Bow-Tie Microstrip Antenna Analysis and Design Using the FDTD Method

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Abstract-This paper presents the design simulation and measured performance of a coplanar-fed bow-tie antenna for monolithic wireless communication applications. 3D-FDTD method is used in the design simulation of this planar antenna structure. Numerical and measured results of the antenna radiation characteristics, including return loss and input impedance are presented and compared. A wideband quarter wave balun is used for impedance transformation of the unbalanced coaxial cable feed line to a balanced coplanar microstrip line. The used substrate is FR4 with ε_r=4.8 and thickness of 1.6 mm. Also some investigations are made on the shape of two poles in order to get different characteristics for antenna.

1. Introduction

In order to gain monolithic transmit /receive modules; antennas should be mounted on a microstrip board that contains MMIC structures. Furthermore, it is difficult to convert a microstrip line on a board to a coaxial cable or waveguide that feed the antenna. Although planar antennas can resolve these requirements, their bandwidths are limited. Many wideband microstrip antennas have been analyzed yet [1-2]. Recently similar printed dipole [3] and bow-tie [4] antennas have been introduced. Regardless the good results for bandwidth and radiation pattern have reported it is desire to analyze microstrip structures. Analyzed bow-tie structure which is mounted on a grounded substrate is shown in Figure1.

Using the FDTD method gives us the possibility of changing the angle of θ and compare characteristics of dipole (θ = 0) and bow-tie microstrip antennas. Other shapes like fractal triangular shapes can also be investigated easily. Also a wideband frequency response can be obtained by this full-wave analysis method. In order to reduce the number of iterations required in FDTD method, the approach proposed by Sheen et al. [5] has been used in this paper. Switchable Mur's first order ABC that surrounds the simulation space has an acceptable accuracy for this type of planar structures. Numerical and measured results of return loss and input impedance are presented and compared. Accuracy of programs has been investigated by simulating the patch antenna introduced in [5]. Measured results of antenna pattern are also presented for horizontal and vertical polarizations. Two types of fractal patterns have been experienced for triangular poles of bow-tie following Romeu [6]. Input characteristics of these structures are obtained numerically and multiband characteristics obtained for the antenna.

2. FDTD Formulation of Microstrip Antenna

FDTD method is a straightforward implementation of discretization procedure to Maxwell equations that can be written as:

$$\nabla \times \vec{E} = -\mu \frac{\partial \vec{H}}{\partial t} \quad (1)$$

$$\nabla \times \vec{H} = \sigma \vec{E} + \varepsilon \frac{\partial \vec{E}}{\partial t} \quad (2)$$

These equations should be satisfied in whole media around the antenna. Generally speaking, we can recognize three distinct materials in microstrip structures; air or free space, conductors of ground plane and top of the microstrip and at last dielectric substrate. We can use a number of identification pointers to designate which material is in a given (i,j,k) location [7]. Further cares should be taken in interfaces between two adjacent media. It can be shown that the interface of free space and dielectric substrate has a dielectric and conductive constants which are average of corresponding constants of the two media [8]. That is:

$$\varepsilon_{interface} = \frac{\varepsilon_{air} + \varepsilon_{substrate}}{2} \quad (3)$$

$$\sigma_{interface} = \frac{\sigma_{air} + \sigma_{substrate}}{2} \quad (4)$$

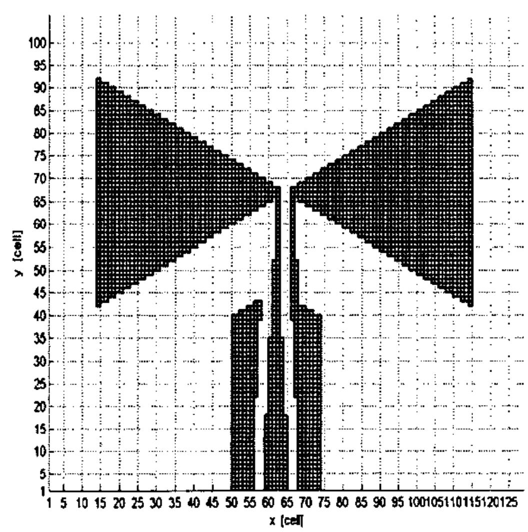


Figure1- Analyzed bow-tie antenna shape

Using central difference approximation scheme and modified Yee cell shown in Figure 2 we can get following FDTD equations which are corresponding approximations of (1) and (2), as in [9]:

