Low Profile, Miniaturized, Inductively Coupled Capacitively Loaded Monopole Antenna

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***Abstract*—A novel high-gain low-profile miniaturized antenna with omnidirectional vertically polarized radiation, similar to a short dipole is presented. The proposed design focus is on increasing the gain and improving the polarization purity of the radiated field in the horizontal plane. The gain and polarization improvement are achieved by isolating the feed structure from a miniaturized resonant radiating structure composed of an in-plane capacitor and a structurally embedded transformer. The antenna topology is developed, based on circuit model and through full-wave simulations the equivalence is established. The equivalent circuit model assists in the initial design, and then minor modifications are required to achieve the desired frequency of operation. The initial topology of the proposed antenna, the so-called Inductively Coupled Capacitively Loaded Monopole Antenna (ICCLMA), consists of two metal layers, a feeding pin and a shorting pin. The performance of the proposed antenna is compared to that of an ordinary inverted F antenna and a more recent low profile vertically polarized antenna (LMMMA) [1]. It is shown that the gain of ICCLMA is 9 dB and 4 dB higher than that of the conventional inverted-F antenna and the LMMMA, respectively. To simplify the fabrication process a modified single-layer ICCLMA topology is presented and optimized. Finally, a design procedure to further reduce the lateral dimension of ICCLMA is presented. A procedure for accurate measurement of antennas with small ground planes is also presented.**

***Index Terms*—Antenna measurement system, electrically small antennas, omnidirectional antennas.**

I. Introduction

VERTICALLY polarized antennas with omnidirectional radiation pattern are highly desirable for many applications including near-ground communications among ad-hoc nodes of wireless devices used in vehicles or unattended ground sensors operating at low frequencies. The need for vertical polarization stems from the fact that near-ground propagation path loss between two near-ground antennas for vertically oriented antennas is by many orders of magnitude lower than any other antenna orientation configurations [2], [3]. In fact, this is the main reason $\lambda_0/4$ monopole antennas with vertical polarization and omnidirectional radiation pattern are prevalent in many communication devices working near the ground. However, as wireless communication devices continue to evolve, the large dimension of the antenna is frequently problematic. Therefore, it is imperative to further investigate methods of realizing extremely short monopole antennas with very small lateral dimensions, while maintaining high radiation efficiency. This will allow ease of integration of such antennas with the package or platform of small wireless devices that are emerging. Recently different types of low-profile antennas with omnidirectional radiation pattern have been proposed. Among these, one approach is to excite radiation from short segments of loaded vertical wires, and the other one is based on exciting a cavity-backed small slot loop antenna [4], [5].

In the studies where short segments of vertical wires are utilized as the intended radiators, the antennas are usually loaded with horizontal resonant structures. Inverted L and Inverted F antennas are the examples of the folded-type resonant loading structures that are commonly used [6], [7]. One drawback of these structures is the energy radiation in the horizontal polarization that stems from relatively large lateral dimension of the loading structures. This, of course, leads to lower gain in the desired polarization channel. In [8]–[13], interesting capacitively loaded monopole antennas with different especial disk geometries are presented as means for reducing the antenna height and improving the bandwidth. The heights of these antennas are in the range of $\lambda_0/10$, presenting excellent operational bandwidth. However, their lateral dimensions are still comparable to the wavelength. For certain applications where the bandwidth can be compromised, it is found that by loading the planar inverted-F antenna (PIFA), the lateral dimension of the antenna can be reduced to $\lambda_0/8$ [14]–[16]. In [16], a via-patch is added inside a conventional PIFA leading to an easier design and fabrication. By changing the height and location of the vias, different degree of size reduction could be obtained, providing more design freedom. Nevertheless, these antennas include a multilayer geometry that cannot be realized using simple PCB fabrication process.

Recently, an extremely short monopole antenna ($\lambda_0/60$ – $\lambda_0/40$), known as LMMMA, with omnidirectional radiation pattern and vertical polarization was introduced [1]. The concept is based on superposition of multiple quarter-wave segments that are meandered and spiraled around to suppress the radiation from horizontal currents above the ground plane. In other words, the cancellation of the horizontal electric current is achieved by introducing another set of electric current that is in the opposite direction in the horizontal plane of the antenna with the original electric current at electromagnetic resonance. As a result, the antenna features a vertically polarized radiation in the horizontal plane. Although the antenna produces almost purely

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vertical polarization, it suffers from low gain due to high ohmic loss of the spiral metallic traces. Hence, it is obvious that more effort will be needed to come up with a design for increasing the gain of extremely short antennas with very small lateral dimensions.

For most monopole low-profile antennas, the size of the ground plane is chosen to be large or moderately large compared with the wavelength [17]–[20]. The size of the ground plane affects the uniformity of the radiation pattern and the gain near the horizontal plane ($\theta = 90^\circ$) due to the edge diffraction [21]. For very small platforms, the size of the ground plane must be comparable with the lateral dimensions of the antenna itself.

In this paper, a novel low profile miniaturized antenna with omnidirectional radiation pattern and vertical polarization is presented. To suppress the radiated fields from horizontal currents over a top load the horizontal currents are distributed uniformly over a large metallic top load. However, the use of this large top load in miniaturized antennas prohibits the use of thin shorting pins because impedance matching to 50 ohm cannot be obtained without the increase in the dimension of the shorting pins. In the proposed antenna, a new in-plane capacitive coupled structure is introduced for compensating inductive coupling between the thin pins. Finally, the use of the large top load, shorting pin and in-plane capacitive coupled structure achieve improved polarization purity and excellent impedance matching with antenna miniaturization. An equivalent circuit model of the proposed antenna is developed for the ease of design. The improved performance of the proposed antenna is compared to that of ordinary inverted F antennas and the more recent low profile vertically polarized antennas. A two-layer antenna is described first to explain the principle of operation. With the help of the equivalent circuit model, this design is then modified to achieve a single-layer counterpart of the same antenna for the ease of fabrication. Furthermore, the design procedure and tradeoff study to reduce the lateral dimension of ICCLMA at the cost of lowering the gain are presented. The proposed antennas are fabricated and measured to validate the design method. It is pointed out that an antenna with an electrically very small ground plane cannot be measured accurately using a common unbalanced coaxial feed. A new gain and pattern measurement method using a balanced version of the monopole (dipole type) antenna is presented.

