Fast electromagnetic characterization of integrated circuit passive isolation structures based on interference blocking

Citation for published version (APA):

DOI:
10.1109/TMTT.2017.2702638

Document status and date:
Published: 01/11/2017

Document Version:
Accepted manuscript including changes made at the peer-review stage

Please check the document version of this publication:

• A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher's website.
• The final author version and the galley proof are versions of the publication after peer review.
• The final published version features the final layout of the paper including the volume, issue and page numbers.

Link to publication

General rights
Copyright and moral rights for the publications made accessible in the public portal are retained by the authors and/or other copyright owners and it is a condition of accessing publications that users recognise and abide by the legal requirements associated with these rights.

• Users may download and print one copy of any publication from the public portal for the purpose of private study or research.
• You may not further distribute the material or use it for any profit-making activity or commercial gain
• You may freely distribute the URL identifying the publication in the public portal.

If the publication is distributed under the terms of Article 25fa of the Dutch Copyright Act, indicated by the "Taverne" license above, please follow below link for the End User Agreement:
www.tue.nl/taverne

Take down policy
If you believe that this document breaches copyright please contact us at:
openaccess@tue.nl
providing details and we will investigate your claim.

Download date: 14. Sep. 2019
Fast Electromagnetic Characterization of Integrated Circuit Passive Isolation Structures Based on Interference Blocking

Mercè Grau Novellas, Ramiro Serra, and Matthias Rose

Abstract—An early characterization of integrated circuit passive isolation structures is crucial to predict their performance and effectiveness in minimizing substrate coupling. In this paper, an electromagnetic modeling methodology is proposed, that can be applied to different types of isolation structures based on interference blocking. It consists in characterizing the different doping profiles by identifying those propagation modes with a greater contribution to substrate coupling and to the aggressor-victim transfer function. This allows a fast electromagnetic characterization, providing insight on the different isolation mechanisms, identification of transitory effects and a fast prediction of isolation effectiveness.

Index Terms—Electromagnetic compatibility, electromagnetic interference, integrated circuits, signal integrity, substrate coupling.

I. INTRODUCTION

SUBSTRATE coupling is a major electromagnetic compatibility (EMC) issue in integrated circuits (ICs). It can have a negative impact on circuit performance or even be a cause of malfunction or failure, especially in mixed-signal applications [1], where different electronic functional blocks (microcontrollers, RF modules, A/D converters, switched-mode power converters, etc.) are integrated in the same substrate. Therefore, it is important to implement techniques to prevent interference from reaching sensitive nodes through the substrate. In this sense, passive isolation structures are commonly used to minimize unwanted coupling [2]. There are two main isolation mechanisms that can be implemented, either separately or combined. The first one consists in providing a low impedance path in order to attract interference currents and route them away from sensitive nodes. In practice, this is normally implemented with a P+ guard ring connected to an off-chip ground [2], [3]. The second one consists in blocking interference propagation by embedding highly resistive or dielectric sections between the aggressor and the potential victim. Most common structures used for interference blocking are silicon dioxide trenches and reverse-biased junctions (e.g., a deep N-well in a P-type substrate), that can be implemented either parallel or transverse to the direction of interference propagation [2], [4], [5].

In practice, passive isolation structures are usually characterized experimentally and/or by means of simulations [2], [4], [6]. An overview of the most common computational techniques used for substrate modeling can be found in [7]. Behavioral equivalent networks of lumped-elements have also been proposed, however they require the use of extraction tools or calibration with on-chip measurements [8]. For this reason, results cannot be easily extrapolated and performing an extensive characterization including all combinations of variables (design, operational, and technological) can be very expensive and time consuming. In addition, IC substrates are able to support several propagation modes, that are determined by their specific doping profile and frequency of operation [9]. Equivalent lumped-element networks are a single-mode representation and therefore they neglect the effects derived from multi-modal interference propagation, such as mode conversion [10].

EMC problems can lead to expensive and time consuming redesigns. Therefore, because of the role of integrated passive isolation structures, it is very important to be able to characterize them and predict their performance as early as possible in the design process. One of the first stages in this design process is the one concerning technology choices. At this early phase, some important operational and physical parameters such as the exact frequency of operation or the circuit layout may not be defined yet. Still, to have information regarding the implications of the technology choice on interference propagation and the effectiveness of protection structures is very valuable, as this is highly linked to the technology with which they are implemented [3].

Ideally, early effectiveness prediction models should be fast in order to allow the performance of multiple parameter sweeps within a reasonable time and still be able to provide relatively accurate behavior trends for the parameters under study and in a wide frequency range. In addition, they should provide insight on the effectiveness of the different isolation mechanisms that contribute to the overall structure transfer function, as this is a key point for designers to prevent potential unwanted coupling and avoid unnecessary design loops.

In this paper, we propose a fast electromagnetic modeling methodology, applicable to all passive isolation structures which are based on interference blocking. The method is based on considering the substrate as a backplaned lossy multilayered dielectric waveguide and describing it by means of
the supported propagation modes. Subsequently, the isolation structures will be treated as inserted sections of a different doping profile and therefore characterized as discontinuities transverse to the direction of interference propagation. In the present paper, a formulation is presented for a general number of propagation modes and optimization strategies are proposed. This novel optimization strategy allows a smart selection of those modes with a more significant contribution to substrate coupling. These two key points enable the application of the proposed methodology to any kind of isolation structure and in any frequency band, while keeping the number of contributing modes to a minimum.

