A Truncated Waveguide Fed by a Microstrip as a Multi-Band WLAN Antenna

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Abstract **─** This paper presents a small truncated waveguide fed by a microstrip line through a transverse coupling slot for multi-band WLAN applications. The analysis of the presented antenna has been performed by replacing the feeding microstrip line with its equivalent Magnetic-Wall Waveguide (MWW) model and using an in-house Method of Moments (MoM) to analyze the structure. The use of entire-domain basis functions both on the slot and on the truncated waveguide apertures allows to obtain a very efficient analysis by exploiting the waveguide modes orthogonality. The proposed radiating element allows to obtain a high radiated power, with a very low cross-polar component in the radiated field, with a -10 dB bandwidth that covers the required frequencies for multi-band WLAN applications (5.2/5.4/5.8 GHz).

Index Terms – Microstrip, waveguide antennas, WLAN antennas.

I. INTRODUCTION

The increasing demands of wireless and short-range and high data rate transmissions, pushed to propose new wireless protocols using different bands of the frequency spectrum in order to support high data rate wireless communications; which led to the request of antennas able to operate at different frequency bands simultaneously (multi-band antennas). In WLAN applications, the most common desirable requirement consists of providing multi-band operations, such as covering 5.2, 5.4 and 5.8 GHz bands for IEEE 802.11a standard.

A number of interesting multi-band antennas for Wireless Local Area Networks (WLAN, IEEE 802.11b/a standard) have been proposed and reported in the literature [1-5]. The most popular solution is the use of microstrip patch antennas, both for their small weight and size, low production cost and ease of fabrication and integration [6]. However, such antennas have a low efficiency, a narrow frequency bandwidth and a relatively high cross-polar component of the radiated field. Moreover, the efficiency of patch antennas is further reduced by the surface waves in the dielectric substrate, which causes power loss and high coupling between the elements in an array environment. All those drawbacks are well known and much literature has been devoted to overcome them. A common choice to improve patch antenna performances is the use of coupling slots in the feeding network in structures with two sandwiched dielectric substrates [7,8]. This solution allows to optimize separately the feeding and the radiating circuit, but increases the production cost and both the design and the realization processes are more difficult.

The narrow frequency band of printed antennas pushed to search for new configurations suitable for broadband applications, which are required for high-speed transfer; such as planar monopoles [9], printed dipoles [10] or slot antennas [11]. The structure proposed in this paper consists of a truncated rectangular waveguide radiator [12], [13], which replaces the patch (and its substrate). This truncated waveguide is fed by the electromagnetic coupling produced by a microstrip transverse slot (Fig. 1). In this way, the

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flexibility (and ease of realization) of the feeding network is retained, but the electromagnetic behaviour of the radiating element is strongly improved. Actually, this configuration allows higher efficiency and a much lower cross-polar component, with respect to a standard patch antenna. Furthermore, the presented structure allows to exploit the advantages of planar technology for the power supply circuit, allowing the realization of matching networks at a low cost and with compact size; especially if compared with the corresponding waveguide network. Another advantage of the described structure is the absence of waveguide-to-microstrip transitions. The performance degradation caused by transitions between planar structures and waveguides can be seen in [14], where a broadband array of circular waveguide radiators is designed with a stripline distribution network fed by a rectangular waveguide. In this structure, the overall efficiency of the array is restricted by unwanted reflections caused by the presence of transitions between the circular waveguide and the stripline and between the stripline and the rectangular waveguide. Finally, the truncated waveguide, as a radiating element, allows to obtain a radiated electromagnetic field with a high polarization purity [13].

The potential drawbacks of the presented structure could be the surface wave, which propagates in the feeding microstrip substrate and the back radiation of the coupling slot. However, the surface wave losses can be minimized by using substrates with suitable dielectric constant and thickness and the back radiation; due to the coupling slot not high since the slot is not resonant (having typical dimensions equal or less than a quarter wavelength) and can be minimized by a suitable choice of the structure parameters.

