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Citation for published version (APA):

DOI:
10.1049/el:20000638

Document status and date:
Published: 01/01/2000

Document Version:
Publisher’s PDF, also known as Version of Record (includes final page, issue and volume numbers)

Please check the document version of this publication:

• A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher's website.
• The final author version and the galley proof are versions of the publication after peer review.
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Modelling of bow-tie microstrip antennas using modified locally conformal FDTD method

J. George

An introduction to bow-tie microstrip antennas has recently attracted a large amount of interest [1 - 4]. In present day communications scenarios, bow-tie microstrip antennas are attractive candidates owing to their compact nature [1] compared to rectangular microstrip antennas. However, only very few attempts have been made to analyse these antennas [5, 6]. The approach, also based on the finite-difference time-domain (FDTD) method, can be used to perform such an analysis, but it requires a variation in the cell size with the central width (w) of the antenna to ensure that the slanting metallic edges of the patch pass exactly through the diagonals of the cell faces. This requirement in turn requires an increased number of cells, both along the antenna length (L) and width (b), than the corresponding rectangular patch antenna of the same length and width. The method is capable of performing the analysis only for bow-tie configurations that can be divided into segments, the Green functions of which are known.

In this Letter, the mentioned inconveniences in the analysis of bow-tie microstrip antennas are overcome by using the modified locally conformal finite-difference time-domain algorithm [7]. The algorithm takes into account the slanted metallic edges of the patch by using slightly modified field update equations, alone or in conjunction with a backward weighted averaging scheme depending on the degree of slanting of the cell face. This method provides the user with the flexibility to determine the number of cells required in different directions independently of w. Simulation results have been compared with the corresponding experimental results to demonstrate the effectiveness of this approach.

Antenna design details and summary of algorithm: The general geometry of the bow-tie antenna and the relevant dimensions are shown in Fig. 1. The figure also includes a hypothetical section of the grid through which the slanting edge passes which is also shown in the figure. All the antennas were fabricated on a substrate of dielectric constant $\varepsilon_r = 4.4$ and thickness $h = 0.16$cm. The feed point (probe feed) was specified in terms of the distance $f_2$ from the radiating edge along the central line as shown in the figure.

The algorithm given in [7] is re-arranged and slightly modified as follows. To begin with, the computational volume is divided into a collection of Yee cubes and then the metal free face areas $(A_j)$ and segment lengths $(l_j)$ at the antenna plane are computed for those cells through which the slanting edges pass. Now, for all the fields associated with them, flag matrices are set up for direct use in the field update equations [7, 8]. For instance, the re-arranged H-field update equation along the z-direction with the corresponding flag matrix $(F_{hz}(z,j,k))$ is given by:

$$
H_{z}^{n+1/2}(i,j,k) = H_{z}^{n-1/2}(i,j,k) + \Delta t \sum_{l} \left\{ E_{z}^{n}(i,j,k)\mu_{A_{z}}(i,j,k) - E_{z}^{n}(i,j,k-1)\mu_{A_{z}}(i,j,k-1) \right\}
$$

where

$$
F_{hz}(i,j,k) = 2 \text{ if } A_{z} < 0.075 \text{ or } \frac{\mu_{A_{z}}}{A_{z}} > 12
$$

$A_{z}$ represents the distorted cell face area normalised with respect to the undistorted cell face area and $\mu_{A_{z}}$ denotes the square of the maximum side length of the distorted cell face. Use of the flag matrix, which is generated in the meshing routine as in [7], directly in the field update equation makes this implementation faster than the method suggested in [7]. This is because the implementation of the update equation itself does both necessary decision making and backward weighted averaging. The other relevant E and H-field components are also updated in a similar manner.

Numerical and experimental results: The bow-tie antenna configurations used in this investigation were categorised depending on whether the length to width ratio $(L/b)$ is equal to, greater than, or less than one. In each category, three different cases of central width $(w = 0.35, 0.58$ and $0.70$) were considered.

The first category considered $(L/b = 1)$ was with $L = 3.5$, and $b = 3.5$cm. The cell dimensions along the x- and y-directions were chosen in such a way that an even number of them exactly matched with ‘L’ and ‘b’ as shown in Fig. 1. The cell dimension along the z-direction was selected so as to make an integral number of cells fill the substrate thickness. The number of cells used along the x-, y-, and z-directions for the three cases of the central width were $n_x = n_y = 40$, and $n_z = 25$, respectively, and the corresponding cell dimensions used were $\Delta_x = \Delta_y = 0.0875$, and $\Delta_z = 0.04$cm. In all the cases investigated in this Letter, excitation was carried out by using a z-directed Gaussian pulse of source resistance $50\Omega$ [9] located just above the ground plane. The number of time steps used in all the cases was 16384 and the size of time step chosen in each case was $75\%$ of the respective Courant limit. Fig. 2r shows the simulated and experimental variations of return.
loss with frequency of the antenna with \( L/b = 1 \) for the three central widths \((w = 0.36, 0.5b \) and 0.7b\) when the experimental feed points were at \( f_3 = 1.5, 1.2, \) and 1.2cm, respectively. The theoretical and experimental TM\(_{30}\) mode resonant frequencies of the three antennas were found to be 1602, 1779, and 1931MHz and 1628, 1784, and 1922MHz, respectively, and the corresponding prediction errors 1.6, 0.28, and 0.47%. From the Figure and the observations it is clear that the algorithm performs excellently in predicting the resonant frequencies of the antennas.

