

PERFORMANCE OF REGULAR PML, CFS-PML, AND SECOND-ORDER PML FOR WAVEGUIDE PROBLEMS

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ABSTRACT: The performance of regular, complex frequency shifted (CFS), and second-order perfectly matched layers (PMLs) is studied for the numerical simulation of waveguide problems. The limitations of each PML is clearly demonstrated. It is shown that the second-order PML is overall the best choice for a general problem, since the regular PML is incapable of absorbing evanescent waves and the CFS-PML is ineffective for the absorption of low-frequency propagating waves. Both two- and three-dimensional problems are considered to confirm the analysis. © 2006 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 48: 2121–2126, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21872

Key words: perfectly matched layers; finite difference time domain (FDTD) method; numerical analysis; computational electromagnetics

1. INTRODUCTION

The basic concept of perfectly matched layers (PMLs) was first introduced by Bérenger [1] as a material-based absorbing boundary condition (ABC) for the finite-difference time-domain (FDTD) simulation of open-region electromagnetic problems. The PML enjoyed a great success because of its great capability of absorbing propagating waves regardless of their frequency. However, in its original form, the PML is incapable of absorbing evanescent waves and various efforts have been attempted to overcome this limitation [2–5]. Among these, the complex frequency shifted (CFS)-PML [5] is shown to be highly effective, especially after the development of a simple convolution-based implementation [6]. The CFS-PML has since been applied to several different problems [7–10].

While the CFS-PML is highly effective in absorbing evanescent waves, it compromises the capability of the original PML in the absorption of low-frequency propagating waves. Therefore, for a general problem where both evanescent and low-frequency propagating waves exist, both the CFS and the regular PMLs fail to provide a highly accurate solution.

To overcome the limitations of the CFS and the regular PMLs, we proposed the second-order PML [11], which retains the advantages of both the CFS and regular PMLs. We have shown that the second-order PML is highly effective in absorbing both evanescent and low-frequency propagating waves in both open-region and periodic problems [11]. In this study, we evaluate the performance of the second-order PML, relative to the regular and CFS-PMLs, for the simulation of waveguide problems with a goal to identify the best choice for various specific problems.

2. REGULAR PML, CFS-PML, AND SECOND-ORDER PML

Following the approach of stretched coordinates [12], we can write the z -projection of Maxwell–Ampere’s law in free-space for the modified Maxwell’s equations as

$$j\omega\epsilon E_z = -\frac{\partial}{\partial y}H_x + \frac{1}{s_x}\frac{\partial}{\partial x}H_y, \quad (1)$$

where s_x is the stretched-coordinate metric coefficient. For waveguide problems, absorption is needed only in one direction, which is assumed to be the x -direction here.

For the second-order PML proposed in [11], we set s_x to be

$$s_x = \left(\kappa_1 + \frac{\sigma_1}{\alpha_1 + j\omega\epsilon} \right) \left(\kappa_2 + \frac{\sigma_2}{\alpha_2 + j\omega\epsilon} \right), \quad (2)$$

where $\kappa_{1,2}$, $\sigma_{1,2}$, and $\alpha_{1,2}$ are the parameters that can be chosen to control the PML performance. Note that the second-order PML is reduced to the regular PML when $\alpha_1 = 0$, $\kappa_2 = 1$, and $\sigma_2 = 0$, and to the CFS-PML when $\alpha_1 \neq 0$, $\kappa_2 = 1$, and $\sigma_2 = 0$. We also note that when $\alpha_1 = \alpha_2 = 0$, the second-order PML is reduced to the one proposed in Ref. 13. In all cases, $\sigma_{1,2} \geq 0$ and $\kappa_{1,2} \geq 1$ only inside the PML region. The numerical implementation of the regular and the CFS-PMLs is described in Ref. 9 and the implementation of the second-order PML is described in Ref. 11 using auxiliary differential equations.

Even though it is not required theoretically that the real part of the metric coefficient (r_{pml}) is 1 at the air/PML interface, in practice setting $r_{\text{pml}} = 1$ at the interface is desirable as it reduces grid reflections. This requirement can easily be satisfied for the regular or the CFS-PML by setting $\kappa = 1$ at the interface. However, it would require a special treatment if the second-order PML is used because the real part of the second-order PML is given by

$$r_{\text{pml}} = \kappa_1\kappa_2 + \frac{\kappa_1\alpha_2\sigma_2}{a_2^2 + (\omega\epsilon)^2} + \frac{\kappa_2\alpha_1\sigma_1}{a_1^2 + (\omega\epsilon)^2} + \frac{\sigma_1\sigma_2[a_1a_2 - (\omega\epsilon)^2]}{[a_1^2 + (\omega\epsilon)^2][a_2^2 + (\omega\epsilon)^2]}. \quad (3)$$

This real part reduces to 1 at the interface when $\sigma_{1,2} = 0$ and $\kappa_{1,2} = 1$ at the interface.

