Modeling of the Influence of Parasitic Coupling Between a Component and Its Feeding Line on Scattering Parameters

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***Abstract*—In this paper, a method is discussed that allows a more precise modeling of the S-parameters of a component by taking into account the way it is connected. The coupling between the feeding line and the component changes the S-parameters of the component by more than just a mere shift in the reference plane. This coupling is approximated by assuming that the feeding line only carries a travelling wave and by using elementary dipoles to model the radiation of the component.**

***Index Terms*—Computer-aided design (CAD), electromagnetic compatibility (EMC), mutual coupling.**

I. Introduction

As MODERN circuits become smaller and the frequencies that they work at become higher, inevitably, parasitic coupling within the circuits will start to influence the behavior of the circuit more and more. Therefore, it is necessary to include the influence of mutual coupling in the circuit simulators that are used to design the circuit.

One way in which the behavior of the circuit is altered by parasitic coupling is by the influence of the coupling between a component and its feeding line. Many regular circuit simulators (A.D.S., M.D.S.) use a black-box S-parameter description [1], [2] of the component and a simple-ideal transmission-line model for the feeding line. The result is that the reference plane of the component is simply shifted over the length of the transmission line, increasing the phase of the S-parameters by the electrical length of the feeding line, while the amplitudes remain constant. In this paper, we will show that, due to parasitic coupling between the feeding line and the component, the amplitude and phase of the S-parameters will differ from the S-parameters that result from the above-mentioned reference plane shift method. This effect is also illustrated in [3] and [4].

In [5]–[7], methods are described that can include this parasitic coupling in circuit simulations in a much faster, though approximate, way than the classically used methods (method of moments (MoM) [8], finite difference time domain [9]). The work in [6] describes a method that approximates the field that is emitted by a discontinuity (component) by using properly excited dipoles. More of these dipoles are needed if the discontinuity's size increases or its radiation pattern has to be modeled at smaller distances (components closer together). The work of [7] describes a method that approximates the radiation behavior of a line by using the specific physical relations (traveling waves) that apply to transmission lines. Both [6] and [7] only model first-order coupling: currents that are induced by induced currents are ignored. This approximation is only valid in cases where the couplings are not too tight.

In the case that is discussed here, the coupling will be very tight because the distance between the component and its feeding line is by definition zero. Therefore, we will consider a small region of the transmission line next to the discontinuity: the region of tight coupling. This tight coupling region is not described in the same way but is taken into account using the classical circuit connection. This region should be made large enough to allow the higher order modes to die out. In this case, only the fundamental mode needs to be taken into account beyond this region.

II. Mixing the S-Parameter and Field Description of a Structure

If we want to use the above-mentioned modules ([6] and [7]) to include the parasitic coupling between the component and feeding line in a circuit simulation, then the tight coupling in the connected region causes major problems. In this section, it is shown how this problem can be solved.

Normally, power transfer is considered to occur in two possible ways: by waves passing along through ports (if they are connected) and by fields that go through the air and substrate. In the case of a component that is connected to a transmission line, a problem arises. In this particular case, the field and wave power transfer will be the same because they are two exchangeable ways of describing the same phenomenon: The connection between the component and the line. The result is that, in this connected case, the power is transferred twice. A simple solution to this problem is to ignore the field contribution in the coupling between the line and the component and to take only the wave passing through the port into account. The disadvantage of this is that information about the parasitic coupling is lost. Another solution could be to take only the field contribution into account and ignore the waves on the port. This would be correct if the field contributions are calculated rigorously (as is the case in the MoM). But the modules discussed in [6] and [7] are not good at predicting tight (close) coupling.

This problem is solved in this paper by using a mixed approach. A transfer section is introduced. It is a small section of

Manuscript received February 5, 2002; revised July 30, 2002. This work was supported by the Flemish Institute for the Advancement of Scientific-Technological Research in Industry (I.W.T.) under a scholarship.

