Proximity coupled fed wideband printed magneto–electric dipole antenna

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Results are presented fora wideband printed magneto-electric dipole antenna. The printed electric dipole and reflector are separated by an air-gap and is fed with a coaxial transmission line and a printed proximity coupled T-shaped feeding strip. The slot between the electric dipole halves forms the magnetic dipole radiating aperture. This feeding mechanism allows for good impedance matching at adjacent resonances of the electric and magnetic dipoles to achieve wideband operation. The two printed electric dipole halves are DC shorted to the reflector. The results show that this antenna can operate over a wide bandwidth with good gain, stable unidirectional radiation patterns and low cross polarization. The effect of the air-gap height and the relative permittivity of the substrate on which the dipole is printed are investigated. Results show that an optimum air-gap height for a specific substrate permittivity can be determined that will result in a gain with a small variation over the entire impedance bandwidth of the antenna. Simulated as well as measured results are presented for a prototype antenna with a small gain variation and overall dimensions of $1.33\lambda_0 \times 1.33\lambda_0 \times 0.19\lambda_0$. A measured impedance bandwidth of 54% with an average gain of 9.2 dBi is achieved.

Introduction: The magneto-electric dipole antenna has been extensively investigated as a wideband and dual-band antenna with very good unidirectional radiation properties, for example, similar radiation patterns in the *E*-plane and *H*-plane, good gain, low cross-polarisation, and low back radiation [1–5]. The most common magneto-electric dipole configuration is a combination of an electric dipole above a planar reflector and a magnetic dipole realised as a quarter-wavelength open-ended waveguide section or slot cavity [1–4]. Recently a magneto-electric dipole was presented as a coaxially-fed printed antenna with the advantage of easier construction and reduced production cost [5]. This printed version of the magneto-electric dipole antenna showed that the magnetic dipole (slot) still radiated efficiently even though the conducting slot cavity side-walls were removed. The printed magneto-electric dipole antenna [5] had reduced height ($0.16\lambda_0$) and less in-band gain variation, but also smaller impedance bandwidth (41%) when compared to some previously published magneto-electric dipole antennas.

An improved version of a printed magneto-electric dipole antenna is presented in this paper. The antenna is also fed with a coaxial transmission line but a printed proximity coupled T-shaped feeding strip is introduced, which allows for good impedance matching at adjacent resonances of the electric and magnetic dipoles to achieve wider bandwidth operation than in [5]. The symmetry of the structure is also improved by DC shorting both halves of the electric dipole to ground [2]. The effect of the air-gap height and the relative permittivity of the substrate on which the dipole is printed are investigated. An optimum air-gap height for a specific substrate permittivity can be determined that will result in a nearly constant antenna gain over the bandwidth. Simulated and measured results for a prototype antenna are also presented.

Printed dipole antenna geometry and design: A front view and cross-section of the new magneto-electric dipole antenna is shown in Figure 1. The two halves of the wide electric dipole are printed on the bottom of the substrate, and the T-shaped feeding strip is printed on the top of the substrate. The substrate and ground are separated by an air-gap. A coaxial feed line comes through the ground plane with the outer ground connected to the inner edge of one halve of the electric dipole, and the centre conductor connected to the printed T-shaped feeding strip. To improve the geometrical symmetry the inner edge of the other halve of the electric dipole is shorted to the ground with a conducting post (with the same dimension as the coaxial feed line). The magnetic dipole is represented by the slot (and cavity beneath the slot, without side-walls, except for the coaxial line and conducting post) between the wide electric dipole halves. The T-shaped feeding strip parameters provide sufficient flexibility to simultaneously achieve impedance matching of the electric and magnetic dipoles at different but closely spaced frequencies, in order to obtain wideband operation of the magneto-electric dipole pair.

