Modeling of Microwave Transmission Lines Considering Frequency-Dependent Current Distribution Effects

by

Diego Mauricio Cortés Hernández

A thesis submitted in partial fulfillment of the requirements for the degree of

DOCTOR OF PHILOSOPHY IN ELECTRONICS

Instituto Nacional de Astrofísica, Óptica y Electrónica.

Tonantzintla, Puebla

Supervised by:

Reydezel Torres Torres, Ph.D.

© INAOE 2017

The author grants INAOE the right to use and reproduce fully or partially the work being presented.
ABSTRACT

This thesis is focused on the modeling and characterization of planar interconnects operating at microwave frequencies covering both, integrated circuit (IC) and printed circuit board (PCB) technologies. In particular, the state-of-the-art of the field is advanced by providing physically-based models incorporating the accurate representation of the series effects (i.e., resistance and inductance). In this regard, the current distribution effects, including the skin and proximity effects, are taken into consideration. The analysis starts with the premise that an accurate modeling of the metal losses in microwave transmission lines for IC and PCB technologies, including frequency-dependent current distribution effects is essential to predict the actual response of a whole transmission channel. From the resulting analysis, three frequency regions were identified, mathematical expressions that relate the cross-section and the electrical characteristics of the structure were developed, and the corresponding accuracy is verified through comparisons with simulations carefully performed on full-wave solvers. Furthermore, dominant effects within different frequency ranges were identified, which allows for the development of new methodologies for the electrical characterization of dielectric materials, and pad de-embedding. All the models and methodologies were applied to single-ended and differential microstrips and striplines, as well as to metal-insulator-metal (MIM) test fixtures used for the characterization of thin dielectric films. In this regard, special attention was paid to appropriately consider the correct propagating modes occurring in the analyzed structures, and also to the scalability of the models for $R$ and $L$. 
ACKNOWLEDGEMENTS

To God for helping me and always be with me.

I would like to thank the Consejo Nacional de Ciencia y Tecnología (CONACyT) and the Instituto Nacional de Astrofísica, Óptica y Electrónica (INAOE) for giving me support during my studies.

To my wife Claudia for his guidance, constant support and encouragement in my life.

To my parents, Hector and Maria Melba by their unconditional love and because always have believed in me.

To my brothers Hector, Angelica, Carlos, Johan, and Sofia by their motivation and be a key part of my life.

To my family, but especially to Norbey, Magda, Patricia, and Jorge by their constant motivation and inspiration.

To my advisor and friend, Dr. Reydezel Torres-Torres, his support and trust gave me the freedom to explore on my own and his guidance gave me the tools to focus in my aim and bring support to others.

This Ph.D project has benefited from discussions with several helpful people, and I want to thank them. Thanks Dr. Roberto Murphy, Dr. Alfonso Torres, Dr. Mónico Linares, Dr. Joel Molina, and Dr. Jorge Carballido.
To my friends Alexander, Diego Felipe, Carlos Eduardo, Mauricio, Laura, Jacqueline, Luz, and Gabriela by always standing by me and give me their unconditional friendship during all this time.
This thesis is dedicated to my parents Maria Melba and Hector for their endless love, guidance and encouragement.
# CONTENT

1. MICROWAVE TRANSMISSION LINES..................................................................................................................8  
  1.1 Modeling interconnection Links..................................................................................................................9  
    1.1.1 Compact modeling ...............................................................................................................................10  
    1.1.2 Equivalent circuits ...............................................................................................................................10  
    1.1.3 Electromagnetic (EM) simulations .........................................................................................................11  
  1.2 Aspects to be considered in the modeling ...............................................................................................12  
    1.2.1 Homogeneous and inhomogeneous transmission lines ....................................................................13  
    1.2.2 Electrical transition and terminations ...............................................................................................13  
  1.3 Object of study ............................................................................................................................................14  
  1.4 Effects associated with the metal traces .................................................................................................15  
    1.4.1 DC losses ............................................................................................................................................15  
    1.4.2 Skin effect ...........................................................................................................................................16  
    1.4.3 Surface roughness ...............................................................................................................................18  
    1.4.4 Proximity effect ...................................................................................................................................19  
  1.5 Single-ended versus differential signaling schemes ................................................................................20  
    1.5.1 Single-mode propagation .....................................................................................................................20  
    1.5.2 Edge-coupled interconnects ................................................................................................................21  
  1.6 Purpose of this project ...............................................................................................................................22  