The imaginary part of the second-order PML, responsible for the attenuation of propagating waves, is given by

$$i_{\text{pml}} = -\frac{\kappa_1\sigma_2\omega\epsilon}{a_2^2 + (\omega\epsilon)^2} - \frac{\kappa_2\sigma_1\omega\epsilon}{a_1^2 + (\omega\epsilon)^2} - \frac{\sigma_1\sigma_2\omega\epsilon(a_1 + a_2)}{[a_1^2 + (\omega\epsilon)^2][a_2^2 + (\omega\epsilon)^2]}. \quad (4)$$

Here, it is important to realize that, if both $a_1 \neq 0$ and $a_2 \neq 0$, the second-order PML will have the same problem as the CFS-PML at low frequencies, since the reflection coefficient for the PML, assuming for simplicity that all the variables are constant, is given by

$$R(\theta) = e^{-2d\eta\omega c \cos \theta i_{\text{pml}}}, \quad (5)$$

where θ is the angle of incidence at the PML interface, η is the free-space impedance, c is the speed of light, and d is the PML thickness. Clearly the reflection coefficient for the PML is 1 when $\omega \rightarrow 0$. A simple way of overcoming this limitation is to set either a_1 or a_2 to be zero. If $a_2 = 0$, the first term in Eq. (5) will be frequency independent and will resemble the absorption of the regular PML.

Even though the second-order PML adds more degrees of freedom that might help to achieve lower reflections, the trade-off between the gain and the computational cost and the time required to adjust all the variables has to be taken into account. If the problem does not require absorption of evanescent waves, the cost

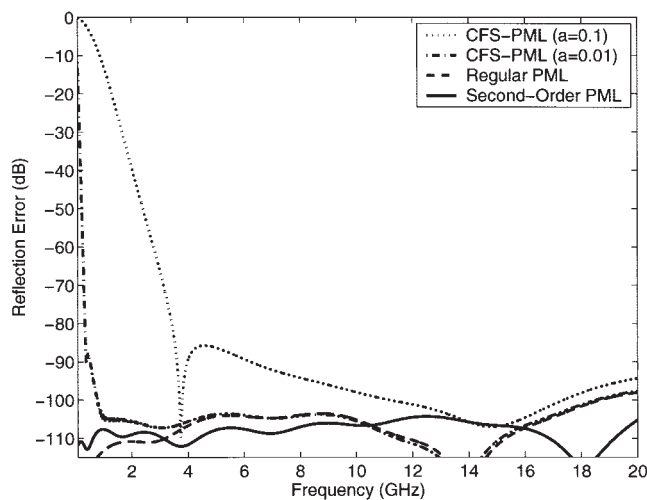


Figure 1 Reflection error as a function of frequency for the TEM mode in a parallel-plate waveguide using a 10-layer regular PML, CFS-PML, and second-order PML

both in memory requirement and in time spent adjusting the parameters for both the CFS-PML and the second-order PML will not pay off. Similarly, if no low-frequency propagating waves are present, the CFS-PML would be sufficient. For general problems, where both evanescent and low-frequency propagating waves are present, the rule-of-thumb is to use the best CFS-PML available as the first term in Eq. (2) and to set $a_2 = 0$ in the second term. Also, $\sigma_{2\max} = 0.075$ seems sufficient to guarantee absorption of low-frequency propagating waves. With this setup, the second-order PML can perform better than both the regular and CFS-PMLs in many different problems.

We begin our analysis of waveguide problems with the simplest 2D-case. An empty parallel-plate waveguide will be simulated. This example will show the difference in the performance of different PMLs. It is suitable for that purpose because such waveguide can support both TEM and higher-order modes. Then we apply the technique to different types of 3D waveguide problems. An empty rectangular waveguide and an inhomogeneous dielectric waveguide will be considered. In both cases, no TEM waves are supported, so the CFS-PML will have a good performance. We finally consider a microstrip problem where the propagating spectrum includes both low-frequency propagating waves, for which the CFS-PML does not perform well, and higher-order modes for which the regular PML will have a poor performance. In all cases, the reflection coefficient is calculated at the PML/air interface 10 cells away from the source. A reference solution is obtained by simulating a waveguide long enough so that no reflection occurs during the simulation period.

3. PARALLEL-PLATE WAVEGUIDE

In the past, several authors have tried to address the problem of absorbing evanescent waves in a waveguide [3, 7, 14, 15]. In all cases, even though the absorption of those waves indeed became better, the PMLs employed were completely ineffective at the cutoff frequencies for the waveguide modes.

