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DESIGN OF A BALUN FOR A BOW TIE ANTENNA IN RECONFIGURABLE GROUND PENETRATING RADAR SYSTEMS

R. Persico

Istituto per il Beni Archeologici e Monumentali
Consiglio Nazionale delle Ricerche
Via Monteroni, Campus Universitario, Lecce 73100, Italy

N. Romano

Dipartimento di Ingegneria dell'Informazione
Seconda Università degli studi di Napoli
Via Roma 29, Aversa 81031, Italy

F. Soldovieri

Istituto per il Rilevamento Elettromagnetico dell'Ambiente
Consiglio Nazionale delle Ricerche
Via Diocleziano 328, Napoli 80124, Italy

Abstract—This paper deals with the design of a reconfigurable antenna that resembles a monolithic UWB bow-tie antenna for Ground Penetrating Radar (GPR) applications. In particular, the attention is focussed on the design of the balun system able to work in the frequency band 0.3–1 GHz; the effectiveness of the design is shown by examining the behaviour of the scattering parameters S_{11} for both the reference monolithic antenna and the designed reconfigurable antenna. Also, an analysis of the radiation pattern of both the monolithic and reconfigurable antennas is presented and confirms the effectiveness of the designed balun system.

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Corresponding author: Raffaele Persico (r.persico@ibam.cnr.it).

1. INTRODUCTION

Reconfigurable antennas are of great interest in communication systems and in imaging and diagnostics, with reference both to single antennas [1–3] and array of antennas [4, 5]. The flexibility of these systems is, in general, exploited to change the work frequency band and/or the direction of the main beam depending on the applicative necessities.

Recently, the idea of the reconfiguration has been translated also with regard to stepped frequency GPR systems with the aim of achieving a large work band with relatively simple (linearly polarised and electrically small) antennas. In this framework, some ideas and also a first investigation of the relevant trade offs are reported in [6, 7]; in those papers, attention is devoted to the feeding path of the antennas, which is switched vs. the frequency in order to improve the matching of the antennas or at least to ensure a sufficient received power on a large working frequency band.

With reference to the antenna to be designed in this work, which is a bow-tie antenna, several interesting efforts have been made in achieving improved performances about the UWB behaviour [8–10].

The possibility to make use of antennas with reconfigurable radiating elements for GPR applications was presented in [11], where the reconfiguration was performed by adopting the total geometry morphing approach [12], The total geometry morphing approach [12] represents the most structurally complicated but also the most versatile method exploited to achieve antenna reconfigurability. Such an approach is based on the interconnection of radiating elements very small in terms of radiated wavelength that singly radiate in a non efficient way. These sub-elements are interconnected by switches (PIN diodes) in order to resemble the geometry of the antenna to be synthesised. In this way, it is possible to change the antenna shape in a controlled way vs. the frequency. In particular, in [11], a bow-tie antenna was designed with the aid of the High Frequency Simulator Software (HFSS), produced by Ansoft [13]. The total geometric morphing allows to pursue the twofold goal of a constant radiation pattern through the whole reconfigurable band and a good matching at the same time, but of course presents also some drawbacks (e.g., the effects of the non-ideal switch connections between the "mosaic tiles" and the effects of the eddy currents excited on the detached elements due to the incident field radiated from the connected elements). For the designed reconfigurable bow-tie antenna, in [14, 15], realistic operative condition, typical of the GPR applications, were considered. In particular, in ref. [14], the designed antenna radiates in presence of a receiving antenna within a configuration in reflection mode, i.e., with the two antennas placed parallel to each other at the interface of a half space.

All the previous papers were concerned with a simplified model of the antenna where the problem of the feeding was not tackled. Here, we deal with the problem of feeding the designed bow tie antenna from a coaxial cable, and so the design of a balun system is needed too.

Therefore, the paper is organized as follow: In the next section, the reference monolithic antenna and the discretized (reconfigurable) version of it are briefly described. In Section 3 the exploited balun system is presented. In Section 4, a comparison in terms of the scattering parameter S_{11} (for different half-spaces) and radiation pattern (in free space) for the monolithic and the discretized antennas are shown. Conclusions follow in Section 5.

2. THE CONSIDERED RECONFIGURABLE ANTENNA

In this paper, we consider the reconfigurable antenna designed in [11, 14] and we briefly report some details necessary for the correct delivering of the paper.

