

Interpolation-Based Parameterized Model Order Reduction of Delayed Systems

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Abstract—Three-dimensional electromagnetic methods are fundamental tools for the analysis and design of high-speed systems. These methods often generate large systems of equations, and model order reduction (MOR) methods are used to reduce such a high complexity. When the geometric dimensions become electrically large or signal waveform rise times decrease, time delays must be included in the modeling.

Design space optimization and exploration are usually performed during a typical design process that consequently requires repeated simulations for different design parameter values. Efficient performing of these design activities calls for parameterized model order reduction (PMOR) methods, which are able to reduce large systems of equations with respect to frequency and other design parameters of the circuit, such as layout or substrate features.

We propose a novel PMOR method for neutral delayed differential systems, which is based on an efficient and reliable combination of univariate model order reduction methods, a procedure to find scaling and frequency shifting coefficients and positive interpolation schemes. The proposed scaling and frequency shifting coefficients enhance and improve the modeling capability of standard positive interpolation schemes and allow accurate modeling of highly dynamic systems with a limited amount of initial univariate models in the design space. The proposed method is able to provide parameterized reduced order models passive by construction over the design space of interest. Pertinent numerical examples validate the proposed PMOR approach.

Index Terms—Delayed systems, interpolation, parameterized model order reduction (PMOR), partial element equivalent circuit method (PEEC).

I. INTRODUCTION

COMPLEX high-speed systems require 3-D electromagnetic (EM) methods [1]–[3] as analysis and design tools. Large systems of equations are usually generated by the use of these methods, and model order reduction (MOR) techniques

Manuscript received June 27, 2011; revised December 18, 2011; accepted December 20, 2011. Date of publication February 07, 2012; date of current version March 02, 2012. This work was supported by the Research Foundation Flanders (FWO) and by the Italian Ministry of University (MIUR) under a Program for the Development of Research of National Interest (PRIN Grant 20089J4SM9-002).

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Digital Object Identifier 10.1109/TMTT.2011.2181858

are crucial to reduce the complexity of EM models and the computational cost of the simulations, while retaining the important physical features of the original system [4]–[7]. Over recent years, the development of methods to build reduced order models (ROMs) of EM systems has been intensively investigated, with applications to interconnects, vias, and high-speed packages [8]–[11]. Among all EM methods, the partial element equivalent circuit (PEEC) method [2] has been found particularly useful for modeling PCBs, interconnects, and power systems. The PEEC method uses a circuit interpretation of the electric field integral equation (EFIE) [12], and it is especially suitable to problems involving both electromagnetic fields and circuits [2], [13], [14]. PEEC equivalent circuits are usually connected to nonlinear circuit devices such as drivers and receivers using a time domain circuit simulator (e.g., SPICE [15]). Complex systems can result in PEEC models where the number of circuit elements can be in the tens of thousands, and therefore the inclusion of these PEEC models directly into a circuit simulator may be computationally intractable.

When signal waveform rise times decrease and the corresponding frequency content increases or the geometric dimensions become electrically large, time delays must be taken into account and included in the modeling. A PEEC formulation that includes delay elements, called the τ PEEC method [16], becomes necessary and leads to systems of neutral delayed differential equations (NDDEs) [17] with constant coefficients and constant delay times in the time domain and to complex algebraic systems of equations with frequency-dependent matrices in the frequency domain. Simply using quasi-static PEEC models can result in significant errors and artifacts in the modeling [18].

Over the years, some techniques for the reduction of NDDE systems have been proposed [19]–[24]. In [21], an equivalent first-order system is computed by means of a Taylor expansion, and then MOR Krylov subspace methods [6], [7] are applied. The reduction process does not preserve the NDDE formulation. In [24], an equivalent first-order system is computed using a single-point Taylor expansion [21], and a corresponding orthogonal projection matrix is computed using a block Arnoldi algorithm [7]. Then, an orthogonal projection matrix for the original NDDE system is extracted, and a reduced NDDE system is obtained. All these previous techniques cannot efficiently handle the reduction of electrically large structures, where delays among coupled elements cannot be neglected or easily approximated by rational basis functions. Recently, a novel MOR method for NDDE systems with large delays has been proposed [25]. It is based on an adaptive multipoint expansion and MOR of equivalent first-order systems [24].

Traditional MOR techniques perform model reduction only with respect to frequency. However, during the circuit synthesis of large-scale digital or analog applications, it is also important to predict the response of the circuit under study as a function of design parameters, such as geometrical and substrate features. A typical design process includes optimization and design space exploration, and thus requires repeated simulations for different design parameter values. Such design activities call for parameterized model order reduction (PMOR) methods that can reduce large systems of equations with respect to frequency and other design parameters of the circuit, such as geometrical layout or substrate characteristics.

Several PMOR techniques have been proposed over the years. Concerning rational systems, multiparameter moment-matching methods presented in [26]–[28] use a subspace projection approach to provide parameterized ROMs, while similar approaches are described in [29]–[31] for NDDE systems. However, the structure of such methods may present some computational problems, and the resulting reduced models usually suffer from oversize when the number of moments to match is high, either because high accuracy (order) is required or because the number of parameters is large. The selection of the multidimensional expansion points and the number of multiparameter moments need to be addressed in these methods. The technique presented in [32] combines traditional passivity-preserving model order reduction methods and interpolation schemes based on a class of positive interpolation operators. A PMOR method based on a parameterization process of matrices generated by EM methods and projection subspaces is proposed in [33]. Overall passivity of parameterized ROMs is guaranteed over the design space of interest in [32] and [33].

This paper proposes a PMOR method for NDDE systems, which is based on an efficient and reliable combination of univariate model order reduction methods, a procedure to find scaling and frequency shifting coefficients and positive interpolation schemes [34]. The PMOR method proposed in this paper starts by computing a set of reduced order τ PEEC models (τ ROMs) using the MOR algorithm [25] for different design parameters values. We note that the MOR method [25] is able to perform reduction only with respect to frequency. Then, a set of scaling and frequency shifting coefficients is computed for the set of τ PEEC reduced models, which are finally interpolated to build a parameterized reduced model. In [32], a set of reduced models is interpolated to build a parameterized reduced model using standard positive interpolation schemes. The new proposed scaling and frequency shifting coefficients enhance and improve the modeling capability of standard positive interpolation schemes [32] and allow accurate modeling of highly dynamic systems with a limited amount of initial univariate models in the design space. The new proposed method does not have to deal with multiparameter moment computations and related issues. The expansion points are chosen only along the frequency axis for the τ ROMs using an adaptive algorithm [25]. Under the assumptions that the original τ PEEC models are passive and the MOR method used to provide the τ ROMs is passivity-preserving, the proposed PMOR method is able to build parameterized reduced models that are passive over the entire design space of interest.