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Digital Object Identifier 10.1109/TMTT.2003.808626

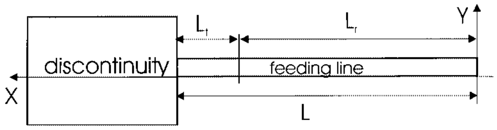


Fig. 1. General component fed by line.

the line that is located right next to the component (L_t in Fig. 1). Coupling with this piece of line is taken into account only by transferring the outgoing wave at the beginning and end of the transfer section to the component and line and vice versa. This transfer section can also be seen as a buffer zone between the component and the line that allows the higher order modes that exist on the line near the component and are caused by the tight coupling to the component, to die out. The length of the transfer section L_t must be chosen in such a way that there are no higher order modes present at the end, thereby allowing the use of an S-parameter description with the reference plane positioned at this end.

The coupling between the remainder of the line (L_r) and the component is low enough to be calculated using [6] and [7].

The effect of the approximation that the field on the transfer section is not considered at all can by shown by solving the structure using a full integral equation model and splitting the field contributions. The following equation shows the integral equation for a planar structure:

$$\int_{x'}\int_{y'} \overrightarrow{J}\,(x',y')\,\overline{\overline{G}}\,(x,y,x',y')\,dx'\,dy' = -\vec{E}_{xi}(x,y) \quad (1)$$

where G is the appropriate Green's dyadic, x' and y' are the source coordinates, and x and y are the observation coordinates. By applying the moment method and assuming that only the line is excited by its fundamental mode, we obtain the following matrix equation:

$$\begin{bmatrix} Z_{cc}Z_{ct}Z_{cr} \\ Z_{tc}Z_{tt}Z_{tr} \\ Z_{rc}Z_{rt}Z_{rr} \end{bmatrix} \begin{bmatrix} J_c \\ J_t \\ J_r \end{bmatrix} = - \begin{bmatrix} 0 \\ 0 \\ E_r \end{bmatrix} \quad (2)$$

where c indicates the component, t is the transfer section, and r is the remainder of the feeding line. The line current can be solved from

$$J_r = -Z_{rr}^{-1}Z_{rc}J_c - Z_{rr}^{-1}E_r - Z_{rr}^{-1}Z_{rt}J_t. \quad (3)$$

Equation (3) shows us that the current on the remaining line section is caused by three field contributions:

1) the field caused by the currents on the component—this can be modeled as component-line coupling;
2) the excitation field caused by an incident wave at the line's port;
3) a field caused by the coupling between transfer and remaining section.

It is this last contribution that will cause problems if we want to use the modules of [6] and [7] to model the component-line coupling.

This last contribution can be approximated using a traveling wave (S-parameter) description, instead of the correct field description. This exchanging of the field description for an S-parameter description is only rigorous in the case where the component would be a half-infinite section of line. In this case, the fields that the traveling waves on the half-infinite line cause on the connected line can be correctly exchanged by an S-parameter description of the connection between both lines. This S-parameter description simply states that the outgoing wave of one line is equal to the incident wave of the other and vice versa. This principle is used in [10] to feed the lines of a structure in the MoM. In this paper, the error that results from the use of a noninfinite line is shown as a function of the length actually used.

If we now define the reflection coefficient at the beginning of the transfer section as the ratio between the incident and reflected wave at that position

$$S_{tt} = \frac{V_l^-(x = L_r)}{V_l^+(x = L_r)} \quad (4)$$

then the current on the line can now be approximated by swapping the field description for terms 2 and 3 in (3) with an S-parameter (travelling wave) description) in

$$J_r = J_{\mathrm{Spar}} - Z_{rr}^{-1}Z_{rc}J_c \quad (5)$$

where J_{Spar} is the current on the remainder of the line, without parasitic coupling, calculated solely based on the component's reflection coefficient and the excitation

$$J_{\mathrm{Spar}}(x) = e^{-\gamma x} - S_{tt}e^{\gamma(x-2(L-L_r))}. \quad (6)$$

The current in (6) is the sum of the incident, feeding, left traveling wave (phase 0 at $x = 0$), and the reflected right traveling wave. The result is an approximation because the transfer section is not infinitely long. In the numerical result section it is shown, for the example of a line-fed patch, that (even for small L_t) the line current (and thus the reflection coefficient at the end of the line) calculated using (5) is a very good approximation for the rigorously calculated current [using (2)].