![Graph](image)

**Fig. 2** Simulated and measured return loss variations with frequency for antenna configurations with three central widths

- **(a)** \( L = 3.5 \) and \( b = 3.5 \)cm
- **(b)** \( L = 3 \) and \( b = 2 \)cm
- **(c)** \( L = 3.5 \) and \( b = 7 \)cm

**Table 1:** Different parameters and characteristics of some of antennas used in investigation

| \( L/b \) | \( w \) | \( f_3 \) | \( \Delta_{\alpha}, \Delta_{\beta}, \Delta_{\gamma} \) | \( n_1, n_2, n_3 \) | TM\(_{30}\) mode resonant frequency [MHz] | Error
<table>
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<td>0.36</td>
<td>1.7</td>
<td>0.1, 0.1, 0.04</td>
<td>60, 40, 25</td>
<td>1528</td>
<td>1513</td>
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<td>1643</td>
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<td>1759</td>
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<tr>
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<td>0.0875, 0.175, 0.044</td>
<td>60, 40, 25</td>
<td>1478</td>
<td>1465</td>
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<tr>
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<td>0.6</td>
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<td>60, 40, 25</td>
<td>1631</td>
<td>1617</td>
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<tr>
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<td>0.5</td>
<td>0.0875, 0.175, 0.044</td>
<td>60, 40, 25</td>
<td>1837</td>
<td>1825</td>
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Investigations were then carried out for the other two categories \((L/b > 1 \) and \( L/b < 1 \)) of bow-tie antenna. Fig. 2b and c show the simulated and experimental variations of return loss with frequency for \( L/b = 2 \) and \( L/b = 0.5 \), respectively. Table 1 gives the different parameters and characteristics of these antennas. Table 1 reaffirms the capability of the algorithm for analysing bow-tie microstrip antennas. Also from Table 1 it is clear that the approach does not require a variation in cell dimensions with antenna central width. This, in turn, indicates the potential flexibility and convenience of the method for using the same number of cells as that of the corresponding rectangular microstrip antenna \((w = 1.0)\) for all possible central widths.

**Conclusions:** The suitability of the modified locally conformal FDTD method for the analysis of bow-tie microstrip antennas has been investigated. All three categories of bow-tie microstrip antennas were considered for investigation and good agreement was observed between the experimental and theoretical results. The present approach enables us to use the same cell dimensions for all derived bow-tie antenna configurations from a given length \((L)\) and width \((b)\).

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**Acknowledgement:** The author wishes to acknowledge the help and encouragement received from P. Mohanan, Dept. of Electronics, Cochin University of Science and Technology, Kerala, India, during the experimental study.

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Electronics Letters Online No. 20000638
DOI: 10.1049/el:20000638

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**References**


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**New class of multibit sigma-delta modulators using multirate architecture**

F. Colodro, A. Torralba, F. Muñoz and L.G. Franquelo

A new multibit sigma-delta modulator is presented where the analogue-to-digital converter in the forward path is replaced by an increase in the clock rate of the integrators in the final stages. Theoretical and simulation results are presented for second- and third-order modulators.

**Introduction:** For high performance converters there is a tradeoff between oversampling ratio \((M)\) and modulator order \((L)\) [1]. High values of \( M \) require high speed integrators, while high values of \( L \) produce instability. In both cases, the power consumption increases. Multibit architectures have been recently used to solve this tradeoff. Multibit modulators offer a direct improvement over one bit topologies of \( 6.n \) dB, where \( n \) is the number of bits. The most important drawback is the high accuracy requirement for the multibit digital-to-analogue converter (DAC) in the feedback path. Several attempts to ameliorate this problem have been made in the past by modifying the classical topology [2], using digital correction [3] or dynamic element matching [4]. Nevertheless, the internal multibit analogue-to-digital converter (ADC) (a flash architecture) is usually ignored, although it makes a non-negligible contribution to the total area and power consumption.

In this Letter a new multibit sigma-delta ADC architecture, based on a multirate-multibit sigma-delta (MM-SD) modulator is presented where the analogue-to-digital converter in the forward path is replaced by an increase in the clock rate of the integrators in the final stages. Theoretical and simulation results are presented for second- and third-order modulators.