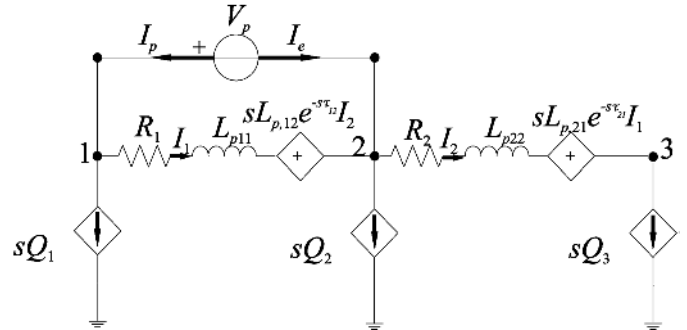


Fig. 1. Illustration of τ PEEC circuit electrical quantities for a conductor elementary cell.

The paper is organized as follows. Section II describes the modified nodal analysis (MNA) equations of the τ PEEC method. Section III describes the proposed PMOR method for NDDE systems. Finally, some pertinent numerical examples based on τ PEEC models validate the proposed technique in Section IV.

II. τ PEEC FORMULATION

The PEEC method [2] stems from the integral equation form of Maxwell's equations. With respect to other EM methods, it is worth pointing out its capability to provide a circuit interpretation of the EFIE equation, thus allowing it to handle complex problems involving both circuits and electromagnetic fields.

In the standard approach, volumes and surfaces are discretized into elementary regions, hexahedra and patches, respectively [16] over which the current and charge densities are expanded into a series of basis functions.

Following the standard Galerkin's testing procedure, topological elements, namely nodes and branches, are generated, and electrical lumped elements are identified modeling both the magnetic and electric field coupling.

Conductors are modeled by their ohmic resistance, while dielectrics require modeling the polarization charge due to the dielectric polarization [35]. Magnetic and electric field coupling are modeled by partial inductances and coefficients of potential, respectively.

An example of τ PEEC circuit for a conductor elementary cell is illustrated, in the Laplace domain, in Fig. 1, where the current controlled voltage sources $sL_{p,ij}I_j$ and the charge controlled current sources sQ_i model the magnetic and electric field couplings, respectively.

Let us assume that the meshing process of conductors and dielectrics has generated n_i volume cells where currents flow and n_n surface cells where charge is located; the resultant number of elementary cells of conductors and dielectrics is n_c and n_d , respectively, and that of electrical nodes is n_n . Furthermore, let us assume to be interested in generating an admittance representation $\mathbf{Y}(s)$ having n_p output currents $\mathbf{i}_p(t)$ under voltage excitation $\mathbf{v}_p(t)$. If the MNA approach [36] is used, the global number of unknowns is $n_u = n_i + n_d + n_n + n_p$, and an admittance representation of the PEEC circuit is obtained [25]:

$$\mathbf{C}(\tau) \frac{d\mathbf{x}(t)}{dt} = -\mathbf{G}(\tau)\mathbf{x}(t) + \mathbf{B}\mathbf{u}(t) \quad (1a)$$

$$\mathbf{i}_p(t) = \mathbf{L}^T \mathbf{x}(t) \quad (1b)$$

where $\mathbf{x}(t) \in \mathbb{R}^{n_u \times 1}$ and $\boldsymbol{\tau} \in \mathbb{R}^{n_\tau \times 1}$ contains all n_τ delays $\{\tau_k\}_{k=1}^{n_\tau}$, which denote the center-to-center delay matrices for the magnetic and electric field coupling. Since this is an n_p -port formulation, whereby the only sources are the voltage sources at the n_p -port nodes, $\mathbf{B} = \mathbf{L}$, where $\mathbf{B} \in \mathbb{R}^{n_u \times n_p}$. Each delayed entry of matrices $\mathbf{C}(\boldsymbol{\tau})$ and $\mathbf{G}(\boldsymbol{\tau})$ act as a delay operator for the corresponding entry of vector $\mathbf{x}(t)$. Hence, (1) can be rewritten in the Laplace domain as

$$s\mathbf{C}(s)\mathbf{X}(s) = -\mathbf{G}(s)\mathbf{X}(s) + \mathbf{B}\mathbf{V}_p(s) \quad (2)$$

$$\mathbf{I}_p(s) = \mathbf{L}^T \mathbf{X}(s) \quad (3)$$

$$\mathbf{C}(s) = \mathbf{C}_0 + \sum_{k=1}^{n_\tau} \mathbf{C}_k e^{-s\tau_k} \quad (4)$$

$$\mathbf{G}(s) = \mathbf{G}_0 + \sum_{k=1}^{n_\tau} \mathbf{G}_k e^{-s\tau_k} \quad (5)$$

which corresponds to the admittance transfer function

$$\mathbf{Y}(s) = \mathbf{L}^T (s\mathbf{C}(s) + \mathbf{G}(s))^{-1} \mathbf{B}. \quad (6)$$

The matrices \mathbf{C}_0 and \mathbf{G}_0 denote the quasi-static contribution.

