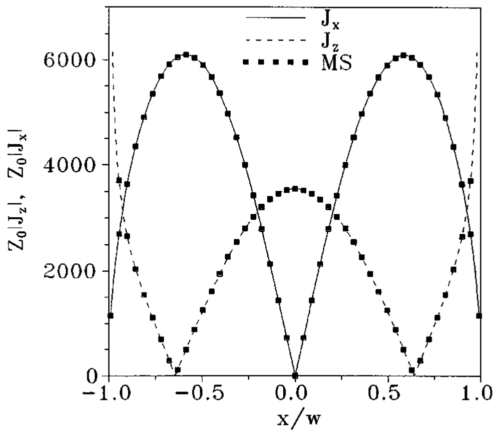

Fig. 4. Modal currents of the third mode of Table II for $f = 30$ GHz ($\beta/k_0 = 1.404470$).

been derived by using method A and reproduced exactly by method B as well. As a further test of the correctness of these results, we also derived analogous curves pertaining to the modified microstrip line (labeled "MS"), with conducting sidewalls, referred to earlier in connection with Table II. As a matter of fact, the curves pertaining to this modified structure were found to be indistinguishable from those obtained earlier by methods A and B.

V. Conclusion

Two independent DSIETs have been used for the exact full-wave analysis of layered microstrip lines. The proposed algorithms combine the simplicity of conventional MoMs with extremely high accuracy both for the propagation constants and modal currents on the strip. In filling up the matrix elements, only rapidly converging real-axis spectral integrals are encountered.

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MIM Shunt-Capacitor Model Using Black Boxes of EM-Simulated Critical Parts

Gregor Gerhard and Stefan Koch

***Abstract*—A new model for metal–insulator–metal shunt capacitors is introduced in this paper. The main difference between the new model and known models is that critical parts of the capacitor's geometry are represented by black boxes. These boxes contain S-parameter files generated with an electromagnetic field solver. The capacitor parts, which depend on the capacitance value, are represented by microstrip and lumped elements. The new model combines the advantages of field simulations with those of lumped- or microstrip-based models. It can easily be used in circuit simulators utilizing their features for design development such as optimizations. The model is compared with two shunt capacitors on microwave monolithic integrated circuits to show the excellent fit.**

***Index Terms*—Capacitor, EM simulation, MIM, model.**

I. Introduction

Metal–insulator–metal (MIM) shunt capacitors are key elements in many microwave and millimeter-wave monolithic integrated circuits (MMICs). DC blocks, matching sections, and biasing circuitry widely utilize this component because of its small space requirement. An accurate model of the structure is, therefore, crucial for any MMIC design. Many monolithic foundries have developed their own proprietary models by means of parameter-extraction methods from experimental data. Other approaches, closer to the physical structure of the MIM capacitor, have been previously presented for series MIM capacitors, resulting in a distributed [1] or lumped [2] equivalent model. Other models calculate the equivalent-circuit parameters with complex formulas based on S-parameter matrices to consider several layout cases [3]. More universal models, as in [4], naturally show less accuracy than

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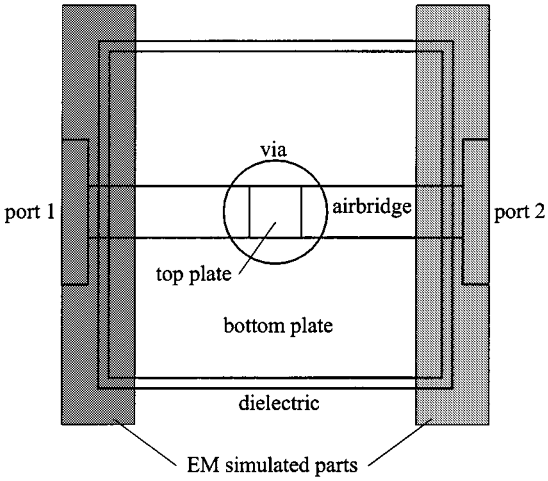

Fig. 1. Top view of a typical shunt capacitor.