The proposed methodology is not intended for modeling specific layout details, as this is not required until more advanced design stages. In such advanced phases, topological and operational parameters of the specific application are defined and typical computational techniques [7], [11], [12] are more suitable for accurate modeling. This comes at a cost of high simulation times. Instead, the proposed methodology is intended to be a fast tool to provide insight on the impact of technology choices on the effectiveness of isolation structures, while avoiding the need of calibration nor the use of extraction tools. Therefore, the developed models are meant for providing information when deciding which technology node or option is better for a new IC, elaborating design guidelines, determining if current design guidelines are still applicable in a new technology and helping with layout planning.

II. DESCRIPTION OF THE MODELING METHODOLOGY

An IC substrate can be approached as a multilayered stack of $m$ lossy dielectric layers in the vertical direction (doping profile), and as practically homogeneous in the horizontal direction. We will model the stack as bounded by a perfect electric conductor (PEC) at the bottom, due to the close presence of a ground plane in either the chip itself or the printed circuit board (PCB). Layer $m + 1$ will be a semi-infinite non-conductive medium, thus the structure is modeled as a semi-open waveguide.

Current density distributions generated by device switching activity and substrate contacts [13] will excite propagation modes that will spread the interference radially throughout the chip. Therefore, the substrate becomes a path for unwanted coupling. The propagation modes will be determined by the size and medium properties of the cross-section doping profile [9].

When a passive isolation structure is implemented, this means that a section with a different doping profile that supports different propagation modes is inserted in the substrate stack. Therefore, in order to establish the basis of the proposed modeling methodology, we will consider three consecutive sections (A-C), that will correspond to the doping profiles where the aggressor (source), blocking structure, and victim are located, respectively. Each of them can have a different number of layers with different heights and material properties (see Fig. 1).

As interference propagates radially, we will use a cylindrical coordinate system to describe the structures under study. For the sake of simplicity, we will consider that the source is centered at the origin and the other sections surround it as coaxial rings. Each section is then characterized by its number of layers $n$ and its absolute radius $r_n$. Subsequently, each layer $n$ is characterized by its absolute height $h_n$, permittivity $\varepsilon_n$ and conductivity $\sigma_n$ (see Fig. 2). The layers are referred using a bottom-top notation, where the PEC is located below layer 1.

![Fig. 2. Layer properties of one multilayered section.](image)

The main propagation mechanism that the structures under study are able to support are transverse magnetic propagation modes TM$^+$, with field components $E_z$, $E_r$ and $H_\phi$ [9]. Due to the fact that there is an open-boundary at the top, the propagation coefficients have both a discrete and a continuous spectrum of values. Discrete propagation coefficients correspond to propagation modes that are confined within the substrate and are obtained by means of an eigenvalue equation [9]. This equation is determined by the size and material properties of the cross-section doping profile, and boundary conditions, when fields are forced to be evanescent in layer $n + 1$. The continuous spectrum provides the solutions for the radiation modes, which can be either propagating or evanescent [14], [15], [16]. The propagating subset accounts for radiation to the open space, while the evanescent subset accounts for energy storage close to the source. The total electric and magnetic fields are the superposition of all contributions.

A. Modeling of a Single Discontinuity

In the presence of a single discontinuity (A-B), when the first section is excited with a certain mode (A1), part of the energy will be transferred to the following section B and part of the energy will be reflected. The energy can be transferred to all the modes supported by section B and, in the same way, can be reflected through all the modes supported by section
A. This phenomena is called mode conversion [10]. Therefore, expressions for electric and magnetic fields in section A can be expressed as

\[ \mathbf{E}_A = a_{A1} \mathbf{e}_{A1} + \sum_{i=1}^{\infty} a_{A1} \rho_{Ai,A1} \mathbf{e}_{Ai} + \int_D a_{A1} \rho_{A1,A1}(k_r) \mathbf{e}_A(k_r) \, dk_r, \]

and

\[ \mathbf{H}_A = a_{A1} \mathbf{h}_{A1} + \sum_{i=1}^{\infty} a_{A1}\rho_{Ai,A1} \mathbf{h}_{Ai} + \int_D a_{A1}\rho_{A1,A1}(k_r) \mathbf{h}_A(k_r) \, dk_r, \]

where \( a_{A1} \) are the modal amplitude coefficients, \( \rho_{Ai,A1} \) is the reflection coefficient of mode \( Ai \) when the structure is excited by mode \( A1 \), the summation corresponds to the reflected discrete modes, and the integral corresponds to the reflected radiation modes, where \( D \) is the integration domain in the complex plane [16].

Similarly, fields in section B can be expressed as

\[ \mathbf{E}_B = \sum_{j=1}^{\infty} a_{Bj} \mathbf{e}_{Bj} + \int_D a_{Bj}(k_r) \mathbf{e}_B(k_r) \, dk_r, \]

and

\[ \mathbf{H}_B = \sum_{j=1}^{\infty} a_{Bj} \mathbf{h}_{Bj} + \int_D a_{Bj}(k_r) \mathbf{h}_B(k_r) \, dk_r, \]

where \( a_{Bj} \) are the modal amplitude coefficients.