III. PARAMETERIZED MODEL ORDER REDUCTION

Considering the influence of the design parameters $\mathbf{g} = (g^{(1)}, \dots, g^{(N)})$ such as layout and substrate features, the MNA formulation (1a)–(1b) becomes

$$\mathbf{C}(\boldsymbol{\tau}, \mathbf{g}) \frac{d\mathbf{x}(t, \mathbf{g})}{dt} = -\mathbf{G}(\boldsymbol{\tau}, \mathbf{g})\mathbf{x}(t, \mathbf{g}) + \mathbf{B}(\mathbf{g})\mathbf{u}(t) \quad (7a)$$

$$\mathbf{i}_p(t, \mathbf{g}) = \mathbf{L}(\mathbf{g})^T \mathbf{x}(t, \mathbf{g}) \quad (7b)$$

while (6) becomes

$$\mathbf{Y}(s, \mathbf{g}) = \mathbf{L}(\mathbf{g})^T (s\mathbf{C}(s, \mathbf{g}) + \mathbf{G}(s, \mathbf{g}))^{-1} \mathbf{B}(\mathbf{g}). \quad (8)$$

In this section, we describe a PMOR algorithm that is able to include, in addition to frequency, N design parameters $\mathbf{g} = (g^{(1)}, \dots, g^{(N)})$ in the reduced order model. The main objective of this PMOR method is to accurately approximate the original scalable system (having a high complexity) with a reduced scalable system (having a low complexity) by capturing the behavior of the original system with respect to frequency and other design parameters.

The proposed algorithm guarantees passivity of the parameterized reduced models over the entire design space of interest, under the assumptions that the original τ PEEC models are passive and the MOR method used to provide the τ ROMs is passivity-preserving. A flowchart that describes the different steps of the proposed PMOR method is shown in Fig. 2.

A. τ ROMs

The proposed PMOR technique starts by computing a set of reduced order models of the PEEC admittance matrix $\mathbf{Y}(s, \mathbf{g}) = \mathbf{L}^T(\mathbf{g})(s\mathbf{C}(s, \mathbf{g}) + \mathbf{G}(s, \mathbf{g}))^{-1}\mathbf{B}(\mathbf{g})$, called τ ROMs, using the MOR method described in [25] for a set of points in the design space, which we call estimation design space grid. The design space $\mathcal{D}(\mathbf{g})$ is considered as the parameter space $\mathcal{P}(s, \mathbf{g})$ without frequency. The parameter space $\mathcal{P}(s, \mathbf{g})$ contains all parameters (s, \mathbf{g}) . If the parameter space is N -dimensional, the

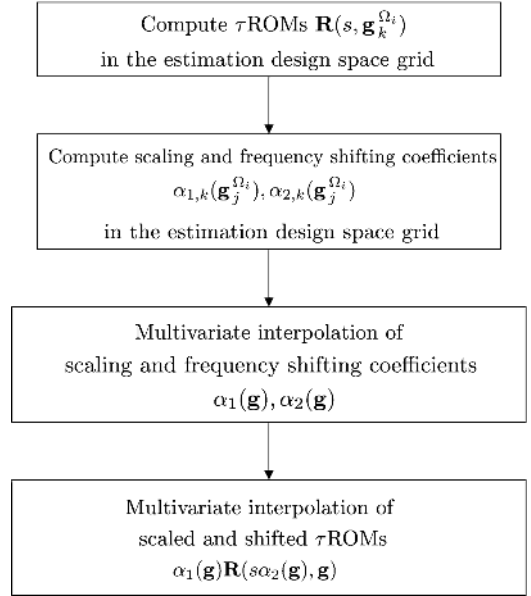


Fig. 2. Flowchart of the proposed PMOR method.

design space is $(N - 1)$ -dimensional. Two design space grids are used in the modeling process: an estimation grid and a validation grid. The first grid is utilized to build the τ ROMs. The second grid is utilized to assess the capability of parameterized reduced models of describing the system under study in a set of points of the design space previously not used for the construction of the τ ROMs. To clarify the use of these two design space grids, we show in Fig. 3 a possible estimation and validation design space grid in the case of two design parameters $\mathbf{g} = (g^{(1)}, g^{(2)})$. A τ ROM is built for each red cross (x) point in the design space. The set of τ ROMs is interpolated, as explained in what follows, to build a parameterized reduced model that is evaluated and compared with original τ PEEC models related to the blue circle (o) design space points. We note that these blue circle (o) points are not used for the generation of the τ ROMs.

N -dimensional and scattered design space grids can also be treated by the proposed PMOR technique that does not impose any constraint on the number of design parameters and the distribution of τ ROMs in the design space.

B. Scaling and Frequency Shifting Coefficients

After the computation of the τ ROMs, the next step is combining them together and building a multivariate representation $\mathbf{R}(s, \mathbf{g})$. The design space is divided into cells using hyperrectangles (regular grids) [37] or simplices (regular and scattered grids) [38]. We note that the proposed PMOR technique can handle regular and scattered design space grids.

Once the design space is divided into cells, a local parameterized model is associated to every cell that is a subdomain of the entire design space. We indicate a cell region of the design space as Ω_i , $i = 1, \dots, P$ (shown in Fig. 4) and the corresponding vertices as $\mathbf{g}_k^{\Omega_i}$, $k = 1, \dots, Q$.

We note that each vertex corresponds to a τ ROM $\mathbf{R}(s, \mathbf{g}_k^{\Omega_i})$. For each cell, an optimization procedure is used to find the scaling and frequency shifting system coefficients that make each vertex an accurate approximant of the other cell vertices.

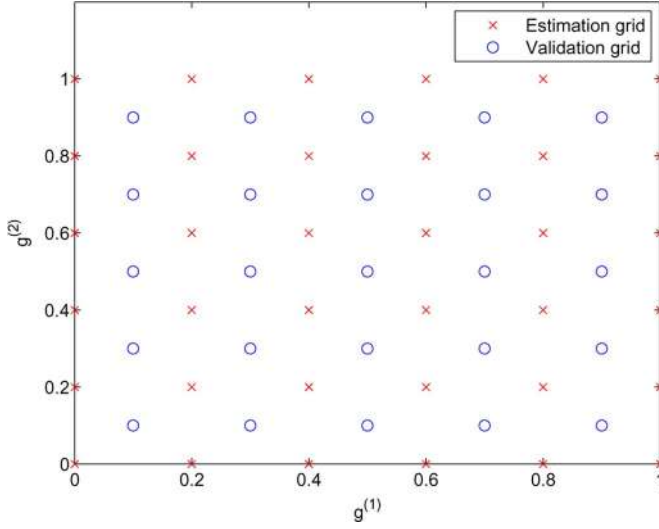


Fig. 3. An example of estimation and validation design space grid.

