formed by the measurement cable and ground plane was orthogonal to the antenna loop) as shown in Fig. 9. RF chip capacitors in the L-section matching network were adjusted until the input impedance was close to 50  $\Omega$  at 433 MHz as seen in Fig. 10. Note that  $C_s$  and  $C_p$  were each implemented via multiple chip capacitors in parallel to reduce losses and allow for fine-tuning. A hemispherical Wheeler cap measurement [56] of the matched loop on ground yielded 40% radiation efficiency (including matching losses).

Next, the VNA cable was disconnected from the antenna and a VNA-connected, sub-resonant magnetic field probe (Beehive Electronics model 100A) was held within a few centimeters of the loop antenna. Normally, such a probe reflects nearly all RF power incident from the VNA. However, at the loop antenna resonant frequency, a small amount of RF power is coupled from the probe; a small dip in the VNA-measured reflection coefficient is evident, indicating the antenna resonance. (This is

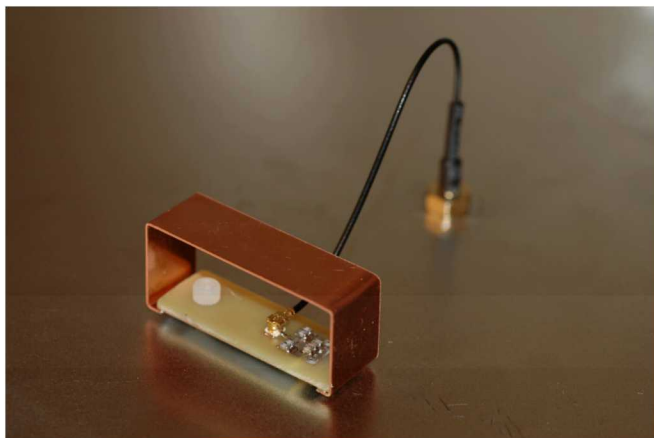


Figure 9. As-built loop antenna with  $t=7.2$  mm located on a ground plane with the VNA measurement cable connected. The bottom face of the loop is spaced 1.27 mm off the ground plane with a nylon screw and nylon washers. The feed gap in the PCB copper is visible in the ground plane reflection of the loop underside.

a classic technique known as “grid dip” [57].) In this case, the probe-indicated resonance was about 2 MHz from that of the initial cabled VNA measurement; this implies the cable did not significantly impact the cabled VNA measurement. Note that experimentally, removing a small amount of PCB feed node copper with a razor can tune  $C_{br}$  to finely adjust the detuning characteristic in accord with (4).

Next, the ground plane was removed, and the probe measurement repeated. The indicated resonance shifted less than 1 MHz, demonstrating that the loop antenna maintains stable tune between the two environments as designed; this data is shown in Fig. 10.

Finally, a miniature, battery-powered swept-frequency RF source with approximately constant-available power in  $50 \Omega$  was set inside the loop. The RF signal transmitted by the loop was monitored with a polarization-matched, sub-resonant monopole probe connected to a spectrum analyzer located about 60 cm away from the loop antenna in the  $\hat{z}$  direction. This was done with the loop in three ground plane configurations: 1)  $z=1.27$  mm away from the ground plane, 2)  $z=75$  mm away from the ground plane, and 3) with no ground plane; results are presented in Fig. 11. When  $z=75$  mm, full-wave modeling shows the loop impedance is close to that in a true free-space environment, however, the radiation pattern remains similar to that of the  $z=1.27$  mm case (i.e., it is approximately a half-toroid). Little difference in the received peak amplitude and frequency was observed between configurations 1 and 2. When the ground plane is removed entirely, the radiation pattern changes significantly—now a full toroid covering  $4\pi$  steradians—and the signal is expected to decrease 3 dB due to this effect. Moreover, the full-wave calculation predicts 1.25 dB decrease in radiation efficiency when moving from  $z=1.27$  mm away from ground plane to free space. The measured signal decreased 4.5 dB, with little frequency shift. This confirms the antenna works as designed; we estimate the change in mismatch loss at the feed between free-space and ground plane environments is about 0.25 dB in this case.

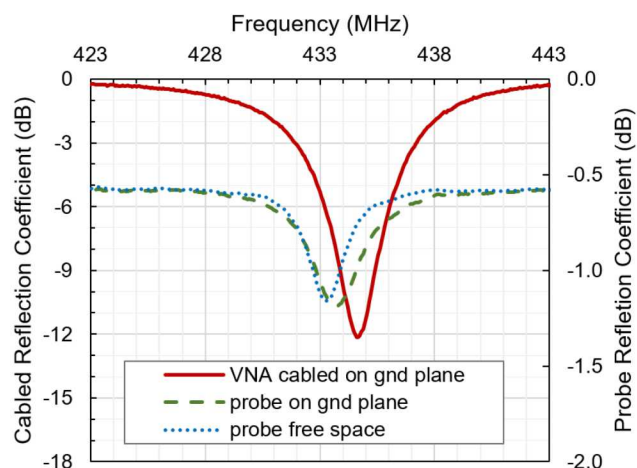


Figure 10. Cabled VNA reflection coefficient of the loop  $z=1.27$  mm away from the ground plane, as well as “grid dip” near-field probe measurements on and off the ground plane. All three resonances are very close to each other, indicating that the measurement cable is not perturbing the measurement and that the antenna has stable tune between the ground plane and free-space environments.

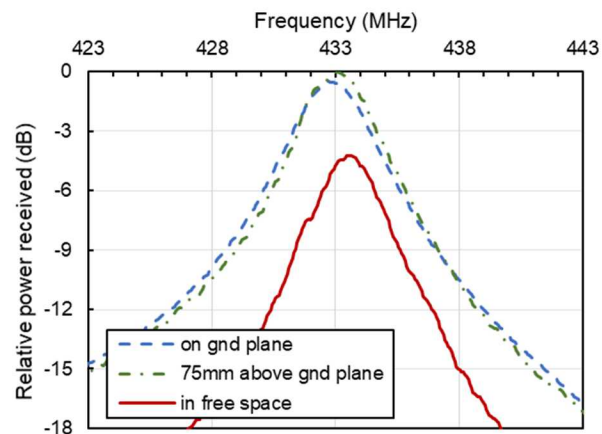


Figure 11. Experimental power received from the transmitting loop antenna with  $t=7.2$  mm in three environments demonstrating minimal frequency shift and limited change in peak amplitude. Accounting for expected changes in radiation pattern and radiation efficiency between the ground plane and free-space environments, we estimate the change in mismatch loss to be about 0.25 dB.

## VI. CONCLUSION

This paper presents a general design principle for preventing detuning when small loop antennas are used both in free space and a fixed distance away from a ground plane. The approach balances the inductive and capacitive parasitics introduced by the ground plane such that the resonant frequency remains unchanged. The design may be easily adapted to commercial passive RFID applications via the design equation (4), flexible inlay fabrication and straightforward adjustment of the matching network to give a complex conjugate match to passive RFID integrated circuits. In addition, a basic tradeoff between loop antenna height and radiation efficiency can be made to tailor the performance for any application.

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