In this section, we will optimize the performance of the regular PML, the CFS-PML, and the second-order PML for each different problem. The idea is to demonstrate the limitations of each metric coefficient and their advantages. The parallel-plate waveguide is the best choice for this study, since we can excite the TEM mode individually, the TM_1 mode individually, or a combination of

those two modes at the same time and compare the performance of each metric coefficient.

Our first case of study is a simple parallel-plate waveguide excited with a TEM mode using the same waveguide as in Refs. 3, 7, and 15. Since the cutoff frequency for the TEM mode is 0 Hz, the propagating spectrum will include frequencies in the entire band. For that reason, the regular PML will perform very well in the entire band, as can be seen in Figure 1. However, the CFS-PML eventually breaks down at low frequencies. The only way to make the CFS-PML work for low-frequency propagating waves is to reduce the value of a , which effectively reduces the CFS-PML to the regular PML. The performance of the second-order PML is as good as that of the regular PML. The parameters used here for the regular PML are $\kappa = 1$ and $\sigma = 5\xi^3$, where ξ is zero at the air/PML interface and 1 at the end. For the CFS-PML, we used the same profile with two different values for a . Finally, for the second-order PML we used $\sigma_1 = 4.5\xi^3$, $\sigma_2 = 2.9\xi^3$, $a_1 = \sigma_2$, $\kappa_1 = \kappa_2 = 1$, and $a_2 = 0$.

The second case is the same parallel-plate waveguide but excited with the TM_1 mode, as was done in Refs. 3 and 7. In this case, the regular PML is totally ineffective at the cutoff frequency of the TM_1 mode. It is possible to improve its performance for evanescent waves at the low-frequency end of the spectrum by setting $\kappa \neq 1$. However, this compromises the absorption of propagating waves at the high-frequency end. Fang and Wu [3] developed a modified PML that can absorb evanescent waves without such a problem. However, the absorption at cutoff is still problematic. Figure 2 shows the performance of the regular PML for this problem with σ having the same profile as in the previous example.

As shown in Ref. 7, the CFS-PML is very effective for the absorption of the evanescent waves below the cutoff frequency for this problem. Indeed, the performance of the CFS-PML is better than the regular PML even if one takes into account the increase in memory requirement from the regular PML to the CFS-PML. Although the simulations in Ref. 7 showed large reflection at the cut-off, this problem can be alleviated by setting $\kappa \neq 1$. Figure 3 shows the results of the CFS-PML using two different values for $\kappa_{\max}$. As is the case with the regular PML, the increase in κ compromises the absorption of high-frequency propagating waves. The profile used for the CFS-PML where $\kappa \neq 1$ is $\sigma = 35\xi^3$, $a = 0.1$, and $\kappa = 1 + 4\xi^2$, with ξ zero at the air/PML interface and one at the end. For $\kappa = 1$, we used $a = 0.1$ and $\sigma = 7\xi^3$.

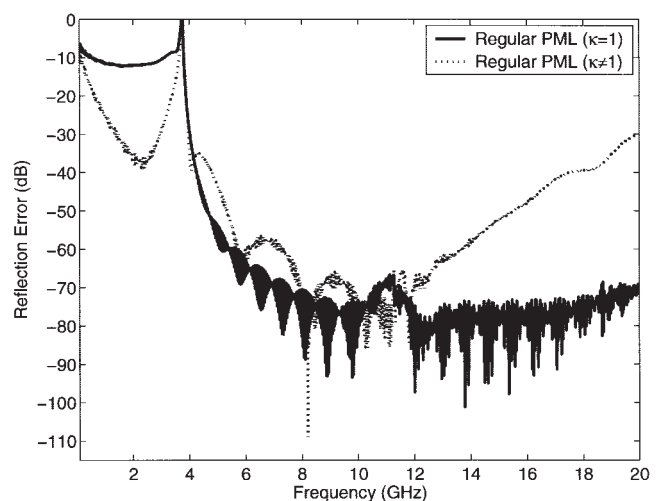


Figure 2 Reflection error for a 10-layer regular PML as a function of frequency for the TM_1 mode in a parallel-plate waveguide for $\kappa_{\max} = 1$ (solid line) and $\kappa_{\max} = 5$ (dotted line)

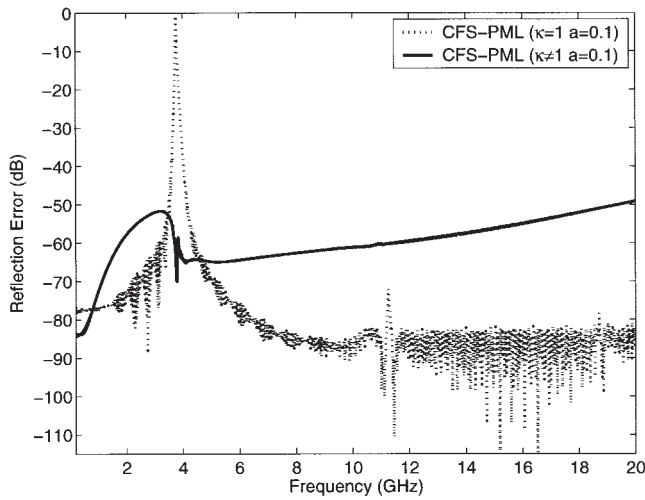


Figure 3 Reflection error for a 10-layer CFS-PML as a function of frequency for the TM_1 mode in a parallel-plate waveguide for different values of κ