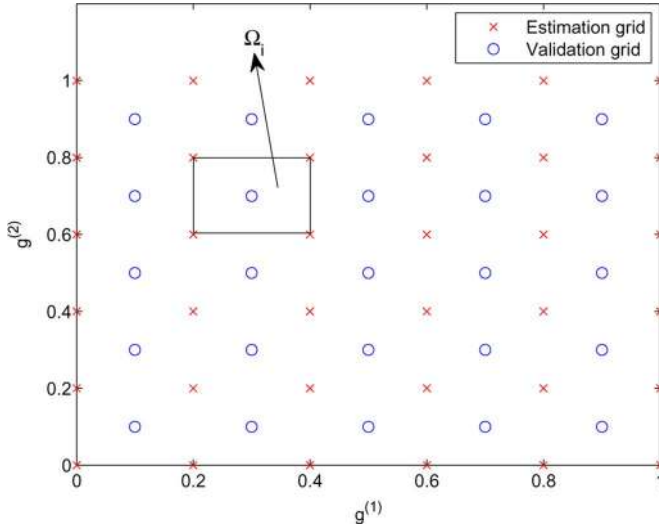


Fig. 4. An example of a design space cell Ω_i .

For each vertex $\mathbf{R}(s, \mathbf{g}_k^{\Omega_i})$, a set of scaling $\alpha_{1,k}(\mathbf{g}_j^{\Omega_i})$, $j = 1, \dots, Q$ and frequency shifting $\alpha_{2,k}(\mathbf{g}_j^{\Omega_i})$, $j = 1, \dots, Q$ real coefficients are found, such that

$$\alpha_{1,k}(\mathbf{g}_j^{\Omega_i}) \mathbf{R}(s\alpha_{2,k}(\mathbf{g}_j^{\Omega_i}), \mathbf{g}_k^{\Omega_i}) \simeq \mathbf{R}(s, \mathbf{g}_j^{\Omega_i}), \quad j \neq k \quad (9)$$

$$\alpha_{1,k}(\mathbf{g}_j^{\Omega_i}) = \alpha_{2,k}(\mathbf{g}_j^{\Omega_i}) = 1, \quad j = k. \quad (10)$$

This optimization problem can be solved using, for example, the Matlab [39] routines *fmincon* and *fminsearchbnd* with $\alpha_{1,k}(\mathbf{g}_j^{\Omega_i}) = 1, \alpha_{2,k}(\mathbf{g}_j^{\Omega_i}) = 1$, $j = 1, \dots, Q$, $j \neq k$ as an initial guess. These routines are able to impose some constraints on the optimized coefficients, which is important to guarantee the passivity of parameterized reduced order models as explained in what follows. If the response of the system under modeling needs to be computed in a specific design space point $\hat{\mathbf{g}}$, a subdomain that contains $\hat{\mathbf{g}}$ is to be found. For each vertex τ ROM of the found subdomain, the

corresponding sets of scaling and frequency shifting coefficients $\alpha_{1,k}(\mathbf{g}_j^{\Omega_i}), \alpha_{2,k}(\mathbf{g}_j^{\Omega_i})$ are interpolated in $\hat{\mathbf{g}}$ and a model $\hat{\alpha}_{1,k} \mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i})$ is built, where $\hat{\alpha}_{1,k} = \alpha_{1,k}(\hat{\mathbf{g}})$ and $\hat{\alpha}_{2,k} = \alpha_{2,k}(\hat{\mathbf{g}})$. Finally, the set of modified τ ROMs $\hat{\alpha}_{1,k} \mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i})$, $k = 1, \dots, Q$, is interpolated at an input–output level as described in [32]. We note that if a generic τ ROM $\mathbf{R}(s, \mathbf{g}_k^{\Omega_i})$ has the admittance representation

$$\mathbf{L}(\mathbf{g}_k^{\Omega_i})^T \left(s\mathbf{C}(s, \mathbf{g}_k^{\Omega_i}) + \mathbf{G}(s, \mathbf{g}_k^{\Omega_i}) \right)^{-1} \mathbf{B}(\mathbf{g}_k^{\Omega_i}) \quad (11)$$

then a corresponding scaled and frequency shifted version $\hat{\alpha}_{1,k} \mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i})$ has the admittance representation

$$\tilde{\mathbf{L}}(\mathbf{g}_k^{\Omega_i})^T \left(s\tilde{\mathbf{C}}(s, \mathbf{g}_k^{\Omega_i}) + \tilde{\mathbf{G}}(s, \mathbf{g}_k^{\Omega_i}) \right)^{-1} \tilde{\mathbf{B}}(\mathbf{g}_k^{\Omega_i}) \quad (12)$$

with

$$\begin{aligned} \tilde{\mathbf{C}}(s, \mathbf{g}_k^{\Omega_i}) &= \mathbf{C}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i}) \\ \tilde{\mathbf{G}}(s, \mathbf{g}_k^{\Omega_i}) &= (\hat{\alpha}_{2,k})^{-1} \mathbf{G}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i}) \\ \tilde{\mathbf{B}}(\mathbf{g}_k^{\Omega_i}) &= \mathbf{B}(\mathbf{g}_k^{\Omega_i}) \\ \tilde{\mathbf{L}}(\mathbf{g}_k^{\Omega_i}) &= \hat{\alpha}_{1,k} (\hat{\alpha}_{2,k})^{-1} \mathbf{L}(\mathbf{g}_k^{\Omega_i}). \end{aligned} \quad (13)$$

The scaling coefficients $\alpha_{1,k}(\mathbf{g}_j^{\Omega_i})$ can be extended to scaling matrices $\mathbf{\Gamma}_{1,k}(\mathbf{g}_j^{\Omega_i})$ such that a modified τ ROM $\hat{\mathbf{\Gamma}}_{1,k}^T \mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i}) \hat{\mathbf{\Gamma}}_{1,k}$ is obtained with

$$\begin{aligned} \tilde{\mathbf{C}}(s, \mathbf{g}_k^{\Omega_i}) &= \mathbf{C}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i}) \\ \tilde{\mathbf{G}}(s, \mathbf{g}_k^{\Omega_i}) &= (\hat{\alpha}_{2,k})^{-1} \mathbf{G}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i}) \\ \tilde{\mathbf{B}}(\mathbf{g}_k^{\Omega_i}) &= \mathbf{B}(\mathbf{g}_k^{\Omega_i}) \hat{\mathbf{\Gamma}}_{1,k} \\ \tilde{\mathbf{L}}(\mathbf{g}_k^{\Omega_i}) &= (\hat{\alpha}_{2,k})^{-1} \mathbf{L}(\mathbf{g}_k^{\Omega_i}) \hat{\mathbf{\Gamma}}_{1,k} \end{aligned} \quad (14)$$

where superscript “ T ” is the transpose operator. Even if the use of scaling matrices can provide more flexibility in the optimization step, we verified that the use of simple scaling coefficients leads to good results in terms of accuracy, while making the optimization faster and independent from the number of ports. In the numerical examples, scalar scaling and frequency shifting coefficients will be used.