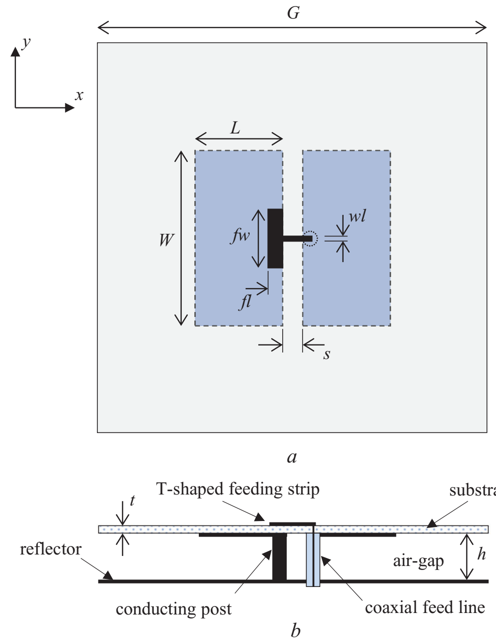


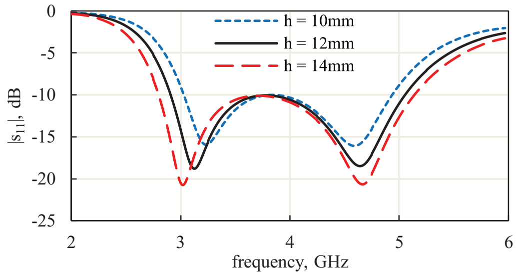
Fig. 1 *Geometry and dimensional parameters of magneto-electric dipole with coaxial proximity coupled feed. (a) Front view of antenna; (b) cross-section through centre of antenna*

*Table 1. Design data for antennas optimised for maximum bandwidth around a centre frequency of 4 GHz. (*G = *100 mm,* s = *4.0 mm,* fl = *1.5 mm,* t = *1.524 mm and* wl = *1.0 mm for all designs.)*

ε_r	*H* [mm]	*L* [mm]	*W* [mm]	*fw* [mm]	BW [%]	Gain variation [dBi]
3.38	10.0	18.2	28.0	5.5	47	9.45 ± 0.75
3.38	12.0	17.2	25.3	6.6	54	9.05 ± 0.35
3.38	14.0	16.6	22.9	7.8	59	7.90 ± 1.10
2.20	12.0	19.2	29.2	8.1	52	9.40 ± 0.70
3.38	12.0	17.2	25.3	6.6	54	9.05 ± 0.35
4.40	12.0	15.9	22.8	5.9	55	8.15 ± 0.95
2.20	14.3	18.7	29.4	7.6	54	9.05 ± 0.45
3.38	12.0	17.2	25.3	6.6	54	9.05 ± 0.35
4.40	10.3	16.5	24.7	5.1	50	9.10 ± 0.40

Several designs were performed using CST Studio Suite, for different substrate permittivities and air-gap heights. A centre frequency of 4 GHz was chosen. A good starting point before optimisation is an electric dipole length ($2L + s$) of $0.5\lambda_0$ and a magnetic dipole length (W) of $0.35\lambda_0$, where λ_0 is the free-space wavelength at the centre frequency. The reflection coefficient can be controlled by the parameters *L*, *W* and *fw*, once appropriate dimensions were chosen for *G*, *s*, *fl*, *t* and *wl*. The electric dipole resonant frequency ($< f_0$) is primarily determined by *L*, the magnetic dipole resonant frequency ($> f_0$) is primarily determined by *W*, and the optimum impedance matching by *fw*. Final design data is presented in Table 1. The impedance bandwidth (SWR $\leq$ 2) and gain variation over the bandwidth are also presented in Table 1.

Air-gap height and substrate permittivity investigation: The simulated reflection coefficient and realised gain as function of frequency for designs with the same substrate permittivity ($\varepsilon_r = 3.38$) and different air-gap heights are shown in Figure 2. The impedance bandwidth improves as air-gap height increases, and the gain variation over the bandwidth is substantially different for different air-gap heights.