The reconfigurable antenna design is based on the total geometry morphing approach and for the case at hand a rectangular grid with 1295 pads and 2141 switches has been exploited. The size of the sub-elements and interconnections has been designed to ensure a dynamic structure easily adaptable to different geometries. In particular, each sub-element is a square with side of 2.5 mm and is driven to the ON or to the /OFF state by means of a ideal circuital model resembling a typical PIN diode, whose resistance is $R = 0.985 * 10^{-3}$ Ohm in the ON state and $R = 4 * 10^{6}$ Ohm in the OFF state. PIN diodes sized $1 \times 0.5\,\mathrm{mm}^2$ have been exploited as interconnection elements.

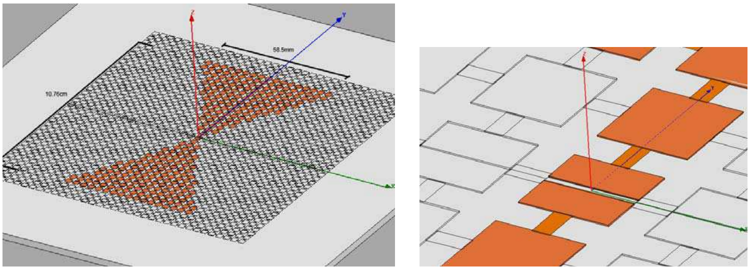
The reconfigurable antenna was designed "to synthesise" a reference monolithic bow-tie antenna that has a metallic element made by copper with a thickness of 36 μm; the metallic element is located on a layer of glass epoxy FR4 with extent 20 cm × 20 cm and 3 mm thickness. The reconfigurable antenna in "bow-tie" configuration is shown in Fig. 1(a); it has a flare angle of 30 degree. Fig. 1(b) depicts a zoom of the feeding point for the reconfigurable antenna.

Figure 2 depicts the corresponding "monolithic" bow tie antenna, whose details are reported in [11], for reference purpose with respect to the comparison to be shown in the next sections. Let us outline that, in principle, the physical differences between the two antennas are not negligible, because the detached metallic elements of the discretized bow tie enters in the field simulation too, as well as the fact that the (many) present switches are not ideal but account for realistic resistance values, as said. Finally, for simplicity, the small curvature of the end of the arms of the monolithic antenna has not be translated into the discretized reconfigurable structure, that ends with a flat bound.

3. RESULTS ABOUT THE DESIGNED BALUN

In this paper both the monolithic and the discretized antenna are fed by means of a coaxial cable, as customarily happens in GPR systems equipped with dipole-like antennas. The balancing of the currents is achieved by means of the linearly tapered balun depicted in Fig. 3, designed in planar technology [16] to achieve the balance of the currents in the range 300–1000 MHz (leading to a fractional bandwidth equal to 107.6%).

In particular, to ensure a gradual adjustment of the input impedance of the antenna and the coaxial in the range 300 MHz–1 GHz, the length of the balun was chosen equal to $L = 0.5\lambda_L$, being λ_L the maximum work wavelength in the structure. This choice has been done having in mind the design of an exponentially tapered transmission line (which is a structure different but with some similarities with the balun at hand). For a tapered transmission line connected to a load impedance Z_L, it is known that the impedance behaves as an exponential function described by $Z(z) = Z_0 e^{az}$, in the range $0 \leq z \leq L$, where Z_0 indicates the characteristic impedance of the line at $z = 0$ and $a = \frac{1}{L}\ln(\frac{Z_L}{Z_0})$ [17]. It is intrinsically supposed that the balun is matched with the impedance of the coaxial cable. The



(a) (b)

Figure 1. (a) Geometry of the reconfigurable antenna with pads sized 2.5 × 2.5mm × 0.036 mm, interconnections by PIN diodes sized 1 × 0.5 mm and dielectric layer made with 16 × 19 cm × 3 mm glass-epoxy FR4 Epoxy. The pads in "ON state" are highlighted in orange. (b) Zoom of the feeding zone.