C. Multivariate Interpolation

Passivity is crucial when the reduced model is utilized in a circuit simulator for transient analysis. Passive systems cannot generate more energy than they absorb through their electrical ports. When the system is terminated on any arbitrary passive loads, none of them will cause the system to become unstable. A linear network described by admittance matrix $\mathbf{Y}(s)$ is passive if [40]:

- 1) $\mathbf{Y}(s^*) = \mathbf{Y}^*(s)$ for all s , where “ $*$ ” is the complex conjugate operator;
- 2) $\mathbf{Y}(s)$ is analytic in $\Re e(s) > 0$;
- 3) $\mathbf{Y}(s)$ is a positive-real matrix, i.e.: $\mathbf{z}^* T (\mathbf{Y}^T(s^*) + \mathbf{Y}(s)) \mathbf{z} \geq 0; \forall s: \Re e(s) > 0$ and any arbitrary vector \mathbf{z} .

Similar results are valid for a linear network described by impedance matrix $\mathbf{Z}(s)$. The interpolated scaling and frequency shifting real coefficients $\alpha_{1,k}(\mathbf{g})$, $\alpha_{2,k}(\mathbf{g})$ have to satisfy the following conditions:

$$\alpha_{1,k}(\mathbf{g}) \geq 0 \quad (15a)$$

$$\alpha_{2,k}(\mathbf{g}) > 0 \quad (15b)$$

to preserve the passivity of the corresponding scaled and frequency shifted τ ROM. If scaling matrices $\mathbf{\Gamma}_{1,k}(\mathbf{g})$ are used instead of scaling coefficients $\alpha_{1,k}(\mathbf{g})$, no particular condition has to be satisfied by $\mathbf{\Gamma}_{1,k}(\mathbf{g})$. Multivariate interpolation schemes based on a class of positive interpolation operators [41] are used to parameterize $\alpha_{1,k}(\mathbf{g})$, $\alpha_{2,k}(\mathbf{g})$. These schemes are able to guarantee the passivity of each scaled and frequency-shifted τ ROM by satisfying the properties (15a)–(15b). The same positive multivariate interpolation schemes are used to interpolate the set of modified τ ROMs $\hat{\alpha}_{1,k}\mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}_k^{\Omega_i})$, $k = 1, \dots, Q$ at an input–output level, which results in a parameterized reduced model passive over the entire design space.

Multivariate interpolation can be realized by means of tensor product [42] or tessellation [38] methods. Any interpolation scheme based on a class of positive interpolation operators can be used.

In the bivariate case (s, g) , each interpolated function $\mathbf{T}(g)$, being in turn $\alpha_{1,k}(g)$, $\alpha_{2,k}(g)$, $\mathbf{R}(s, g)$, can be written as

$$\mathbf{T}(g) = \sum_{k=1}^{K_1} \mathbf{T}_{g_k} \ell_k(g) \quad (16)$$

where K_1 represents the number of τ ROMs vertices of a specific subdomain, and each interpolation kernel $\ell_k(g)$ is a scalar function satisfying the following constraints:

$$\ell_k(g) \geq 0 \quad (17)$$

$$\ell_k(g_i) = \delta_{k,i}. \quad (18)$$

A possible choice is to select $\ell_k(g)$ as in piecewise linear interpolation.

In the general multivariate case, multivariate interpolation methods that belong to the general class of positive interpolation operators can be used, e.g., the piecewise multilinear and multivariate simplicial methods [37]. We note that the interpolation kernel functions of these methods only depend on the design space grid points, and their computation does not require the solution of a linear system to impose an interpolation constraint. These positive interpolation schemes have already been used in [32]. In the proposed new PMOR technique, a powerful novelty is introduced by the interpolation process of a set of scaling and frequency shifting system coefficients, which increase the modeling capability of the proposed algorithm with respect to [32], where the interpolation process were only applied to the ROMs treated as input–output systems.

D. Passivity Preserving Interpolation

Under the assumptions that the original τ PEEC models are passive and the MOR method used to provide the τ ROMs is passivity-preserving, the proposed PMOR method is able to build

parameterized reduced models that are passive over the entire design space of interest.

A scaling coefficient α_1 or matrix $\mathbf{\Gamma}_1$ is applied at the input–output level of the system, while a frequency shifting coefficient α_2 is a compression or expansion term for the Laplace variable s . Considering admittance and impedance representations, if α_1, α_2 satisfy (15a)–(15b), passivity is preserved when these coefficients are applied to a passive system. It is straightforward to prove that if a nonnegative scalar coefficient is applied to a passive system, it preserves the three passivity conditions. A positive frequency shifting coefficient is a compression or expansion term for the Laplace variable s , and therefore if a system is passive in the s -domain, it is also passive in a compressed or expanded s -domain.

Once a set of scaled and frequency-shifted τ ROMs $\hat{\alpha}_{1,k}\mathbf{R}(s\hat{\alpha}_{2,k}, \mathbf{g}^{\Omega_i})$, $k = 1, \dots, Q$, which are passive, is built for each cell in the estimation design space grid, the next step of the proposed PMOR method is focused on combining together these τ ROMs by a multivariate interpolation scheme to obtain a parameterized ROM $\mathbf{R}(s, \mathbf{g})$ with overall passivity. Conditions 1)–2) are preserved in (16) and corresponding multivariate extensions, as they are weighted sums with real nonnegative weights of systems respecting these two conditions. Concerning Condition 3), we refer to the following theorem [43].

Theorem 1: Any nonnegative linear combination of positive real matrices is a positive real matrix.