We will consider that the source of interference is a certain current density distribution \( \mathbf{J}(r, \phi, z) \) located in section A. Each modal amplitude coefficient exciting the discontinuity is related to the source according to [9]

\[ a_{Ai} = -\int_{S} \mathbf{J} \cdot \mathbf{e}_{Ai} \, dV \cdot \int_{S} \mathbf{r} \cdot (\mathbf{e}_{Ai} \times \mathbf{h}_{Ai}) \, dS. \]

In practice, \( \mathbf{J}(r, \phi, z) \) will be determined by the set of components and structures generating the interference (e.g. switching devices, substrate contacts, etc) [13]. In addition, the presence of low impedance structures able to attract interference currents will also play a role in defining the source pattern [3]. For the sake of simplicity, we will consider that the source is a vertically oriented current distribution centered at the origin of coordinates. However, any kind of current distribution can be used to characterize the discontinuity.

The general expressions of the field components that are tangential to the discontinuity interface, for a mode or spectral component \( i \) and in the \( n \)th layer, are [9]

\[ E_{zni} = \frac{-j}{\omega \mu \varepsilon_n} [a_{ni} H_0^2(k_{ri}r) + \sum_{j=1}^{\infty} a_{nj} \rho_{ij} H_1^1(k_{ri}r)] + k_{ri}^2 f_{ni}(z), \]

\[ H_{\phi ni} = \frac{1}{\mu} [a_{ni} H_1^2(k_{ri}r) + \sum_{j=1}^{\infty} a_{nj} \rho_{ij} H_1^1(k_{ri}r)] k_{ri} f_{ni}(z), \]

where the \( z \)-dependent term \( f_{ni}(z) \) corresponds to

\[ f_{ni}(z) = \begin{cases} \cos(k_{zni}(z - h_{n-1}) + D_{zni}) & n \leq m \\ e^{-j k_{zni}z} & n = m + 1 \end{cases}, \]

where \( k_{ri} \) and \( k_{zni} \) are the radial and vertical propagation coefficients respectively, related to the wavenumber in each layer as \( k_n^2 = k_{ri}^2 + k_{zni}^2 \). Backward and forward propagation is described with Hankel functions of the first and second kind, respectively [17]. The infinite summation in the backward propagation term corresponds to the contribution of all possible reflections from other modes due to mode conversion. In (6)-(8) we consider that the continuous spectrum is discretized and its contributions to the backward term are included in the summation.

The condition that the total tangential electric and magnetic fields are continuous at the discontinuity interface (A-B) provides two initial equations. Following the procedure described in [10], we take the cross product of the electric continuity equation with each magnetic modal component in section A and, analogously, we take the cross product of the magnetic continuity equation with each electric modal component in section B. Subsequently, we integrate all resulting equations over the discontinuity interface. By applying the orthogonality property of the set of discrete modes and spectral components of radiation modes in each section [15], several terms will vanish and equations can be derived to determine the discontinuity unknowns.

**B. General Model of a three-Section Discontinuity**

The basic model of the structures under study consists of a three-section continuity with two consecutive interfaces (see Fig. 1). For a given excitation in the aggressor section, the complete set of unknowns will consist of the scattering parameters in sections A and B, and the amplitude coefficients in sections B and C. We will consider that there is no backward propagation in section C. In order to be able to solve the complete system, the second discontinuity needs to be solved first.

For the B-C discontinuity, considering \( p \) modes in section B and \( q \) modes in section C, and that the structure is excited by mode \( B1 \), we obtain the following system of equations

\[ \begin{bmatrix} BC1 \mid BC2 \\ p \times p \mid p \times q \end{bmatrix} \begin{bmatrix} \rho_{B1B1} \\ \vdots \\ \rho_{BpB1} \end{bmatrix} = \begin{bmatrix} BC5 \\ p \times 1 \end{bmatrix}, \]

where \( BC1 \) and \( BC4 \) are diagonal matrices. The elements of submatrices \( BC1 \) to \( BC6 \) are

\[ BC1_{jj} = -k_{rBj}^2 Y_{BjBj} H_0^0(k_{rBj}r_B), \]

\[ BC2_{ij} = k_{rCj}^2 Y_{CjCj} H_0^2(k_{rCj}r_B), \]

\[ BC3_{ij} = k_{rBj} Y_{CjBj} H_1^1(k_{rBj}r_B), \]

\[ BC4_{jj} = k_{rCj} Y_{CjCj} H_1^1(k_{rCj}r_B), \]

\[ BC5_{ij} = \frac{1}{\mu} [a_{ni} H_1^2(k_{ri}r) + \sum_{j=1}^{\infty} a_{nj} \rho_{ij} H_1^1(k_{ri}r)] k_{ri} f_{ni}(z), \]

\[ BC6_{ij} = \frac{-j}{\omega \mu \varepsilon_n} [a_{ni} H_0^2(k_{ri}r) + \sum_{j=1}^{\infty} a_{nj} \rho_{ij} H_1^1(k_{ri}r)] + k_{ri}^2 f_{ni}(z), \]
where combinations. the overall solution will be the superposition of all possible amplitude coefficients for all modes in section B, therefore

\[ Y_{\alpha i} = \sum_{n=1}^{m+1} \int_{h_{n-1}}^{h_n} \frac{a_{n\alpha} f_{n\alpha}(z)}{a_{1\alpha} \varepsilon_{na}} \frac{a_{n\beta j}}{a_{1\beta j}} f_{n\beta j}(z) \, dz, \]

where \( \varepsilon_{na} \) is the permittivity in layer \( n \) of section \( \alpha \) and the subscripts \( \alpha i \) and \( \beta j \) refer to mode \( i \) in section \( \alpha \) and mode \( j \) in section \( \beta \), respectively.