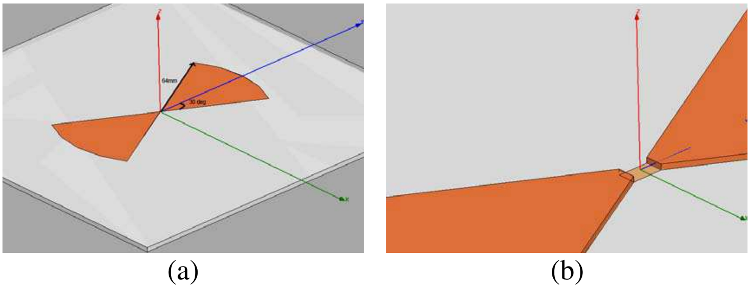


Figure 2. (a) Reference bow-tie antenna: thickness 36 mm, radius 64 mm and flared angle 30 degree, placed on substrate epoxy FR4 with extent 20 cm × 20 cm × 3 mm. (b) Zoom of the feeding zone.

reflection coefficient at distance L from the load is given by [17].

$$\Gamma = \frac{\ln Z_L/Z_0}{2} e^{-j\beta L} \frac{\sin \beta L}{\beta L} \tag{1}$$

where $\beta = (2\pi/\lambda)$ is the wave-number of the guided mode propagating in the balun. Eq. (1) is based on the theory of the small reflections, and so it is valid rigorously for small mismatching between load and tapered line.

The modulus of the reflection coefficient in (1) is depicted in Fig. 4 and behaves as a sinc function; accordingly, if we increase the electrical length of the taper so to guarantee $\beta L \geq \pi$ in correspondence of the minimum work frequency, this guarantees a low reflection all over the work band, in the meaning that the reflection coefficient is attenuated at least 13 dB compared to its value for a short length of the line. This justifies the choice $L = 0.5\lambda_L$, where λ_L is the maximum wavelength of the mode propagating along the tapered line.

In particular, we have made use of a substrate for the taper with relative dielectric permittivity $\varepsilon_r = 20$ in the work frequency range 0.3–1 GHz, so to make shorter the wavelength within the taper; this choice leads to a physical length L equal to 26.8 cm.

At this stage, however, we have still to match the taper to the coaxial cable, i.e., we have to synthesise a value Z_0 (the characteristic impedance of the line) close to the intrinsic impedance of the coaxial cable, which is supposed here equal to 50 ohms. This aim is achieved by designing suitably the widths of the arms of the taper. Let us note that the proposed balun is similar to a *micro-strip line* on the unbalanced side (see Fig. 3(a)) and to *a double strip line* at the gap of the bow-tie antenna (see Fig. 3(b)). Therefore, the size of the *microstrip line* on the unbalanced side (from the side of the coaxial cable) was set

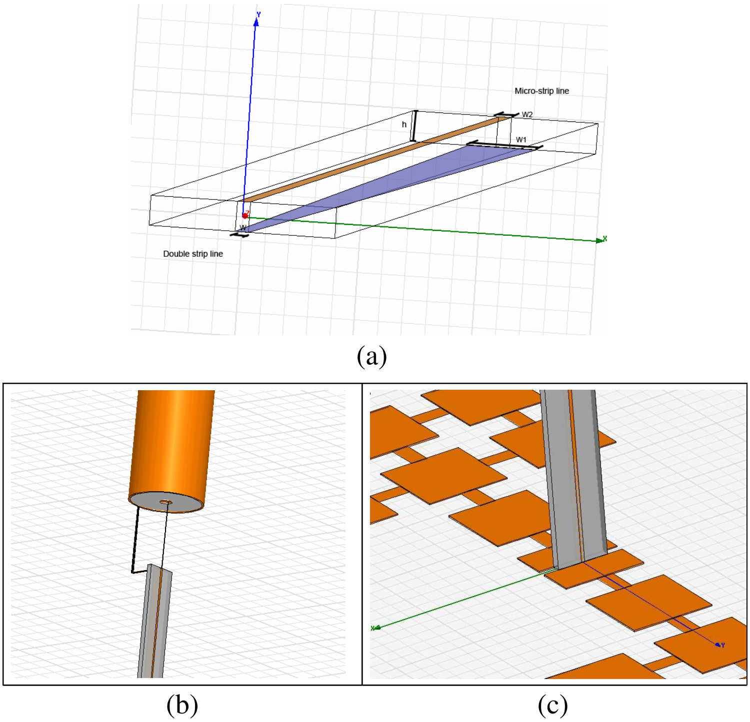


Figure 3. (a) Geometry of the linearly tapered balun. The double stripline is connected to the antenna terminals; the microstrip line is connected to the coaxial cable. (b) Detail of the connection coaxial cable-balun. (c) Detail of the connection antenna-balun.