Since (16) and the multivariate extensions are weighted sums with real nonnegative weights of passive systems, condition 3) is satisfied by construction over the entire design space of interest. We have proven that all the three passivity conditions for admittance (and impedance) representations are preserved in our PMOR algorithm.

E. Method Complexity and Parallelization

As shown in Fig. 2, the proposed method is composed of the following four steps with a related complexity.

- 1) The construction of the τ ROMs is performed using the MOR method [25].
- 2) The computation of scaling and frequency-shifting coefficients is performed using optimization routines. We recall that using scalar scaling and frequency-shifting coefficients leads to a good accuracy of parameterized ROMs, while keeping the optimization process computationally cheap and fast.
- 3) The multivariate interpolation of scaling and frequency-shifting coefficients uses positive local interpolation methods with interpolation kernel functions that only depend on the estimation grid points.
- 4) Comments similar to those of the previous point are valid for the multivariate interpolation of scaled and shifted τ ROMs. We recall that the estimation design space grid is divided into cells using hyperrectangles (regular grids) [37] or simplices (regular and scattered grids) [38]. Each cell is defined by a specific number of vertices that are used to perform interpolation in the cell itself.

All four steps of the proposed technique have a certain degree of parallelism, as follows.

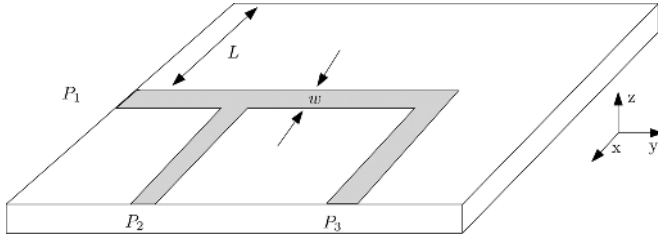
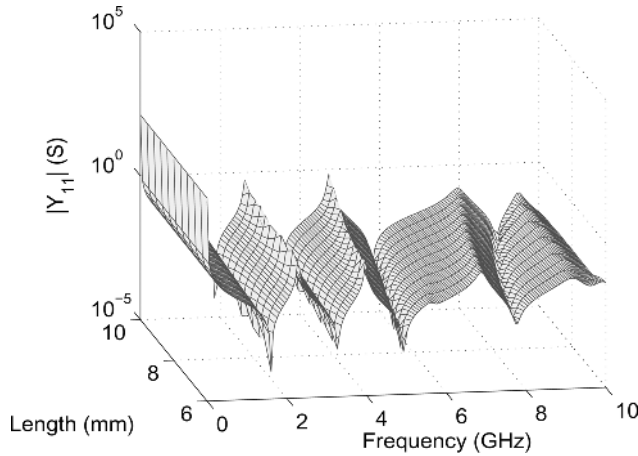


Fig. 5. Structure of the three-port microstrip power-divider circuit.

TABLE I
PARAMETERS OF THE THREE-PORT MICROSTRIP POWER-DIVIDER CIRCUIT

Parameter	Min	Max
Frequency (<i>freq</i>)	100 kHz	10 GHz
Length (<i>L</i>)	6 mm	10 mm

Fig. 6. Magnitude of the bivariate τ ROM of $\mathbf{Y}_{11}(s, L)$.

- 1) Since each τ ROM is independent from the others, the construction of a single τ ROM can be performed in parallel with the others.
- 2) The computation of scaling and frequency-shifting coefficients can be parallelized cell by cell.
- 3) A parallelization cell by cell is also feasible for the multivariate interpolation of scaling and frequency-shifting coefficients.
- 4) Comments similar to those of the previous point are valid for the multivariate interpolation of scaled and shifted τ ROMs.

IV. NUMERICAL RESULTS

This section presents two numerical examples that validate the proposed PMOR method. Let us define the weighted RMS-error as

$$\text{Err}(\mathbf{g}) = \sqrt{\frac{\sum_{i=1}^{N_i N_o} \sum_{k=1}^{K_s} |w_{Y_i}(s_k, \mathbf{g})(Y_{r,i}(s_k, \mathbf{g}) - Y_i(s_k, \mathbf{g}))|^2}{N_i N_o K_s}} \quad (19)$$

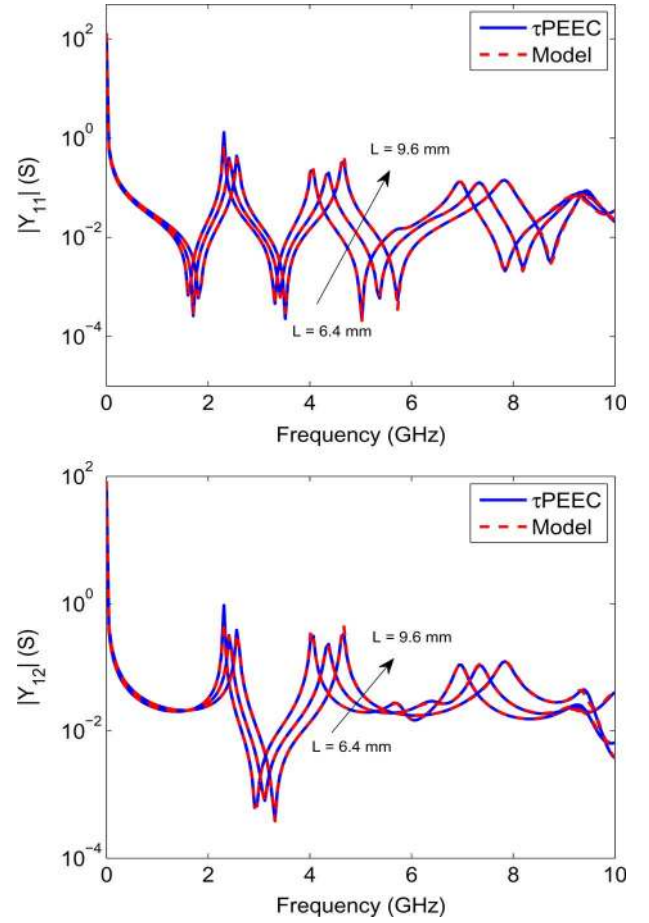
Fig. 7. Magnitude of the bivariate τ ROMs of $\mathbf{Y}_{11}(s, L)$ and $\mathbf{Y}_{12}(s, L)$ ($L = \{6.4, 8, 9.6\}$ mm).