2. DESCRIPTION OF THE RLGC MODEL .........................................................................................................23  
  2.1 C and G Parameters ..................................................................................................................................24  
  2.2 R and L Parameters ..................................................................................................................................25  
    2.2.1 Single-ended transmission lines .........................................................................................................25  
    2.2.2 Edge-coupled transmission lines .........................................................................................................28  
  2.3 Characteristic impedance and propagation constant ..............................................................................30  
  2.4 Summary ...................................................................................................................................................31
<table>
<thead>
<tr>
<th>Chapter</th>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>3. ANALYSIS ON IC</td>
<td>3.1 MIM (Metal-Insulator-Metal)</td>
<td>33</td>
</tr>
<tr>
<td></td>
<td>3.2 Single-ended homogeneous transmission lines</td>
<td>34</td>
</tr>
<tr>
<td></td>
<td>3.3 Pad de-embedding</td>
<td>40</td>
</tr>
<tr>
<td></td>
<td>3.4 Summary</td>
<td>47</td>
</tr>
<tr>
<td>4. ANALYSIS ON PCB</td>
<td>4.1 Single-ended homogeneous transmission lines</td>
<td>61</td>
</tr>
<tr>
<td></td>
<td>4.1.1 Standard and low-loss laminates</td>
<td>62</td>
</tr>
<tr>
<td></td>
<td>4.1.2 Ultra-low-loss laminates</td>
<td>64</td>
</tr>
<tr>
<td></td>
<td>4.2 Differential homogeneous lines</td>
<td>69</td>
</tr>
<tr>
<td></td>
<td>4.3 Summary</td>
<td>77</td>
</tr>
<tr>
<td>5. CONCLUSIONS</td>
<td>5.1 Modeling of microwave transmission channels</td>
<td>78</td>
</tr>
<tr>
<td></td>
<td>5.2 Electrical characterization of materials</td>
<td>80</td>
</tr>
<tr>
<td></td>
<td>5.3 Pad de-embedding</td>
<td>81</td>
</tr>
<tr>
<td>6. PUBLICATION LIST</td>
<td>6.1 Journal</td>
<td>82</td>
</tr>
<tr>
<td></td>
<td>6.2 Conference Publications</td>
<td>83</td>
</tr>
<tr>
<td>BIBLIOGRAPHY</td>
<td></td>
<td>84</td>
</tr>
</tbody>
</table>
LIST OF FIGURES

Fig. 1. Interconnection channel formed between two devices. The transitions mostly introduce reflections whereas the main path introduces the losses and the delay in the signal. ........................................................................................................................................................................ 9

Fig. 2. Photographs of the cross section of a) single-ended and b) coupled transmission lines. .................................................................................................................................................................................................................. 10

Fig. 3. Some modeling approaches for representing the behavior of transmission lines: a) analytical modeling [26], b) equivalent electrical model, and c) full-wave simulations. ........................................................................................................................................................................ 11

Fig. 4. Homogeneous and inhomogeneous transmission lines. .................................................................................................................................................................................................................. 13

Fig. 5. Examples of discontinuities: bends and pads. .................................................................................................................................................................................................................. 14

Fig. 6. Change in the current distribution from low (DC region) to high frequencies within the cross section of a conductor with circular cross section. Notice that the current is more intense near the surface due to the skin effect. ........................................................................................................................................ 16

Fig. 7. Sketch depicting the surface roughness at the metal-to-dielectric interface of a microstrip transmission line. .................................................................................................................................................................................................................. 18

Fig. 8. Proximity effect occurring between two adjacent wires carrying opposite current flows. Notice that more current is confined in the inner faces as the spacing decreases. ........................................................................................................................................................................................................ 19