II. The Inductively Coupled Capacitively Loaded Monopole Antenna (ICCLMA)

A. Concept of ICCLMA

As mentioned before, one of the drawbacks of the very short miniaturized monopole antenna (LMMMA) is its low radiation efficiency resulted from the strong electric currents concentrated on the horizontal thin metallic traces. However, establishing this current is necessary to establish the required high current level on the short vertical pin which is the main radiating component of the antenna. As the height of the antenna decreases, higher level of current is required and thus the antenna radiation efficiency gets lower. In order to increase the gain, we need to suppress the radiated fields from horizontal currents over the resonating structure, while keeping the lateral dimensions small. This can be accomplished by using a large capacitive top-load as well as high Q lumped-type resonant structures built by separating the radiating structure (= the secondary circuit) from the feed structure (= the primary circuit) using a magnetic coupling and an additional capacitive coupling.

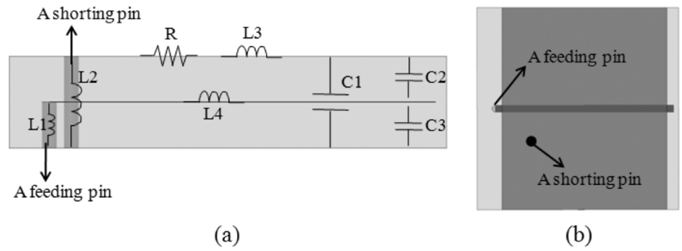

Fig. 1. (a) Equivalent circuit model drawn in side view and (b) top view of a multilayer ICCLMA.

The high-Q resonant structure consists of magnetically-coupled two pins acting as an transformer, a relatively wide metallic patch acting as a capacitor and an additional capacitive coupling between the feed structure and the radiating structure. The desired operation of the transformer can be realized by the proper placement of a feeding pin and the shoring pin in the resonator which is a vertical element responsible for vertically polarized radiation. The salient feature of the proposed antenna is the fact that the internal elements of the antenna structure are used for the desired functionality and impedance matching without the need for lumped elements or external matching network. Otherwise, the use of external lumped elements will render poor radiation efficiency. As will be shown, the transformer coupling, capacitance of the top load and other capacitances can be adjusted to achieve the impedance matching as well as obtaining the resonant condition at a proper frequency.

B. Multilayer ICCLMA and Equivalent Circuit Model

The topology of ICCLMA designed initially consists of three layers as shown in Fig. 1(a). This includes a ground plane, a thin metallic trace, and a wide metallic patch that are, respectively, placed in the bottom, the middle and the top layers. Fig. 1(a) also depicts an equivalent circuit model superimposed over the side view drawing of the proposed multilayer ICCLMA. This circuit model helps the arrangement of different structural components of the antenna. A vertical feeding pin and a shorting pin are inductively coupled. While the feeding pin is connected to a narrow metallic trace in the middle layer, the shorting pin is attached to a wide metallic plate in the top layer. The top view of the structure shown in Fig. 1(b) indicates the position of the shorting pin to be offset from the symmetry plane in the center and a short distance away from the feeding pin. In other words, the shorting pin does not touch the metallic trace in the middle layer and thus the radiating structure can be coupled to the feed magnetically.

The resistance (R) in the equivalent circuit represents the sum of the radiation resistance, ohmic loss in metals, dielectric loss and surface wave loss. The top wide metallic surface is represented by a shunt capacitor in the secondary circuit of the transformer in the circuit model. The uniform current distribution

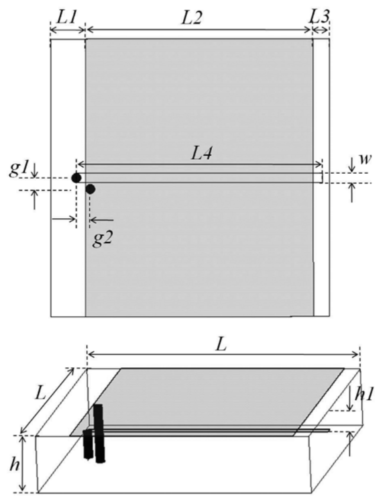

Fig. 2. Design parameters of a multilayer ICCLMA.

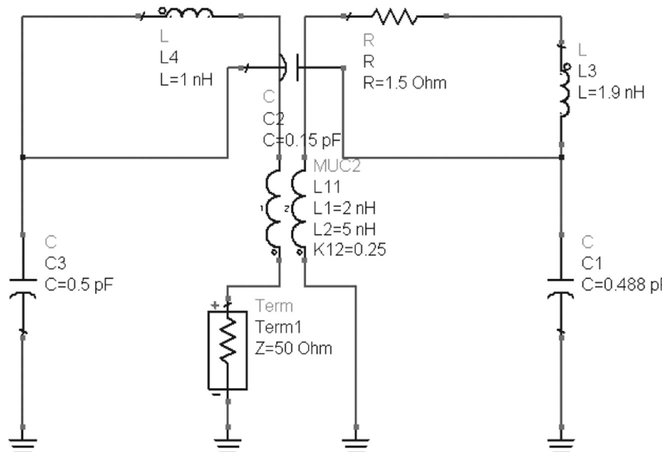

Fig. 3. Equivalent circuit model of a multilayer ICCLMA.