The radiated power can be modulated by acting on various parameters, such as the length of the coupling slot, its width, the truncated waveguide length and transverse dimensions. Moreover, a thin dielectric slab lying on the slot can be placed inside the radiating element to increase the degrees of freedom in the design process.

The proposed radiating element is designed in this work for multi-band WLAN applications, allowing to obtain a high radiated power, a very low cross-polar component in the radiated field and with a -10 dB bandwidth that covers the required frequency bandwidths (5.2/5.4/5.8 GHz). The analysis of the antenna in Fig. 1 has been performed by replacing the feeding microstrip line with its equivalent Magnetic-Wall Waveguide (MWW) model [15] (as indicated in Fig. 2) and using the well-assessed Method of Moments (MoM) procedure described by the authors in [12], [16-20]. The use of entire-domain basis functions both on the slot and on the truncated waveguide apertures, allows to obtain a very efficient analysis, since their choice has been made in order to exploit the waveguide modes orthogonality.

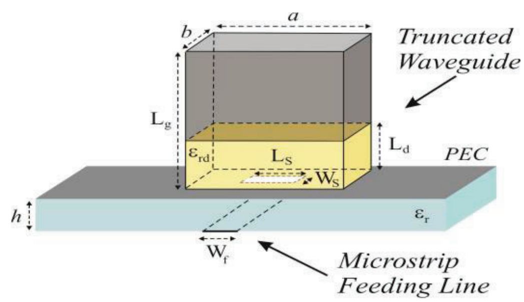

Fig. 1. Radiating element geometry.

II. NUMERICAL ANALYSIS OF THE ANTENNA

The proposed radiating element can fulfill a large range of requirements, thanks to the relatively large number of degrees of freedom. However, the effective use of this flexibility in the design calls for an efficient and accurate analysis procedure. Therefore, our choice has fallen on the Method of Moments (MoM) in the Galerkin formulation using entire domain basis functions (EDBF) [16,21]. As a matter of fact, though it requires some efforts to be devised, compared with differential approaches implemented into general purpose EM solvers, its accuracy and computational effectiveness can hardly be paralleled. Of course, a suitable model of the structure must be devised in order to build-up an in-house software for the analysis of it. The two more critical points are the requirements that the antenna aperture is cut into an infinite PEC plane and the modeling of the feeding microstrip. The former is quite standard [22] and will not be further discussed. On the other hand, an effective modeling of the microstrip line has been achieved by representing the microstrip line with an equivalent Magnetic-Wall Waveguide (MWW) [23]. This equivalence is a popular tool for analyzing microstrip discontinuities, but it is accurate also for microstrip slots, as long as those slots are well within the MWW [15]. Therefore, the structure that has been actually analyzed is shown in Fig. 2. It consists of a magnetic wall waveguide replacing the microstrip feeding line (Fig. 1), with a transverse coupling slot cut in its ground plane, which feeds a truncated waveguide. The truncated waveguide can be partially filled with a dielectric slab lying on the slot. The slot between the apertures Σ^i and Σ^e radiates into the region bounded by the waveguide and the aperture Σ^r is the antenna element, which radiates into free space.

The MoM full-wave analysis of the structure in Fig. 2 starts by replacing each aperture with an unknown magnetic current. Let M^e, M^i the currents on the two sides of the slot, and M^r the current on the external aperture Σ^r. The continuity of the tangential magnetic field is then enforced on each aperture:

$$\begin{cases} \underline{H}_s[\underline{M}^i,\underline{M}^e]-\underline{H}_w[\underline{M}^i]=\underline{H}_{inc} & \text{on } \Sigma^i \\ \underline{H}_s[\underline{M}^i,\underline{M}^e]+\underline{H}_{wt}[\underline{M}^e,\underline{M}^r]=0 & \text{on } \Sigma^e \\ \underline{H}_{wt}[\underline{M}^e,\underline{M}^r]+\underline{H}_{fs}[\underline{M}^r]=0 & \text{on } \Sigma^r \end{cases}, \quad (1)$$

where $\underline{H}_{inc}$ is the impressed magnetic field incident under the slot in the feeding microstrip line, $\underline{H}_w$ is the magnetic field in the microstrip region, $\underline{H}_s$ is the magnetic field in the slot region, $\underline{H}_{wt}$ is the magnetic field in the truncated waveguide region and $\underline{H}_{fs}$ is the magnetic field in the free-space region.