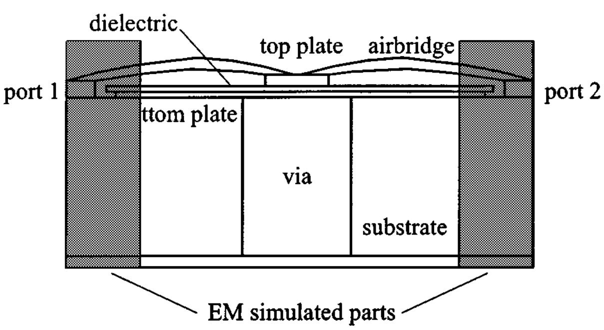

Fig. 2. Side view of a typical shunt capacitor.

those developed for very special cases. Also, some fundamental investigations on capacitor calculations utilizing analytic formulas can be found in the literature [5]. Most of these models are designed for series MIM capacitors. This might be due to the fact that, for MMICs, only a few foundries allow to place capacitors directly over a via to get a compact shunt capacitor. On the other hand, if offered, such a shunt capacitor is a very space-saving element and, therefore, used by many designers. From a modeling point-of-view such capacitors over vias are much harder to model than series capacitors. Small value capacitors are especially very difficult to model because the via impedance comes into the same range as the capacitor impedance, and the fringing capacitance plays a major role. Additionally, the complex geometrical structure of a shunt capacitor makes it difficult to find a model valid for the whole Ka-band. Therefore, shunt-capacitor models are rarely found in literature.

In Section II, a new shunt-capacitor model will be introduced, which is designed close to the physical structures. The critical parts are simulated with the electromagnetic (EM) simulator IE3D [6], which is based on the method of moments. These parts do not change with the capacitance value and, therefore, can be represented in the model by black boxes.

II. New Shunt-Capacitor Model

The new shunt-capacitor model introduced here is designed close to the physical dimensions of the shunt capacitor. Both the capacitor's bottom plate and the capacitor's top plate including the air bridges and ports are modeled. During the modeling process, it turns out that the transition from the ports to the air bridges and the bottom plate is very critical and hard to model. For this reason, the transition is simulated with an EM simulator. The EM simulator uses the method of moments. The EM simulated part includes the port, a small part of the bottom plate, and the appropriate part of the air bridge (Figs. 1 and 2). The achieved S-parameters are put into the model by two black boxes. It is very important for the usefulness of the model that the EM simulated parts do not change when varying the capacitance value by changing the capacitor's top plate size.

Between the two black boxes with the EM simulated parts, the shunt capacitor is modeled with microstrip and lumped elements (Fig. 3). All the microstrip-element dimensions are chosen with identical dimensions as in the existing geometry. The air bridges are modeled by transmission lines considering that the substrate parameters are those of a transmission line in air with a several micrometer thick air substrate. This means that, for the air bridge, the dimensions of the bottom plate are infinite. The cap top, which physically determines the capacitance value, is modeled with a transmission line. In the middle of the transmission line, a capacitor is placed with a resistor in line leading to the bottom plate. The bottom plate is modeled by a transmission line and a linear tapered transmission line leading from the black boxes to the capacitor. The inner bottom plate tapered transmission lines are due to the fact that the current density near the via behaves approximately like that of a tapered transmission line and the upper via dimensions are much smaller than those of the bottom plate. Between the two tapered transmission lines, an inductance and a series resistor is connected to ground. These two lumped elements represent the via.