$$H_x^n(i,j,k)=H_x^{n-1}(i,j,k)+\frac{\Delta t}{\mu}\left[\frac{E_y^{n-1}(i,j,k+1)-E_y^{n-1}(i,j,k)}{\Delta z}-\frac{E_z^{n-1}(i,j+1,k)-E_z^{n-1}(i,j,k)}{\Delta y}\right] \quad (5\text{-}1)$$

$$H_y^n(i,j,k)=H_y^{n-1}(i,j,k)+\frac{\Delta t}{\mu}\left[\frac{E_z^{n-1}(i+1,j,k)-E_z^{n-1}(i,j,k)}{\Delta x}-\frac{E_x^{n-1}(i,j,k+1)-E_x^{n-1}(i,j,k)}{\Delta z}\right] \quad (5\text{-}2)$$

$$H_z^n(i,j,k)=H_z^{n-1}(i,j,k)+\frac{\Delta t}{\mu}\left[\frac{E_x^{n-1}(i,j+1,k)-E_x^{n-1}(i,j,k)}{\Delta y}-\frac{E_y^{n-1}(i+1,j,k)-E_y^{n-1}(i,j,k)}{\Delta x}\right] \quad (5\text{-}3)$$

$$E_x^n(i,j,k)=C_a(M)E_x^{n-1}(i,j,k)+\frac{\frac{\Delta t}{\varepsilon\Delta y}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_z^{n-1}(i,j,k)-H_z^{n-1}(i,j-1,k)\right]-\frac{\frac{\Delta t}{\varepsilon\Delta z}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_y^{n-1}(i,j,k)-H_y^{n-1}(i,j,k-1)\right] \quad (5\text{-}4)$$

$$E_y^n(i,j,k)=C_a(M)E_y^{n-1}(i,j,k)+\frac{\frac{\Delta t}{\varepsilon\Delta z}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_x^{n-1}(i,j,k)-H_x^{n-1}(i,j,k-1)\right]-\frac{\frac{\Delta t}{\varepsilon\Delta x}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_z^{n-1}(i,j,k)-H_z^{n-1}(i-1,j,k)\right] \quad (5\text{-}5)$$

$$E_z^n(i,j,k)=C_a(M)E_z^{n-1}(i,j,k)+\frac{\frac{\Delta t}{\varepsilon\Delta x}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_y^{n-1}(i,j,k)-H_y^{n-1}(i-1,j,k)\right]-\frac{\frac{\Delta t}{\varepsilon\Delta y}}{1+\frac{\sigma\Delta t}{2\varepsilon}}\left[H_x^{n-1}(i,j,k)-H_x^{n-1}(i,j-1,k)\right] \quad (5\text{-}6)$$

So that the constant of $C_a(M)$ for each media that is indicated by M can be written as:

$$C_a(M)=\frac{1-\frac{\sigma\Delta t}{2\varepsilon}}{1+\frac{\sigma\Delta t}{2\varepsilon}} \quad (6)$$

Constant of M is defined by identification pointers before beginning of the main field refreshment procedure.

These formulas are used in iteration loops to renew field components in entire computational domain. Tangential electric fields on zero thickness and perfect conductors are assumed to be zero separately in main timing loop. This treatment can be made by assuming a large value for conducting constant of conductors when we introduce material types for the first time.

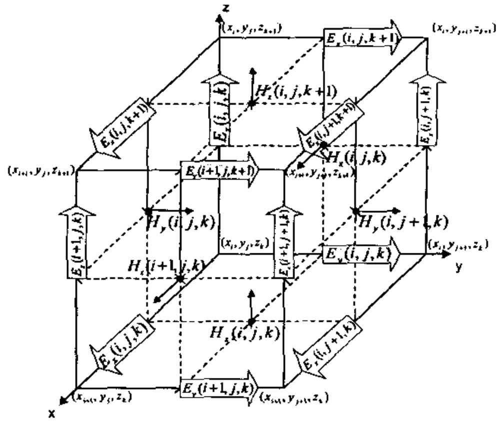


Figure2- Modified Yee cell

Bottom layer of the substrate is a ground plane and the other five walls are truncated by Mur's first order absorbing boundary conditions [10-11]. This ABC satisfies stability of computations in planar problems that the incident wave to the walls is normal. Using Mur's second order ABC or combination of these two boundary conditions causes unstable criteria.