Equations (1) are actually a system of integral equations in the unknown currents. In order to solve it, we expanded all unknown in a suitable set of EDBF. The slot can be considered as a waveguide with transverse sections L_S and W_S. Therefore, our EDBF are on all apertures, the lowest-order modes of the relevant waveguides. More precisely, since the slot is narrow we have neglected the longitudinal component of the electric field on it. Therefore, only the magnetic currents directed along x are used as unknowns and are expressed as truncated sinusoidal series with respect to x:

$$\underline{M}^{(i,e)}=\sum_{p=1}^{N}a_p^{(i,e)}\sin\left[\frac{p\pi}{L_S}\left(x+\frac{L_S}{2}\right)\right]\underline{i}_x = \sum_{n=1}^{N}a_p^{(i,e)}\,\underline{m}^p(x)\,. \quad (2)$$

The magnetic current on the aperture Σ^r is expressed as a linear combination of the truncated waveguide modes. With this choice, in fact, it is possible to exploit the orthogonality of the modal functions (each basis function will excite only one mode of the waveguide), strongly simplifying the calculations of the MoM matrix elements:

$$\underline{M}^r=\sum_{k,j}^{N,M}a_{k,j}^r\sin\left[\frac{k\pi x}{a_T}\right]\cos\left[\frac{j\pi y}{b_T}\right]\underline{i}_x + b_{k,j}^r\cos\left[\frac{k\pi x}{a_T}\right]\sin\left[\frac{j\pi y}{b_T}\right]\underline{i}_y = \sum_{k,j}^{N}a_{k,j}^r\,\underline{m}_x^{k,j}(x,y)+b_{k,j}^r\,\underline{m}_y^{k,j}(x,y)\,. \quad (3)$$

In equations (2) and (3), $a_p^{(i,e)}$, $a_{k,j}^r$ and $b_{k,j}^r$ are the unknown coefficients of the linear combination.

In the region of the magnetic wall waveguide equivalent to the microstrip feeding line and in the free-space region, the fields are calculated through the Green function of the vector potential $\underline{F}$ in the spectral domain, using a Fourier representation for the currents and potentials [15].

In the two waveguide regions, the fields are simply computed using a model expansion; since every unknown excites a single mode, because of the mode orthogonality.

Expansions (2) and (3) are inserted into (1), which are then projected on the same EDBF used to express the unknown, to get the final linear system of the MoM.

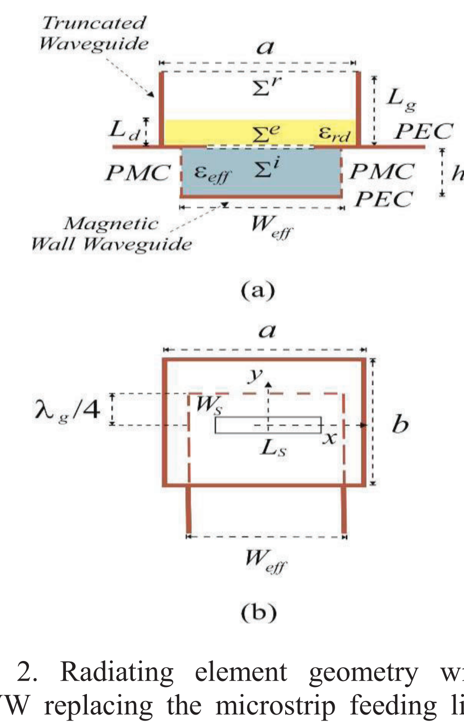

Fig. 2. Radiating element geometry with the MWW replacing the microstrip feeding line: (a) front view and (b) top view.