The new model shall be compared with measurements of two shunt capacitors of the above-described type with different capacitance values. The first shunt capacitor has a capacitance value of 700 fF (Cap A), the second shunt capacitor has a capacitance value of 300 fF (Cap B). Both shunt capacitors are realized on a 100-μm-thick GaAs substrate. Due to the symmetry and reversibility of the capacitor and its model, S_{11} is equal to S_{22} and S_{21} is equal to S_{12}. These S-parameters are shown in Figs. 4 and 5. It can be seen that the simulated data is very close to the measured data. To get a closer look on the really small deviations, error vectors have been calculated (Fig. 6). The error vector connects a frequency point of the measured data with the corresponding frequency point of the model in the Smith chart. Due to the fact that the Smith chart's radius is 1.0, the magnitude of the error vector times 100 can be interpreted as percentage deviation. The error vectors for both capacitors are shown in Fig. 7. It can be seen that the magnitudes of error vectors of S_{21} are smaller than 0.01 nearly in the whole frequency range from dc to 45 GHz. The magnitudes of the error vectors of S_{11} increase slightly with frequency, but are smaller than 0.05 up to 44 GHz. It should be pointed out that nothing is fitted to obtain these results. To get this excellent agreement, only the capacitance value and geometrical properties are set according to the real physical dimensions.

III. Conclusion

A new model based on microstrip elements and EM simulated parts for a MIM shunt capacitor has been presented in this paper. The model fits very well to the measured data over the range under study, from dc to 45 GHz. Even for very small and, therefore, hard to model capacitance values, such as the 300-fF shunt capacitor investigated here, the model compares very well to the measurement. The greatest advantage of the new model is that only the capacitance value and geometry data

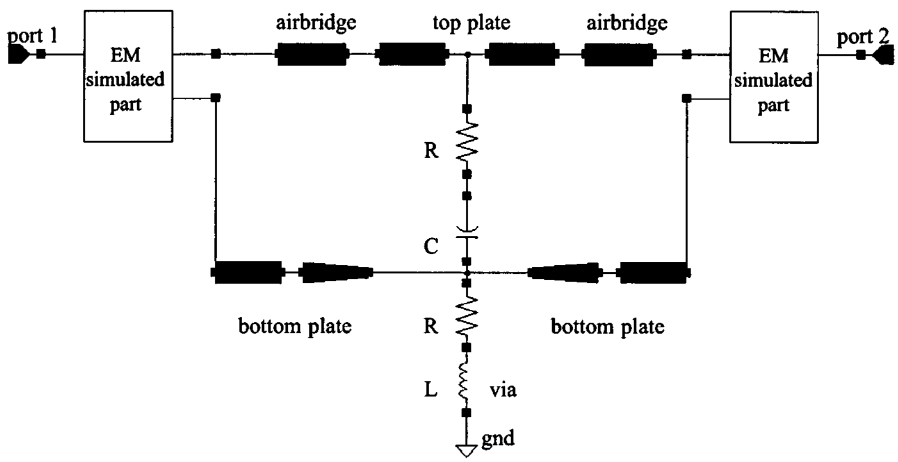

Fig. 3. Schematic of the new shunt-capacitor model using black boxes with EM simulated parts, lumped elements, and distributed elements.

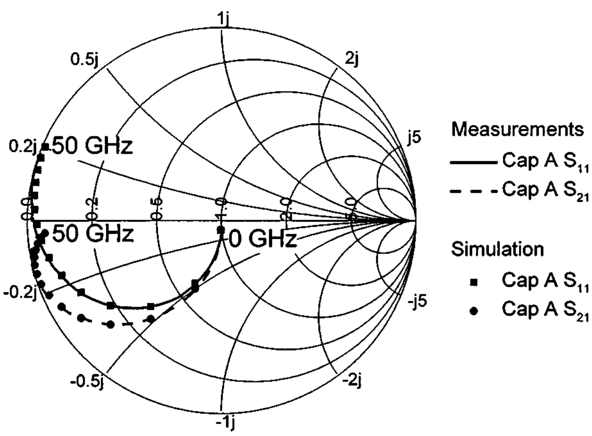

Fig. 4. Comparison of measurements and simulated data for capacitor Cap A.