Fig. 9. Configuration of the electric and magnetic fields in: a) single-ended stripline and b) ground-coplanar waveguide (G-CPW). ........................................................................................................................................................................................................ 20

Fig. 10. Configuration of the electric and the magnetic field in edge-coupled striplines presenting: a) odd, and b) even propagation modes. ........................................................................................................................................................................................................ 21

Fig. 11. RLGC model of a single-ended transmission line. .................................................................................................................................................................................................................. 24
Fig. 12. Full-wave simulation showing the operating frequency regions for a microstrip TL. 25

Fig. 13. Frequency-dependent current distribution in edge-coupled stripline when operated in odd mode. 29

Fig. 14. Frequency-dependent current distribution in equally-spaced edge-coupled striplines when operated in even mode. 30

Fig. 15. Sketch representing three-quarters of the MIM structure used for dielectric characterization; the corresponding equivalent circuit is also shown. 34

Fig. 16. Regression of experimental $\omega \text{Re}[Z_{DUT}]$ versus $\omega$ data corresponding to Al$_2$O$_3$ to extract the parasitic series resistance $R$. 37

Fig. 17. Regression of experimental $\omega \text{Im}[Z_{DUT}]$ versus $\omega^2$ data for Al$_2$O$_3$ to extract the parasitic series inductance $L$. 37

Fig. 18. Frequency-dependent extracted parameters for HfO$_2$. Notice that $\varepsilon_r$ and tan$\delta$ are decreasing with the thermal treatment. 38

Fig. 19. $S_{11}$ parameter of HfO$_2$ without TT, notice that a good agreement between experimental data and model is obtained. 39

Fig. 20. Conceptual representation of the current concentration within the medium-frequency region; the parameters used to represent the corresponding frequency dependence of the current distribution are included. 40

Fig. 21. a) Plot of $R$ versus $f$, and b) experimental curve of $L$ used to extract the parameters in the medium-frequency region. 43

Fig. 22. Experimental data used to obtain $L_{\text{ext}}$ through a linear regression. 44

Fig. 23. $R$ and $L$ versus frequency curves showing the validation with full-wave simulation. 44
Fig. 24. $R$ and $L$ versus frequency curves showing the confrontation of the proposed model with experimental data. ................................................................. 45

Fig. 25. Sketch detailing a test fixture for probing a DUT. Notice the conceptual depiction of the current path in the ground plane at microwave frequencies. .......... 47

Fig. 26. Schematic representation of the current crowding effect within the cross section of the trace connecting the signal pad with the DUT. The current distribution within the ground plane laterally decreases due to the skin effect, impacting the conductor laterally ($\delta_w$) and vertically ($\delta_t$). ................................................................. 48

Fig. 27. Simulated current distribution once the effective width of the ground plane becomes frequency dependent. Three different heights for the signal trace and two different frequencies were considered. The simulations were performed considering $w_t = 20 \mu m$, $t_t = 0.5 \mu m$, and $t_g = 1.6 \mu m$. ................................................................. 51

Fig. 28. Equivalent circuit representation of a measured device. ........................................ 56

Fig. 29. S-parameters and intrinsic inductance for the prototyped inductor processed with different de-embedding methods................................................................. 57

Fig. 30. Complex propagation constant data and characteristic impedance for the fabricated transmission line obtained after processing with different de-embedding methods. ................................................................. 58

Fig. 31. Sketch of single-ended a) microstrip, and b) stripline interconnects. ............. 64

Fig. 32. Effective tan$\delta$ and $\varepsilon_r$ for the microstrip line........................................ 65

Fig. 33. tan$\delta$ and $\varepsilon_r$ for the stripline. ................................................................. 65

Fig. 34. $L$ and $C$ extracted from experimental data. Notice the resonances impacting experimental data................................................................. 68

Fig. 35. Plot of $\alpha f$ vs square root of $f$; linear curve fitting is used in order to extract $k$ and tan$\delta$ using (63). .................................................................................. 68
Fig. 36. Comparison between experimental data and the transmission line model. Notice that if tan\(\delta\) is neglected, the insertion loss at high frequencies is smaller than the obtained by experimental results. .............................................................. 69

