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PAPER

Modal-Expansion Analysis of Electromagnetically Coupled Coaxial Dipole Antennas

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SUMMARY This paper presents a modal-expansion analysis of the electromagnetically coupled coaxial dipole antenna. The analysis of the antenna problem is initially simplified using the even-odd mode excitation and then the resultant half structure is divided into two parts; one is the characterization of a coaxial feeding network and the other is the modeling of a sleeve monopole antenna driven by a coaxial line. The formally exact modal-expansion method is employed to analyze both parts. The analysis of the sleeve monopole antenna is facilitated by introducing a perfectly conducting boundary at a distance from the monopole's top end. The current distribution and input impedance of the electromagnetically coupled coaxial dipole antenna are obtained by finding expansion coefficients through enforcing the continuity of tangential field components across regional interfaces and cascading the two parts together. Numerical results for the coaxial dipole antenna's radiation characteristics are presented and discussed.

key words: *antennas, dipole antenna, electromagnetically coupled coaxial dipole, modal-expansion method*

1. Introduction

The dipole antenna has received extensive attention over the past decades because of its structural simplicity and desirable omni-directional radiation pattern in the horizontal plane [1]. Recently, a new type of collinear antenna was proposed by Miyashita et al. [2] and it has the electromagnetically coupled coaxial dipole as the radiating element fed by an annular ring slot cut on the outer conductor of the feeding coaxial cable. A Wiener-Hopf analysis was presented in [2] to characterize the proposed collinear antenna. Although the analysis is exact, it is rather complicated and cannot take the effect of conductors' finite thicknesses into account. Rao, Hashiguchi and Fukao [3] described an efficient numerical method to analyze the far-field of an electromagnetically coupled coaxial dipole element and applied the numerical technique for array applications. A similar problem of radiation from a slot cut on the cylindrical surfaces or the outer conductor of a coaxial cable was extensively studied by many researchers, such as Chang [4], Kiang [5], [6], and Eom et al. [7].

This paper applies the rigorous modal-expansion method [8]–[10] to the analysis of electromagnetically coupled coaxial dipole antennas. First, the structure of the antenna problem is reduced to half by the use of the even-odd mode excitation. Second, the resultant half structure is then divided into two parts; one is the coaxial feeding network and the other is a sleeve monopole antenna driven by a coaxial line. Both parts are characterized by the formally exact modal-expansion method. For the modeling of the sleeve monopole antenna, a perfectly conducting boundary at the top end of the monopole is introduced in order to the confine the domain of interest and in turn to make possible the field expansion in the open region into the summation of discrete modes. The introduction of a conducting wall in the top does not have much effect on the antenna characterization since its radiation is null in the axial direction of the antenna. After that, the resultant structure is subsequently divided into a number of regions and the modal-expansion method is employed to determine the electromagnetic fields in all the regions of interest. Finally, the radiation performance parameters of the electromagnetically coupled coaxial dipole antenna can be obtained from the computed field components and scattering matrices.

The paper is organized as follows. In Sect. 2, the model we employ to characterize the electromagnetically coupled coaxial dipole antenna is first described. The structures involved in the model consist of the coaxial feeding network and a sleeve monopole antenna driven by a coaxial line. The modal-expansion method for obtaining the scattering parameters of the coaxial feeding network is then presented in Sect. 3. Section 4 provides the brief formulation for the modal-expansion analysis of the sleeve monopole antenna because the detailed formulation was published in [9]. Numerical results are shown and compared to available experimental data in Sect. 5. Concluding remarks are given in Sect. 6.

2. The Analysis Model

The geometry of an electromagnetically coupled dipole antenna fed through an annular ring slot by a coaxial transmission line is shown in Fig. 1. Electromagnetic energy is coupled to the outer conducting pipe, which is the main radiation source, from the coaxial feed line through an annular ring slot cut on its outer conductor. In the array application, a small portion of the energy is radiated by one element and the remaining will be transmitted to subsequent elements through the coaxial feed line. In this paper, we

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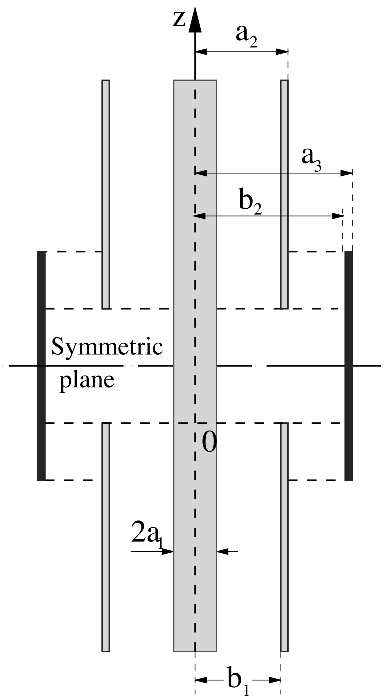


Fig. 1 Geometry of the electromagnetically coupled dipole antenna.

only deal with the modal-expansion analysis of a single element. Characterization of mutual coupling in the array configuration by the modal-expansion method will be reported in a forthcoming paper.