TABLE II
(ORDER, DELAYS) OF τ PEEC MODELS AND τ ROMS

	τ PEEC models	τ ROMs
L_1	(2667,732)	(60,732)
L_2	(2667,730)	(60,730)
L_3	(2667,727)	(60,727)
L_4	(2667,730)	(60,730)
L_5	(2667,742)	(48,742)
L_6	(2667,737)	(60,737)

with

$$w_{Y_i}(s, \mathbf{g}) = |(Y_i(s, \mathbf{g}))^{-1}| \quad (20)$$

where N_i and N_o are the number of input and output system ports, respectively, and K_s is the number of frequency samples. The worst case RMS error over the validation grid is chosen to assess the accuracy and the quality of parameterized ROMs

$$\mathbf{g}_{\max} = \underset{\mathbf{g}}{\text{argmax}} \text{Err}(\mathbf{g}), \mathbf{g} \in \text{validation grid} \quad (21)$$

$$\text{Err}_{\max} = \text{Err}(\mathbf{g}_{\max}) \quad (22)$$

behavior of the system, and it is used in the numerical examples. The proposed PMOR algorithm was implemented in Matlab R2009A [39], and all experiments were carried out on Windows

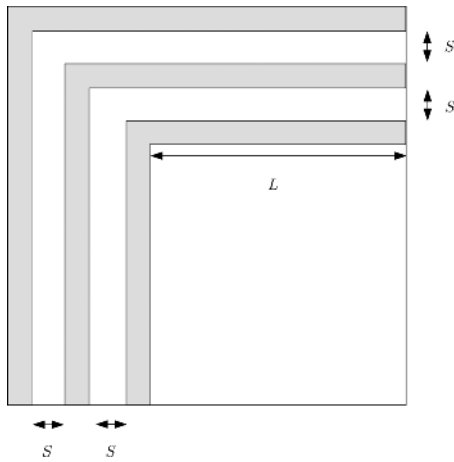
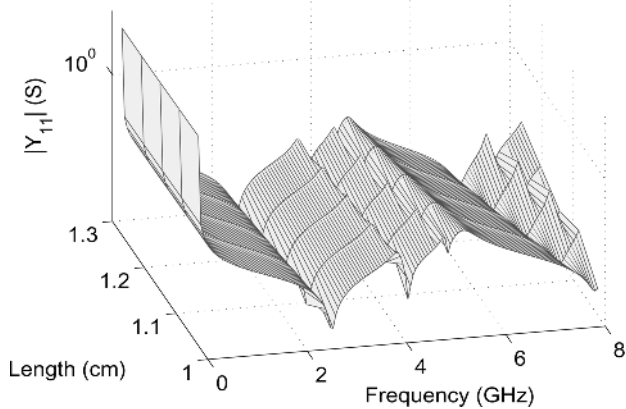
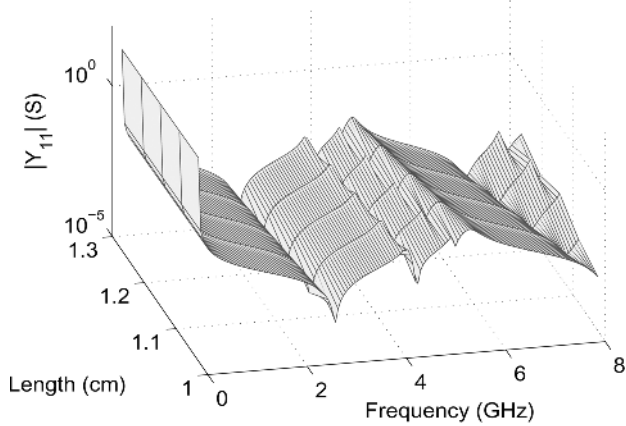
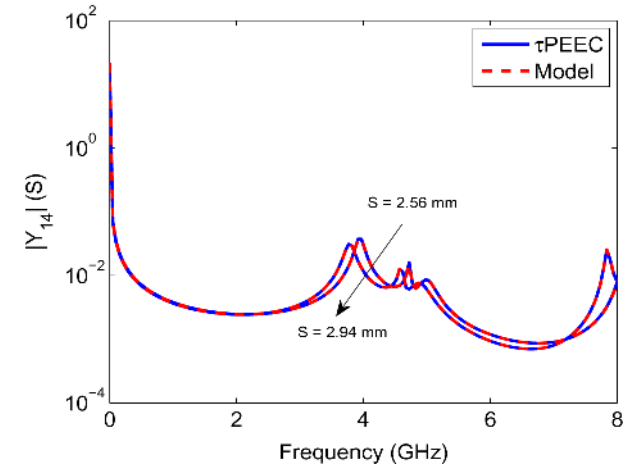
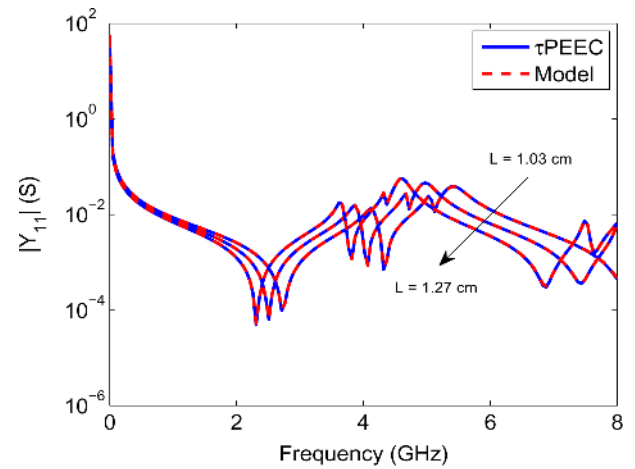


Fig. 8. Structure of the three bends.

 TABLE III
 PARAMETERS OF THE THREE BENDS

Parameter	Min	Max
Frequency (<i>freq</i>)	100 kHz	10 GHz
Length (<i>L</i>)	1 cm	1.3 cm
Spacing (<i>S</i>)	2.5 mm	3 mm


 Fig. 9. Magnitude of the trivariate τ ROM of $\mathbf{Y}_{11}(s, L, S)$ for $S = 2.56$ mm (top) and $S = 2.94$ mm (bottom).