Fig. 37. A modeling approach to representing the current distribution and inductance loops for the case of a differentially operated interconnect. ................................................................. 70

Fig. 38. A modeling approach to representing the current distribution and inductance loops for the case of common mode operation. ................................................................. 72

Fig. 39. Experimentally determined a) \(R\) and b) \(L\) for coupled stripline pairs when operated in differential mode. ................................................................................................................... 74

Fig. 40. Experimentally determined a) \(R\) and b) \(L\) for coupled stripline pairs when operated in common mode. ................................................................................................................... 75
LIST OF TABLES

Table 1. Extraction of the model parameters for $R$ and $L$ in differential mode............ 74

Table 2. Extracted $R$ and $L$ data in common mode............................................. 75
MUCH OF THE PROGRESS in IC technology is related to the scaling of CMOS technology. In this regard, decreasing the feature size of semiconductor devices is what has allowed increasing the clock frequencies following Moore's Law [1]. Furthermore, some of the processed signals present microwave frequencies [2], [3], and propagate within the IC, the package substrate, and the PCB laminate. In general, the interconnects used for guiding signals at these frequencies are analyzed as microwave transmission lines [2], and the corresponding performance is determined by their composing materials (i.e. metal and dielectrics), and certainly by the geometry of its structure. In consequence, for carrying out a successful design of the interconnects in electronics systems, electromagnetic models are used for different types of simulations. From a physically-based perspective, the most accurate representations include very complex phenomenon such as current distribution and dispersion effects. In fact, the effects related with the dielectric materials different experimental techniques have been developed to directly extract the associated parameters from measurements [4], [5], [6], [7], [8], whereas for the case of the effects associated with the conductors, important research is being currently more intensely performed. This is because the recent changes in the manufacturing processes and the increase in the operating frequency, which impact the current flow in the metal structures (i.e. signal strip and ground plane), strongly determine the performance of the interconnects [9], [10], [11], [12]. In this chapter, different considerations in the modeling of microwave transmission channels and the goal of the research project is presented: The premise is that an accurate modeling of the metal losses in microwave transmission lines for IC and
1. MICROWAVE TRANSMISSION LINES

*PCB technologies, including frequency-dependent current distribution effects helps to predict the actual response of a whole transmission channel.*

### 1.1 Modeling interconnection Links

Fig. 1 presents a sketch of a typical transmission channel; it is formed to communicate two devices, Device 1 and Device 2. Notice that the signal is propagating through the integrated circuit (IC), the package, and the printed circuit board (PCB). Typically, the interconnection can be represented by means of segmented transmission lines and electrical transitions such as vias, bends, and solder balls. Even though the transitions introduce undesired signal reflections and may also dissipate some power, the planar transmission lines represent the main contribution of losses and delay in the signal path due to considerable dispersion and dissipation effects. This is due to the fact that, although the transitions are impacting the performance of the interconnect due to undesired changes of impedance, the signal propagates through relatively long distances within the transmission lines. In this regard, the interconnect channels are formed by structures such as single-ended and edge-coupled transmission lines, as those illustrated in Fig. 2. The representation of these interconnects for modeling purposes can be obtained by applying different approaches, such as those illustrated in Fig. 3, which are briefly discussed hereafter.

![Interconnection channel sketch](image)

**Fig. 1.** Interconnection channel formed between two devices. The transitions mostly introduce reflections whereas the main path introduces the losses and the delay in the signal.
1.1.1 Analytical modeling

Although planar transmission lines can be successfully analyzed using commercial electromagnetic field simulators, this approach is reserved for detailed analysis of particular sections of an interconnect very long simulation time is required. In contrast, using closed-form expressions for calculating the electrical response of lines (such as that shown in Fig. 3a) can help to significantly speed up the analysis and simulation process for complex circuits [13], [14], [15]. For this reason, designers prefer using this type of approximations when the validity for a wide frequency range and causality in the time domain is guaranteed [9], [16]. Therefore, many analytic equations have been obtained in order to characterize and model uniform interconnections; these equations are important when deriving synthesis formulas that can be used in the electromagnetic design.