Due to its structural symmetry, we can consider only half of the structure by invoking the even and odd mode excitations at the coaxial ports. The symmetrical plane can either be an electric wall for an odd mode excitation or a magnetic wall for an even mode excitation. Since the formulations for both excitations are very similar, we hereby only detail the formulation with an electric wall at the symmetrical plane. Once the reflection coefficients for both excitations are obtained, we can then calculate the scattering parameters of the two-port feeding network:

$$S_{11} = S_{22} = \frac{1}{2}(\Gamma_e + \Gamma_o) \tag{1a}$$

$$S_{21} = S_{12} = \frac{1}{2}(\Gamma_e - \Gamma_o) \tag{1b}$$

where Γ_e and Γ_o are the reflection coefficients in the coaxial feed line of the half structure for even and odd mode excitations, respectively.

For the analysis of the half structure with an electric or a magnetic wall at the symmetric plane, we can invoke the following two-step approach. One is to characterize the coaxial feeding network shown in Fig. 2 using the formally exact modal-expansion method and the other is to obtain the refection coefficient in the coaxial feed line of the sleeve monopole antenna shown in Fig. 4. Once both parts are analyzed by the modal-expansion method, one can then readily combine them using the well-known scattering matrix cascading technique [11] to obtain all scattering parameters of

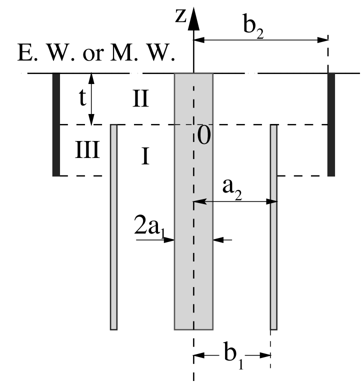


Fig. 2 Geometry of half of the coaxial feeding network.

the structure and the radiation characteristics of the electromagnetically coupled coaxial dipole antenna.

3. Analysis of the Coaxial Feeding Network

Figure 2 shows the half of the coaxial feeding network for the electromagnetically coupled dipole antenna. There are three coaxial waveguides involved in the feeding network: Waveguide I of inner and outer radii a_1 and b_1, Waveguide II of inner and outer radii a_1 and b_2, and Waveguide III of inner and outer radii a_2 and b_2, respectively. Since the feeding network is axi-symmetric, only TM_{0n} modes can be excited [12] and therefore will be considered in all three waveguides. The cylindrical coordinate system (ρ, ϕ, z) is employed in the paper and is also shown in Fig. 2 with the z-axis being along the antenna's axial direction. The length of waveguide II is t, half of the width of the annular ring slot. The transverse electromagnetic fields in three waveguides are expressed by the summation of their eigen modes weighted with unknown coefficients as follows.

Waveguide I

$$E_\rho^I = \sum_{n=0}^{N_I} [A_{1n}^+ e^{-\gamma_{1n} z} + A_{1n}^- e^{\gamma_{1n} z)}] e_{In\rho} \tag{2a}$$

$$H_\phi^I = \sum_{n=0}^{N_I} [A_{1n}^+ e^{-\gamma_{1n} z} - A_{1n}^- e^{\gamma_{1n} z}] Y_{In} e_{In\rho} \tag{2b}$$

where $e_{1n\rho}$ is the transverse electric field for the eigenmode TM_{0n}, Y_{1n} and γ_{1n} are its modal admittance and propagation constant, respectively. A_{1n}^+ and A_{1n}^- are the incident and reflected modal amplitudes of the TM_{0n} mode in the feed waveguide I.

$$e_{1n\rho} = \begin{cases} \dfrac{1}{\sqrt{2\pi \ln(b_1/a_1)}} \dfrac{1}{\rho}, & n = 0 \\ \dfrac{\sqrt{\pi}(-x_{1n}/a_1) Z_1(x_{1n}\rho/a_1)}{2\sqrt{[Y_0(x_{1n})/Y_0(x_{1n}b_1/a_1)]^2 - 1}}, & n > 0 \end{cases}$$

$$Z_1\left(\frac{x_{1n}\rho}{a_1}\right) = J_1\left(\frac{x_{1n}\rho}{a_1}\right) Y_0(x_{1n}) - Y_1\left(\frac{x_{1n}\rho}{a_1}\right) J_0(x_{1n})$$

$$Y_{1n} = \frac{j\omega\varepsilon_0\varepsilon_{rI}}{\gamma_{1n}}$$

$$\gamma_{1n} = \sqrt{(x_{1n}/a_1)^2 - k_0^2 \varepsilon_{rI}}$$

with $x_{10} = 0$, and x_{1n} being the root of the equation $J_0(x_{1n}b_1/a_1)Y_0(x_{1n}) - Y_0(x_{1n}b_1/a_1)J_0(x_{1n}) = 0$. J_0 and Y_0 are Bessel functions of the first and second kinds of order zero, respectively. ε_{rI} is the relative permittivity of the material in the feed waveguide I.