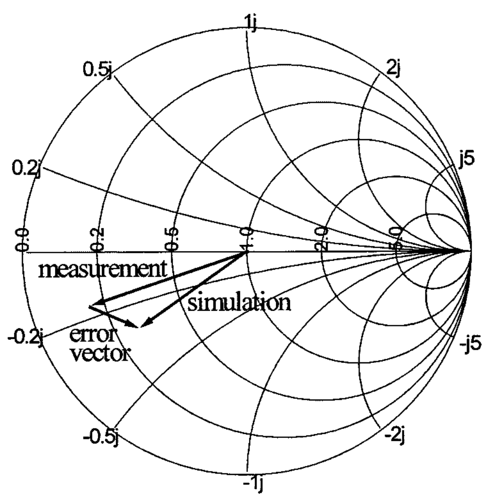

Fig. 6. Error vector is the difference between a frequency point of the measured and simulated data.

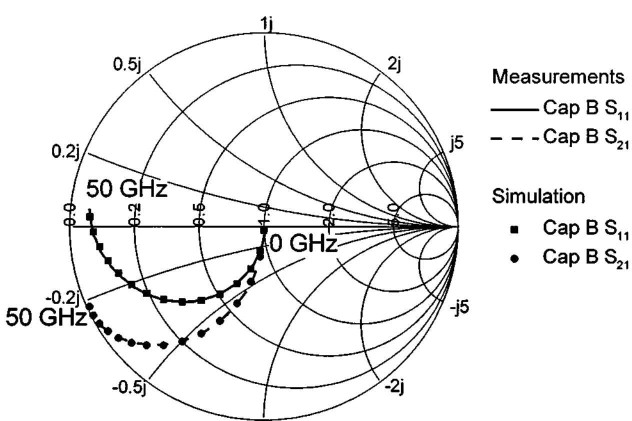

Fig. 5. Comparison of measurements and simulated data for capacitor Cap B.

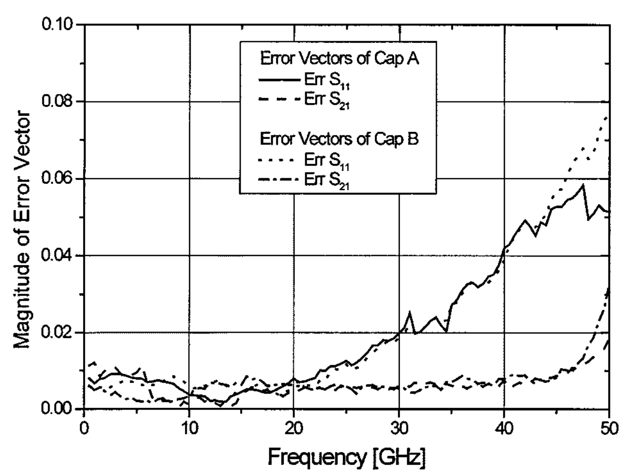

Fig. 7. Magnitude of the error vector of S_{11} and S_{21} for both capacitors.

have to be adapted to the real capacitor layout. No other model parameters have to be tuned, as it has to be done in the case of fitted models. Compared with pure EM-field simulations, the new model's advantage is that it can easily be used in circuit simulators utilizing their features for design development.

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A Simple Procedure for Impedance Matching and Tuning of Microwave Couplers for an Electron Linear Accelerator

Marcos A. R. Franco, Valdir A. Serrão, Celso Fuhrmann, and Sílvio B. Herdade

***Abstract*—A simple experimental procedure to match and tune "doorknob"-type microwave couplers is presented in this paper. The procedure is suitable for accelerating structures with both input and output couplers and allows a fast convergence to the minimum reflection condition for a cavity coupler with fixed phase shift. The standing-wave ratio and the coupling cavity phase shift as functions of the coupler dimensions and frequency are also reported.**

***Index Terms*—Accelerator RF systems, electron linear accelerators, impedance matching, microwave measurements, waveguide couplers.**

I. Introduction

In an RF electron linear accelerator, the injected beam is accelerated by the longitudinal electric field of the pulsed electromagnetic wave, which propagates along the symmetry axis of an accelerating periodic structure. The microwave power is carried by rectangular waveguides in the fundamental propagation mode TE_{10}, while in the periodic structure, the fundamental mode is the TM_{01}. Therefore, to minimize reflections at the waveguide-to-accelerating structure junction, an impedance-matching and mode-transforming device is required. The device, called a "microwave coupler," must minimize the standing-wave ratio (SWR) between the waveguide and coupler cavity without changing the phase of the wave launched along the accelerating structure. For a safe high-power amplifier operation, only SWR values less than 1.1 are tolerated.