The reader is referred to section II of [12] for further details of the MoM procedure.

The computational time is mainly due to the matrix filling, while the solution is quite fast. Since the MoM matrix using EDBF is small and relatively well conditioned, a further increase in the computational efficiency can be gained if the matrix is computed only in few points and then interpolated to build-up the full antenna response [16, 24].

Note that the use of the equivalence between a microstrip line and a magnetic-wall waveguide (MWW) [15] is a key point. As a matter of fact, the microstrip line needs not to be discretized and therefore we have to solve a MoM system with about a dozen of unknowns, with respect to the hundreds of unknowns required by a standard planar MoM.

III. ANTENNA DESIGN AND RESULTS

In the analysis of the proposed antenna, the designer has to take into account some specific properties of the structure, in order to obtain an efficient and performing WLAN antenna. In particular, the geometrical dimensions of the radiating element must fulfill the following requirements:

1. The microstrip feeding line must be designed providing that, at the operating frequencies only the fundamental quasi-TEM mode propagates.
2. The transverse dimension a of the truncated waveguide must be chosen in order that only the fundamental TE_{10} mode is excited and the higher modes attenuation is large enough to prevent higher-order mode coupling between the slot and the aperture.

The MoM procedure described in this work has been widely assessed in [12], [16] and [17]; while in [13], both the equivalence between the MWW and the microstrip for our application and the MoM code have been validated by comparison with a general purpose FEM commercial software, Ansys HFSS. In Fig. 3 (a) of [13], a comparison has been performed between the structure fed by the microstrip line and the structure fed by its equivalent magnetic wall waveguide; both simulated with a general purpose FEM commercial software, Ansys HFSS. The results of our MoM procedure have been also reported for a matter of completeness. These comparisons validate both the equivalence between the MWW and the microstrip for our application and the MoM code, since the results of the MoM code and HFSS are virtually equivalent. Since the computational time required by HFSS is about two orders of magnitude greater than the one required by our MoM code [12], [13], for the parametric analysis of the proposed radiating element, we have used the MoM procedure.

In the presented simulations, the design frequency is 5.5 GHz, the dielectric substrate is Roger Duroid 5800 with $\varepsilon_r = 2.2$ and h = 2 mm, the dielectric substrate partially filling the truncated waveguide is Rogers TMM4 with $\varepsilon_r = 4.5$ and the width of the coupling slot is $W_S = 1$ mm.

In order to satisfy the previous requirements 1 and 2 at the design frequency of 5.5 GHz, *a* has been fixed to the value of 50 mm and the width of the microstrip feeding line is $W_f = 6$ mm; which corresponds to a characteristic impedance of 50 Ω. The feeding microstrip line ends with a quarter wavelength termination beyond the coupling slot, as indicated in Fig. 2.

The circuital and radiating properties of the proposed antenna depend on the interaction between the aperture of the truncated waveguide and the coupling slot. The results presented in [12], [13], suggest that the radiated power can be modulated within a very wide range by adequately choosing the geometrical parameters of the structure.

In order to characterize the radiating element, we show its radiated power simulated using the MoM procedure described in section II; as a function of the slot length L_S and of the height L_d of the dielectric substrate partially filling the truncated waveguide for a length L_g of the truncated waveguide equal to 20 mm (Fig. 3). Similar curves are obtained by choosing different lengths for the truncated waveguide. The radiated power has been calculated by the scattering matrix of the structure obtained feeding the radiating element in Fig. 2 using two ports as, $1-|S_{11}|^2-|S_{21}|^2$. The results presented in Fig. 3 suggest that the radiated power can be modulated within a very wide range by adequately choosing the geometrical parameters of the structure. On the other hand, so as to match the radiating element in the required frequency bandwidths (namely, 5.2, 5.4 and 5.8 GHz bands following IEEE 802.11a standard for WLAN applications), its frequency response as a function of the geometrical parameters is shown in Fig. 4; where the reflection coefficient simulated using the MoM procedure described in section II, is shown for different slot lengths, L_S with a length of the dielectric substrate partially filling the truncated waveguide equal to L_d = 7.5 mm and with a length L_g of the truncated waveguide equal to 20 mm. Similar curves are obtained by choosing different lengths for L_d and L_g.