Waveguide II

$$E_\rho^{II} = \sum_{n=0}^{N_{II}} A_{2n} \frac{\sinh[\gamma_{2n}(t-z)]}{\sinh(\gamma_{2n}t)} e_{2n\rho} \tag{3a}$$

$$H_\phi^{II} = \sum_{n=0}^{N_{II}} A_{2n} \frac{\cosh[\gamma_{2n}(t-z)]}{\sinh(\gamma_{2n}t)} Y_{2n} e_{2n\rho} \tag{3b}$$

Waveguide III

$$E_\rho^{III} = \sum_{n=0}^{N_{III}} (A_{3n}^+ e^{-\gamma_{3n}z} + A_{3n}^- e^{\gamma_{3n}z}) e_{3n\rho} \tag{4a}$$

$$H_\phi^{III} = \sum_{n=0}^{N_{III}} (A_{3n}^+ e^{-\gamma_{3n}z} - A_{3n}^- e^{\gamma_{3n}z}) Y_{3n} e_{3n\rho} \tag{4b}$$

where $e_{2n\rho}$, Y_{2n}, γ_{2n}, $e_{3n\rho}$, Y_{3n}, and γ_{3n} are defined in the same way as $e_{1n\rho}$, Y_{1n}, γ_{1n} using their respective radii and eigen values. Application of the boundary conditions that the tangential electric and magnetic field components must be continuous across the regional interface ($z = 0$) leads to the following matrix equations.

$$\mathbf{A_2} = \mathbf{M_{21}}(\mathbf{A_1^+} + \mathbf{A_1^-}) + \mathbf{M_{23}}(\mathbf{A_3^+} + \mathbf{A_3^-}) \tag{5a}$$

$$\mathbf{Y_1}(\mathbf{A_1^+} - \mathbf{A_1^-}) = \mathbf{M_{21}^T}\left[\frac{Y_{2n}}{\tanh(\gamma_{2n}t)}\right]\mathbf{A_2} \tag{5b}$$

$$\mathbf{Y_3}(\mathbf{A_3^+} - \mathbf{A_3^-}) = \mathbf{M_{23}^T}\left[\frac{Y_{2n}}{\tanh(\gamma_{2n}t)}\right]\mathbf{A_2} \tag{5c}$$

where $[.]^T$ is the transpose operator of a matrix, $\mathbf{M_{21}}$ and $\mathbf{M_{23}}$ are the E-field mode-matching matrices defined, respectively, as

$$M_{21(n,m)} = 2\pi \int_{a_1}^{b_1} e_{2n\rho} e_{1m\rho} \rho d\rho$$

$$M_{23(n,m)} = 2\pi \int_{a_2}^{b_2} e_{2n\rho} e_{3m\rho} \rho d\rho$$

Eliminating $\mathbf{A_2}$ in (5) and rearranging the modal expansion coefficient matrices yields the following scattering matrix.

$$\begin{bmatrix} \mathbf{A_1^-} \\ \mathbf{A_3^-} \end{bmatrix} = \begin{bmatrix} \mathbf{S_{11}} & \mathbf{S_{12}} \\ \mathbf{S_{21}} & \mathbf{S_{22}} \end{bmatrix} \begin{bmatrix} \mathbf{A_1^+} \\ \mathbf{A_3^+} \end{bmatrix} \tag{6}$$

where

$$\begin{bmatrix} \mathbf{S_{11}} & \mathbf{S_{12}} \\ \mathbf{S_{21}} & \mathbf{S_{22}} \end{bmatrix} = 2\begin{bmatrix} \mathbf{I} + \mathbf{Y_1}^{-1}\mathbf{Y_{L11}} & \mathbf{Y_1}^{-1}\mathbf{Y_{L12}} \\ \mathbf{Y_3}^{-1}\mathbf{Y_{L21}} & \mathbf{I} + \mathbf{Y_3}^{-1}\mathbf{Y_{L22}} \end{bmatrix}^{-1} - \mathbf{I} \tag{7a}$$

