# A Short-Open Deembedding Technique for Method-of-Moments-Based Electromagnetic Analyses

Marco Farina, *Member, IEEE*, and Tullio Rozzi, *Fellow, IEEE*

***Abstract*—A short-open calibration (SOC) technique for deembedding structures with an arbitrary, possibly coupled, number of ports is introduced in this paper. While deembedding algorithms used in commercial software packages require the analysis of two "standard" structures for each set of ports, the proposed solution requires only one standard to be analyzed, with a significant reduction in the overall computation time. Moreover, unlike other deembedding techniques, the SOC technique does not rely on specific assumptions about the nature of the port discontinuities and of the feeding lines. This fact circumvents ambiguities linked to the definition of the characteristic impedance when hybrid modes are involved. Implementation-ready formulas are provided.**

***Index Terms*—Calibration, electromagnetic analysis, moment methods.**

## I. Introduction

THE method of the moments (MoM) is a widely employed tool for the analysis of planar structures [1]–[3]. In fact, several three-dimensional (3-D) commercial software packages exploit the MoM in a spectral-domain framework. Recent advances have also highlighted the ability of some MoM approaches to model active devices [4].

A common problem in 3-D MoM techniques is to select an appropriate excitation mechanism and, hence, to correctly extract the network parameters for the structure being modeled. To this aim, two major excitation schemes were proposed, namely, the traveling-wave excitation [5] and the delta-gap source excitation [6]. The first scheme has also been recently used in order to characterize planar structures involving lossy "thick" conductors [7]; however, due to its simplicity, flexibility, and inherent reliability, the delta-gap source excitation is the most popular technique, generally adopted by the majority of the commercial packages.

In this technique, an impressed electric field at some position along the feeding line identifies a port, leading to an explicit integral equation to be solved by the MoM. Its solution provides the current distribution and, the electric field at the ports being known, the network parameters for the desired ports. Often, the delta-gap source is placed near a ground wall so as to have a ground-referenced network description: this is quite a natural way to deal with boxed structures, but in [3], image theory was used in order to analyze open structures in a similar way.

The impressed source is responsible for a port discontinuity: in [8], it was shown that such a discontinuity in a lossless structure behaves much like a 0.1-pF shunt capacitance in a 50-$\Omega$ microstrip line. More generally, the port discontinuity has to be modeled as an error network superimposed to the circuit being analyzed. This error network does not fundamentally alter results for electrically large structures like filters or resonators, but it does strongly affect results for small discontinuities, e.g., as shown in [9].

The problem of evaluating and removing the error network in RF measurements is addressed by deembedding algorithms, such as the thru-reflection-line (TRL) technique, where by putting well-behaved calibration elements—the "standards"—in place of the device-under-test (DUT), the error networks may be evaluated. Generally, error networks may also include parts of the feeding lines, this allows specifying a given reference plane for the network parameters being measured.

In [8], a specific deembedding algorithm for electromagnetic (EM) numerical tools was introduced. As indicated, in the product documentation of Sonnet's **em** or AWR's Microwave Office, two "standards" are analyzed for each set of ports. The first standard is a set of lines having the same length as the distance between the reference plane and the delta-gap source, whereas the second set is the same, but of double length. A maximum of four sets of ports are defined, one for each side of the enclosing box, leading to a maximum of eight standards. Note that each "standard" requires a separate structure to be analyzed.

In [9], a novel short-open calibration (SOC) scheme was introduced, taking advantage of an important feature of the MoM approach, namely the knowledge of the current distribution at a given plane that may even be physically inaccessible. This technique requires just *one* standard for each port set. This leads to a significant reduction of the computation time in the deembedding procedure. Note that the computation time in the deembedding procedure of some structures, like those involving "thick" and lossy conductors, may be rather large.

A key feature of the SOC scheme consists in not relying on any specific assumption about the nature of the port discontinuity and feeding lines: as compared with existing deembedding algorithms, the SOC does not require defining a characteristic impedance [10] either, which is still a subject of debate when hybrid modes are involved, as for microstrip and coplanar structures.

Unfortunately, in [9], only expressions for two-port error networks were derived. While uncoupled multiports may also be treated by defining a separate calibration standard for each port, no coupled ports were dealt with.

Manuscript received July 26, 2000; revised October 25, 2000.