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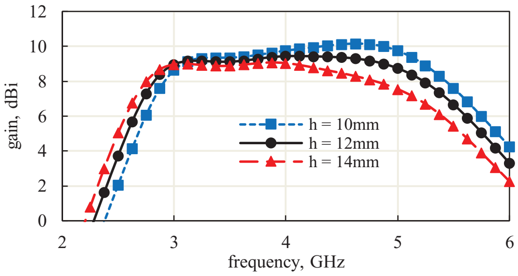

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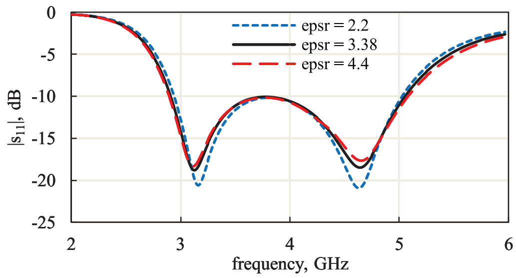
Fig. 2 *Simulated results for designs with the same substrate permittivity (ε_r = 3.38) and different air-gap heights (a) reflection coefficient; (b) realised gain*



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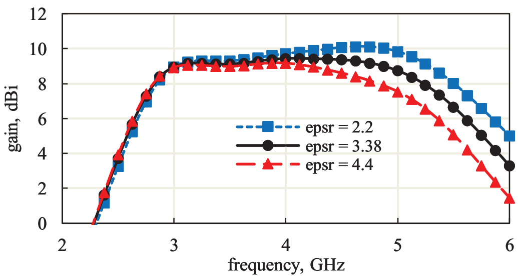

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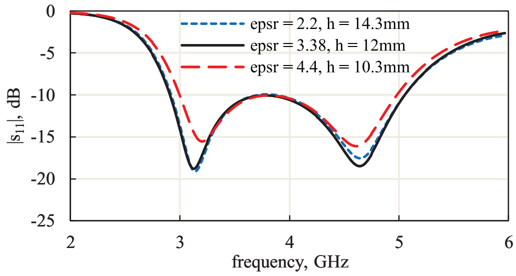
Fig. 3 *Simulated results for designs with the same air-gap height (h = 12 mm) and different substrate permittivities (a) reflection coefficient; (b) realised gain*

The results for designs with the same air-gap height ($h = 12$ mm) and different substrate permittivities are shown in Figure 3. The impedance bandwidth improves only slightly as ε_r increases, but the gain variation changes substantially for different ε_r values.

The results for designs with different dielectric substrates and optimised air-gap heights for constant gain over the bandwidth are shown in Figure 4. There is less than 1 dB gain variation over the impedance bandwidth for all three the designs. The impedance bandwidths for the $\varepsilon_r = 3.38$ ($h = 12$ mm) and $\varepsilon_r = 2.2$ ($h = 14.3$ mm) designs are almost exactly the same (54%), which is an indication that a converged optimum bandwidth solution (for constant gain) has been reached.

Figure 5 shows electric fields represented as isolines in the *xy*-plane just beneath the substrate (for the $h = 12$ mm and $\varepsilon_r = 3.38$ design). The expected field patterns associated with the electric dipole and magnetic dipole (slot) is clearly evident at the respective resonant frequencies.

Simulated radiation patterns: Simulated radiation patterns for the antenna design with $h = 12$ mm and $\varepsilon_r = 3.38$ are shown in Figures 6



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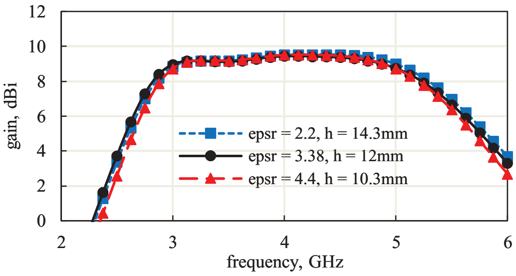

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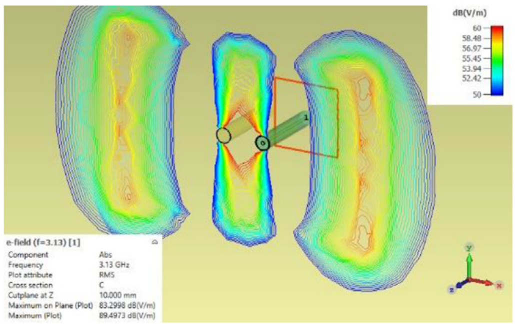
Fig. 4 *Simulated results for designs different dielectric substrates and optimised air-gap heights (a) reflection coefficient; (b) realised gain*