We assume that the transfer section is long enough to ensure that the current profile on the discontinuity does not change much as a function of the total length of the line. This means that the reflection at the end of the transfer section (S_{tt}) becomes independent of the length of the remaining line and can be stored in a model file for the component, along with the excitation data for the dipoles, which model the radiation of the current on the component (see [6]).

III. Practical Implementation

In this section, we will show how the modules from [6] and [7] can be used to include parasitic coupling between the components and their feeding lines.

These modules are used in a library-based approach: the components that are used in the circuit that has to be analyzed first have to be simulated by themselves using a plain MoM simulation. During this library-building simulation, the S-parameters and the proper excitation to the dipoles (so they generate the same field pattern as the original component) of the model are determined (see [6]). During this library-building simulation, the component is fed by lines of length L. This is shown for the one-port case in Fig. 2.

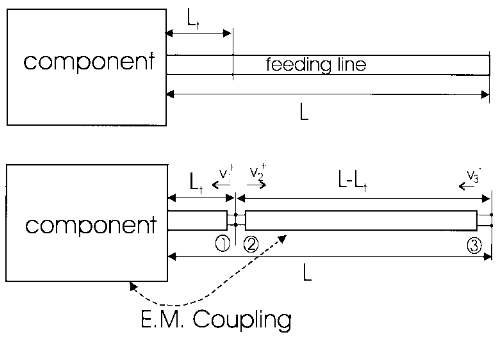


Fig. 2. One-port fed by line and equivalent circuit.

The result of this simulation is the reflection coefficient $S_{\rm ex}$ at the feed end of the line (port 3 of the circuit equivalent at the bottom of Fig. 2). The length (L) of the feeding lines that are used during the library-building simulation must be large enough to allow proper deembedding of the reflection parameter $(S_{\rm ex})$. This means that the lines must be at least a quarter of a wavelength if [10] is used for deembedding.

The model of the component must include the transfer section between the component itself and the reference plane. This section is shown as L_t in the figure. This section is needed for two reasons:

1) to keep the coupling between the component and the remaining line low enough to use the modules described in [5] and [6];
2) to allow higher order modes to die out so a monomode S-parameter description can be used at the reference plane (at the right end of L_t).

The length of this included line section (L_t) should be as small as possible to allow flexible use of the component's model when it is used to build circuits. To satisfy 1) and 2), however, ideally, it must be as long as possible. It is shown in Section IV that even for a fairly short L_t section 1) and 2) can by sufficiently satisfied.

The S-parameter data that should be stored in the model (S_{11}) has its reference plane at the end of L_t. The S-parameters that were calculated during the library-building simulation $(S_{\rm ex})$, however, have their reference plane at the end of the line (at length L). The obvious way to calculate S_{11} from $S_{\rm ex}$ is to use a plain reference plane shift (increasing the phase of $S_{\rm ex}$ with $2\beta L$). By doing this, however, we ignore the coupling between the component and its feeding line.

A better approach is to use the equivalent circuit in Fig. 2 and try to eliminate the feeding line from the circuit. It must be stressed here that, due to the parasitic coupling between the component and the line, both must be regarded together as a three-port and not as a cascade of a one- and a two-port. This means that the operation is no longer just a simple reference plane shift because ($S_{12} \neq 0$ and $S_{13} \neq 0$).

If S_{11} is known, than the reflection at port 3 $(S_{\rm ex} = v_3^- / v_3^+)$ can be calculated from the following set of equations with $v_3^+ = 1$ (reciprocity is used already):

$$\begin{bmatrix} S_{11} S_{12} S_{13} \\ S_{12} S_{22} S_{23} \\ S_{13} S_{23} S_{33} \end{bmatrix} \begin{bmatrix} v_1^+ \\ v_2^+ \\ v_3^+ \end{bmatrix} = \begin{bmatrix} v_2^+ \\ v_1^+ \\ v_3^- \end{bmatrix}. \quad (7)$$