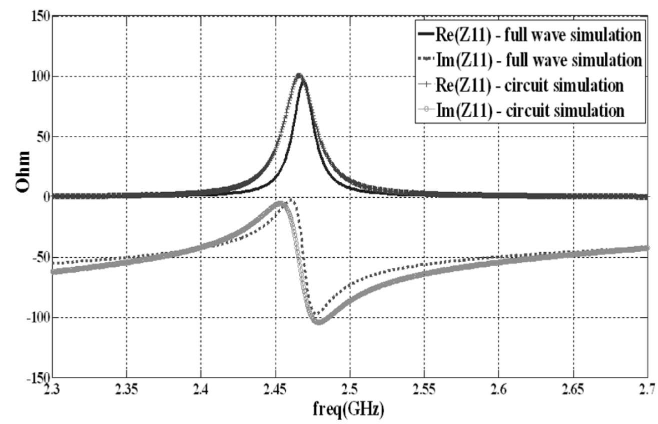

Fig. 4. Real and imaginary parts of input impedance simulated by full wave and circuit simulators.

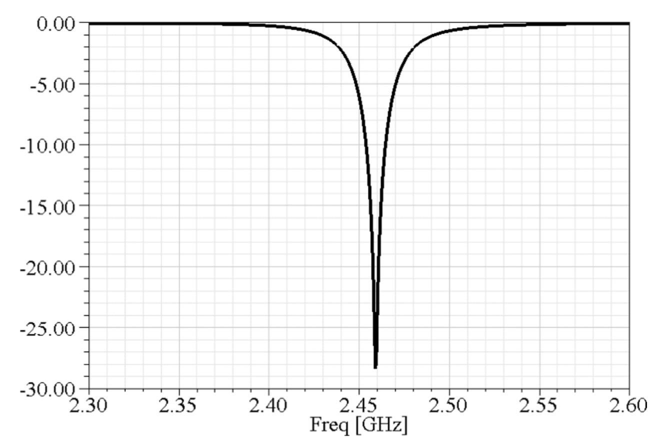

Fig. 5. Simulated S_{11} when $L = \lambda_0/8$ and $h = \lambda_0/40$.

supported by the wide copper sheet on the top layer leads to significant reduction of the ohmic loss. Additional capacitances are introduced due to the placement of a narrow metallic trace in the middle layer. One very important feature of the new design is that the antenna input impedance can match to almost any desired value between 10 and 100 Ohm. Impedance matching can be achieved by adjusting the parameters of the transformer and the values of capacitors and inductors shown in the circuit model. The coupling coefficient of the transformer corresponds to the distance between a feeding pin and a shorting pin. The values of the inductance and capacitances are related to the diameter of two pins and the width, length and height of metal sheets, respectively.

Design parameters are shown in Fig. 2. Their value are given by $L = 15\ \text{mm} = \lambda_0/8$, $h = 3.14\ \text{mm} = \lambda_0/40$, $L1 = 1.875\ \text{mm}$, $L2 = 12.25\ \text{mm}$, $L3 = 0.875\ \text{mm}$, $L4 = 13.25\ \text{mm}$, $h1 = 1.57\ \text{mm}$, $w = 0.5\ \text{mm}$, $g1 = 0.6\ \text{mm}$ and $g2 = 0.75\ \text{mm}$ where λ_0 is free-space wavelength at the resonant frequency. The diameters of the two pins are chosen to be 0.5 mm. The substrate used in this design has a dielectric constant of 2.2 and dielectric loss tangent of 0.0009. In order to consider ohmic loss, conductivity of copper is used in all metallic traces and vertical pins in the full-wave analysis.

It is interesting to examine whether the equivalent circuit model can predict the behavior for the proposed antenna structure. The values of the lumped elements in Fig. 3 are first derived from the design parameters in Fig. 2 and slightly adjusted, based on full wave simulation results. Fig. 4 shows the real and imaginary parts of input impedance which are derived from a circuit (Advanced Design System 2009) and full wave (Ansoft HFSS 12) simulator. It is shown that the two results are in very good agreement, meaning that the equivalent circuit model can predict the performance perfectly. Simulated S_{11} in Fig. 5 indicates good impedance matching and fractional 10-dB return loss bandwidth of 0.45% due to high quality factor.

The vertically (co-) and horizontally (cross-) polarized radiation pattern in the E-plane and H-plane are presented in Fig. 6(a) and (b). The co-polarized radiation pattern is omnidirectional pattern with the direction of maximum gain occurring at $\theta = 90°$. While the gain of co-polarized radiation at $\theta = 90°$ near ground observation is 0.46 dBi, the gain of cross-polarized radiation at $\theta = 90°$ is less than -25 dBi. As expected, the proposed antenna provides excellent suppression of horizontally polarized radiation.

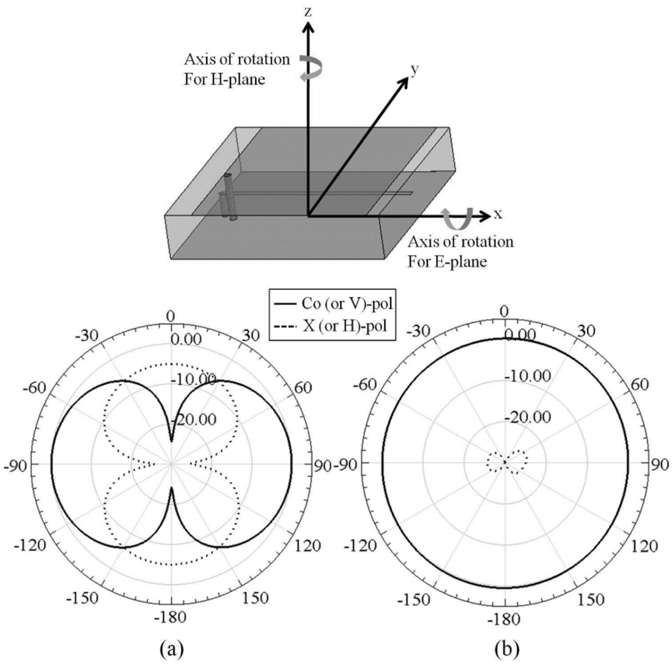

Fig. 6. Simulated (a) E-Plane and (b) H-Plane radiation patterns when $L = \lambda_0/8$ and $h = \lambda_0/40$.