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Publisher Item Identifier S 0018-9480(01)02425-5.

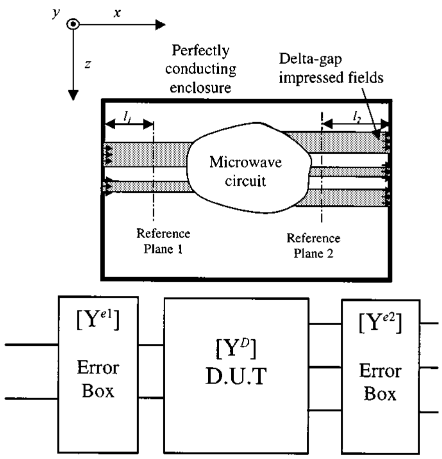


Fig. 1. Physical layout and network representation of a general deembedding problem.

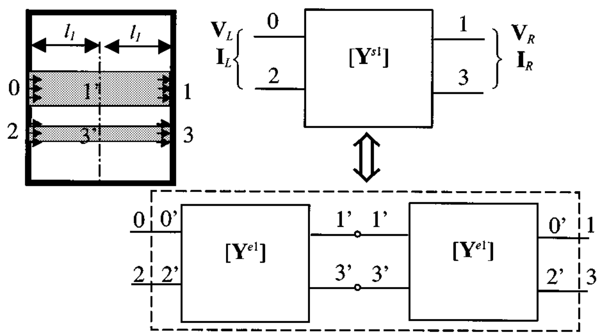


Fig. 2. “Standard” and its network representation $(\mathbf{Y}^{s1})$ in order to calculate the error network $\mathbf{Y}^{e1}$.

The aim of this paper is to present general expressions leading to a generalized SOC scheme for MoM-based approaches, which would be able to process an arbitrary number of ports, including coupling effects.

## II. Theory

The problem, as well as its network representation, is shown in Fig. 1. The objective is to calculate the network parameters $\mathbf{Y}^e$ of the error networks in order to obtain the network parameters $\mathbf{Y}^D$ of the DUT, once simulated the whole circuit, including parasitics and unwanted possibly coupled and lossy transmission lines.

Let us consider the left-hand-side ports. According to the SOC technique, the only “standard” to be analyzed is the one shown in Fig. 2, namely, a set of coupled lines having length $2l_1$. It should be stressed that the algorithms generally adopted in MoM-based software packages require the additional analysis of the same structure, but involving line lengths $l_1$.

Hence, the first step of the deembedding algorithm is to determine the error network parameters $\mathbf{Y}^e$ from the network parameters of the standard, say, $\mathbf{Y}^s$. The problem for two coupled ports is depicted in Fig. 2, where primes have been used for the error port numbering, while unprimed numbers identify the standard ports.

The standard network $\mathbf{Y}^s$ is symmetric, while all networks are reciprocal. Fig. 3 shows how this network is related to the unknown error network $\mathbf{Y}^e$, highlighting two internal nodes where currents $I_0$ and $I_1$ may be calculated. Note that these “internal” ports may be accessed by MoM, corresponding to the currents in the symmetry plane of the standard—this feature allows saving one standard.

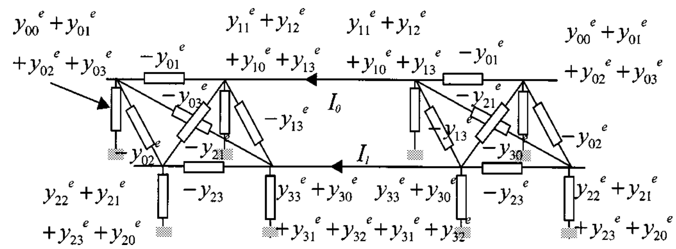


Fig. 3. Details of how the standard network is related to the error networks.

In order to lighten the notation, primes will not be used when writing the elements of $\mathbf{Y}^e$ in the remainder of this paper. The reader should consider as “primed ports” those referred to $\mathbf{Y}^e$. This consideration also applies to Fig. 3.