The performance of the second-order PML is shown in Figure 4, which is just slightly better than the performance of the CFS-PML while both significantly outperform the regular PML. It should be noted that the regular PML performs better at high frequencies because we were mainly concerned about the maximum reflection coefficient in the entire frequency band. Both the CFS-PML and the second-order PML can be made to better absorb high-frequency propagating waves, but at the expense of a higher reflection around the cutoff frequency. The profile for the second-order PML is $\sigma_1 = 12\xi^2$, $\sigma_2 = 12\xi^4$, $\kappa_1 = \kappa_2 = 1 + \xi^2$, $a_1 = 0.1 + 0.65\sigma_2$, and $a_2 = 0.1$.

The last case of study for the parallel-plate waveguide has a simultaneous excitation of both the TEM and the TM_1 modes. In this case, neither the regular PML nor the CFS-PML are expected to have a good performance at low frequencies. The regular PML will be limited by its poor absorption around the cutoff frequency and its poor absorption of evanescent waves. The CFS-PML will be limited by its poor performance for low-frequency propagating waves. Figure 5 shows the results of the regular PML using

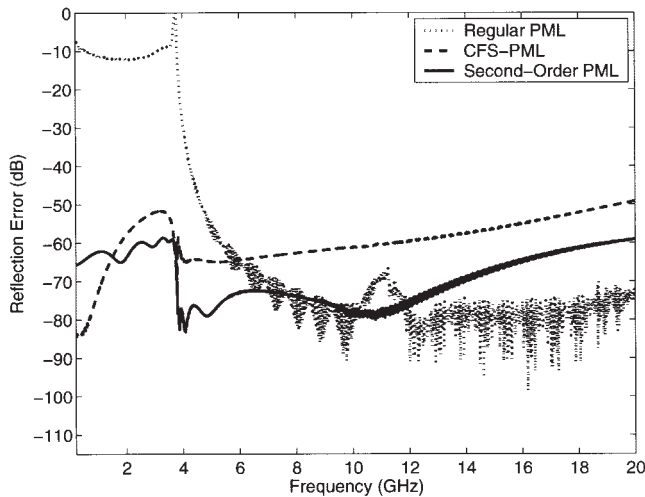


Figure 4 Reflection error for a 10-layer regular PML, CFS-PML, and second-order PML as a function of frequency for the TM_1 mode in a parallel-plate waveguide

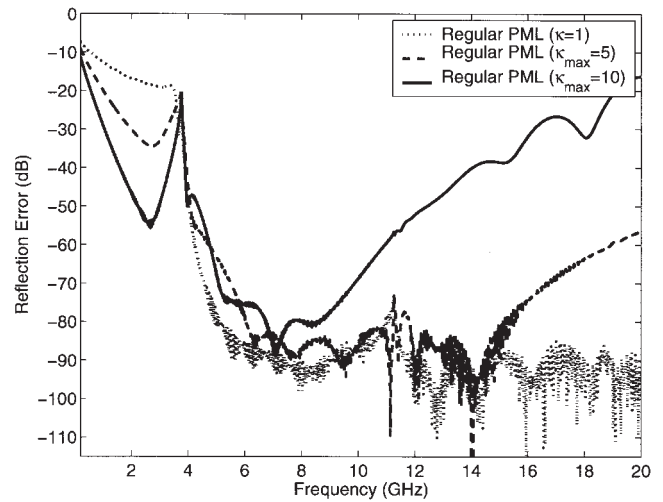


Figure 5 Reflection error for a 10-layer regular PML as a function of frequency for the TEM and TM_1 modes in a parallel-plate waveguide for different values of $\kappa_{\max}$. The values for κ start from 1 at the interface and grow with a quadratic profile to the value of $\kappa_{\max}$ at the end

different values for $\kappa_{\max}$, all having a quadratic profile. Next, we analyze the CFS-PML performance. Figure 6 shows the results of using the CFS-PML with a constant $a = 0.05$ and a linear decay for $a = 0.05(1 - \xi)$, both with $\sigma = 17\xi^3$ and $\kappa = 1 + 4\xi^2$. As can be expected, the CFS-PML fails to absorb low-frequency waves, even when a has a linear decay.