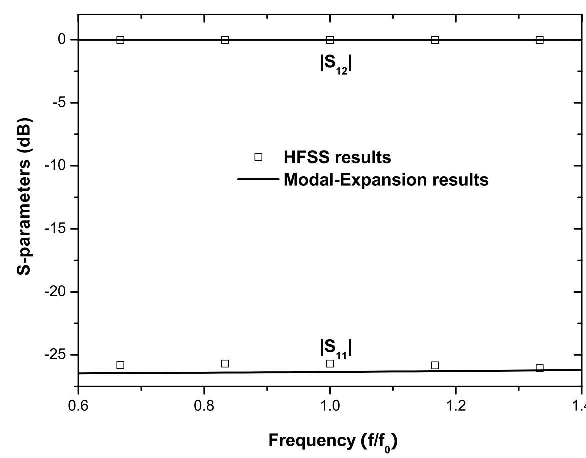
$$\mathbf{Y_{L11}} = \mathbf{M_{21}^T}\left[\frac{Y_{2n}}{\tanh(\gamma_{2n}t)}\right]\mathbf{M_{21}} \tag{7b}$$

$$\mathbf{Y_{L12}} = \mathbf{Y_{L21}^T} = \mathbf{M_{21}^T}\left[\frac{Y_{2n}}{\tanh(\gamma_{2n}t)}\right]\mathbf{M_{23}} \tag{7c}$$

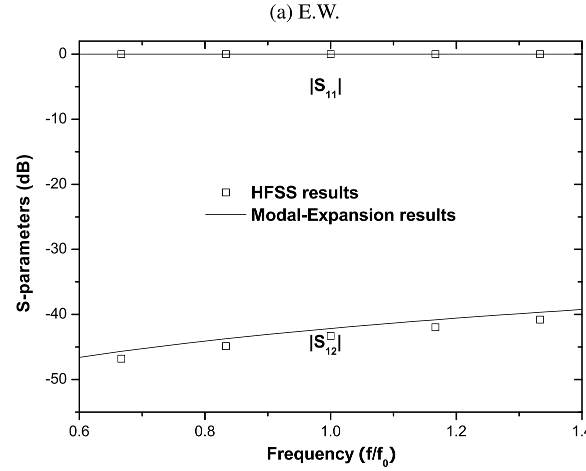
$$\mathbf{Y_{L22}} = \mathbf{M_{23}^T}\left[\frac{Y_{2n}}{\tanh(\gamma_{2n}t)}\right]\mathbf{M_{23}} \tag{7d}$$

with $\mathbf{I}$ in (7a) being the identity matrix.

Figure 3 shows the comparison of our modal-expansion results with those obtained with Ansoft's HFSS for the S-parameters of the model in Fig. 2. Good agreement can be seen from Fig. 3, which validates our formulation and the Fortran code. In order to ensure that the edge condition at $z = 0$ can be satisfied, the number of modes considered in these three waveguides must follow [11] $N_{II} = N_I \frac{b_2-a_1}{b_1-a_1} = N_{III} \frac{b_2-a_1}{b_2-a_2}$. It should be mentioned that the computation time



(a) E.W.



(b) M.W.

Fig. 3 S-parameters of half of the coaxial feed network shown in Fig. 2 ($a_1 = 0.0039\lambda_{f0}$, $b_1 = 0.009\lambda_{f0}$, $a_2 = 0.0092\lambda_{f0}$, $b_2 = 0.023\lambda_{f0}$, $t = 0.0082\lambda_{f0}$, $\varepsilon_{rI} = 1.0$).

of our modal-expansion method is less than a second for obtaining the curves in Fig. 3, while it takes more than 183 seconds for HFSS to get the results for the same structure on the same computer.

4. Analysis of the Sleeve Monopole Antenna

Figure 4 shows the geometry of the other part in the half structure of the electromagnetically coupled coaxial dipole antenna, which is a sleeve monopole antenna driven by a coaxial transmission line. The modal-expansion method previously introduced for monopole [8], sleeve monopole [9] and asymmetric dipole antenna [10] is employed here to obtain the reflection coefficient in the coaxial feed line. Similar to [8]–[10], we introduce a conducting plane on the top of the monopole antenna to confine the domain of interest and facilitate the modal-expansion analysis. As expected, the introduced conducting boundary has negligible effect on the calculated input impedance because the antenna's radiation is null in the axial direction. The inner conductor of the coaxial feed line is of radius a_2. The interior and exterior radii of the circular pipe are b_2 and a_3, respectively.

Since the formulation for the model in Fig. 4 is the same as that in [9], we will only give the final equation here. The detailed derivation and expressions are available in [9].