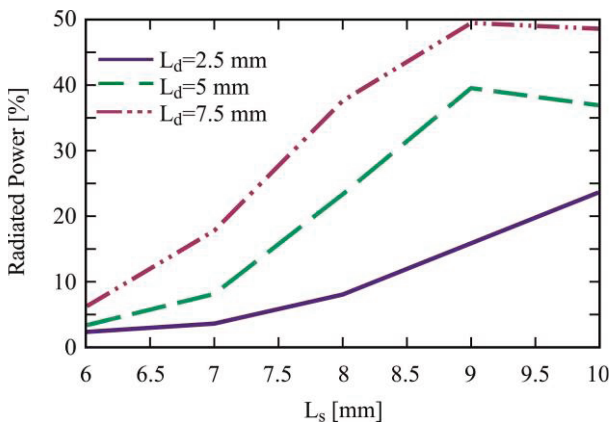

Fig. 3. Simulated (MoM) radiated power (percentage of incident power) for the antenna in Fig. 1, with L_g = 20 mm.

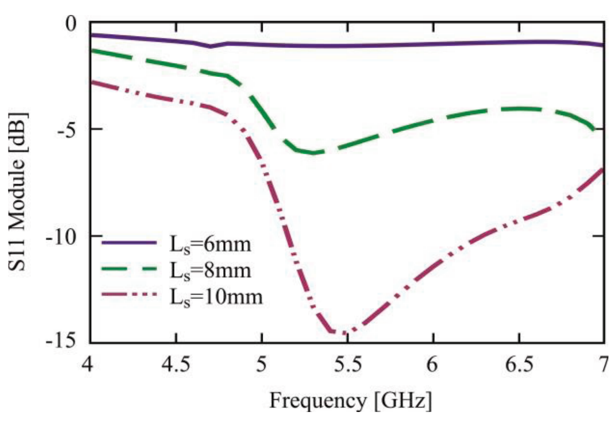

Fig. 4. Simulated (MoM) frequency response for the antenna in Fig. 1, with L_d = 7.5 mm and L_g = 20 mm.

A good compromise between the radiated power, the input matching and the dimensions of the structure is obtained with the following choice of the parameters: L_S = 10 mm, L_d = 7.5 mm and L_g = 20 mm. This choice corresponds to a radiated power around 50% and to a -10 dB bandwidth from 5.1 GHz to 6.4 GHz, as shown in Figs. 4 and 5; where the frequency response of the designed antenna simulated using the MoM procedure described in section II, is reported. In Fig. 5 the reflection coefficient is shown also for different values of the dielectric substrate partially filling the truncated waveguide. It is apparent that the best choice is the use of a dielectric with ε_r = 4.5, namely Rogers TMM4, as we already claimed at the beginning of this section. Figure 6 shows the E-plane and H-plane of the radiated fields for this antenna at the operating frequencies of 5.2, 5.4 and 5.8 GHz, computed using the MoM procedure described in section II. The cross-polar component of the radiated field is very low, with a value below -40 dB, with respect to the co-polar component and the gain of this element is about 6.5 dB in the whole operating bandwidth.

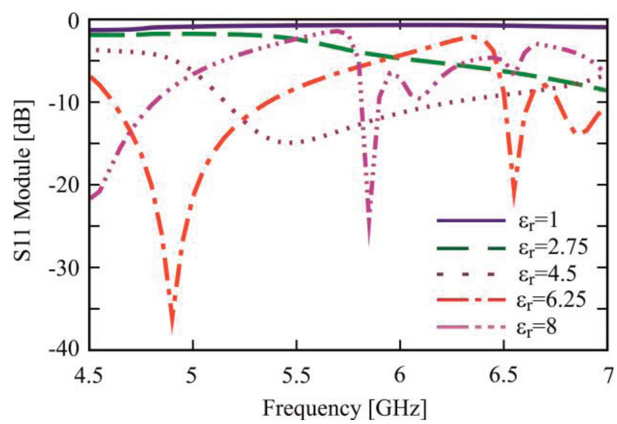

Fig. 5. Simulated (MoM) frequency response for the designed antenna (L_S = 10 mm, L_d = 7.5 mm and L_g = 20 mm).