Following the same procedure, we can write a similar system of equations for discontinuity A-B. However, now there are forward and backward waves at both sides of the interface. Considering \( t \) modes in section A and \( p \) modes in section B, and that the structure is excited by mode \( A_1 \), we obtain

\[
\begin{bmatrix}
AB1 \\
t \times t
\end{bmatrix}
\begin{bmatrix}
AB2 \\
t \times p
\end{bmatrix}
\begin{bmatrix}
\rho_{A1,A1} \\
\rho_{A1,A1}
\end{bmatrix}
= \begin{bmatrix}
AB5 \\
t \times 1
\end{bmatrix},
\]

\[
\begin{bmatrix}
AB3 \\
p \times t
\end{bmatrix}
\begin{bmatrix}
AB4 \\
p \times p
\end{bmatrix}
\begin{bmatrix}
a_{B1} \\
a_{B1}
\end{bmatrix}
= \begin{bmatrix}
AB6 \\
p \times 1
\end{bmatrix},
\]

where \( AB1 \) is a diagonal matrix. The elements of submatrices \( AB1 \) to \( AB6 \) are

\[ AB1_{ij} = -k_{rA}^2 Y_{A j-A j} H_0^1(k_{rA} r_A), \]

\[ AB2_{ij} = k_{rB j} Y_{B j-A i} H_0^2(k_{rB} r_A) + \sum_{s=1}^{p} k_{rB s} Y_{B s-A i} \rho_{B s-B j} H_1^1(k_{rB} r_A), \]

\[ AB3_{ij} = -k_{rA} Y_{B j-A j} H_1^1(k_{rA} r_A), \]

\[ AB4_{ij} = \begin{cases} k_{rB j} Y_{B j-B j} H_1^2(k_{rA} r_A) + \rho_{B j-B j} H_0^1(k_{rB} r_A) & \text{if } i = j, \\ k_{rB i} Y_{B i-B j} H_1^2(k_{rA} r_A) & \text{if } i \neq j \end{cases}, \]

\[ AB5_{i1} = \begin{cases} k_{rA}^2 Y_{A1-A1} H_0^2(k_{rA} r_A) & i = 1, \\ 0 & i > 1 \end{cases}, \]

\[ AB6_{i1} = k_{rA} Y_{B i-A1} H_1^2(k_{rA} r_A). \]

Every possible excitation in \( A \) will result in a set of amplitude coefficients for all modes in section B, therefore the overall solution will be the superposition of all possible combinations.

C. Efficient Characterization of Passive Isolation Structures

In practice, the source current density distribution will excite all supported modes in section A to a greater or lesser extent, depending on its size, shape, frequency, orientation, etc. Therefore, the previous system of equations has to be solved \( t \times p \) times. Taking into account the complete modal expansion of fields in semi-open waveguides, described by (1)-(4), a discretization of the continuous spectrum should be performed and a subsequent truncation of the number of equations is required to be able to tackle the problem. Nevertheless, the purpose of the proposed model is not to obtain an accurate representation of the fields in each section, but to obtain the transfer function between aggressor and victim. For this reason, there is no need to consider the whole set of supported modes. Instead, it is sufficient to consider those discrete modes and spectral components in each section that effectively contribute to substrate coupling and/or account for the maximum power transfer between sections A and C.

The required number and type of modes will vary depending on the substrate type. A study of when propagation in IC substrates can be described only with discrete modes, in terms of frequency, doping profile, and distance to the source can be found in [9]. Generally speaking, the higher the conductivity in layer \( m \), the longer the distance to the source in which the dominant modes are the ones confined in the substrate stack and, therefore, interference propagation can be described with a few discrete modes.

The layer in which circuits are implemented (active layer), that corresponds to layer \( m \) in sections A and C in Fig. 1, is normally highly-doped. Therefore, the doping profile where both aggressor and victim are located can usually be characterized with discrete modes. On the other hand, the isolation stack top layer is a charge-free or dielectric medium, as its function is precisely to interrupt the surrounding active layer. Thus the dominant modes in section B consist of radiation modes. Although non-dominant, other propagation mechanisms also supported by this type of doping profiles are discrete guided complex modes [18].

Our approach for a fast electromagnetic characterization of passive isolation structures is to perform a smart selection of the propagation modes with the maximum contribution to the transfer function between aggressor and victim. If very few modes are used to characterize each section, the system of equations proposed in Sec. II.B is easily solvable. Because of the relatively high conductivity of the active layer, the selection of modes in sections A and C is straightforward. The corresponding eigenvalue equation [9] will provide the propagation coefficients of the relevant propagation modes, as those that do not propagate belong to the evanescent radiation subset.

Analogously, in section B one can derive an eigenvalue equation that will provide the propagation coefficients of the discrete guided complex modes. Even though non-dominant within the typical size range of passive isolation structures, these will contribute to the transfer function, as they effectively propagate through the structure. In addition to discrete complex modes, we will also consider the spectral component that
accounts for the maximum power transfer. Because passive isolation structures are located very close to the source of interference, we can neglect the propagating radiation modes and only consider evanescent spectral components.