according to the following formula [14]:

$$\frac{W_1}{h} = \left\{ \begin{array}{l} \frac{8e^A}{e^{2A}-2} per \frac{W_1}{h} \leq 2 \\ \frac{2}{\pi}\left\{B-1-\log_e(2B-1)+\frac{\varepsilon_r-1}{2\varepsilon_r}C\right\} per \frac{W_1}{h} \geq 2 \end{array} \right\}$$

where

$$\begin{aligned} A &= \frac{Z_c}{60}\sqrt{\frac{\varepsilon_r+1}{2}}+\frac{\varepsilon_r-1}{\varepsilon_r+1}\left(0.23+\frac{0.11}{\varepsilon_r}\right) \\ B &= \frac{377\pi}{2Z_c\sqrt{\varepsilon_r}} \\ C &= \log_e(B-1)+0.39-\frac{0.61}{\varepsilon_r} \end{aligned} \tag{2}$$

where W_1 is the width of the narrower conductor (the larger one does not affect the impedance if it is sufficiently larger than W_1), h is the transverse distance between the two conductors of the microstrip line and Z_c is the impedance of the coaxial cable (equal to 50 ohms as said above). The distance h has been set to 2.54 cm (it is the thickness of the gap of the antenna), so W_1 has been chosen equal to 1.15 cm according with Eq. (2). Finally W_2 has been chosen equal to $5W_1$ thus not affecting the validity of the formula in (2).

In order to test the designed balun, first the reflection coefficient at the unbalanced end has been simulated, in the range 0.3–1 GHz. The result is reported in Fig. 5 and shows that the matching with the coaxial cable is very good all over the band 0.3–1 GHz and even beyond.

Figure 6 shows instead the modulus of the reflection coefficient at the end of the coaxial cable looking towards the antenna. More precisely, the antenna is replaced by several test loads, in order to check that the balun does not introduce (by itself) strong reflections back into the coaxial cable. From this figure, it can be seen that the balun does not introduce strong reflections by itself, even if it is not able to match very well impedances strongly different from that of the coaxial cable.

Of course Figs. 5 and 6 do not show the correct working of the structure with regard to the balance of the currents between the arms. In the next section, however, the comparison between the radiation pattern of the monolithic and of the discretized structure shall implicitly show the good quality of the balancing all over the considered frequency band, inferred from the symmetric shape of the radiation patterns.

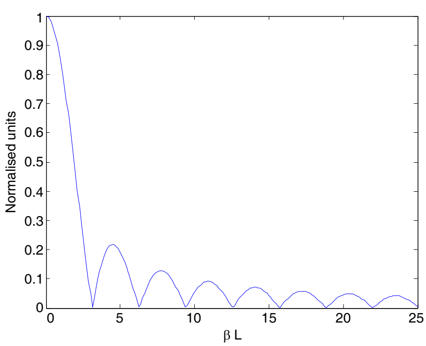


Figure 4. Modulus of the reflection coefficient vs. βL according to Eq. (1).

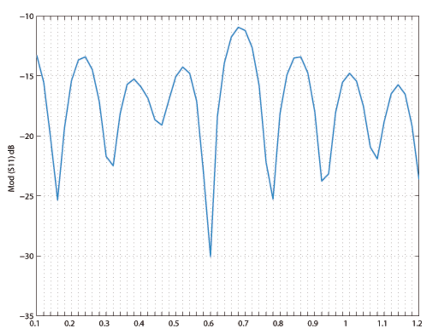


Figure 5. Frequency response of the reflection coefficient at the feed unbalanced port of the balun.