Two basic procedures to match and tune the coupling system are presented in the literature. First, the modified nodal-shift method (Gallagher's method) is used to match the coupler, while the phase between the cavities of the periodic structure is checked by the nodal-shift method [1]–[3]. A second approach is the time-domain reflectometry method [4], [5] in which the coupler is fed by a pulsed microwave signal, with width shorter than the structure filling time. The time difference between the reflected waves from the input and output couplers appears because of the time required for the microwave to travel through the structure. The matching adjustments are performed by varying the geometrical parameters of the couplers. After this step, the nodal-shift method is used to check the final tuning of the structure with the couplers.

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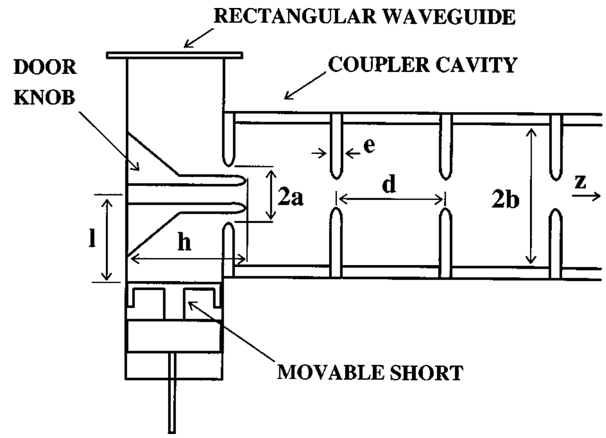

Fig. 1. Schematic view of the doorknob microwave coupler and disk-loaded waveguide structure, where **e** is the disk thickness, **d** is the cavity length, **2b** is the cylindrical cavity diameter, **2a** is the disk-hole diameter, **h** is the doorknob height, and **l** is the movable short position.

In this paper, an experimental procedure to match and tune two doorknob-type couplers with a 1300-MHz $2\pi/3$-mode disk-loaded structure is presented [6]. This procedure, which includes the optimal characteristics of the two previously discussed methods, is a simple and fast way to improve the matching and tuning of the couplers.

II. Doorknob Microwave Coupler

In the doorknob-type coupler, a cylindrical cone and movable short are used to match the impedance between the waveguide and accelerating structure (Fig. 1). The doorknob coupler permits a simple mechanical adjustment and, because of the cylindrical symmetry, it does not introduce transversal asymmetries in the axial electric field.

To match and tune a doorknob coupler, it is necessary to adjust three geometric parameters: the disk-hole diameter **(2a)** of the coupler cavity, the doorknob height **(h)**, and the movable short position **(l)**.

III. Experimental Methods and Procedure

Three complementary experimental methods were used to tune and match the couplers: the time-domain reflectometry method to determine approximate values of the geometric parameters (**h**, **l**, **2a**); the nodal-shift method to measure the phase shift between the cavities of the accelerating structure, and the Gallagher's method to obtain more accurate measurements of the SWR values.

In the reflectometry method, a microwave pulse is injected into the assembly composed by the doorknob couplers and the accelerating structure (Fig. 2). Although the reflections can be reduced (SWR values near 1.3) by varying the parameters **h**, **l**, and **2a**, the phase shift cannot be measured. The values of **h**, **l**, and **2a** determined by this method are the starting data to get more accurate measurements and to really minimize the SWR.

In Gallagher's method, the SWR is measured only for the coupler in which the microwave pulse is injected. These SWR measurements are