III. ISOLATION CHARACTERIZATION

Having obtained the modal amplitude coefficients at aggressor and victim sections, there are different strategies to characterize the isolation capabilities of a structure. Each of them takes into account different isolation mechanisms and their comparison provides a valuable insight on the role that they play. These are:

- The voltage ratio calculated as the integral of the total electric field at aggressor and victim locations \( V(\text{victim})/V(\text{source}) \) provides a transfer function that takes into account the effect of the attenuation due to the distance between them, the parallel and transverse isolation layers, and the transitory effects due to the interaction between the modes considered. This transfer function is equivalent to a simulated or measured S21.
- The ratio of the fundamental mode amplitude coefficient at victim and aggressor sections \( (a_{C1}/a_{A1}) \) accounts for the effect of the parallel and transverse isolation layers. For this reason it allows us to de-embed the effect of the attenuation due to distance, which can vary significantly with frequency, especially in substrates with parallel isolation layers [9].
- The ratio of the fundamental mode amplitude coefficient at victim section with and without isolation structure \( (a_{C1}/a_{C1}) \) is defined as its isolation effectiveness. This provides information about the isolation capabilities of the transverse discontinuity as a function of the substrate in which it is embedded. In addition, it allows us to de-embed the effect of other isolation structures, if present.

IV. SYMMETRIC STUDY CASE: MODELING OF TRENCHES IN SOI SUBSTRATE

We define a symmetric structure as the one in which aggressor and victim sections have the same doping profile, such as the case of a SiO2 trench implemented in a silicon over insulator (SOI) substrate. The frequency range of interest is defined as from DC up to 10 GHz, in order to cover the majority of EMC sensitive applications (e.g., RF, mixed-signal). We will study the isolation capabilities of trenches as a function of the surrounding doping profile in which they are implemented. In order to do this, we first start from a standard case [9] and obtain the relevant propagation coefficients for this substrate type in the frequency range of interest (see Fig. 3). The absolute height, conductivity and relative permittivity used are shown in Table I. Section B consists of two layers, however, the second layer has been artificially split in order to make the calculation of coefficients described by (16) easier.

From the values shown in Fig. 3, one can conclude that for almost all the frequency range of interest, the fundamental mode (referred as mode 1 in the figure) is the dominant one. There is a second mode of higher order and with higher negative imaginary part of the propagation coefficient that, consequently, is attenuated more rapidly. This mode starts to become more relevant from approximately 200 MHz. We can observe different resonances due to the presence of the isolation layer. In section B, and for the frequency range of interest, only a discrete complex mode (the fundamental mode) is obtained, so we will characterize this section with one discrete mode plus the spectral component responsible for the maximum energy transfer. According to this, we will use two modes to characterize each section \( (t = p = q = 2) \).

Because the overall system of equations is considerably reduced, it is possible to solve it for different values of spectral components in section B, in order to determine which is the one that contributes the most to the overall transfer function or, in other words, the spectral component that is better matched with modes in sections A and C. In addition, one should expect that the value of this mode propagation coefficient would be of a similar magnitude to the ones in the adjacent sections. For this reason, there is no need to discretize the whole spectrum of possible values, but just to establish a window centered at the value of the imaginary part of the propagation coefficients that we want to match.

Nevertheless, in order to simplify this procedure and avoid iterations, one can take advantage of the symmetry of the structure to optimize the search. In this case, it is logical to assume that the mode responsible for the maximum power transfer at the first interface A-B, will also account for the maximum contribution at the second interface B-C. Therefore, the problem is simplified to just one discontinuity, in the following manner: the discontinuity A-B is excited only with the fundamental mode A1, and we will assume that all the values shown in Fig. 3, one can conclude that for almost all the frequency range of interest, the fundamental mode (referred as mode 1 in the figure) is the dominant one. There is a second mode of higher order and with higher negative imaginary part of the propagation coefficient that, consequently, is attenuated more rapidly. This mode starts to become more relevant from approximately 200 MHz. We can observe different resonances due to the presence of the isolation layer. In section B, and for the frequency range of interest, only a discrete complex mode (the fundamental mode) is obtained, so we will characterize this section with one discrete mode plus the spectral component responsible for the maximum energy transfer. According to this, we will use two modes to characterize each section \( (t = p = q = 2) \).

Because the overall system of equations is considerably reduced, it is possible to solve it for different values of spectral components in section B, in order to determine which is the one that contributes the most to the overall transfer function or, in other words, the spectral component that is better matched with modes in sections A and C. In addition, one should expect that the value of this mode propagation coefficient would be of a similar magnitude to the ones in the adjacent sections. For this reason, there is no need to discretize the whole spectrum of possible values, but just to establish a window centered at the value of the imaginary part of the propagation coefficients that we want to match.