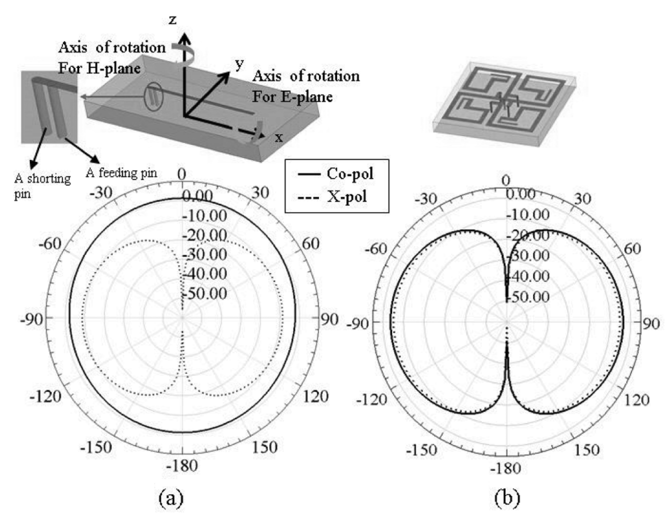

Fig. 7. E-Plane radiation patterns of (a) a conventional inverted-F antenna and (b) multi-element monopole antenna with the same vertical profile as the height of the proposed antenna ($=\lambda_0/40$).

C. Bench Marking

In this section, the performance of the proposed antenna is compared with those of other low-profile antennas reported in the literature. This comparison includes the size, gain, and polarization purity of the proposed antenna against a conventional inverted F antenna and the multi-element monopole antenna (MMA) introduced in [1]. Two antennas with the same height ($= 3.14\ \mathrm{mm} = \lambda_0/40$) as that of the proposed antenna, are designed as shown in Fig. 7. To address a general drawback of low profile antennas designed using a $\lambda_0/4$ open circuited transmission line, the conventional inverted-F antenna with a $\lambda_0/4$ thin metallic trace is chosen. The drawback is poor polarization purity caused by very small height of the antenna ($=\lambda_0/40$) and horizontal current flowing along the thin meta-llic trace in one direction. While MMA has the same ground plane size as that of the proposed antenna, the conventional inverted-F antenna has larger ground plane since no miniaturization technique is applied. The simulated gain and the dimension of these antennas and the proposed antenna are presented in Table I. For the case of an inverted-F antenna, horizontal polarized gain ($=-2.1$ dBi) on H-plane is dominant over the expected vertical polarized gain ($=-8.7$ dBi). Although MMA has the improved gain due to the efficient cancellation of horizontal current, its gain is still low as -3.6 dBi because of high ohmic loss generated from narrow metallic traces. For the proposed antenna, the gain is substantially improved ($= 0.46$ dBi) and is 9 dB higher than that of the conventional inverted F antenna. Moreover, the new antenna has 4 dB higher gain than the MMA with the same physical dimensions, as shown in Table I. At $\theta = 90^\circ$ (near ground observation), the ratio of horizontally to vertically polarized gain is less than -25 dB, suggesting excellent suppression of the radiation in horizontal polarization.

TABLE I
GAIN AND DIMENSION OF A CONVENTIONAL INVERTED-F ANTENNA, MMA AND PROPOSED ANTENNA

Antenna type	Lateral dimension (mm X mm)	Height (mm)	Radiation Efficiency (%)	Gain (Co-pol.) on H-plane (dBi)	Gain (X-pol.) on H-plane (dBi)	Co-pol to X-pol (dB)	BW(%)
Inverted F Antenna	15 X 36	3.14	61.46	-8.7	-2.1	-6.6	0.62
MMA	15 X 15	3.14	44.57	-3.6	-5.2	1.6	0.49
ICCLMA	**15 X 15**	**3.14**	**70.89**	**0.46**	**-25.9**	**26.3**	**0.47**

III. Single-Layer ICCLMA and Additional Size Reduction

A. Single-Layer ICCLMA

In the previous section, it was shown that the gain of a multilayer ICCLMA is greatly higher than that of other existing low-profile miniaturized antennas. However, the complexity of the multilayer geometry cannot be ignored. This architecture requires complex multilayer-printed circuit fabrication and is subject to alignment errors and higher cost. This is perhaps the reason why multilayer low-profile miniaturized antennas are not popular, although they have better performance [14], [16]. In this section, a modified single-layer ICCLMA having the same equivalent circuit model is presented. The multilayer structure had capacitances C2 and C3 in its equivalent circuit that are responsible for miniaturization as well as impedance matching. It turns out that the same capacitances can be realized in single-layer ICCLMA architecture as shown in Fig. 8. The series capacitor C2 between the primary circuit and the secondary circuit is created by an in-plane interdigital capacitance, and C3 is the shunt capacitance between the strip attached to the primary circuit (left side of the interdigital capacitor) and the ground plane. Fig. 9 shows the simulated radiation pattern of single-layer ICCLMA having the same dimension as a multilayer ICCLMA. This antenna has a slightly lower gain of 0 dBi at the same resonant frequency.

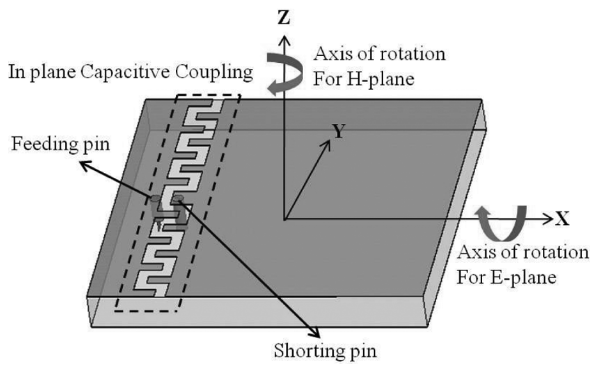

Fig. 8. Topology of single-layer ICCLMA.