 Fig. 10. Magnitude of the trivariate τ ROMs of $\mathbf{Y}_{11}(s, L, S)$ for $L = \{1.03, 1.15, 1.27\}$ mm, $S = 2.69$ mm and $\mathbf{Y}_{14}(s, L, S)$ for $L = 1.15$ mm, $S = \{2.56, 2.94\}$ mm.

A. Three-Port Microstrip Power-Divider Circuit

A three-port microstrip power-divider circuit [44], [45] has been modeled in this example. The structure is shown in Fig. 5. The dimensions of the circuit are $[20, 20, 0.5]$ mm in the $[x, y, z]$ directions, and the width of the microstrips is equal to 0.8 mm. The relative dielectric constant is equal to $\epsilon_r = 2.2$. A bivariate τ ROM is built as a function of the length L in addition to frequency. Their corresponding ranges are shown in Table I.

The τ PEEC method is used to compute the $\mathbf{C}, \mathbf{G}, \mathbf{B}, \mathbf{L}$ matrices in (1a)–(1b) for 11 values of L . Then, we have built reduced models for six values of L by means of the MOR algorithm described in [25]. Table II shows the order and the number of delays of these 6 τ PEEC models and corresponding τ ROMs.

A bivariate τ ROM is obtained using the proposed PMOR method with the piecewise linear interpolation scheme. Fig. 6 shows the magnitude of the parameterized reduced model of $\mathbf{Y}_{11}(s, L)$. Fig. 7 shows the magnitude of $\mathbf{Y}_{11}(s, L)$ and $\mathbf{Y}_{12}(s, L)$ for the length values $L = \{6.4, 8, 9.6\}$ mm. These specific L values have not been used in the τ ROMs generation process; nevertheless, an excellent agreement between reduced and PEEC models can be observed. The worst case RMS-error defined in (22) is equal to 0.08. As clearly seen, the parameterized reduced model captures very accurately the b

TABLE IV
(ORDER, DELAYS) OF τ PEEC MODELS AND τ ROMs

	τ PEEC models	τ ROMs
(L_1, S_1)	(2124,510)	(64,510)
(L_2, S_1)	(2124,508)	(64,508)
(L_3, S_1)	(2124,510)	(64,510)
(L_4, S_1)	(2124,511)	(64,511)
(L_5, S_1)	(2124,517)	(64,517)
(L_6, S_1)	(2124,519)	(64,519)
(L_1, S_2)	(2124,509)	(64,509)
(L_2, S_2)	(2124,511)	(64,511)
(L_3, S_2)	(2124,513)	(64,513)
(L_4, S_2)	(2124,512)	(64,512)
(L_5, S_2)	(2124,519)	(64,519)
(L_6, S_2)	(2124,520)	(80,520)
(L_1, S_3)	(2124,511)	(64,511)
(L_2, S_3)	(2124,514)	(64,514)
(L_3, S_3)	(2124,518)	(64,518)
(L_4, S_3)	(2124,515)	(64,515)
(L_5, S_3)	(2124,518)	(64,518)
(L_6, S_3)	(2124,522)	(64,522)
(L_1, S_4)	(2124,511)	(64,511)
(L_2, S_4)	(2124,512)	(64,512)
(L_3, S_4)	(2124,519)	(64,519)
(L_4, S_4)	(2124,519)	(64,519)
(L_5, S_4)	(2124,519)	(64,519)
(L_6, S_4)	(2124,527)	(64,527)
(L_1, S_5)	(2124,510)	(64,510)
(L_2, S_5)	(2124,513)	(64,513)
(L_3, S_5)	(2124,518)	(64,518)
(L_4, S_5)	(2124,520)	(64,520)
(L_5, S_5)	(2124,519)	(80,519)
(L_6, S_5)	(2124,524)	(64,524)

B. Three Bends

Three bends over a ground plane with differential ports have been modeled in this example. The structure is shown in Fig. 8. The width of the bends and the distance of the bends from the ground plane are equal to 1 mm. A trivariate τ ROM is built as a function of the length L and the spacing between the conductors S in addition to frequency. Their corresponding ranges are shown in Table III.

The τ PEEC method is used to compute the \mathbf{C} , \mathbf{G} , \mathbf{B} , \mathbf{L} matrices in (1a)–(1b) over the estimation grid (six values of L and five values of S) and the validation grid (five values of L and four values of S), as shown in Fig. 3. Then, we have built τ ROMs in the estimation grid by means of the MOR algorithm described in [25]. Table IV shows the order and the number of delays of these 30 τ PEEC models and corresponding τ ROMs. A trivariate τ ROM is built by means of the presented PMOR approach with the piecewise multilinear interpolation scheme. Fig. 9 shows the magnitude of the parameterized reduced model of $\mathbf{Y}_{11}(s, L, S)$ for the spacing values $S = \{2.56, 2.97\}$ mm. Fig. 10 shows the magnitude of $\mathbf{Y}_{11}(s, L, S)$ for the length and spacing values $L = \{1.03, 1.15, 1.27\}$ mm, $S = 2.69$ mm and of $\mathbf{Y}_{14}(s, L)$ for the length and spacing values $L = 1.15$ mm, $S = \{2.56, 2.94\}$ mm. Even if these specific L and S values have not been used in the τ ROMs generation process, the model accurately describes the system under study in these design space points. The worst case RMS-error defined in (22) is equal to 0.065. The parameterized reduced model is able to accurately describe the parameterized behavior of the system.

V. CONCLUSION

We have presented a new PMOR technique applicable to NDDE systems. It is based on an efficient and reliable combination of univariate model order reduction methods, a procedure to find scaling and frequency shifting coefficients and positive interpolation schemes. An innovative passivity-preserving interpolation of reduced systems at an input-output level is proposed, which allows accurately modeling of highly dynamic systems. The proposed method does not have to deal with multiparameter moment computation and related issues. The expansion points are chosen only along the frequency axis using an adaptive algorithm. Under the assumptions that the original τ PEEC models are passive and the MOR method used to provide the τ ROMs is passivity-preserving, the proposed PMOR method is able to build parameterized reduced models that are passive over the entire design space of interest. Numerical examples have validated the proposed PMOR approach on practical application cases, showing that it is able to build very accurate parameterized τ ROMs of dynamic EM systems.

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