As shown in Fig. 4, the structure is divided into four regions: 1, 2, 3 and 4. It should be mentioned that the structure considered here is axi-symmetrical and therefore there are only three non-zero field components: E_z, E_ρ, and H_ϕ. The field components in all four regions can be expressed in cylindrical harmonic field expansions. After having obtained all the field expansion expressions, we can apply the boundary conditions that the tangential electromagnetic field components must be continuous across the regional interfaces to derive the relationships between all the expansion coefficients [9]. After some manipulations, one can obtain the reflection coefficient matrix in the coaxial feed line.

$$\mathbf{A_{1r}} = \mathbf{S_{11}A_{1i}} = [2(\mathbf{I} - \mathbf{Y_1^{-1}Y_{L0}})^{-1} - \mathbf{I}]\mathbf{A_{1i}} \tag{8}$$

where $\mathbf{Y_{L0}}$ was provided in [9]. We can then calculate the reflection coefficients for all the TM_{0n} ($n \geq 0$) modes in the coaxial feed waveguide, which include one dominant TEM mode and N_1 higher-order TM modes. Finally, we can combine the reflection coefficient matrix in this section and the scattering matrices of the two-port coaxial feeding network in the previous section to obtain the overall reflection coefficient in the coaxial feed waveguide for the electromagnetically coupled dipole with an electric wall at its symmetric plane. The same process can be repeated with minor changes for computing the overall reflection coefficient in the feed waveguide with a magnetic wall at its symmetric plane. Scattering parameters of the electromagnetically coupled coaxial dipole can then be calculated from (1).

Once the scattering parameters are determined, one can also calculate the radiation pattern of the dipole antenna by assuming an incident wave at one port of the coaxial feed waveguide and then determine all the expansion coefficients in all the waveguides and regions. The radiation pattern can be obtained by an integration involving the electric surface current on the outer conducting surface ($\rho = a_3$) [13]. The current distribution on the outer surface of the circular pipe

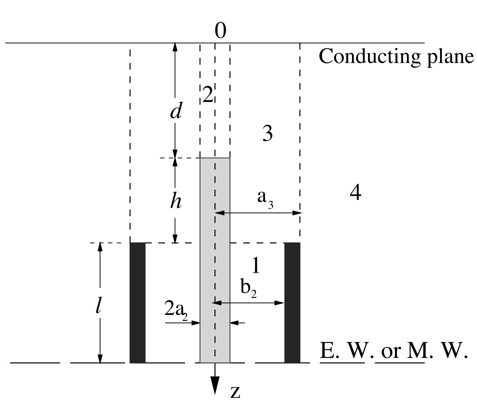
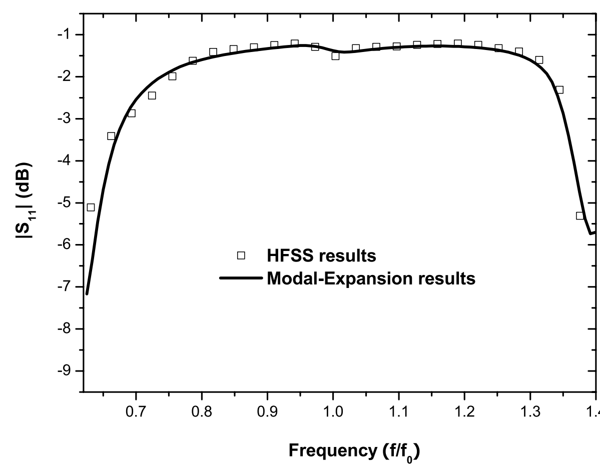
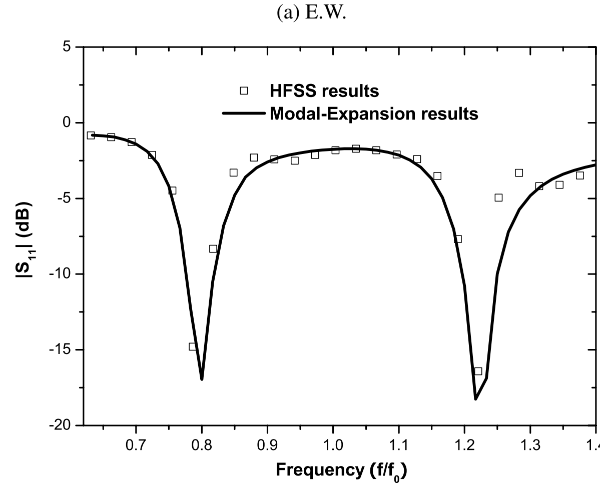


Fig. 4 Geometry of a sleeve monopole antenna driven from a coaxial transmission line.



(a) E.W.



(b) M.W.

Fig. 5 Comparison of our modal-expansion results with those of HFSS for the reflection coefficients in the feed line of the sleeve monopole shown in Fig. 4 (a_2 = $0.0092\lambda_{f0}$, b_2 = $0.023\lambda_{f0}$, a_3 = $0.023\lambda_{f0}$, l = $0.21\lambda_{f0}$, h = $0.5\lambda_{f0}$).

where $\rho = a_3$ and $-l < z < l$ is,

$$J_z = \frac{1}{2}\sum_{n=0}^{N_4}\left[A_{4n}^e \cos\frac{(2n+1)\pi z}{2L} Y_{4n}^e + A_{4n}^o \cos\frac{n\pi z}{L} Y_{4n}^o\right] \tag{9}$$

where A_{4n} is the expansion coefficient in Region 4, $L = d + h + l$, $Y_{4n} = \frac{\varepsilon_n H_1^{(2)}(\gamma_{4n}a_3)}{L\gamma_{4n}H_0^{(2)}(\gamma_{4n}a_3)}$ and the superscripts "e" and "o" represent "even mode" and "odd mode" excitations, respectively [9].