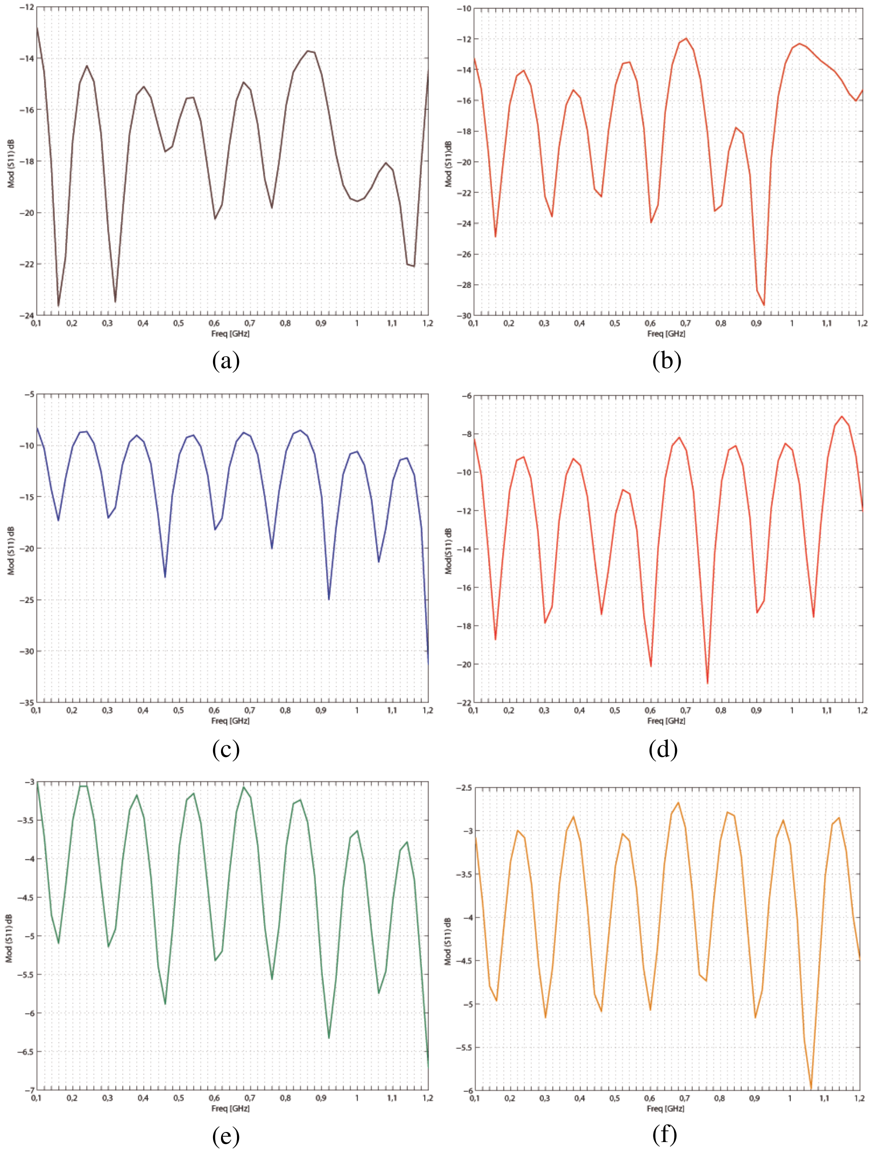


Figure 6. Frequency response of the Modulus of S_{11} at the feed port of the balun, when balun's balanced load impedance is respectively: (a) 50 Ohms. (b) $50 + j50$ Ohms. (c) 75 Ohms. (d) $75 + j75$ Ohms. (e) 200 Ohms. (f) $200 + j200$ Ohms.