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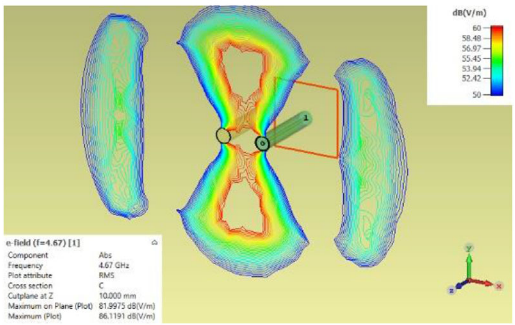

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Fig. 5 *Isoline electric field magnitude representation in the* xy*-plane (a) at the electric dipole resonance,* f = *3.13 GHz; (b) at the magnetic dipole resonance,* f = *4.67 GHz*

and 7. *E*- and *H*-plane patterns are shown at the centre frequency of 4 GHz, and two more frequencies (3 and 5 GHz) close to the impedance bandwidth edges. The patterns are very stable and show very little variation between the different frequencies, and exhibit very good front-to-back ratios (better than 17 dB) and very good cross-polarisation levels in both planes (below −25 dB).

Experimental results: A prototype antenna (with $h = 12$ mm) was manufactured and measured. A Rogers RO4003C substrate with $\varepsilon_r = 3.38$, $\tan\delta = 0.0021$ and $t = 1.524$ mm was used. A semi-rigid coaxial cable with an outer radius of 1.1 mm was used as feed. Figure 8 shows the simulated and measured reflection coefficient and realized gain as function of frequency. Good agreement between the simulated and measured results was obtained. A 54% impedance bandwidth (SWR $\leq$ 2) was

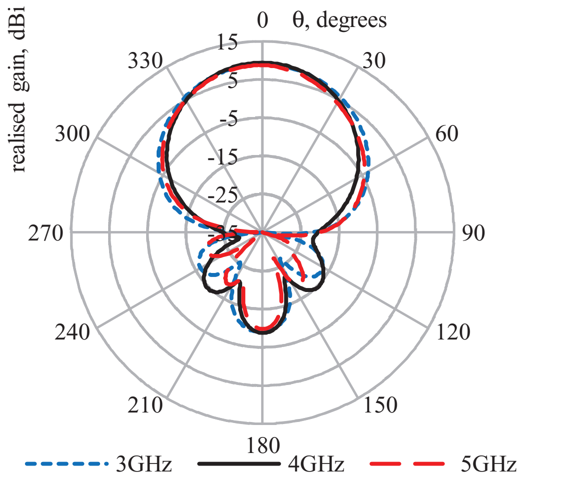

Fig. 6 *Simulated E-plane co-polarisation radiation patterns (cross-polarization levels < −35 dBi) for design with* h = *12 mm and* ε_r = *3.38*

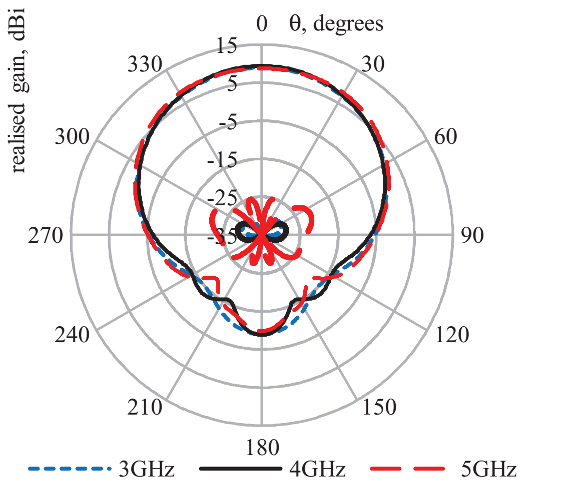

Fig. 7 *Simulated H-plane co- and cross-polarisation radiation patterns for design with* h = *12 mm and* ε_r = *3.38*