In order to evaluate the performance improvement of the broadband antenna proposed in this paper, we can compare it with existing solutions for the same WLAN applications. In [25], the proposed multiband E-shaped printed monopole antenna for Multiple-Input–Multiple-Output (MIMO) system, has a good input matching over the operating frequency band; although, the radiated fields are not satisfactory if compared to the radiating element presented in our work showing a high cross-polar component and a significantly lower gain (around 3 dB). The same considerations can be made if we compare the antenna proposed in this paper with [2], [26] and [27]. These comparisons demonstrate that the presented antenna can be a very competitive candidate as a WLAN antenna.

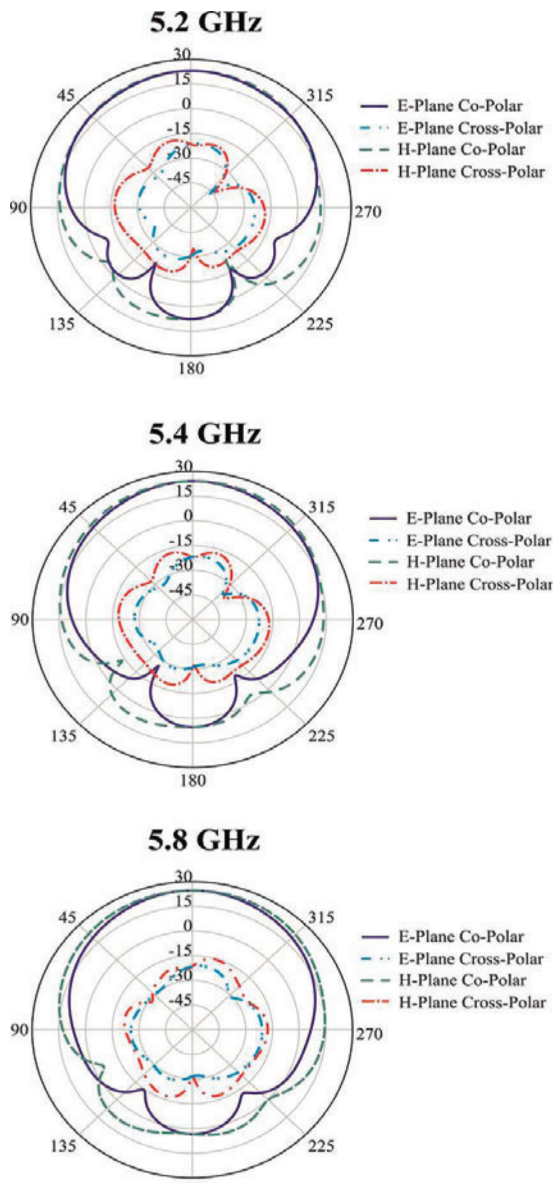

Fig. 6. Far field pattern of the designed antenna, computed using MoM (L_S = 10 mm, L_d = 7.5 mm and L_g = 20 mm).

VI. CONCLUSION

In this work, a small truncated waveguide fed by a microstrip line through a transverse coupling slot for multi-band WLAN applications is presented. An effective and accurate analysis procedure has been devised, using the Method of Moments (MoM), with entire domain basis functions to analyze the structure and replacing the feeding microstrip line with its equivalent Magnetic-Wall Waveguide (MWW) model. This choice allows to fully exploit the flexibility of the proposed radiating element.

A radiating element able to work as a multi-band WLAN antenna, from 5.1GHz to 6.4 GHz, has been obtained. It covers this frequency range with a very good input matching and a negligible cross-polar component in the radiated field and allows a higher radiated power than patch antennas.

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