Let us apply a set of odd unitary excitations to $\mathbf{Y}^s$, i.e.,

$$\nu_k = (-1)^k, \qquad k = 0, 1, \ldots, (N-1) \tag{1}$$

with $N$ being the total number of ports belonging to the standard, $\nu$ being the amplitude of the delta-gap excitation, and $k$ being the port where $\nu$ is applied. This way, an ideal “short” at the symmetry plane is obtained. The middle nodes of the network in Fig. 3 are at ground potential and, consequently, the currents at this plane, accessible to the MoM by integration of the expansion function along the middle plane of the structure, are

$$-I_0^O = y_{01}^e + y_{21}^e \quad -I_1^O = y_{23}^e + y_{03}^e. \tag{2}$$

In (2), the superscript $O$ is used to remind the $I^O$ they are currents calculated under the hypothesis of odd excitation.

Let us apply a second set of odd excitations, differing from the previous one only for the fact that the amplitudes at ports 2 and 3 are $+2$ and $-2$, respectively. More generally, we may indicate a similar set of odd excitations as

$$\begin{aligned} v_k^{(q)} =& (-1)^k, \qquad k = 0, 1, \ldots, N-1; \quad k \neq 2q, 2q+1 \\ =& 2(-1)^k, \qquad k = 2q, 2q+1 \\ q =& 1, \ldots, \frac{N-2}{2} \end{aligned} \tag{3}$$

where $q$ is an auxiliary index introduced to identify at which port of $\mathbf{Y}^s$ excitations of amplitude $+2$ and $-2$ are applied.

According to this notation, the “short-circuit” currents at the internal ports corresponding to excitation $v_k^{(1)}$ are given by

$$-I_0^{O(1)} = y_{01}^e + 2y_{21}^e \quad -I_1^{O(1)} = 2y_{23}^e + y_{03}^e. \tag{4}$$

By combining (2) and (4), one obtains

$$\begin{aligned} y_{21}^e =& -I_0^{O(1)} + I_0^O \\ y_{01}^e =& -I_0^O - y_{21}^e \\ y_{23}^e =& -I_1^{O(1)} + I_1^O \\ y_{03}^e =& -I_1^O - y_{23}^e. \end{aligned} \tag{5}$$

More generally, for any number of ports, it can be verified that

$$\begin{aligned} y^e_{2q,2k+1} &= y^e_{2k+1,2q} = y^e_{2k,2q+1} = -I_k^{O(q)} + I_k^O, \qquad k \neq q \\ y^e_{2k,2k+1} &= y^e_{2k+1,2k} = -I_k^O - \sum_{l\neq k} y^e_{2l,2k+1}, \qquad k = q \\ &\text{for } k = 0, 1, \ldots, \frac{N-2}{2} \quad q = 1, \ldots, \frac{N-2}{2}. \end{aligned} \tag{6}$$

By using (6), the whole set of cross admittances linking the internal nodes to the external ones are obtained.

It actually is not necessary to calculate the responses to excitations (1) and (3), as we are assuming to have calculated the parameters of the whole standard, namely, $\mathbf{Y}^s$, and the excitation sets (1) and (3) are not independent from the ones involved in the computation of $\mathbf{Y}^s$. In fact, when computing $\mathbf{Y}^s$, a set of excitations is applied so that

$$y^s_{ij} = I_i\Big|_{v_k = \begin{cases} 1, & \text{if } k = j \\ 0, & \text{otherwise} \end{cases}}, \qquad k, i, j = 0, \ldots, N-1.$$

The current at the internal ports may be obtained and stored during this computation. If we define $i_l^{(j)}$ as the current at the internal port $l$ when just a unitary voltage gap is applied to the standard port $j$, the internal currents required by (6) are just a linear superposition of the $i_l^{(j)}$ as follows:

$$\begin{aligned} I_l^O &= \sum_{j=0}^{N-1} (-1)^j i_l^{(j)} \\ I_l^{O(q)} &= \sum_{j=0, j\neq 2q, 2q+1}^{N-1} (-1)^j i_l^{(j)} + 2 i_l^{(2q)} - 2 i_l^{(2q+1)}, \\ &\qquad l = 0, \ldots, \frac{N-2}{2}. \end{aligned} \tag{7}$$

It is noted that computing the response to a particular excitation is not particularly time consuming, for, as long as the structure is not modified, its moment matrix is unchanged.