Finally, Figure 7 shows the performance of the second-order PML as compared to those of the regular and the CFS-PMLs. For the regular PML, we have $\sigma = 5\xi^3$ and $\kappa = 1$. For the CFS-PML, we have $\sigma = 17\xi^3$, $a = 0.075$, and $\kappa = 1 + \xi^2$. Finally, for the second-order PML, we have $\sigma_1 = 17\xi^3$, $\kappa_1 = 1 + \xi^2$, $\sigma_2 = 0.05\xi^3$, $a_1 = 0.075 + \sigma_2$, $\kappa_2 = 1$, and $a_2 = 0$. By doing so, we recover in the second-order PML the same behavior as the CFS-PML for absorbing evanescent waves while achieving a much better absorption at low frequencies.

4. RECTANGULAR WAVEGUIDE

In the previous section, we had different profiles for the regular PML, the CFS-PML, and the second-order PML, depending on the

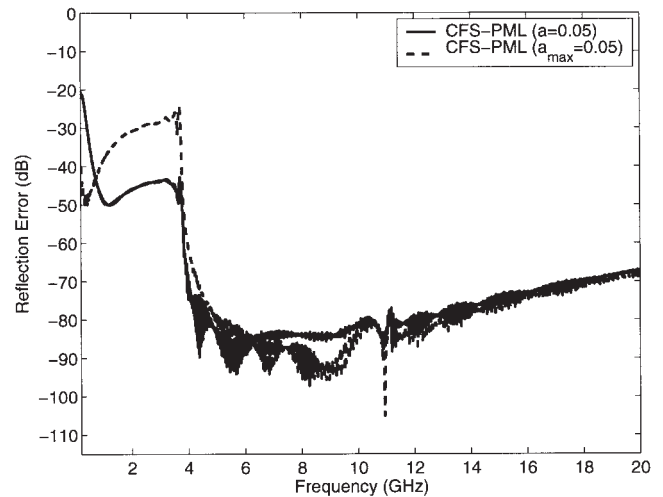


Figure 6 Reflection error for a 10-layer CFS-PML as a function of frequency for the TEM and TM_1 modes in a parallel-plate waveguide for constant a (solid line) and a linear decay in a (dashed line)

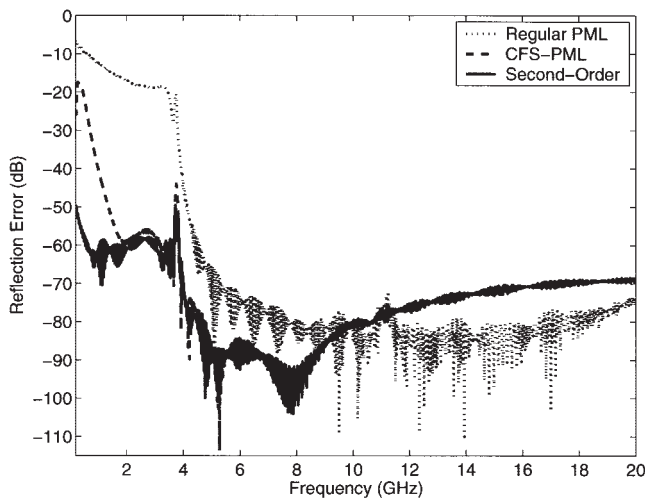


Figure 7 Reflection error for a 10-layer regular PML, CFS-PML, and second-order PML as a function of frequency for the TEM and TM_1 modes in a parallel-plate waveguide

type of problem we were solving. The idea was to optimize the three different metric coefficients for each problem. This approach, however, is not recommended for practical purposes. It is always desirable to have a PML robust enough to absorb waves in a general problem. Since the most general case was the one with both the TEM and TM_1 modes, we use that profiles for all the 3D problems shown here. Hence, in all cases here we have, for the regular PML, $\sigma = 5\xi^3$ and $\kappa = 1$. For the CFS-PML, we have $\sigma = 17\xi^3$, $a = 0.075$, and $\kappa = 1 + \xi^2$. Finally, for the second-order PML, we have $\sigma_1 = 17\xi^3$, $\kappa_1 = 1 + \xi^2$, $\sigma_2 = 0.05\xi^3$, $a_1 = 0.075 + \sigma_2$, $\kappa_2 = 1$, $a_2 = 0$, where ξ varies from 0 at the air/PML interface to 1 at the end.

4.1 Empty Waveguide

Our first 3D problem is an empty rectangular waveguide having a cross-section of 40 mm $\times$ 10 mm. We excite this waveguide with TE_{10} and TE_{20} modes. As stated in Ref. 15, the higher-order evanescent waves attenuate faster than the lower-order ones, so one needs to set the best parameters for the first higher-order mode.