Because we assume that the higher order modes have died out at port 1, this matrix is a simple monomode matrix. For the transmission line between ports 2 and 3, the S-parameters are: $S_{22} = 0$, $S_{33} = 0$, and $S_{23} = e^{-j\gamma(L-L_t)}$. The S_{12} and S_{13} coupling parameters are calculated using the modules described in [6] and [7]. To solve the reverse problem ($S_{\rm ex}$ is known, S_{11} is wanted), we can derive the following equation from (7):

$$S_{11} = ((1 - S_{12})B - S_{13})C^{-1}$$

with

$$B = (1 + S_{23}^{-1} S_{13}(1 - S_{12})^{-1} S_{22}) \times S_{23}^{-1}(S_{ex} - S_{13}(1 - S_{12})^{-1} S_{23} - S_{33})$$

and

$$C = (1 + (1 - S_{12})^{-1} S_{22} S_{23}^{-1} S_{13})^{-1}(1 - S_{12})^{-1} \times (S_{22} S_{23}^{-1}(S_{ex} - S_{33}) + S_{23}). \quad (8)$$

Equation (8) is also valid for n-ports. In this case, the S-parameters will become matrices instead of scalars.

The calculated S_{11} matrix is stored in the model file along with the dipole excitation data. During a circuit calculation with this model file, (7) will be used to attach the feeding line (which may now have a different length) to the model and calculate the reflection parameters for the entire structure (components + lines). It is straightforward to see that, in the special case where the line that is attached in this phase has the same length as the one that was used in the library-building calculation, the result will be exact, because (7) and (8) are then inverse operations.

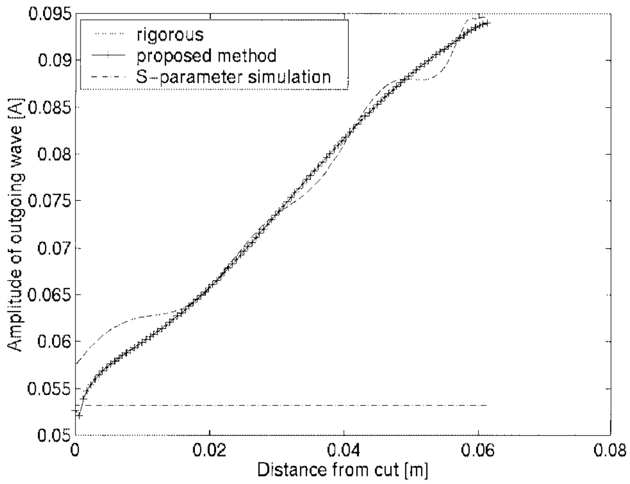
IV. Numerical Results

In this section we will test the proposed method on a line-fed patch. We will show the influence of the length of the feeding line on the behavior of the patch, which is neglected in normal S-parameter calculations. We will show how the proposed method results in an improvement of the prediction of the structure's behavior while needing only a small calculation time.

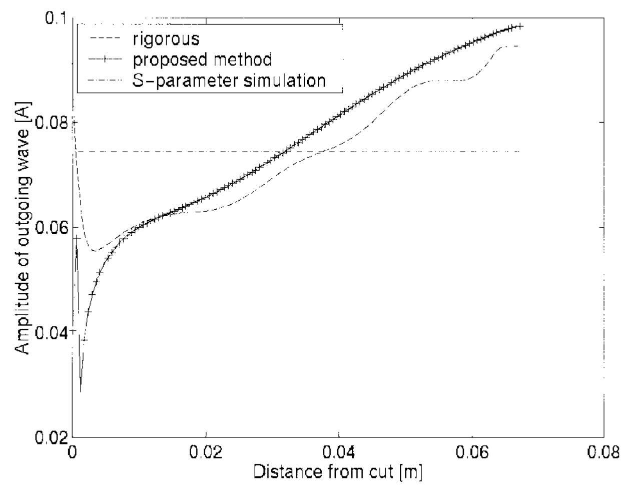
First we will show the validity of (5) by comparing it with the rigorous result calculated using (2). The component in this case is a 14.3 × 14.3 mm big patch, fed at the edge by a 1.3-mm-wide, 70-mm-long line on top of a 1.575-mm-high substrate with a relative permitivity of 2.2. Fig. 3 shows the right traveling (outgoing) component of the current on the remaining part of the line at 6.86 GHz ($x = 0$ to $x = L - L_t$, for both the rigorous solution and the approximation using (5). The current can be split in to left- and right-traveling waves using [7, eq. (10)]. The length of the transfer section is 7.6 mm in Fig. 3(a) and 1.8 mm in Fig. 3(b).