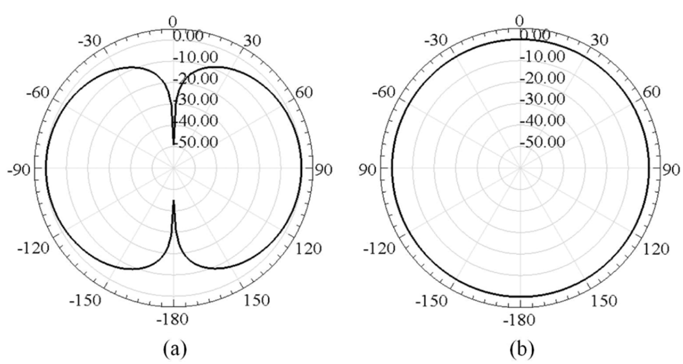

Fig. 9. Simulated (a) E-Plane and (b) H-Plane radiation patterns of a single-layer ICCLMA when $L = \lambda_0/8$ and $h = \lambda_0/40$.

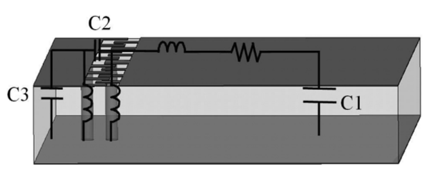

Fig. 10. Equivalent circuit model of single-layer ICCLMA.

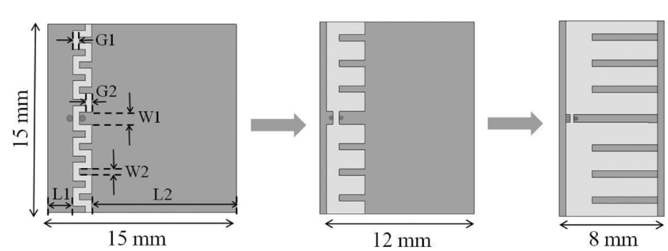

Fig. 11. Design procedure for additional size reduction of single-layer ICCLMA.

B. Additional Size Reduction of Single-Layer ICCLMA

As discussed in Section II-B, the introduction of the series capacitance C2 in the proposed ICCLMA antenna allows for the ease of impedance matching at lower frequency. This property also allows for obtaining additional size reduction of single-layer ICCLMA. Fig. 10 shows equivalent circuit model of single-layer ICCLMA. When C2 is changed by modifying the interdigital structure as shown in Fig. 11, C1 and C3 are also changed at the same time because they are proportional to the area of metal sheet on the top layer.

It is interesting to note that the proper combination of C1, C2 and C3 values can achieve excellent impedance matching of smaller antennas. Fig. 11 describes how this principle is applied for additional size reduction. The suitable combination of C1, C2 and C3 values can be found by changing the gap distance between two metallic sheets on the top layer (G1 and G2), the width of protruding parts of the interdigital shape (W1 and W2), as well as L1 and L2. For example, the center prong in Fig. 11 determines the series capacitance C2 and other prongs are modified to get the required capacitance C1. The impedance matching for three different dimensions is well achieved at the same resonant frequency as shown in Table II. Finally, the size can be reduced from 15 mm × 15 mm to 8 mm × 15 mm, which corresponds to almost 50% size reduction. However, the size reduction comes at the expense of reduction in gain from 0 dBi to -1.9 dBi.

One important advantage of this design procedure is that smaller antennas can be perfectly matched to 50 ohm without modifying parameters related to pins such as the diameter of two pins and the gap distance between two pins, while folded low-profile antennas such as PIFA usually require changing the area of the rectangular sheet shorting the top plate when additional miniaturization is required. Furthermore, in the proposed antenna, thin pins with diameter as small as 0.5 mm are used, while most folded low-profile antennas have a wide rectangular shorting sheet that is comparable to the other antenna dimensions. This property is desirable when commercial PCB technology is used in fabrication of low-profile vertically polarized antennas. Basically, metalized via holes can be used to realize the shorting pins very accurately.

IV. Measurements

A. Measurement Issues for Antennas With Electrically Small Ground Plane ($\ll \lambda_0$)

In order to validate the design approach discussed in the previous section, the proposed antenna was fabricated and its input impedance and radiation pattern was measured. One of the design goals is to fabricate an antenna with very small dimensions including the size of the ground plane ($\ll \lambda_0$). However, in general there is a problem with characterizing the input impedance and radiation pattern of such antennas with unbalanced feed. For measuring the input impedance using a network analyzer or the radiation pattern using a spectrum analyzer, a long coaxial cable must be connected to the antenna. The problem is that the antenna near-field is coupled to the coaxial cable over which an electric current can get excited. As a result, both the input impedance and the radiation pattern of the antenna are changed. Fig. 12 depicts the electric field distribution around the proposed antenna fed by a coaxial cable. Significant electric current exists on the external layer of the coaxial cable that unexpectedly contributes to radiation. This causes a shift in the resonant frequency, poor impedance matching and radiation pattern. Although omni-directional radiation pattern is expected (see Fig. 13(a)), deformed radiation pattern is actually measured as shown in Fig. 13(b). At $\theta = 90^\circ$, which is near ground observation, the gain is dropped from the expected 0 dBi to -8 dBi.

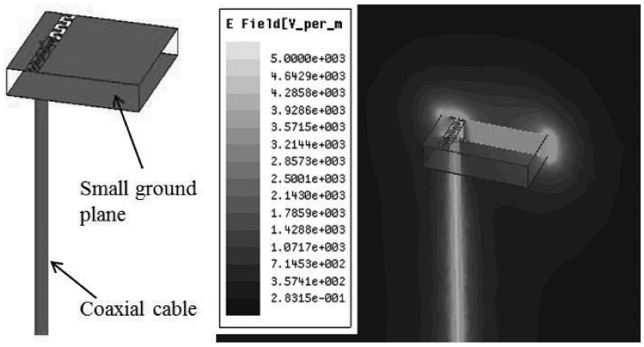

Fig. 12. Electric field distribution in the proposed antenna fed by coaxial cable.