3. Source Implementation

Excitation conditions of coplanar-fed microstrip structure are shown in Figure 3. This structure has three terminals. E_z component of electric field underneath these three terminals is assumed to be Gaussian pulse and treated as a hard source. The time relation of source is as follows:

$$E_z(t)=\exp\left[-\frac{(t-t_o)^2}{T^2}\right] \quad (7)$$

In order to reduce the source distortion and the number of iterations those are intrinsic features of FDTD method, a changeable source wall is assumed following Sheen *et al*[5]. This wall is a magnetic wall all over the source plane and when the excitation pulse vanishes to zero is substituted by Mur's first order ABC. This approach causes computations get converged in time steps less than 4500.

Meshing characteristics and structural parameters of the antenna are given in table1.

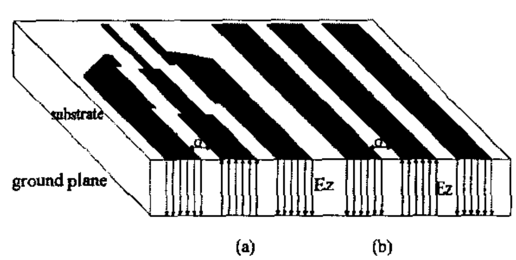


Figure3- Excitation scheme of coplanar microstrip;
a) matching network,
b) coplanar microstrip line

Table1. Structural parameters and characteristics

Dielectric constant	$\varepsilon_r = 4.8$
Substrate thickness	H = 1.6 mm
Length of the antenna	L = 51.5 mm
Gap length of terminals	G = 1.5 mm
Cells dimension	$\Delta x = 0.5, \Delta y = 1, \Delta z = 0.4$ [mm]
Number of cells	$N_x = 135, N_y = 105$ $N_z = 16$
Number of iterations	N = 8192
Time step size	$\Delta t = 0.944$ ps
Gaussian pulse parameters	$T = 32*\Delta t$, $t_0 = 4*t$
Source wall switch time	N=260

4. Frequency Related Return Loss and Input Impedance

Computation of return loss is done according to the following relation:

$$S_{11}(f) = \frac{E_{ref_1}(f)}{E_{inc_1}(f)} \tag{8}$$

It shows that we should have incident and reflected waves in the whole domain separately. When we use port excitation that mentioned earlier there will be a combination of these waves. In order to separate incident wave from reflected wave, in the first step the simulation is done on the three microstrip coupled lines as shown in Figure3-b. Electric field components on the surface of the top layer are stored for further computations. These fields can be supposed approximately as incident waves. Then antenna structure including matching and balun network is analyzed. The fields that are obtained in this part are combination of incident and reflected fields. We can get reflected field by subtracting the incident field in previous step from the total field in this one. Since excitation of the antenna is done by means of E_z component of the electric field, all the computations are just done on this component. After finding S_{11} we can get input impedance by using the following relation:

$$Z_{in} = Z_0 \frac{1 + S_{11}e^{j2kl}}{1 - S_{11}e^{j2kl}} \tag{9}$$

Observation point for calculating S_{11} parameter should be at least 10 cells off the first discontinuity in the microstrip feed line. Input impedance of the antenna is numerically obtained and shown in Figure 4.

5. Design of the Wideband Balun

A coplanar quarter wave impedance transformer is used to transform antenna input impedance to 50Ω coaxial cable feed line. This network should also act as a wideband balun. Some bondings that are used in the place of steps can perform the action of transforming balance network to unbalance. To design this network we should know the input impedance of the antenna in the center frequency of interest. We used a 50Ω coupled microstrip line long enough to feed the antenna and then simulated the structure and found the input impedance. An input impedance of 120Ω is obtained and then design of the matching network was done according to the works of Matthaei [12] and Rizzi [13]. In order to get a maximally flat characteristic, binomial transformer formulation is used in a four step configuration. Since uniform meshing is used in simulations, some differences will exist in widths of the lines calculated and widths that are used in simulation. This network with a resistive load of 120Ω is also simulated separately using Microsoft Office™ software.

6. Results

Design of the antenna has performed according to design and performance considerations presented for dipole antennas in [14] and numerical results obtained by FDTD method. Measurements on input characteristics of the antenna have been performed using HP8714E RF Network Analyzer in frequency range of 2 GHz to 3 GHz. Numerical and measured results of S_{11} parameter is shown in Figure5 and Figure6 respectively.