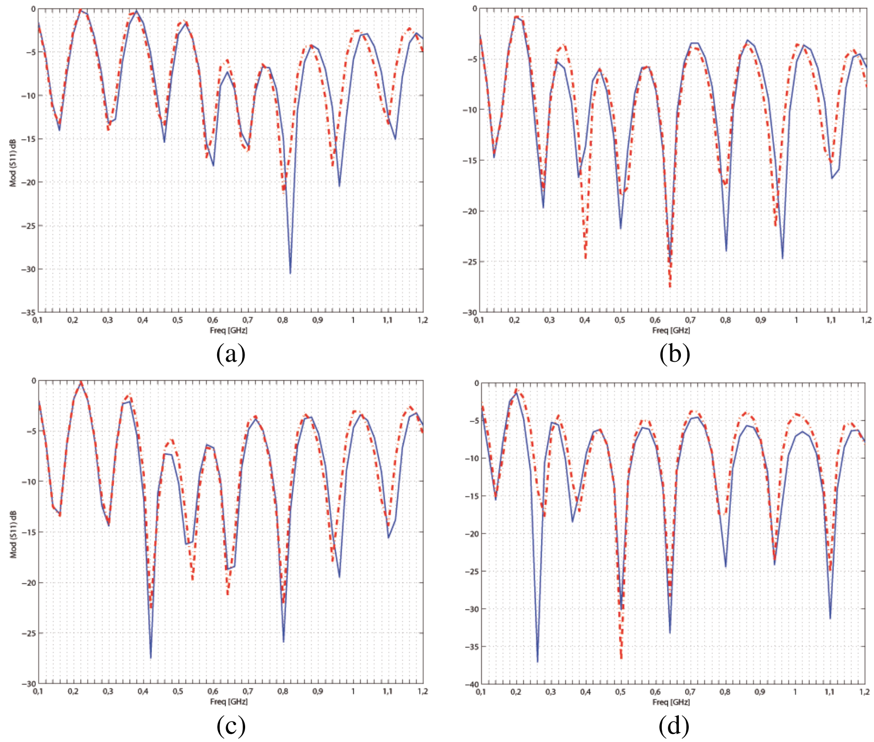


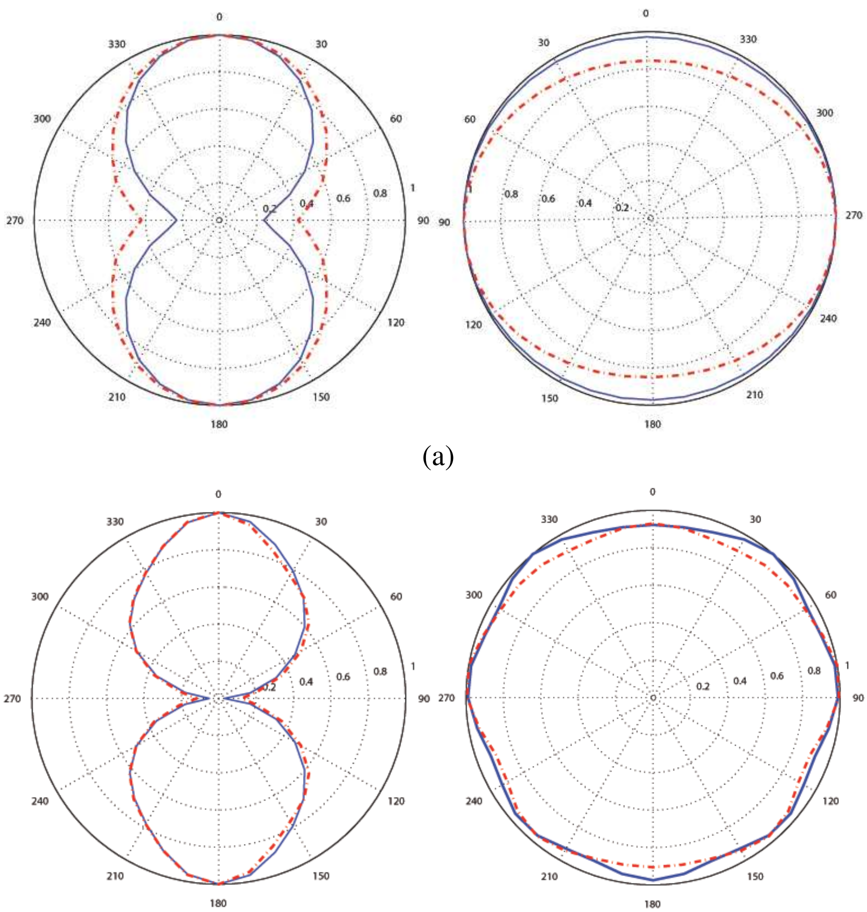
Figure 7. Frequency response of the Modulus of S_{11} at the feeded port of balun (fed in unbalanced way by a coaxial cable). Blu line: Monolithic antenna; Red line: Discretised reconfigurable antenna. (a) Antenna in free-space. (b) Antenna on half space with relative dielectric permittivity 4. (c) Antenna on half-space with relative dielectric permittivity 9. (d) Antenna on half-space with relative dielectric permittivity 16.