1.1.2 Equivalent circuits

Equivalent circuits (see Fig. 3b) are obtained studying the topology of the interconnection. Basically, this representation is a good alternative to establish a relationship between parasitics, geometry, and the signal frequency. Although equivalent circuit models have been used as a workhorse in integrated-circuit design due to the intuitive concepts required for the corresponding development and use [17], [18], [19], conventional simplified models cannot capture the underlying physics
involved in high frequencies. Nevertheless, when appropriately developed, equivalent circuits can be used for advanced characterization and modeling at these frequencies provided that the appropriate incorporation of these effects is carried out [20], [21].

1.1.3 Electromagnetic (EM) simulations

EM simulators numerically solve Maxwell's equations when a complex structure is divided into thousands and even millions of cells within a meshed model. For this purpose, differential, integral, and even both of these types of equations are

\[ Z_0 = \frac{\eta}{2\pi} \ln \left[ \frac{\xi h}{w} + \sqrt{1 + \left( \frac{2h}{w} \right)^2} \right], \quad \xi = 6 + (2\pi - 6)e^{-(30.666h/w)^{0.7528}} \text{ and } \eta = \frac{377}{\sqrt{\varepsilon_{eff}}} \]

\[ (a) \]

\[ (b) \]

\[ (c) \]

\[ (d) \]

**Fig. 3.** Some modeling approaches for representing the behavior of transmission lines: a) analytical modeling [26], b) equivalent electrical model, and c) full-wave simulations.
transformed into matrix equations and iteratively solved by applying matrix inversion techniques. Currently, different EM simulators are available for the electromagnetic analysis of the transmission structures. In fact, EM simulators are classified by the number of space coordinates or dimensions (D) considered when solving a given structure under certain conditions. This classification lists the type of simulators as follows:

1D. These approaches are used for solving problems where the field and source functions are assumed to depend on one space dimension. Typical examples based on transmission line problems are those dealing with uniform plane wave propagation with radial dependence [22].

2D. These approaches are used for solving problems where the fields and source functions can be assumed to depend on two dimensions. Typical applications include planar structures such as cross section problems in transmission lines and waveguides, coaxial transversal EM (TEM) problems, spherical problems depending only on radius and azimuth or radius and elevation.

2.5D. In this case, problems are solved considering the dependence of the EM fields on three dimensions. Nevertheless their sources are assumed to be confined within conductive planes with two space dimensions. An arbitrary number of infinite dielectric layers with conductive interconnections among them are allowed.

3D (full wave). This is a method for solving problems where the fields and source functions are considered to vary in three space dimensions. This category involves full-wave general-purpose formulations [23].

1.2 Aspects to be considered in the modeling

For a typical transmission interconnect within electronics systems, different phenomena may impact the signal integrity. Thus, in order to better understand the structure of interconnects in actual applications, a description of some of the possible structures forming a transmission line channel is presented in this section.
1.2.1 Homogeneous and inhomogeneous transmission lines

Fig. 4 shows homogeneous and inhomogeneous transmission lines. Notice that for the homogeneous transmission line, a traveling signal would experience the same dielectric properties along the structure; consequently, the propagation constant ($\gamma$), and the characteristic impedance ($Z_c$) is the same along the interconnect. On the other hand, the electrical parameters in an inhomogeneous transmission line vary with position. In Fig. 4, the inhomogeneous transmission line experiences the effect of two dielectric materials presenting different dielectric permittivity (i.e., $\varepsilon_1$ and $\varepsilon_2$) along its length. In this case, the dielectric permittivity seen by the traveling signal varies with position; therefore, $\gamma$ and $Z_c$ also depend on the position. Fortunately, in many practical cases effective values for these fundamental parameters can be used to simplify the analysis.