Figure 5 gives the modal-expansion results for the S-parameters of the model shown in Fig. 4. Also given are those obtained by Ansoft's HFSS for comparison. It can be seen that the agreement between them is very good.

5. Numerical Results

A Fortran program has been written to calculate the reflection coefficient and transmission coefficient in the coaxial feed line as well as the radiated field/power pattern based on the formulation described in the previous sections. As mentioned in Sect. 3 and [9], numbers of TM modes considered in regions I, II, III, 1, 2, 3, and 4 must follow the well-known criterion [11] and is numerically studied here to yield convergent results.

The convergence behavior of the input impedance of a sleeve monopole antenna with respect to N_4 is shown in Table 1. It can be seen that $N_4 = 80$ is a reasonable value to ensure convergent results. Obviously, the value N_4 varies with the length of L. The bigger the length L, the larger is the number N_4. The influence of the introduced artificial conducting plane has been examined in Table 2. It is found that its effect is negligible when the length of d is more than λ_{f0}.

Figure 6 presents the reflection and transmission coefficients of an incident wave propagating in the coaxial feed line of the electromagnetically coupled half-wavelength dipole antenna. Also shown in Fig. 6 are the measured data from [2] and HFSS results for magnitude of the reflection and transmission coefficients for comparison. It is seen that they agree well, which validates our formulation and code developed. The difference between our modal-expansion results and measured ones may be attributed to the small difference in the structure dimensions used in simulation and the actual experimental model. We can also see from Fig. 6 that the radiation is relatively weak because most of the incident power is transmitted to the other port of the the coaxial feed line. In order to enhance the radiation and increase the antenna's gain, an array of multiple coaxial dipoles can be formed along the axis [2].

It should be mentioned that our modal-expansion method is highly efficient, compared to Ansoft's HFSS. For modeling the whole dipole antenna including both even and odd mode excitations, the computation time of our modal-expansion method is 40 seconds, while it requires more than 108 minutes for HFSS to simulate the whole structure on the same computer.

Table 1 Convergence of the input impedance of a sleeve monopole antenna with respect to N_4, the number of the radial-line TM modes in Region IV ($a_2 = 0.0092\lambda_{f0}$, $b_2 = 0.023\lambda_{f0}$, $a_3 = 0.023\lambda_{f0}$, $l = 0.21\lambda_{f0}$, $h = \lambda_{f0}$).

N_4	Input resistance (Ohms)	Input reactance (Ohms)
10	58.51	-27.96
20	59.03	-19.54
40	58.81	-16.26
80	58.44	-14.63
160	58.24	-13.76

Table 2 Effect of the distance from the introduced conducting boundary to the monopole end in Fig. 4 on the sleeve monopole antenna's input impedance ($a_2 = 0.0092\lambda_{f0}$, $b_2 = 0.023\lambda_{f0}$, $a_3 = 0.023\lambda_{f0}$, $l = 0.21\lambda_{f0}$, $h = \lambda_{f0}$).

d	Input resistance (Ohms)	Input reactance (Ohms)
$0.354\lambda_{f0}$	58.65	-12.98
$0.707\lambda_{f0}$	57.89	-13.26
$1.414\lambda_{f0}$	58.19	-13.54
$2.121\lambda_{f0}$	57.95	-13.15

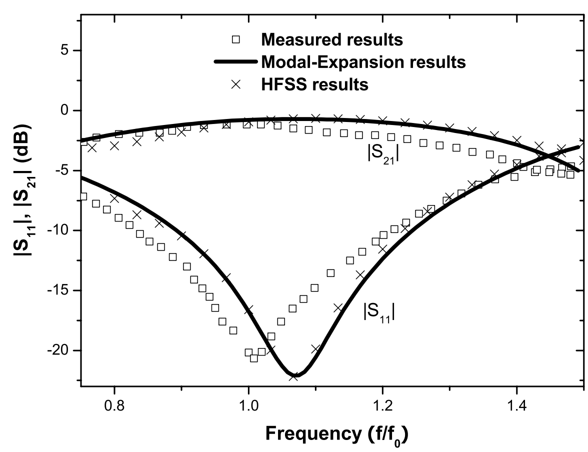


Fig. 6 S-parameters of a single electromagnetically coupled coaxial dipole antenna ($a_1 = 0.0039\lambda_{f0}$, $b_1 = 0.009\lambda_{f0}$, $a_2 = 0.0092\lambda_{f0}$, $b_2 = 0.022\lambda_{f0}$, $a_3 = 0.023\lambda_{f0}$, $l = 0.21\lambda_{f0}$, $t = 0.0082\lambda_{f0}$, $\varepsilon_{rI} = 1.0$, $h = 0.5\lambda_{f0}$).