Nevertheless, in order to simplify this procedure and avoid iterations, one can take advantage of the symmetry of the structure to optimize the search. In this case, it is logical to assume that the mode responsible for the maximum power transfer at the first interface A-B, will also account for the maximum contribution at the second interface B-C. Therefore, the problem is simplified to just one discontinuity, in the following manner: the discontinuity A-B is excited only with the fundamental mode A1, and we will assume that all the values shown in Fig. 3, one can conclude that for almost all the frequency range of interest, the fundamental mode (referred as mode 1 in the figure) is the dominant one. There is a second mode of higher order and with higher negative imaginary part of the propagation coefficient that, consequently, is attenuated more rapidly. This mode starts to become more relevant from approximately 200 MHz. We can observe different resonances due to the presence of the isolation layer. In section B, and for the frequency range of interest, only a discrete complex mode (the fundamental mode) is obtained, so we will characterize this section with one discrete mode plus the spectral component responsible for the maximum energy transfer. According to this, we will use two modes to characterize each section \( (t = p = q = 2) \).

Because the overall system of equations is considerably reduced, it is possible to solve it for different values of spectral components in section B, in order to determine which is the one that contributes the most to the overall transfer function or, in other words, the spectral component that is better matched with modes in sections A and C. In addition, one should expect that the value of this mode propagation coefficient would be of a similar magnitude to the ones in the adjacent sections. For this reason, there is no need to discretize the whole spectrum of possible values, but just to establish a window centered at the value of the imaginary part of the propagation coefficients that we want to match.

Nevertheless, in order to simplify this procedure and avoid iterations, one can take advantage of the symmetry of the structure to optimize the search. In this case, it is logical to assume that the mode responsible for the maximum power transfer at the first interface A-B, will also account for the maximum contribution at the second interface B-C. Therefore, the problem is simplified to just one discontinuity, in the following manner: the discontinuity A-B is excited only with the fundamental mode A1, and we will assume that all

![Fig. 3](image-url)  
Fig. 3. Radial propagation coefficients of sections A and C in the symmetric case: (a) Real part; (b) Imaginary part.

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>SYMMETRY STUDY CASE PARAMETERS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Section</td>
<td>A</td>
</tr>
<tr>
<td>Radius</td>
<td>200 μm</td>
</tr>
<tr>
<td>Layer 1</td>
<td>( h=250 ) μm</td>
</tr>
<tr>
<td>σ</td>
<td>10 S/m</td>
</tr>
<tr>
<td>εr</td>
<td>11.9</td>
</tr>
<tr>
<td>Layer 2</td>
<td>( h=252 ) μm</td>
</tr>
<tr>
<td>σ</td>
<td>0 S/m</td>
</tr>
<tr>
<td>εr</td>
<td>3.9</td>
</tr>
<tr>
<td>Layer 3</td>
<td>( h=253 ) μm</td>
</tr>
<tr>
<td>σ</td>
<td>1000 S/m</td>
</tr>
<tr>
<td>εr</td>
<td>11.9</td>
</tr>
</tbody>
</table>
reflected power goes to the same mode. The energy transferred
to section B will go to two modes: B1 is the fundamental
discrete complex mode of the section and B2 is the spectral
component to optimize. In this scenario, we will select the
propagation coefficient for B2 that minimizes the reflection
coefficient of A1, given by the following expression:

$$\rho_{A1\rightarrow A1} = \frac{QH_r^2(k_{rB1}r_{A1}) - k_{B1}Y_{A1A1}H_0^2(k_{rA1}r_A)}{k_{rA1}Y_{A1A1}H_0(k_{rA1}r_A) - QH_r^2(k_{rA1}r_A)}.$$  \hspace{0.5cm} (24)

where

$$Q = \sum_{i=1}^{n} \frac{Y_{B1B1}^2(k_{rB1}r_A)}{Y_{B1B1}^2(k_{rB1}r_A)}.$$  \hspace{0.5cm} (25)

In order to assess the validity of the assumptions made, finite
element method (FEM) simulations have been performed using
the electromagnetic (EM) full-wave software CST-Microwave
Studio. We choose this simulation tool, as it allows us to obtain
fields induced inside lossy dielectric volumes. Moreover, pre-
vious studies have shown very good agreement between FEM
simulations and measurements of on-chip passive isolation
structures [2]. Fields induced in the substrate due to a current
distribution of \( J = I \delta(r) \hat{z} \) from \( z = 0 \) to \( z = h_3 \) are simulated
and compared to the ones obtained with the proposed model.
Results obtained for the parameters in Table I are shown in
Fig. 4, where the trench location is highlighted with a vertical
line.