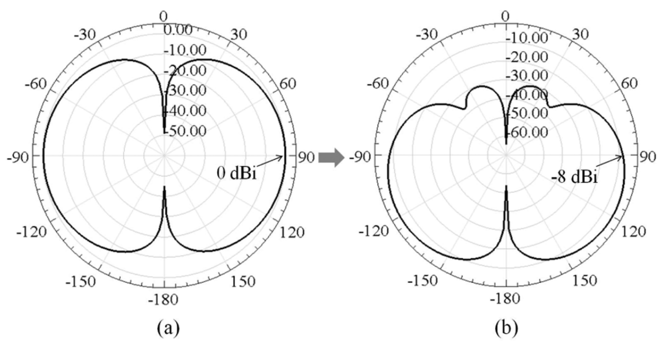

Fig. 13. Simulated E-Plane radiation pattern of single-layer ICCLMA (a) without and (b) with a coaxial cable.

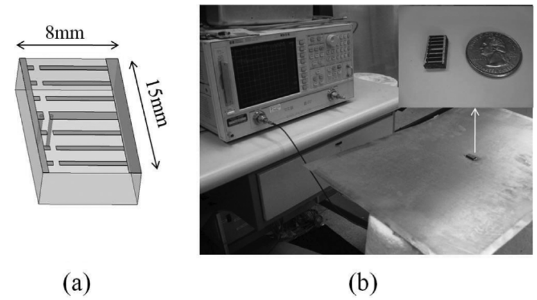

Fig. 14. (a) Miniaturized single-layer ICCLMA and (b) measurement set up using a large ground plane ($3\lambda_0 \times 3\lambda_0$).

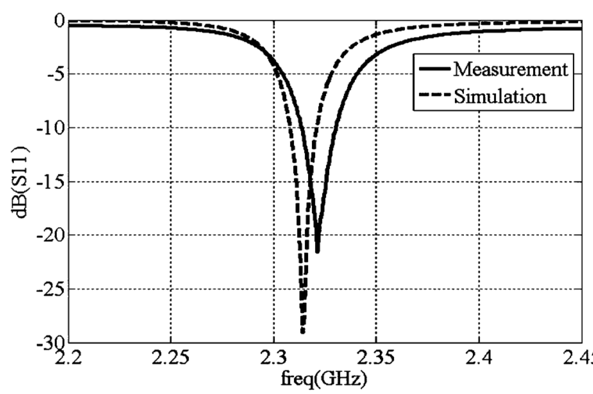

Fig. 15. Measured and simulated S_{11} of a miniaturized single-layer ICCLMA on a large ground plane.

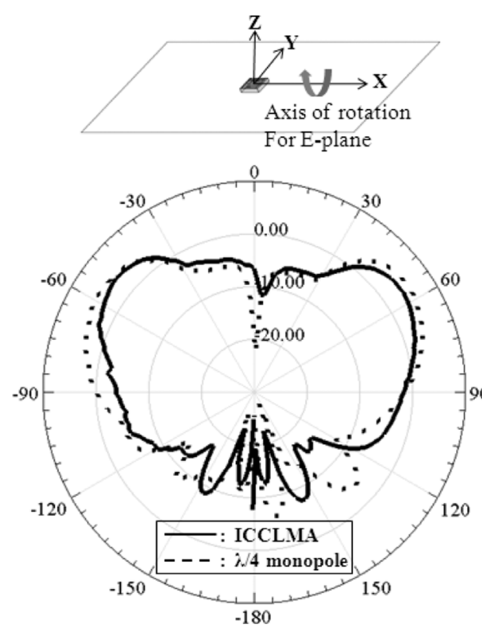

Fig. 16. Measured E-Plane radiation patterns of a ICCLMA and $\lambda_0/4$ monopole antenna.

B. Validation of the Proposed Design Method Using a Large Ground Plane

In order to avoid the coupling to the feed line and the resulted undesirable radiation problem from the feed line connected to the antenna with electrically small ground plane, one can use a very large ground plane. The proposed antenna placed on a large ground plane is measured by a vector network analyzer. It should be pointed out that the design parameters of the antenna on a large ground plane are slightly different from those of the same antenna on a small ground plane. A miniaturized single-layer ICCLMA on $3\lambda_0 \times 3\lambda_0$ ground plane is designed, fabricated and measured as shown in the measurement setup of Fig. 14. Fig. 15 shows the measured reflection coefficient, compared with the simulation result. A good agreement is observed. Measured radiation pattern of this antenna is compared to that of a $\lambda_0/4$ monopole antenna on the same ground plane. Fig. 16 shows the overall gain of the proposed antenna operating at the resonant frequency is comparable to that of a $\lambda_0/4$ monopole antenna.

C. Antenna Measurement Using a Balanced Architecture

Miniaturized monopole antennas are considered for small wireless nodes with small platforms where the back of the antenna ground plane can support the active elements. As described in the previous section, the characterization of small antennas by connecting a long coaxial cable leads to uncertainties in the antenna radiation parameters. One approach is to use large ground planes ($>2\lambda_0$ on each side) as was shown in the previous section, but at low frequencies the size of the ground plane may become prohibitively large. To avoid this problem, an alternative method to measure the input impedance and radiation pattern of electrically small antennas is presented. This can be done using a balanced architecture or simply the dipole version of the monopole antenna (see Fig. 17). A balanced architecture produces a null surface in the plane bisecting the dipole structure. In this plane, any metallic structure like the antenna feed can be inserted without affecting the antenna

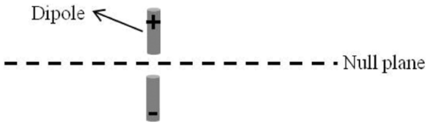

Fig. 17. Measurement method using a balanced architecture.