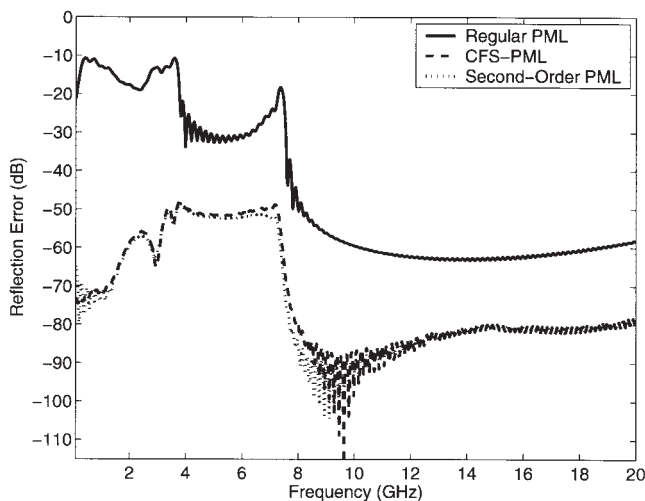


Figure 8 Reflection error for a 10-layer regular PML, CFS-PML, and second-order PML as a function of frequency for the TE_{10} and TE_{20} modes in a rectangular waveguide

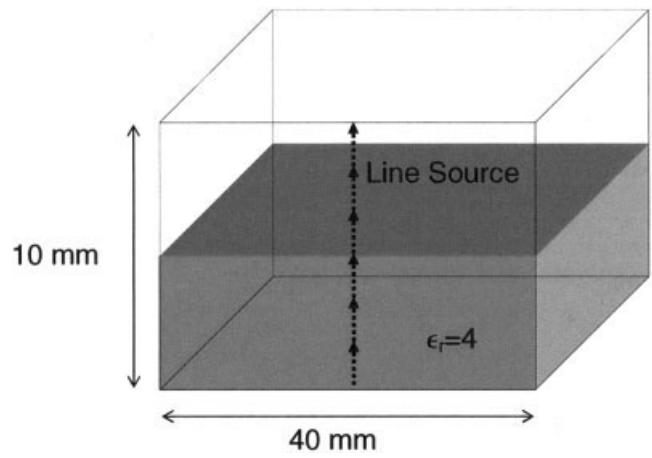


Figure 9 Computational domain for the inhomogeneous waveguide problem excited by a line source

But contrary to Ref. 15, here we launch both TE_{10} and TE_{20} modes simultaneously to check the performance of the PML. The cutoff frequencies for the two modes are 3.75 and 7.5 GHz, respectively. For this problem, no TEM wave is supported. We expect the result to be similar for the TM_1 mode in the parallel-plate waveguide problem. The CFS-PML is expected to perform better than the regular PML and the second-order PML to be similar to the CFS-PML. This is indeed the case as shown in Figure 8.

4.2 Inhomogeneous Waveguide

The next two examples are made of a partially filled waveguide. In both cases, the upper half of the waveguide is filled with air and the lower half with a dielectric having $\epsilon_r = 4$. In the first example, we used a line source to excite multiple modes, as shown in Figure 9. The spectrum of the excited incident field is plotted in Figure 10.

The performance of the regular PML, CFS-PML, and second-order PML is exhibited in Figure 11. We can clearly see that the regular PML fails to absorb the waves around the cutoff frequencies. By using the same CFS-PML, as in the example for the empty waveguide, the reflection error drops by 40 dB throughout the frequency band with its maximum value being around -30 dB

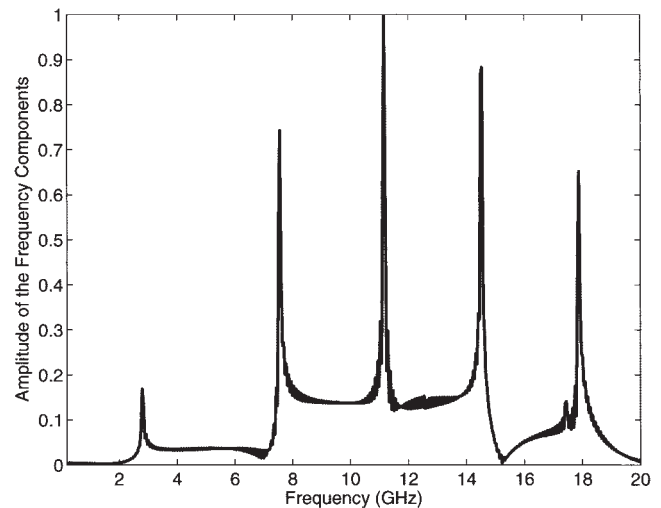


Figure 10 Spectrum of the incident field 10 cells away from the source for the inhomogeneous waveguide problem