The remaining elements of $\mathbf{Y}^e$ have to be recovered by the knowledge of $\mathbf{Y}^s$. To this aim, it is convenient to rearrange the elements of $\mathbf{Y}^e$ and $\mathbf{Y}^s$ so as to first consider the ports to the left-hand side—the ones identified by even indexes, in the following indicated by a subscript $L$—and then those to the right-hand side (subscript $R$). By virtue of this arrangement, $\mathbf{Y}^s$ may be partitioned as follows:

$$\mathbf{I} = \begin{bmatrix} \mathbf{I}_L \\ \mathbf{I}_R \end{bmatrix} = \begin{bmatrix} \mathbf{Y}^s_{LL} & \mathbf{Y}^s_{LR} \\ \mathbf{Y}^s_{RL} & \mathbf{Y}^s_{RR} \end{bmatrix} \begin{bmatrix} \mathbf{V}_L \\ \mathbf{V}_R \end{bmatrix} = \mathbf{Y}^s \mathbf{V} \tag{8}$$

with the same partitioning applied to $\mathbf{Y}^e$. Due to the symmetry of the standard and to the reciprocity of all networks involved, the relationships

$$\mathbf{Y}^s_{LL} = \mathbf{Y}^s_{RR} \quad \mathbf{Y}^s_{LR} = \mathbf{Y}^s_{RL} \quad \mathbf{Y}^e_{LR} = \mathbf{Y}^e_{RL} \tag{9}$$

are satisfied. Let us apply a set of odd excitations to the standard network, by putting in (8)

$$\mathbf{V}_R = -\mathbf{V}_L \tag{10}$$

so that

$$\mathbf{I}_L = (\mathbf{Y}^s_{LL} - \mathbf{Y}^s_{LR})\mathbf{V}_L. \tag{11}$$

Note that we can build $(N-2)/2+1$ independent vectors $\mathbf{V}_L$, namely, the dimension of a partitioning block in $\mathbf{Y}^e$, satisfying relationships (10) and (11).

With the assumed excitations, the internal ports in Fig. 2 are at ground potential, i.e., using primed symbols for voltages and currents at the ports of $\mathbf{Y}^e$

$$\mathbf{V}_R{}' = 0. \tag{12}$$

Hence,

$$\mathbf{I}_L{}' = \mathbf{Y}^e_{LL}\mathbf{V}_L{}' = \mathbf{Y}^e_{LL}\mathbf{V}_L. \tag{13}$$

Considering that $\mathbf{I}_L{}' = \mathbf{I}_L$, and comparing (11) and (13), for every excitation $\mathbf{V}_L$ satisfying choice (10), one has

$$\mathbf{Y}^e_{LL}\mathbf{V}_L = (\mathbf{Y}^s_{LL} - \mathbf{Y}^s_{LR})\mathbf{V}_L \tag{14}$$

so that it has to be

$$\mathbf{Y}^e_{LL} = \mathbf{Y}^s_{LL} - \mathbf{Y}^s_{LR} \tag{15}$$

which is

$$y^e_{2k,2l} = y^s_{2k,2l} - y^s_{2k,2l+1}, \qquad k, l = 0, \ldots, \frac{N-2}{2}. \tag{16}$$

At this point, we have calculated the elements of the block $\mathbf{Y}^e_{LR}$ by means of (6), and those of $\mathbf{Y}^e_{LL}$ by means of (15). We still need a way compute the self and mutual admittances of the internal nodes, namely, the block $\mathbf{Y}^e_{RR}$. To this aim, let us apply a whole set of even excitations to the standard network

$$\mathbf{V}_R = \mathbf{V}_L. \tag{17}$$

In this case too, we can build $(N-2)/2+1$ independent vectors satisfying (17), namely, a complete basis for a block of $\mathbf{Y}^e$. Due to the even excitation, no current flows through the internal nodes, producing an ideal open-circuit for the network $\mathbf{Y}^e$ as follows:

$$\mathbf{I}_R{}' = 0. \tag{18}$$

Hence, being $\mathbf{V}_L{}' = \mathbf{V}_L$, one obtains

$$\begin{aligned} \mathbf{I}_L{}' &= \mathbf{Y}^e_{LL}\mathbf{V}_L + \mathbf{Y}^e_{LR}\mathbf{V}_R{}' \\ 0 &= \mathbf{Y}^e_{LR}\mathbf{V}_L + \mathbf{Y}^e_{RR}\mathbf{V}_R{}'. \end{aligned} \tag{19}$$