Figure 5 shows the results obtained for the same structure
but with a thicker active layer (\( h_3 = 254 \mu m \)). As can be seen,
there is very good agreement between calculated results and
simulations. After the trench, one can observe a transitory de-
crease of the field magnitude. This is due to the fact that higher
order modes are excited due to mode conversion, that interact
destructively with the fundamental mode and, therefore, the
total electric field magnitude is transitorily decreased. Because
they are attenuated faster, the effect disappears after a certain
distance, where the level of interference is then determined
only by the fundamental mode. The length and magnitude of
the electric field drop is frequency dependent, being more
significant at lower frequencies, where evanescent modes play
a much more significant role. For a thicker active layer (Fig. 5),
the substrate resonance is slightly shifted to the left compared
to the one obtained for the first study case (Figs. 3 and 4),
being now centered at 100 MHz. Due to this, the electric
field drop at 100 MHz is more significant. If additional modes
from the continuous spectrum are included in section C, the
proposed model allows the representation of this phenomena.
However, this would increase the size and complexity of the
system of equations. After the transitory behavior, considering
only two modes in each section, the deviation of the predicted
electric field for the 1 \( \mu m \) thick active layer case is less than
1 dB. As for the 2 \( \mu m \) thick active layer case, the worst case
deviation of the predicted value is 4 dB. Full-wave simulation
time was approximately three hours for five frequency points,
while computation time was one minute and 15 seconds for
50 frequency points in the same computer.
Figure 6 shows the trench transfer function calculated as the voltage ratio at different locations, and as the ratio of the fundamental mode amplitude coefficient at aggressor and victim sections. Results show that the attenuation due to the distance is frequency dependent. This is due to the resonances caused by the presence of the parallel isolation layer. In addition, at higher frequencies the transfer function value calculated as the ratio of the fundamental mode amplitude coefficient at aggressor and victim sections (distance de-embedding) is above 0 dB. This is due to the fact that energy that previously belonged to mode 2 and therefore was rapidly attenuated, is transferred to the fundamental mode because of the mode conversion phenomena. This mode is able to propagate interference more efficiently and thus, in this frequency range, the trench is not blocking but enhancing substrate coupling.

Figure 7 shows the isolation effectiveness of a SiO$_2$ trench for different active layer conductivities ($\sigma_a$). One can observe that the trench isolation effectiveness remains almost constant at low frequencies, due to the presence of the parallel SiO$_2$ layer. Small variations of the active layer conductivity have a significant impact on the isolation effectiveness at higher frequencies, where the effect of mode conversion is more important. As it can be seen, the comparison of the three contribution of the different isolation mechanisms involved.

In addition, one can identify in which cases the isolation is provided by a transitory phenomena. When transitory isolation peaks occur due to destructive mode interaction, increasing the distance to the source of interference would decrease the isolation, which is a counterintuitive behavior.

V. NON-SYMMETRIC STUDY CASE: MODELING OF DEEP N-WELLS

A non-symmetric structure is one in which the doping profiles of aggressor and victim sections are not equal, as in the case of a deep n-well inserted in a p-type substrate. In this study case, we will consider a reverse-biased single junction structure [2]. Thus, section A doping profile will have a parallel depleted layer between the substrate itself and the active layer, while section B will consist of a transverse depleted region on top of the substrate, interrupting the active layer. The parameters used in this study case are shown in Table II. Even though sections A and C do not have the same doping profile, they both have a highly-doped active layer. This implies that the same type of modes will describe propagation in both sections, having similar propagation coefficients in the frequency range far from the resonances caused by the depleted layer (see Fig. 8).

![Graph](image1)

**Fig. 6.** Transfer function of a 1 $\mu$m wide SiO$_2$ trench at r=200 $\mu$m in a SOI substrate

![Graph](image2)

**Fig. 7.** Isolation effectiveness of a 1 $\mu$m wide SiO$_2$ trench at r=200 $\mu$m in a SOI substrate

![Graph](image3)

**Fig. 8.** Radial propagation coefficients of sections A and C in the non-symmetric case: (a) Real part; (b) Imaginary part.

Following the same approach as in the symmetric case, we will characterize section B with the fundamental discrete complex mode plus the spectral component of the evanescent radiation modes that contributes the most to the structure transfer function. However, the considerations made in Sec. III in order to simplify the search, may not be accurate for the whole
frequency range, as the present study case is not symmetrical. For this reason, a different approach is considered in order to optimize the procedure for non-symmetrical structures. In this case, we consider a single discontinuity A-B. Section A is characterized with $t$ modes and section B is characterized only with the spectral component to be optimized. The amplitude coefficient of a spectral component $k_{rB}$ for a certain excitation mode $A_e$ corresponds to the following expression

$$C_{B \rightarrow Ae} = \frac{C_{Ae}}{\sqrt{Y_{\delta A \rightarrow \delta Ae}}} \sum_{i=1}^{t} Y_{A_i \rightarrow A} \frac{Y_{B \rightarrow A_i} H_1^{i}(k_{rB}r_{Ai}^{A})}{H_0^{i}(k_{rB}r_{Ai}^{A}) - Q k_{rB} H_0^{i}(k_{rB}r_{Ai}^{A})} \left( H_1^{i}(k_{rB}r_{A}^{A}) - H_0^{i}(k_{rB}r_{A}^{A}) \right),$$

where

$$Q = \sum_{i=1}^{t} \frac{Y_{B \rightarrow A_i} H_1^{i}(k_{rB}r_{A}^{A})}{Y_{A_i \rightarrow A} k_{rB} H_0^{i}(k_{rB}r_{A}^{A})}.$$  

(26)

In this scenario, we will select the spectral component $k_{rB}$ with the maximum amplitude coefficient, when all modes exciting the discontinuity are considered simultaneously

$$\max \left( \sum_{i=1}^{t} C_{B \rightarrow A_i} \right).$$  

(28)

EM simulations have been performed using CST-Microwave Studio, in order to assess the validity of the assumptions made in this study case. The source and simulation configurations are the same as the one used in Sec. III. Fields induced in the substrate have been obtained and compared to the ones calculated with the proposed model. Figure 9 corresponds to

the results obtained with the parameters shown in Table II. The location of the transverse depleted region has been highlighted with a line.