Since cell quantization of the matching network force some restrictions on actual size of the lines in the simulation, and also the same diameters are used in implemented antenna, better results of bandwidth obtained numerically. Numerical bandwidth is 17% that is roughly in good agreement with measured result when wire bonding is used. Using ground plane too close to the structure is caused impedance mismatch in the half band of the frequency of interest. Omitting the ground plane underneath the bow-tie shape exceeds the bandwidth up to 25%. Using none uniform meshing techniques results accurate models for matching network. None uniform meshing is also useful in modeling of bondings since their width are very thinner than cells size that is used in the simulation.

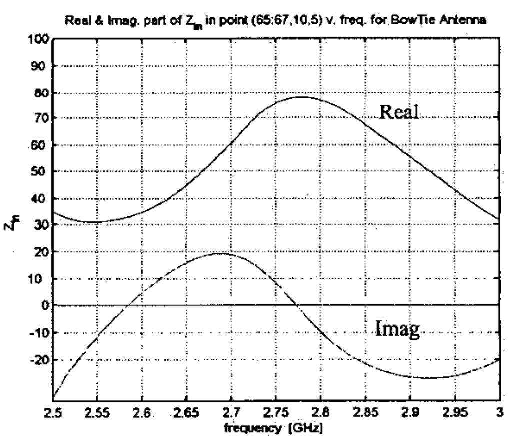


Figure4- Real and imaginary part of input impedance obtained numerically

Antenna is implemented on FR-4 substrate with a dimension of 100mm×110mm. Then ground plane can not be considered as an infinite plane. In simulations, ground plane is continued to Mur's ABC and constructed a model of infinite ground plane. These are some of error sources that cause a less bandwidth comparing to printed bow-tie antennas.

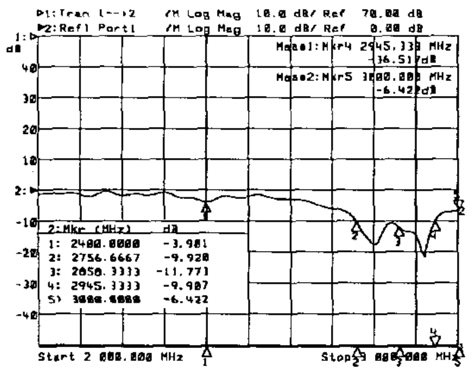


Figure5- Numerical result obtained for S_{11} parameter

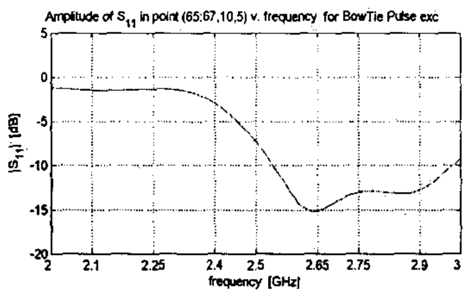
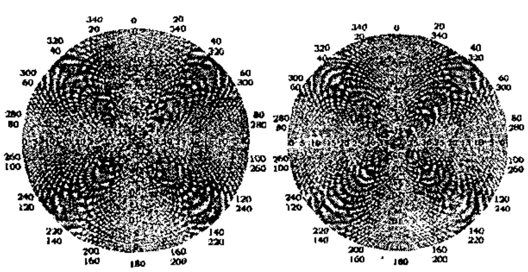


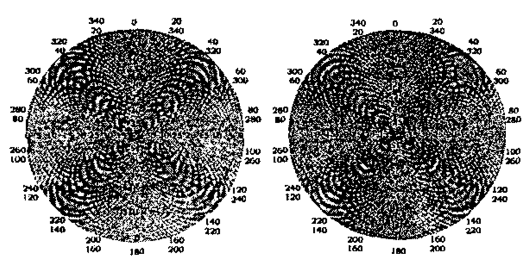
Figure6- Measurement result obtained for S_{11} parameter

Radiation pattern of the antenna is measured for vertical and horizontal polarizations and is shown in Figures 7 and 8. Measurements are done in the vicinity of two edge frequencies of Figure5.



(a) (b)

Figure7- Radiation pattern in frequency of 2.7 GHz
a) Horizontal polarization, b) Vertical polarization



(a) (b)

Figure8- Radiation pattern in frequency of 2.9 GHz
a) Horizontal polarization, b) Vertical polarization

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