From Fig. 3, we can conclude that there is only a small error that becomes larger closer to the component. The accuracy of the current approximation is only important at the fed end of the line, where the approximation is most accurate, because it is used there to calculate the S-parameters for the entire structure.

The next example illustrates the use of (7) and (8) together with the modules of [6] and [7] to calculate the S-parameters of a patch-bent structure on a substrate with a height of 1.575 mm

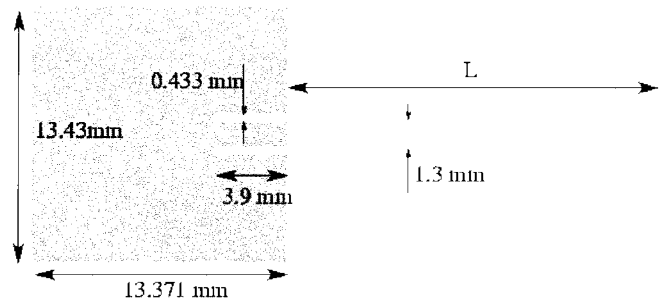


(a)

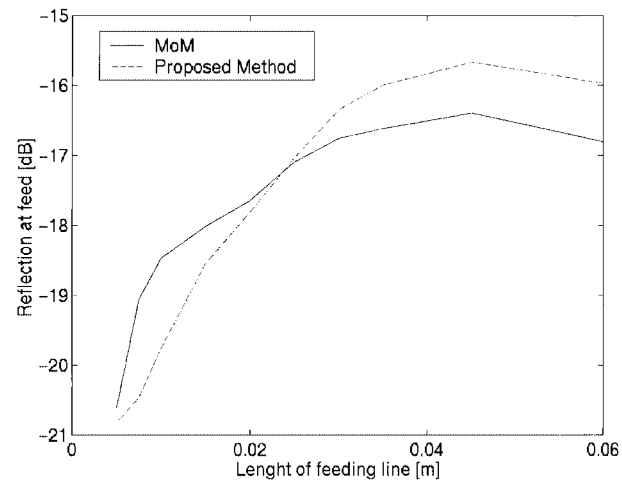


(b)

Fig. 3. Right-traveling component of the current on the remaining line that is feeding the patch. Shown are the rigorous (MoM) result, the proposed method, and the (constant) current calculated using a classical S-parameter (using the reflection coefficient at the end of the transfer section) circuit simulator. The X axis indicates the distance from the end of the transfer section. (a) $L_t = 7.6$ mm. (b) $L_t = 1.8$ mm.



(a)



(b)

Fig. 4. (a) Edge-fed patch fed by a line with length L. (b) Reflection at fed port calculated with MoM (continuous line) and the proposed method (dashed line) as a function of the feeding line length.