1.2.2 Electrical transition and terminations

An electrical transition or discontinuity is originated by a change in impedance along a transmission channel; this impacts $\gamma$ and $Z_c$. These discontinuities are mainly due to bends, vias, pads, and other structures that negatively impact the propagation of the signal (see Fig. 5). In this regard, the discontinuities can be accurately modeled using full-wave simulators. However, models based on equivalent circuits and closed-form formulas can be used to quantify the impact on the signal integrity in a simpler way [24], [25], [26].
1.3 Object of study

This work is mainly focused on the study of homogeneous planar transmission lines and the associated attenuation introduced by the metal losses ($\alpha_c$). The importance of this topic relies on the fact that most of the total routing in microwave channels are formed by homogeneous transmission lines. In order to show the reader, some basic concepts regarding signal attenuation are briefly discussed afterward.

The total attenuation ($\text{Re}(\gamma) = \alpha$) in transmission lines is due to the separate contribution of the metal losses, the dielectric losses ($\alpha_d$) and radiation ($\alpha_r$); mathematically, it is represented as:

$$\alpha = \alpha_c + \alpha_d + \alpha_r$$  \hspace{1cm} (1)

For PCB and IC interconnects in current technologies, $\alpha_r$ noticeably impacts the total attenuation at frequencies higher than 100 GHz; therefore, for the operation frequencies of applications these days, it is reasonable to assume that $\alpha_r$ is negligible when compared to the other attenuation contributions [27]. In this case, the total attenuation can be calculated by:

$$\alpha = \alpha_c + \alpha_d$$  \hspace{1cm} (2)

*Fig. 5.* Examples of discontinuities: bends and pads.
Bear in mind that both, conductor and dielectric attenuation, simultaneously affect the signal in a microwave interconnect. Hence, it is necessary to consider the properties of these two types of materials and the corresponding geometry when assessing the impact of each one of the attenuation contributions to the signal degradation. For the case of the conductor attenuation, it can be written in the following form [26]:

\[
\alpha_c = \frac{R}{2Z_c}
\]  

(3)

whereas the dielectric attenuation is given by:

\[
\alpha_d = \frac{GZ_c}{2}
\]  

(4)

where \( R \) and \( G \) are the per-unit-length resistance and conductance associated with the metal and dielectric properties and geometry of the transmission line, respectively [28], [27], [26].

1.4 Effects associated with the metal traces

Electromagnetic effects on transmission lines depend on frequency and on the electrical properties of the dielectric and the conductors forming the structure [29]. At microwave frequencies, the electric and magnetic fields propagating in TEM mode impact the current distribution within the cross section of the interconnect. A description of these effects is presented below.

1.4.1 DC losses

At low frequencies, the current uniformly flows within the cross section of the metal traces and ground planes. Thus, in this frequency region, the resistance remains almost constant because the area through which the current is flowing does not change. Hence, the DC resistance \( R_{DC} \) in a conductor can be calculated by:
\[ R_{DC} = \frac{l}{\sigma A} \]  \hspace{1cm} (5)

where \( l \) is the length of the conductor, \( A \) is the area through which the current is flowing, and \( \sigma \) is the metal conductivity. The left side of Fig. 6 shows a sketch depicting the current flow at low frequencies.

Regarding the inductance at low frequencies \((L_{DC})\), the calculation is performed similarly as for the resistance case by accounting for the geometry and the properties of the conductors \([9]\).

### 1.4.2 Skin effect

As frequency increases, the cross-sectional current distribution in the conductor tends to flow at the internal surface or "skin" of the conductor. The right side of Fig. 6 shows the change in the effective area through which the current flows within the cross section of a circular metal trace from low (DC region) up to high frequencies. This change in the current distribution causes an increase of the resistance, which is proportional to the square root of frequency \((f)\) for a perfectly smooth conductor. For the case of the inductance, a gradual decrease is observed as frequency rises; in fact, at very high frequencies, the inductance tends to reach a minimum which corresponds to what is referred to as the external inductance. The parameter used to characterize this