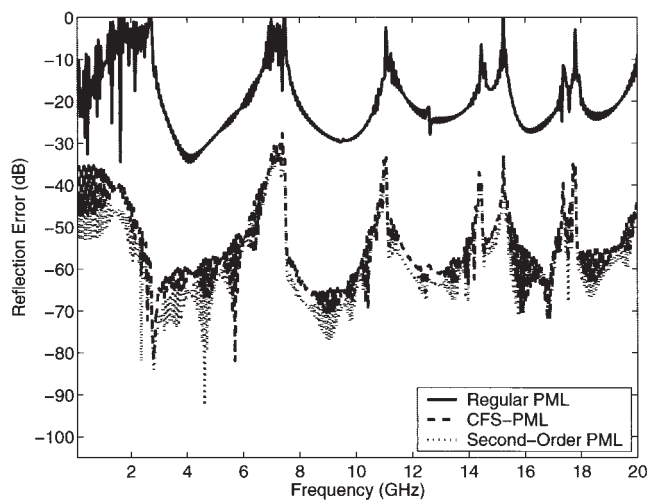


Figure 11 Reflection error for a 10-layer regular PML, CFS-PML, and second-order PML as a function of frequency for the inhomogeneous waveguide problem excited by a line source

compared to 0 dB for the regular PML. The second-order PML has a similar performance to that of the CFS-PML.

In the last example, we simulate a microstrip line inside the waveguide. In this case, the structure supports a quasi-TEM mode. Hence, low-frequency propagating waves can be excited. Similarly to the previous problem, we use a line source located in the dielectric to excite both the quasi-TEM and waveguide modes, as shown in Figure 12. Figure 13 shows the reflection error for this problem. Similar to the 2D problem with low-frequency evanescent and propagating waves, the regular and the CFS-PMLs have a poor performance at low frequencies. However, the second-order PML performs significantly better.

5. CONCLUSION

We studied the performance of three different types of metric coefficients for the PML. The regular PML is the best choice when only propagating waves are present. The CFS-PML is the best choice if strong evanescent waves are present but low-frequency propagating waves are absent. In more general cases, where both low-frequency propagating waves and strong evanescent waves are present, the second-order PML is the best choice. Both the

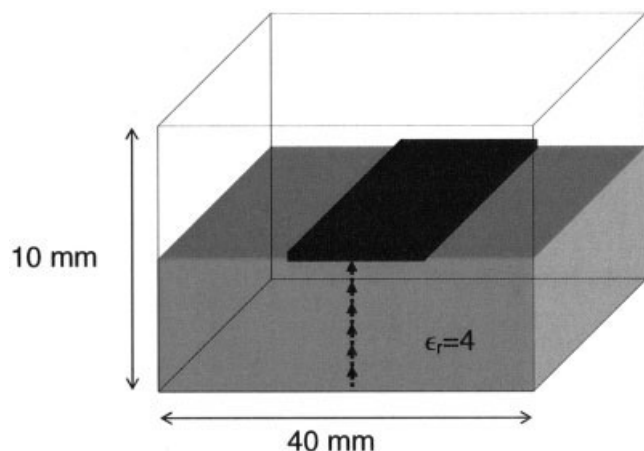


Figure 12 Computational domain for the microstrip problem excited by a line source

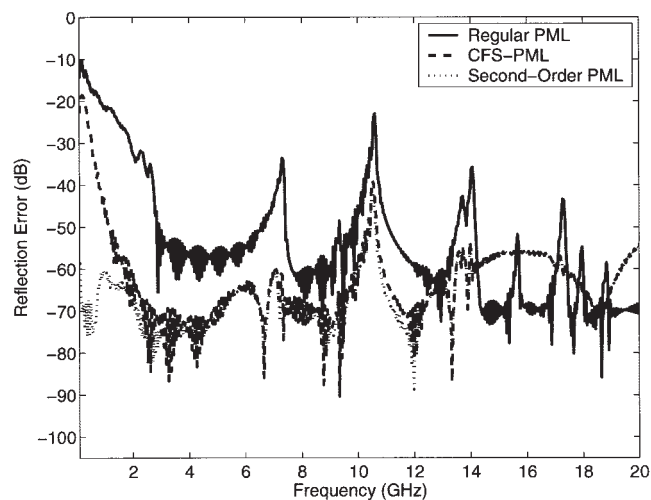


Figure 13 Reflection error for a 10-layer regular PML, CFS-PML, and second-order PML as a function of frequency for the microstrip problem excited by a line source

regular and the CFS-PMLs are special cases of the second-order PML; hence, an implementation of the second-order PML includes both of them at the expense of some memory and computation time increase.