By eliminating the internal node voltages $\mathbf{V}_R{}'$ in (19), the relationship

$$\mathbf{Y}^e_{RR}\mathbf{Y}^e_{LR}{}^{-1}\left[\mathbf{I}_L - \mathbf{Y}^e_{LL}\mathbf{V}_L\right] = -\mathbf{Y}^e_{LR}\mathbf{V}_L \tag{20}$$

is found. However, substituting condition (17) in the definition of $\mathbf{Y}^s$ yields

$$\mathbf{I}_L = \left(\mathbf{Y}^s_{LL} + \mathbf{Y}^s_{LR}\right)\mathbf{V}_L. \tag{21}$$

By virtue of (20) and (21), the equation

$$\mathbf{Y}^e_{RR}\mathbf{Y}^e_{LR}{}^{-1}\left[\mathbf{Y}^s_{LL} + \mathbf{Y}^s_{LR} - \mathbf{Y}^e_{LL}\right]\mathbf{V}_L = -\mathbf{Y}^e_{LR}\mathbf{V}_L \tag{22}$$

holds for every element $\mathbf{V}_L$ of the basis so that

$$\mathbf{Y}^e_{RR} = -\mathbf{Y}^e_{LR}\left[\mathbf{Y}^s_{LL} + \mathbf{Y}^s_{LR} - \mathbf{Y}^e_{LL}\right]^{-1}\mathbf{Y}^e_{LR} \tag{23}$$

gives an expression for the unknown $\mathbf{Y}^e_{RR}$ in terms of known quantities.

The above set of relationships yields the whole $\mathbf{Y}^{e1}$ matrix. A correction block may be calculated by simply changing the sign of the calculated parameters and exchanging odd and even ports. By interconnecting such a correction network before the first error network of Fig. 1, the deembedding procedure of the left-side-hand ports is completed. The same procedure has to be followed for the remaining sides. In synthesis, the whole algorithm is required to:

1) generate the standard structure of Fig. 2 and recover $\mathbf{Y}^s$;
2) apply to the standard the excitations (1) and, if $N > 2$, the $(N-2)/2$ excitations (3), obtaining $I_k^O$ and $I_k^{O(q)}$, respectively, in order to use (6); alternatively $I_k^O$ and $I_k^{O(q)}$ may be directly obtained by (7);
3) calculate values of the self and mutual admittances at the external ports by (15);
4) calculate the remaining self and mutual admittances at the internal ports by means of (23);
5) build up the correction matrix.

## III. Results

As a first step, the above relationships were tested by the circuit simulator provided in the Microwave Office tool. Several arbitrary structures were designed to be error networks and were connected in order to simulate the “standard.” By means of the described excitations, currents at the desired circuit nodes were measured and successfully employed to test the algorithm.

Subsequently, the SOC algorithm was included in our in-house MoM program. The technique is a 3-D version of the approach described in [7] as generalized transverse resonance-diffraction approach, and will be the subject of a future paper. Its features include the ability to treat lossy conductors of finite thickness, but for our purposes, it is applied to thin ideal conductors in order to compare results with the ones obtained by commercial software packages. To this aim, the conductor thickness is reduced down to a fraction of a micrometer. In this case, the field formulation reduces to a standard spectral-domain approach with delta-gap excitations. All the structures are enclosed in a metallic box.

The first structure analyzed is an ideal coupled stripline as, for striplines, an exact solution is known [11]. Formulas for this case are also available at the web site by Sonnet Software. The two striplines are 20-mm long and 5-mm wide, while being spaced 1 mm. The 2-mm distance between the ground planes is filled with a medium of permittivity $\varepsilon_{\mathrm{r}} = 3.0$. The enclosing box is 80-mm wide, in order to reduce its influence down to a minimum; the reference planes are fixed at 4 mm from the excitation plane, hence the deembedded structure are two 12-mm-long coupled striplines. Data are obtained at six frequency points between 10 and 15 GHz and Fig. 4 shows a comparison for the magnitude of $S11$ and $S13$ with and without deembedding, while Fig. 5 compares the $S12$ and $S14$ phase. Exact results are highlighted by marks.