Figure 7 shows the radiation pattern of the electromagnetically coupled coaxial dipole antenna. It is seen that the radiation is maximum in the horizontal plane and is almost a null in the axial (vertical) direction, which is expected. This also affirms our assertion that the introduction of a conducting wall above the monopole antenna has little influence on the analysis of the coaxial dipole antenna, but it makes possible the modal expansion of the field expressions in terms of the summation of discrete eigen-mode functions.

Figures 8, 9 and 10 show the variation of the antenna's return loss characteristics with respect to the dimensions h, b_2 and d of the coaxial dipole antenna. It is seen that the length h in Fig. 4 has insignificant effect on the dipole antenna's radiation, which validates our method of using a sleeve monopole in the model to approximate the half structure in Fig. 1. The size of b_2 has a large influence on the amplitude of the return loss and can therefore be used to ad-

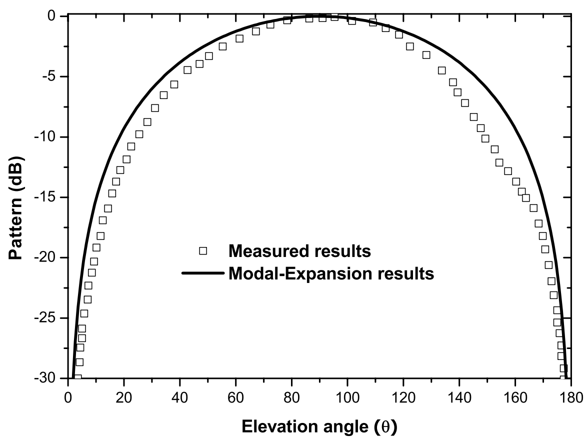


Fig. 7 Calculated radiation patterns of the electromagnetically coupled coaxial dipole antenna. ($a_1 = 0.0039\lambda_{f0}$, $b_1 = 0.009\lambda_{f0}$, $a_2 = 0.0092\lambda_{f0}$, $b_2 = 0.022\lambda_{f0}$, $a_3 = 0.023\lambda_{f0}$, $l = 0.21\lambda_{f0}$, $t = 0.0082\lambda_{f0}$, $\varepsilon_{rI} = 1.0$, $h = 1.0\lambda_{f0}$).

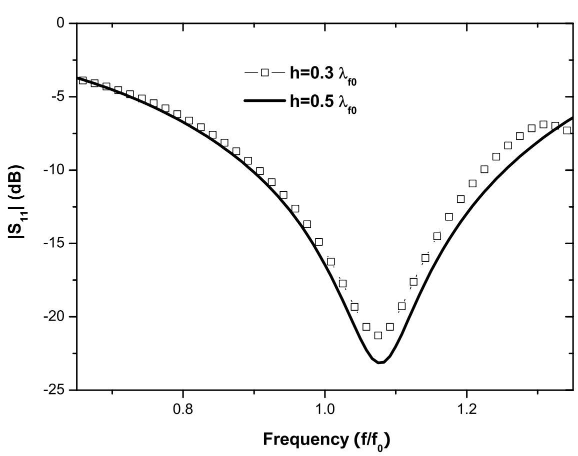


Fig. 8 Calculated return loss results of an electromagnetically coupled coaxial dipole antenna for different values of h (other parameters are the same as those in Fig. 6).

just the matching performance of the feed coaxial-line. It is also observed that it has a noticeable effect on the resonant frequency of the dipole antenna. A large radius b_2 is equivalent to a longer dipole, which in turn implies a lower resonant frequency. These characteristics can be useful to guide us to design the single coaxial dipole element and the electromagnetically coupled dipole array. It is seen from Fig. 10 that the resonant frequency is very sensitive to the length of the single-dipole, as we expected.

As mentioned previously, one advantage of our modal-expansion method for analyzing the electromagnetically coupled coaxial dipole antenna is that it can take the thickness of conductors into account to obtain more accurate results. The effect of the thickness of inner coaxial cable and outer pipe on the dipole's return loss are examined and shown in Figs. 11 and 12, respectively. It can be seen that the resonant frequency shifts when the conductors' thickness varies, especially for the inner conductor, which may explain the difference between our modal-expansion results

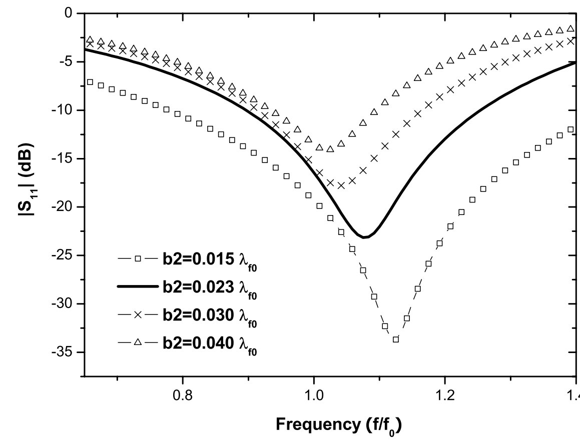


Fig. 9 Calculated return loss results of an electromagnetically coupled coaxial dipole antenna for different values of b_2 (other parameters are the same as those in Fig. 6).