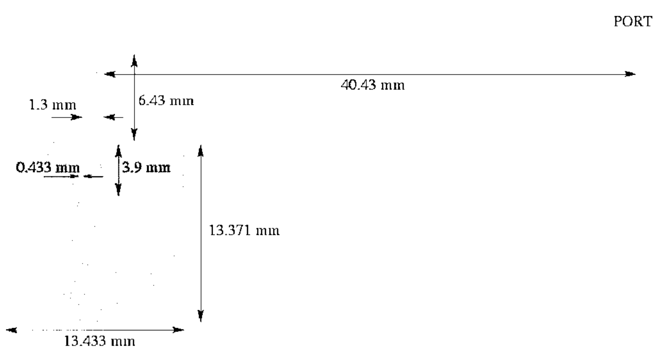


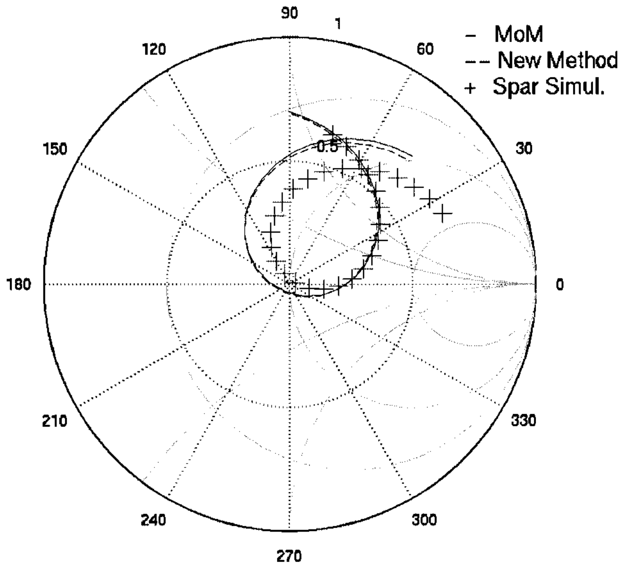
Fig. 5. Analyzed edge-fed patch fed by line with a bent.

and $\varepsilon_r = 2.2$. We begin by investigating the correlation between the S-parameters, at 7.2 GHz (the resonant frequency), calculated rigorously (MoM) and with the proposed method for the patch-line structure in Fig. 4(a), as a function of the length of the feeding line (which has three segments across its length). The length of the transfer section is 5 mm.

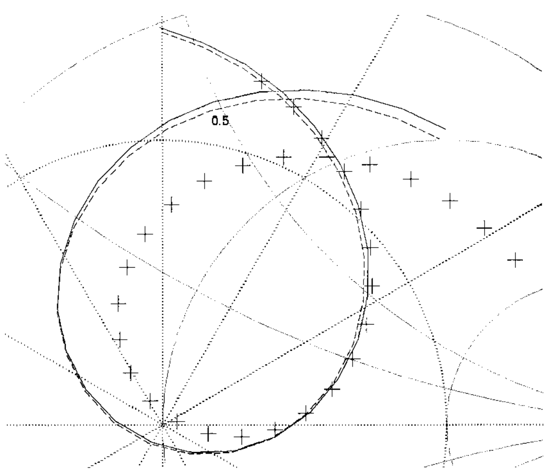
The graph in Fig. 4(b) shows the result for MoM and the new method. The reflection at the transition (S_{11} in Fig. 2) for the patch was calculated with a line length of 25 mm, and the error therefore becomes zero at this length, proving that (8) is the exact inverse of (7). The increasing error for small feeding lengths could be due to the deembedding procedure of MoM, which becomes inaccurate for small feeding line lengths. It can also be explained by the inaccuracy for smaller line lengths that was demonstrated in the previous example (Fig. 3) and by the model that is used to calculate line coupling [7], which is theoretically only exact for infinitely long lines.

A 90° bent is now inserted in the feeding line at a distance of 6.43 mm from the patch (Fig. 5). The bent model is calculated in the same way as the path model. The length of the transfer section for the bent is 4 mm for both lines. The bent itself is made up of six square segments.

Fig. 6 shows a comparison of the reflection for the total structure as a function of frequency (6.8–7.6 GHz.). The proposed method is clearly more accurate than the plain S-parameter simulation. For this plain S-parameter simulation, the S-parameters of the components (blackbox description for patch and bent) are obtained by simulating it with a long line (40 mm) and the transmission lines that connect the components are ideal lines that do not take the parasitic coupling in the circuit into account. The MoM needs 35 s for one frequency point and the proposed method only needs 0.26 s (both) on an HP J-6000 (550-MHz



(a)



(b)

Fig. 6. Reflection at feed for the structure in Fig. 5. Comparison between regular MoM, proposed methodm and simple S-parameter circuit simulator. (a) This figure is a magnification of (b).

processor) machine. (This does not include the calculation time for the Green's functions.)

V. Conclusion

A new technique to include the parasitic coupling between a component and its feeding lines into the circuit calculations has been demonstrated. The results have been compared to the MoM and are in good agreement. A large increase of accuracy is observed compared to a plain S-parameter circuit simulation. The method is much faster than MoM and needs only a fraction of the computer memory.

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