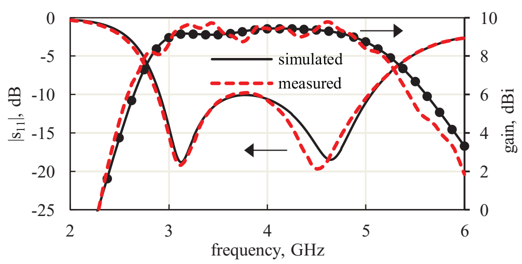

Fig. 8 *Simulated and measured reflection coefficient and realised gain for design with* h = *12 mm and* ε_r = *3.38*

measured. The gain was measured in a compact antenna range with a time gate to eliminate reflections from the reflector and sidewalls – the maximum deviation of about 0.6 dB within the antenna impedance bandwidth is within the accepted measurement accuracy of the facility. The average measured gain within the impedance bandwidth was 9.2dBi, compared to a simulated average of 9.1 dBi.

The measured radiation patterns are shown in Figures 9 and 10, for the centre frequency of 4 GHz, and the two bandwidth edge frequencies (3 and 5 GHz). The correlation with the simulated patterns in Figures 6 and 7 is very good. The measured front-to-back ratio is better than 17 dB, and the boresight cross-polarisation is below −20 dB at all three frequencies. The pattern shapes are remarkably constant over the impedance bandwidth.

Conclusion: A proximity coupled T-shaped probe feed allows wide-band impedance matching of printed magneto-electric dipole antennas.

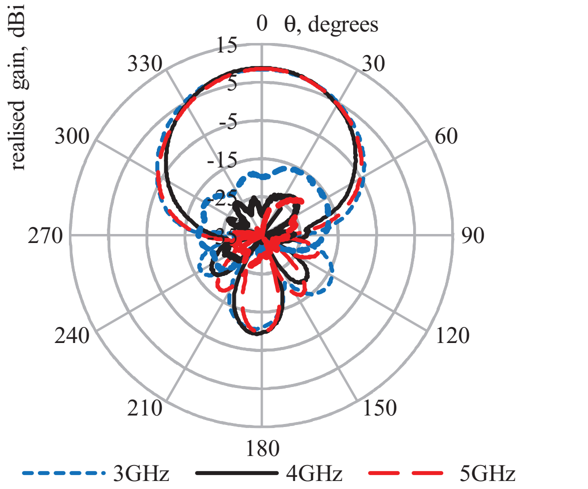

Fig. 9 *Measured E-plane co- and cross-polarisation radiation patterns for design with* h = *12 mm and* ε_r = *3.38*

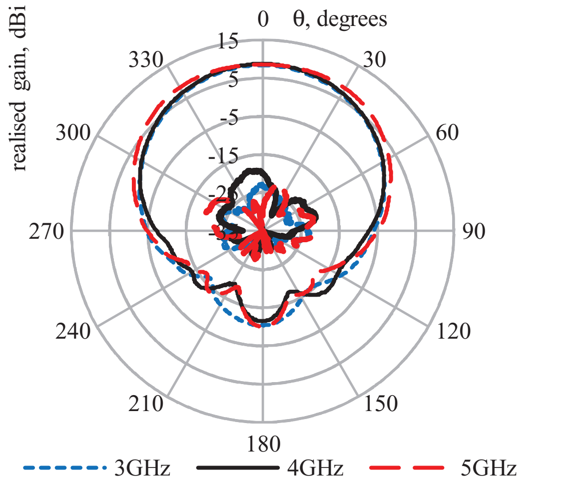

Fig. 10 *Measured H-plane co- and cross-polarisation radiation patterns for design with* h = *12 mm and* ε_r = *3.38*

For optimum impedance bandwidth a large air-gap and high permittivity substrate should be used. For constant gain over the bandwidth an optimum air-gap height has to be used for a specific substrate permittivity. The upper limit of achievable "small gain variation" bandwidth for this specific geometry is around 54%. Shaping of the electric and magnetic dipole geometries may result in improved bandwidth performance.

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