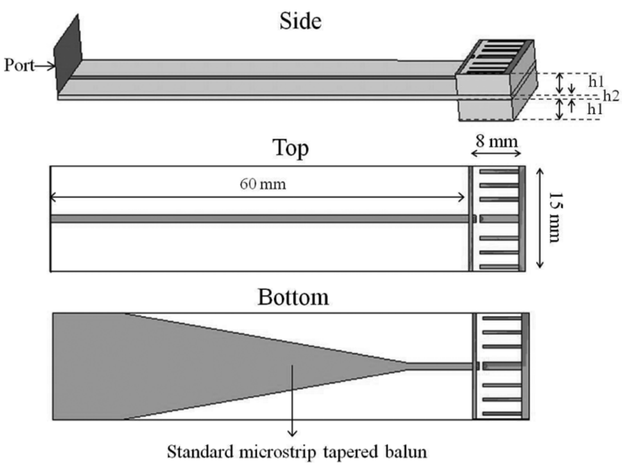

Fig. 18. Side, top and bottom views of measurement set up of a balanced miniaturized antenna compatible with unbalanced measurement system.

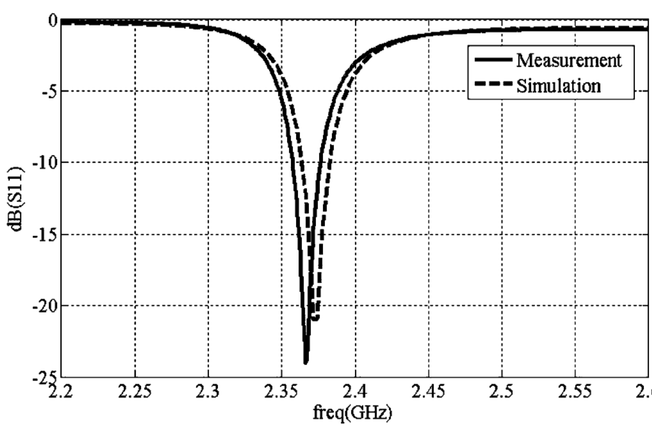

Fig. 19. Measured and simulated S_{11} of the proposed balanced architecture.

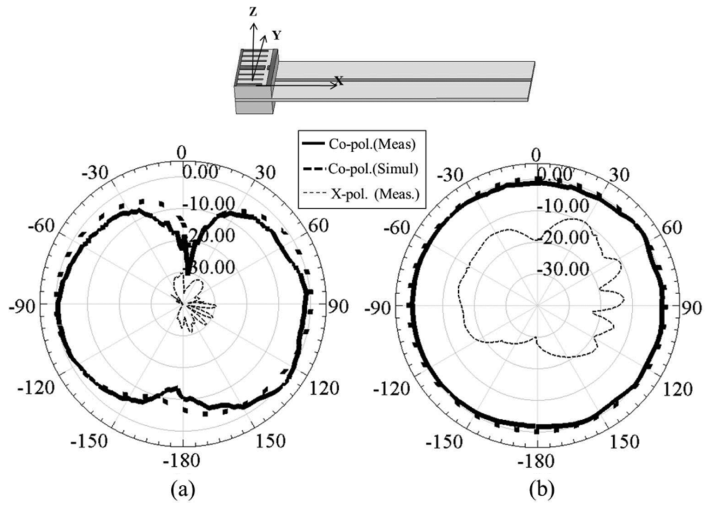

Fig. 20. Measured and simulated (a) E-Plane ($=\mathrm{XZ\text{-}Plane}$) and (b) H-Plane ($=\mathrm{XY\text{-}Plane}$) radiation patterns of the proposed balanced architecture.

TABLE II
GAIN AND RESONANT FREQUENCY OF SINGLE-LAYER ICCLMAS WITH THREE DIFFERENT DIMENSIONS

Size (mm X mm)	fc (GHz)	Gain (dB)	BW (%)	G1	G2	W1	W2	L1	L2
15 X 15	2.39	0	0.46	0.5	0.5	1	0.5	2	11.5
12 X 15	2.358	-0.6	0.38	1	3	1	0.5	0.5	8.5
8 X 15	2.362	-1.9	0.36	2	7	0.5	0.5	0.5	0.5

characteristics. The only difficulty is that a balun structure must be used. Although measuring a balanced version of the antenna is not representative of the performance of the antenna on small ground, by redesigning and testing the antenna on small ground as a balanced topology it can be examined whether the simulation results of the antenna previously designed on small ground is trustable or not.

Fig. 18 shows a balanced low-profile miniaturized antenna connected to a microstrip balun in the null plane. The standard microstrip tapered balun is connected to two back-to-back ICCLMAs. The vertical dimensions of the antenna and balun set up shown in Fig. 19 are given by $h1 = 3.175$ mm and $h2 = 0.787$ mm. Fig. 20 shows the measured and simulated reflection coefficients of a balanced architecture. The simulated fractional 10-dB return loss bandwidth is 0.78%, while that of 8 mm $\times$ 15 mm single ICCLMA with small ground plane is 0.36% (see Table II). The simulated and measured gains in Fig. 20 are 0.66 dBi and 0.36 dBi, which are higher than that of 8 mm $\times$ 15 mm single ICCLMA with small ground plane in Table II. This is not a surprise because the volume of the dipole ICCLMA is twice that of the monopole version. In Fig. 20, measured radiation pattern shows good agreement with the simulated results. We also tested the antenna with different length of balun and feed line, and found out that the antenna input impedance and radiation pattern remain unchanged regardless of the length of the feeding lines. Therefore, placing active components with differential (= balanced) outputs in the null plane, this balanced architecture can be used for small RF platforms requiring purely vertically polarized radiation.