There is very good agreement between calculated and simulated results. One can also see the transitory decrease of the electric field magnitude after the discontinuity, due to destructive mode interaction. However it is not as significant as the decrease shown for the trench study case. After the transitory behavior, the worst case deviation of the predicted electric field is less than 2 dB. Simulation and computation times are similar to the ones in the previous section.

Due to the fact that aggressor and victim sections are not characterized by the same modes, the transfer function of the structure cannot be obtained by calculating the ratio between modal amplitude coefficients in sections A and C. Therefore, it is obtained by means of the voltage ratio (see Fig. 10). However, one should take into account that the attenuation due to the distance is also included. As it can be seen in Fig. 10, for different victim locations and when there are no isolation layers, as in the case of section C, the difference in attenuation levels between curves is almost constant over the whole frequency range of interest. This is not the case for SOI substrates, where the attenuation difference at different victim locations is highly dependent on frequency (see distance de-embedding procedure in Fig. 6). In order to obtain the isolation effectiveness, we calculate

Fig. 9. Amplitude of the induced z-component of the electric field due to a
current of I=1 A, at x=240 µm, in a deep N-well with a 1 µm thick depleted
region, for (a) f=100 MHz, (b) f=1 GHz, (c) f=10 GHz.

Fig. 10. Transfer function of a deep N-well with a 1-µm wide transverse
deprecated region at $r=99$ µm.
structure transfer function and isolation effectiveness behavior. The closer that the transverse discontinuity is to the source, the more significant is the effect of mode interaction, as the higher order mode (mode 2) reaches the discontinuity less attenuated. This has an effect on the isolation peak amplitude, however the peak remains centered at the same frequency point, as this depends mainly on the depleted region width. In addition, for high frequencies it can be seen that the deep N-well stops being effective, and interference is enhanced due to mode conversion. Figure 12 shows the calculated isolation effectiveness for different thicknesses of both parallel and transverse depleted regions. As expected, the thickness impacts the cutoff frequency of the deep N-well. Furthermore, for thicker depleted regions the effect of destructive mode interaction is more significant and also occurs at higher frequencies.

VI. CONCLUSION

In this paper we have developed a complete modeling methodology that enables fast electromagnetic characterization of integrated passive isolation structures based on interference blocking. It is based on the characterization of the different doping profiles by means of the supported propagation modes. We have modeled isolation structures as waveguide discontinuities transverse to the direction of interference propagation, presenting a formulation for a general number of propagation modes in each section. Subsequently, those propagation modes that dominantly contribute to the aggressor-victim transfer function have been considered to define the system of equations to be solved.

The proposed model has been applied to two study cases. One in which the victim and the aggressor cross-sections have equal doping profiles and a second one with different doping profiles. We have shown that with a smart selection of just a few of the most representative modes in each section, the problem can be significantly simplified and the model is still able to quickly provide reasonably accurate transfer functions. However, if a more accurate representation of fields in all sections is required, more modes can easily be included. In addition, we have proposed strategies to optimize the mode selection in those regions in which the continuous spectrum dominates. With these, we can perform an early prediction of the isolation effectiveness of these type of structures avoiding the need of calibration by means of extraction tools or experimental tests. The proposed model provides insight on the different existing isolation mechanisms such as attenuation due to distance, and parallel and transverse isolation layers, allowing us to quantify the contribution of each of them.

The calculated results have been compared to full-wave simulations, obtaining very good agreement in both symmetric and non-symmetric cases. We have also shown that effects due to the multimodal characteristic of interference propagation in integrated circuit substrates already occur at a few hundred MHz. These can cause a transitory increase of the isolation effectiveness, but also enhance interference coupling, especially for frequencies above 1 GHz.

The proposed generalized case is based on three consecutive sections (aggressor, isolation and victim). In future work, this will be extended to a general number of sections, in order to characterize more complex structures (such as triple wells) or cascading multiple substrate sections with isolation structures. This will allow us to represent more realistic scenarios, enabling the experimental validation of the developed methodology.

REFERENCES


Mercé Grau Novellas received the Bachelors degree in telecommunications engineering and the Masters degree in micro and nanoelectronics engineering both from Universitat Autonoma de Barcelona (UAB), Barcelona, Spain. She is currently working toward the Ph.D. degree at the Eindhoven University of Technology, Eindhoven, Netherlands, working on the implications of the technology choice on the EMC performance in automotive IC products and applications. During her Master studies, she joined the Antenna and Microwave Systems Group, UAB, working on design and characterization of metamaterial structures.

Ramiro Serra received the B.S. degree in electronic engineering from the Instituto Tecnologico de Buenos Aires, Argentina, in 2000, the postmaster degree specializing in technological applications of nuclear energy from Instituto Balseiro, Bariloche, Argentina, in 2004, and the Ph.D. degree in electronics and communications engineering from Politecnico di Torino, Torino, Italy, in 2009. He is currently an Assistant Professor within the Electrical Energy Systems group at the Eindhoven University of Technology, Eindhoven, Netherlands. Dr. Serra is the Chair of the International Union of Radio Science (URSI) Commission E (Noise and Interference) and the Secretary of the General National URSI Committee for The Netherlands. He is a member and work package champion of the SC 77B/CISPR-A joint working group for the standard IEC 61000-4-21 on Reverberation Chambers and member of the CIRED/CIGRE joint working group C4.31 on EMC between communication circuits and power systems.