Assuming the reference system of Fig. 1, volume currents $J_x$ and $J_z$ have been expanded by a set of functions that are piecewise sinusoidal along their direction ($x$ and $z$, respectively), and piecewise constant along the transverse direction and the $y$-direction. On the other hand, $J_y$ is expanded by means of a

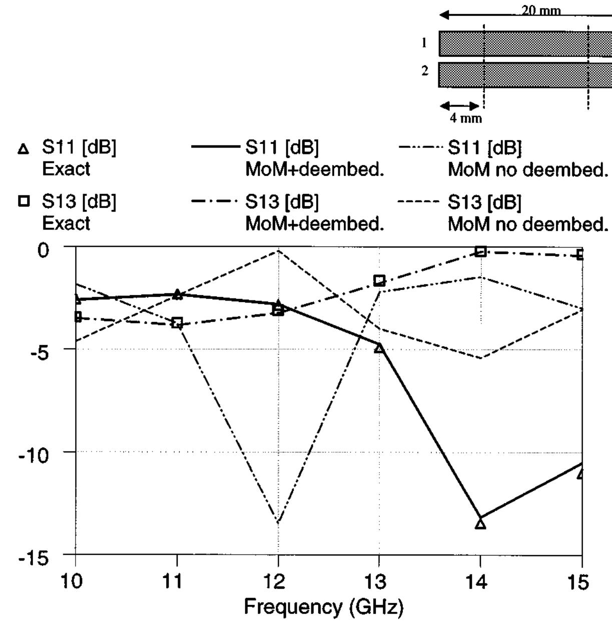


Fig. 4. Comparison between results from the MoM, deembedded MoM, and exact values of $S11$ and $S13$ magnitude for two coupled striplines (see text for parameters).

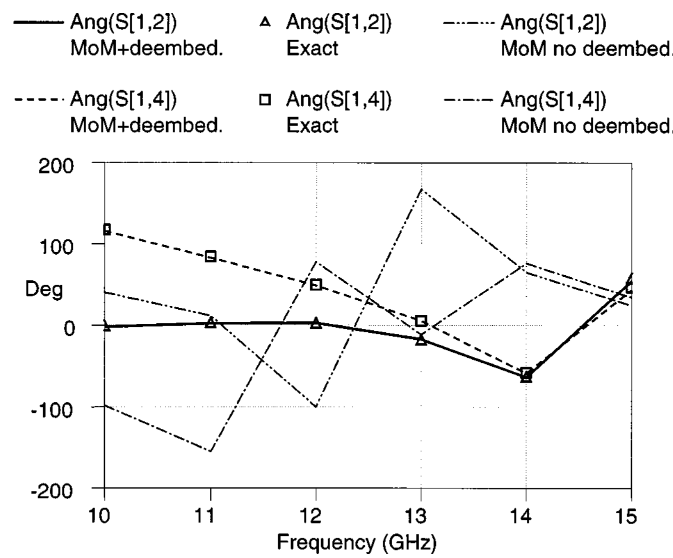


Fig. 5. Comparison between results from the MoM, deembedded MoM, and exact values of $S12$ and $S14$ phase for two coupled striplines.

set of piecewise constant functions along all directions. Conductors are modeled by 20 subsections in the $x$-direction and five subsections in the $z$-direction. The total numbers of expansion functions for the structure are $NJ_x = 210$, $NJ_y = 200$, and $NJ_z = 160$, while the “standard” required $NJ_x = 100$, $NJ_y = 80$, and $NJ_z = 64$.

The test on the striplines, while being simple and reliable, does not verify how the algorithm works when dealing with structures supporting hybrid modes, namely, involving inhomogeneous media. In order to check this case, as shown in Fig. 6, we have applied our approach to two open-ended coupled microstrip lines, and compared the results to those obtained from Emsight in Microwave Office, a commercial