V. CONCLUSION

A novel miniaturized low-profile vertically polarized antennas is presented. The gain and polarization improvement are achieved by isolating the feed structure from the miniaturized resonant radiating structure and using inductive, top loaded and in-plane capacitive couplings appropriately. Using this approach, a high gain low profile miniaturized antenna with omnidirectional and vertically polarized radiation, similar to a short dipole is designed and fabricated. The gain of this antenna is 9 dB and 4 dB higher than those of the conventional inverted-F and the LMMMA [1], respectively. In addition, the proposed antenna shows excellent suppression of horizontally polarized radiation in the horizontal plane (near ground observation). A procedure is also outlined to achieve further size reduction which, of course, results in lower antenna gain. Measurement method and set up for antennas with electrically small ground planes are discussed, and it is shown that a dipole version of a monopole design can be used to easily characterize the antenna performance.

REFERENCES

[1] W. B. Hong and K. Sarabandi, "Low-profile, multi-element, miniaturized monopole antenna," *IEEE Trans. Antennas Propag.*, vol. 57, no. 1, pp. 72–80, Jan. 2009.

[2] D. Liao and K. Sarabandi, "Terminal-to-terminal hybrid full-wave simulation of low-profile, electrically-small, near-ground antennas," *IEEE Trans. Antennas Propag.*, vol. 56, no. 3, pp. 806–814, Mar. 2008.

[3] D. Liao and K. Sarabandi, "Optimization of low-profile antennas for applications in unattended ground sensor networks," *IEEE Trans. Antennas Propag.*, vol. 53, no. 11, pp. 3747–3756, Nov. 2005.

[4] N. Behdad and K. Sarabandi, "Bandwidth enhancement and further size reduction of a class of miniaturized slot antennas," *IEEE Trans. Antennas Propag.*, vol. 52, pp. 1928–1935, Jan. 2004.

[5] W. Hong, N. Behdad, and K. Sarabandi, "Size reduction of cavity-backed slot antennas," *IEEE Trans. Antennas Propag.*, vol. 54, pp. 1461–1466, 2006.

[6] T. Taga and K. Tsunoda, "Analysis of a planar inverted-F antenna by spatial network method," *IEICE, B-2*, vol. J74-B-2, no. 10, pp. 538–545, Oct. 1991.

[7] S. Sekine, T. Ito, N. Odachi, Y. Murakami, and H. Shoki, "Design method for a broadband inverted-F antenna by parallel resonance mode," *IEICE, B*, vol. J86-B, no. 9, pp. 1806–1815, Sep. 2003.

[8] J. McLean, H. Foltz, and G. Crook, "Broadband, robust, low profile monopole incorporating top loading, dielectric loading, and a distributed capacitive feed mechanism," in *Proc. IEEE Int. Symp. Antennas Propag.*, Jul. 11–16, 1999, vol. 3, pp. 1562–1565.

[9] S. Tokumaru, "Multiplates: Low profile antennas," in *IEEE Int. Symp. Antennas Propag.*, Oct. 1976, vol. 14, pp. 379–382.

[10] N. Herscovivi and E. Dziadek, "Omnidirectional antennas for wireless communication," in *Proc. IEEE Int. Symp. Antennas Propag.*, Jul. 11–16, 1999, vol. 1, pp. 556–559.

[11] T. Noro and Y. Kazama, "Low profile and wide bandwidth characteristics of top loaded monopole antenna with shorting post," in *Proc. IEEE Int. Workshop on Antenna Technol. Small Antennas and Novel Metamater.*, Mar. 6–8, 2006, pp. 108–111.

[12] C. Delaveaud, P. Levegue, and B. Jecko, "Small-sized low-profile antenna to replace monopole antennas," *Electron. Lett.*, vol. 34, pp. 716–717, Apr. 1998.

[13] G. Goubau, "Multielement monopole antennas," in *Proc. Workshop on Electrically Small Antennas ECOM, Ft.*, Monmouth, NJ, May 1976, pp. 63–67.

[14] C. R. Rowell and R. D. Murch, "A capacitively loaded PIFA for compact mobile telephone handsets," *IEEE Trans. Antennas Propag.*, vol. 45, pp. 837–842, May 1997.

[15] R. L. Li, G. DeJean, M. M. Tentzeris, and J. Laskar, "Development and analysis of a folded shorted-patch antenna with reduced size," *IEEE Trans. Antennas Propag.*, vol. 52, pp. 555–562, Feb. 2004.

[16] C. Y. Chiu, K. M. Shum, and C. H. Chan, "A tunable via-patch loaded PIFA with size reduction," *IEEE Trans. Antennas Propag.*, vol. 55, no. 1, pp. 65–71, Jan. 2007.

[17] A. S. Meier and Summers, "Measured impedance of vertical antennas and effects of finite ground planes," *Proc. IEEE*, vol. 37, pp. 609–616, 1969.

[18] K. H. Awadalla and Maclean, "Input impedance of a monopole antenna at the center of a finite ground plane," *IEEE Trans. Antennas Propag.*, vol. AP-26, pp. 244–248, 1978.

[19] Weiner, "Monopole element at the center of a circular ground plane whose radius is small or comparable to a wavelength," *IEEE Trans. Antennas Propag.*, vol. AP-35, pp. 488–495, 1987.

[20] Bhattacharyya, "Effects of ground plane and dielectric truncations on the efficiency of a printed structure," *IEEE Trans. Antennas Propag.*, vol. 39, pp. 303–308, 1991.

[21] M. C. Huynh and W. Stutzman, "Ground plane effects on the planar inverted-F antenna (PIFA) performance," *IEE Proc. Microwave Antennas Propag.*, vol. 150, no. 4, pp. 209–213, Aug. 2003.

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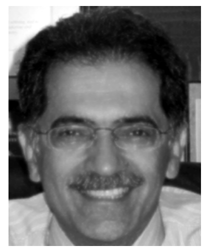

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