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EFFICIENT AND SYSTEMATIC EXTRACTION METHOD FOR ACCURATE DETERMINATION OF HBT TEMPERATURE-DEPENDENT DC MODEL PARAMETERS

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ABSTRACT: This paper presents an efficient and systematic parameter extraction method of HBT temperature-dependent DC model, suitable for monitoring device/process development. All parameters of the model current sources, including self-heating parameters, were extracted using a developed simple algorithm based on a set of algebraic nonlinear equations. The modeling methodology was successfully applied to predict accurately the DC characteristics of a $2 \times 25 \mu\text{m}^2$ InGaP/GaAs HBT device. © 2006 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 48: 2126–2131, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21867

Key words: HBT device; DC model; thermal effect; extraction method

1. INTRODUCTION

Heterojunction bipolar transistors (HBTs) have demonstrated high-power output densities at microwave frequencies and are increasingly utilized for design of RF circuits, such as power amplifiers, oscillators, and mixers [1]. Thus, an accurate large-signal model for HBT devices is of great importance in designing such circuits especially when the transistor is operated in nonlinear regions where self-heating effect becomes significant. Although the temperature sensitivity of transistor parameters is significant for all types of power transistors, it is particularly important for the HBT with its relatively poor thermal conductivity. Efforts have been made to include this effect in HBT DC equivalent circuit models [2–7], and close agreements with measured data have been obtained for $I_c - V_{ce}$ characteristics. To simulate accurately the DC characteristics of the device, several other effects have to be included in the HBT DC model such as Kirk effect. Beside this, the DC model extraction process is as important as the DC model topology and formulations. This is because the end-user of a model must know how to extract the parameters of that model. Several researchers have worked extensively in the field of modeling of HBT devices, and different model topologies with corresponding parameter extraction methodologies were proposed earlier [2–12]. However, there are still some difficulties in extracting accurately and directly some of the DC model parameters such as diode parameters, self-heating parameters, and thermal resistance.

In this paper, an efficient, systematic, and detailed method for the extraction of the HBT DC model parameters is presented. A simple and reliable procedure was developed for the determination of the thermal resistance R_{th} , which required only forward Gummel data at different ambient temperatures and different collector–emitter bias voltages. The developed HBT DC model was implemented in ADS as an SDD (symbolically defined device) and validated using measured $I-V$ characteristics.

2. MODEL DESCRIPTION

2.1. Model Parameters

The adopted HBT large-signal model is based on the equivalent electric circuit shown in Figure 1. The thermal equivalent circuit included in Figure 1 is used to calculate temperature rise at the emitter junction. R_{th} is the thermal resistance. The configuration of the large-signal model equivalent electric circuit is similar to that used for small-signal model presented in Ref. 13.

The extrinsic elements describing parasitic capacitances and inductances of the device are not shown in Figure 1, and they were de-embedded from device measurements using the technique published in Ref. 14. R_b , R_c , and R_e are the extrinsic resistances. R_{bb} is the intrinsic base resistance. C_{be} is the base-emitter capacitance. C_{bc} and C_c are respectively the extrinsic and intrinsic base-collector capacitance. This paper focuses only on the extraction and validation of the DC part of the presented HBT large-signal model.

Four diodes were used in the model to account for the currents flowing through the device junctions. $Dbf3$ and Dbr represent base-emitter and base-collector junctions respectively. $Dbf1$ and $Dbf2$ model base-emitter leakage and recombination at the device surface and in the emitter space-charge region respectively. Collector–emitter current source I_{CT} in the model is calculated as follows:

$$I_{CT} = I_{cf} - I_{cr}, \quad (1)$$

where I_{cf} is the forward collector–emitter current component and I_{cr} is the reverse collector–emitter current component.

The total forward base current is calculated from the sum of the current components I_{bf1} , I_{bf2} , and I_{bf3} that correspond to the diodes $Dbf1$, $Dbf2$, and $Dbf3$ respectively. The reverse base-collector current component I_{br} corresponds to the diode Dbr .

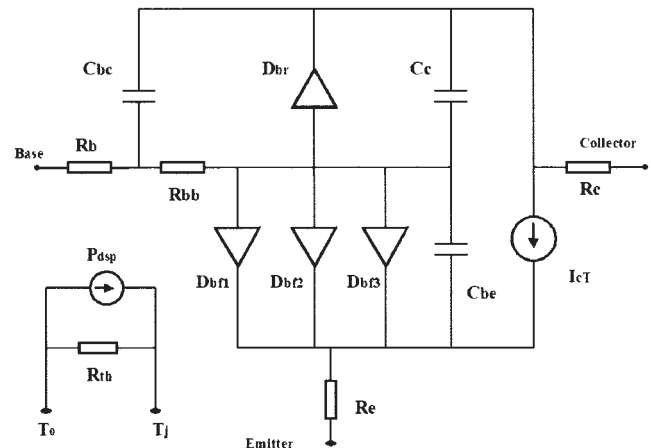


Figure 1 Equivalent circuit of the adopted HBT large signal model