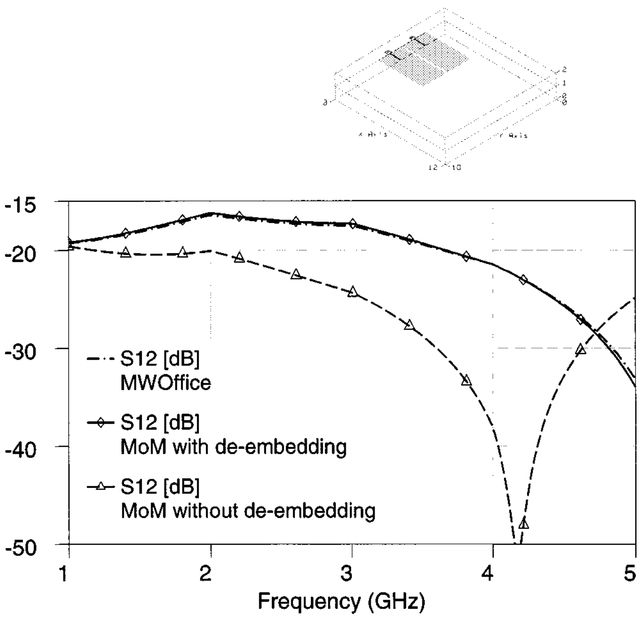


Fig. 6. Comparison between results from the MoM, deembedded MoM, and results from Microwave Office for two open-ended coupled microstrip lines.

spectral-domain MoM using a double-standard deembedding technique. The substrate thickness is 1 mm, its permittivity $\varepsilon_r$ is 12.9, the strips are 2-mm wide and 4-mm long, and the spacing is 0.2 mm. The box is 10-mm wide and 10-mm long; the strip closer to the wall is distant 1.8 mm from it. The reference plane is placed at 1 mm from the excitation plane. The two strips are electrically short over the whole frequency range. The total numbers of expansion functions for the structure are $NJ_x = 48$, $NJ_y = 48$, and $NJ_z = 32$, while for the standard, $NJ_x = 42$, $NJ_y = 36$, and $NJ_z = 24$. Also in this case, the agreement is satisfactory.

## IV. Conclusions

This paper has presented an SOC algorithm for numerical deembedding in MoM approaches, which is able to handle arbitrary number of ports, possibly coupled. Implementation-ready formulas have also been included.

The proposed technique requires just one "standard" structure to be analyzed for each set of coupled ports in order to extract the correction network. This feature saves a significant amount of time when compared with existing deembedding algorithms, requiring the characterization of at least two standard structures. Moreover, the present algorithm evaluates the $Y$ matrices of the error networks with virtually no *a priori* assumption about their nature and their internal structure.

## Acknowledgment

The authors are particularly grateful to one of the unknown reviewers for his very thoughtful remarks and suggestions that actually led to considerable overall improvements in this paper.

## References

[1] J. C. Rautio and R. F. Harrington, "An electromagnetic time–harmonic analysis of shielded microstrip circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 726–730, Aug. 1987.

[2] G. V. Eleftheriades and R. Mosig, "On the network characterization of planar passive circuits using the method of moments," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 438–445, Mar. 1996.

[3] L. Zhu and K. Wu, "Characterization of unbounded multiport microstrip passive circuits using an explicit network-based method of moments," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 2114–2124, Dec. 1997.

[4] M. Farina and T. Rozzi, "Electromagnetic modeling of linear pHEMT's for millimeter waves," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, June 2000, pp. 1801–1804.

[5] R. W. Jackson and D. M. Pozar, "Full-wave analysis of microstrip open-end and gap discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1036–1042, Oct. 1985.

[6] L. P. B. Katehi and N. G. Alexopoulos, "Frequency-dependent characteristics of microstrip discontinuities in millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1029–1035, Oct. 1985.

[7] T. Rozzi and M. Farina, Eds., *Advanced Electromagnetic Analysis of Passive and Active Planar Structures*. London, U.K.: IEE Press, 1999, pp. 222–227.

[8] J. C. Rautio, "A deembedding algorithm for electromagnetics," *Int. J. Microwave Millimeter-Wave Computer-Aided Eng.*, vol. 1, no. 3, pp. 282–287, 1991.

[9] L. Zhu and K. Wu, "Unified equivalent-circuit model of planar discontinuities suitable for field theory-based CAD and optimization of M(H)MIC's," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1589–1603, Sept. 1999.

[10] ——, "Network equivalence of port discontinuity related to the source plane in a deterministic 3-D method of moments," *IEEE Microwave Guided Wave Lett.*, vol. 8, pp. 87–89, Feb. 1998.

[11] J. C. Rautio, "An ultrahigh precision benchmark for validation of planar electromagnetic analyses," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2046–2050, Nov. 1994.

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