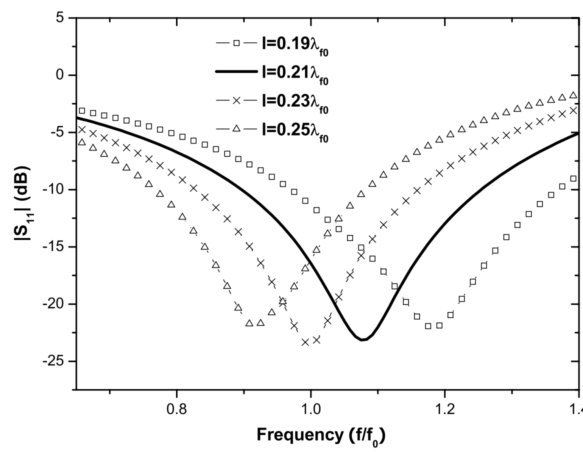


Fig. 10 Calculated return loss results of an electromagnetically coupled coaxial dipole antenna for different values of l (other parameters are the same as those in Fig. 6).

with measured ones in [2], as shown in Fig. 6.

6. Concluding Remarks

This paper has described an efficient modal-expansion analysis of an electromagnetically coupled coaxial dipole antenna. The even- and odd-mode excitations have been employed initially to simplify the modeling of the whole antenna structure. The resultant half structure has then been characterized by the modal-expansion method using a two-step approach; one is to model the coaxial feeding network and the other is to analyze a sleeve monopole antenna. Numerical results for the reflection and transmission coefficients in the coaxial feed line and the antenna's radiation pattern have been presented and compared with available data. The presented modal-expansion method has no limitation on the radius of dipole antennas and can take the conductors' finite thicknesses into account. The described analysis is currently being generalized to treat electromagnetically coupled coaxial dipole arrays.

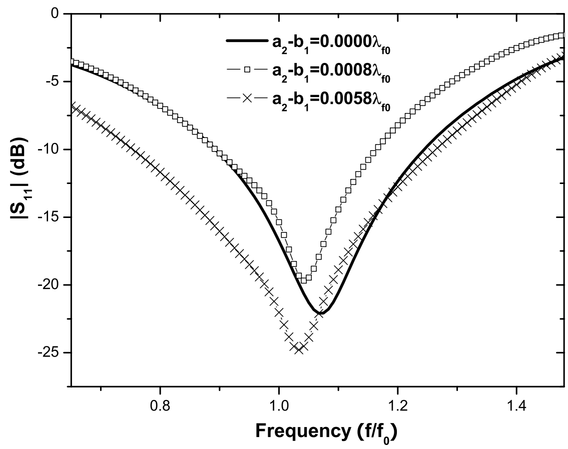


Fig. 11 Calculated return loss results of an electromagnetically coupled coaxial dipole antenna for different thicknesses of the inner coaxial cable (other parameters are the same as those in Fig. 6).

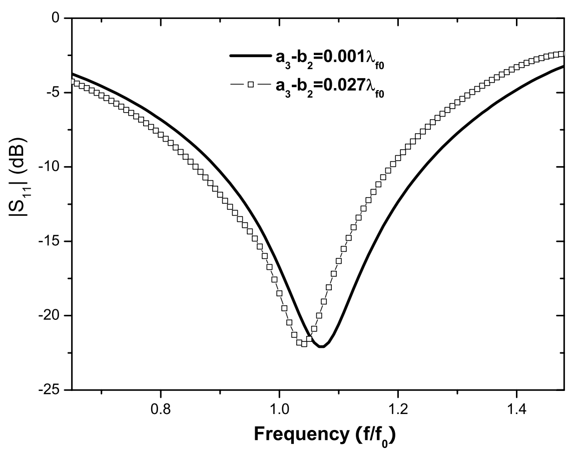


Fig. 12 Calculated return loss results of a coaxial dipole antenna for different thicknesses of the outer pipe (other parameters are the same as those in Fig. 6).

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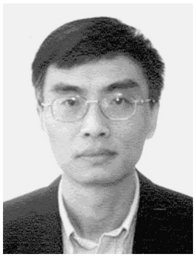

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