4. RESULTS ABOUT THE RECONFIGURABLE ANTENNA

In this section some numerical results about both the discretized and the monolithic antennas are shown, both with regard to the impedance matching on a half space and the radiation pattern in free space. The simulations have been performed making use of the HFSS code.

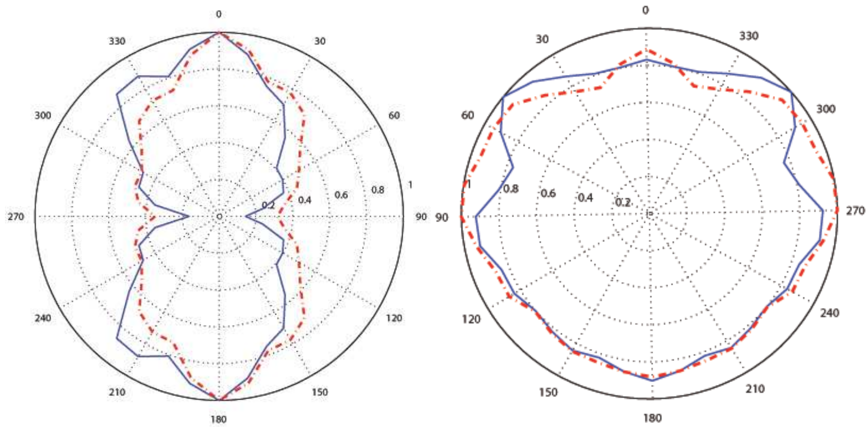
Figure 7 depicts the modulus of the reflection coefficient at the input port of the balun, both for the discretized and the monolithic antennas. As can be seen, the behaviour of the discretized antenna is quite similar to that of the monolithic antenna. Noticeably, moreover, the minima of all the reflection coefficients occurs about at the same frequencies, which makes it possible to think of a relatively robust (with respect to uncertainties about the permittivity of the soil) equalization procedure.

Figure 8 shows the radiation pattern in free space of both the monolithic and the discretised antennas. From Fig. 8, it can be appreciated that the two patterns are in good agreement at 300 and 650 MHz. Some discrepancy arises at 1 GHz, where also some asymmetric behaviour of the pattern (especially for the monolithic antenna) can be seen. This is probably due to some current induced by the field radiated by the arms on the external part of the coaxial cable. In fact, the upper and lower lobe of the pattern on the E-plane remains



(a)

(b)



(c)

Figure 8. Radiation patterns in E-plane and H-plane normalized to $E_{\max}$ in free space for the monolithic antenna (blue line) and the discrete antenna (Red line). The feeding of the antenna comprises the balun. (a) 300 MHz. (b) 650 MHz. (c) 1 GHz.

substantially equal to each other, and symmetrical with respect to the 270–90 degree axis, even if they are bent toward the direction of the coaxial cable. The pattern at 300 and 650 MHz, instead, show the typical shape of the pattern of a dipole like antennas. This is an indirect proofs of the fact that the balance of the currents provided by the balun is satisfying all over the band 300–1000 MHz.

5. CONCLUSIONS

In this paper the discrete version of a bow tie antenna has been proposed and compared with the corresponding monolithic antenna. It has been shown that the discrete geometry provides results not much different from the corresponding monolithic one. As a drawback, the smallness of the elements of the discretised antenna makes it somehow more complicated than the monolithic one, but it is also important with regard to the mitigation of the spurious effects of the currents excited on the detached elements of the discrete geometry.

This is a preparatory (and in progress) work in view of the possibility to implement a reconfigurable antenna for GPR prospecting, where the shape and the size of the discrete active elements can be changed vs. the frequency in a controlled way by programming the combined aperture of the switches. In this way one can extend the equivalent work frequency band antenna both in terms of impedance matching and in terms of radiation pattern. In fact, based on the preliminary results shown here, we can expect that a suitable variation vs. the frequency of the configuration of the active elements can improve the matching meaningfully and can make the radiation pattern less variable vs. the frequency.

As future research activity, we plan to implement the antenna solution and perform realistic measurements to verify the effectiveness of the proposed solution. For the realistic implementation of the proposed solution it is necessary to design the DC network feeding the PIN diodes and the cavity for the back lobe suppression.

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