![Fig. 6](image-url)  \hspace{1cm} \textit{Fig. 6. Change in the current distribution from low (DC region) to high frequencies within the cross section of a conductor with circular cross section. Notice that the current is more intense near the surface due to the skin effect.}
effect is called the skin depth and is the distance from the metal inner surface where the magnitude of the electromagnetic fields drops down to 37% of the value at the surface [26]. Mathematically, a widely used model for the skin depth is given by [12]:

$$\delta = \frac{2}{\sqrt{\sigma \omega \mu}}$$

(6)

where $\omega = 2\pi f$ is the angular frequency, and $\mu$ is the magnetic permeability. Thus, once the skin effect modifies the effective area through which the current is flowing, (6) can be used to obtain the frequency-dependent resistance of a metal trace. For instance, for the signal trace forming a microstrip transmission line, when the current is assumed to be confined to the bottom part (i.e., the closest face to the ground plane), the resistance can be approximately calculated from:

$$R_t = \frac{i}{\omega \delta}$$

(7)

Now, replacing (6) into (7) yields an expression where the frequency is explicitly included; this is:

$$R_t = \frac{\pi \mu}{\sigma w} \sqrt{f} = k_\delta \sqrt{f}$$

(8)

where $k_\delta = \sqrt{\pi \mu / \sigma w^{-1}}$ is the skin effect constant. On the other hand, the internal inductance of a conductor at high frequencies is calculated by involving the skin effect as [26]:
1.4.3 Surface roughness

For current technologies, the surface of the metal foils is roughened to promote the adhesion to the dielectric when manufacturing PCB laminates. The resulting rough surface looks very irregular presenting peaks and valleys of variable height. Nevertheless, the root-mean-square surface height ($h_{rms}$) is the standard parameter used in industry to characterize the surface profile. In this regard, when $h_{rms}$ is comparable to $\delta$, a substantial percentage of the current flow suffers from the effect of the irregular form of the metal surface. This occurs at microwave signals for current technologies (see Fig. 7). In fact, the impact of this effect on the electrical performance of interconnects is seen as an increase of the losses associated with the conductor material. Thus, it can be accounted for by modifying the expression for representing the series resistance to [26]:

$$R_{AC} = K_H R_t$$  \hspace{1cm} (10)

where $R_{AC}$ is the frequency-dependent resistance, and $K_H$ is the roughness coefficient.

---

**Fig. 7.** Sketch depicting the surface roughness at the metal-to-dielectric interface of a microstrip transmission line.
In packaging technology, the surface roughness of metal foils may vary from 0.5 up to 10 μm, depending on the manufacturing process.

Now, regarding the inductance of the interconnect, the internal inductance is also impacted by the metal surface roughness. Thus, based on (10) and (11), the inductance can be calculated by:

\[ L_{AC} = \frac{R_{AC}}{2\pi f} \]  

(11)

### 1.4.4 Proximity effect

In Fig. 8 a sketch of the proximity effect occurring between two metal layers carrying opposite current flows is presented [30]. Basically, as the spacing between the metals decreases, more current tends to confine closer to the adjacent sections. In this case, the effect consists in the induction of magnetic fields from the first wire to the second wire, redistributing the current on the surface. Thus, if the spacing between the conductors decreases, this field is more intense and the effect is more evident. The proximity effect constricts the current, increasing the effective resistance above the value that would exist for a uniform flow throughout the trace width. Basically, the resistance tends to increase by the proximity effect in the conductor and the

![Proximity Effect Diagram](image)

**Fig. 8.** Proximity effect occurring between two adjacent wires carrying opposite current flows. Notice that more current is confined in the inner faces as the spacing decreases.
inductance tends to decrease slowly up to an asymptote established by the external inductance [27].

1.5 Single-ended versus differential signaling schemes

Within today’s electronics systems, two differently configured interconnects are mainly used for guiding signals in ICs, packages, and PCBs: single-ended and edge-coupled interconnects. The main characteristics of each one of these two types of interconnects are determined by the geometry and the configuration of the electric and magnetic fields. Likewise, the distribution of the electromagnetic fields impacts the current distribution of the conductors. Next, some advantages and disadvantages of these structures are described.

1.5.1 Single-mode propagation

In Fig. 9, the cross-section of two types of single-ended transmission lines is shown illustrating the configuration of the magnetic and electric fields. In plain words, a single-ended interconnect is that presenting only one signal trace and using ground planes that may be common to other interconnects. In this case, the signal trace guides the signal using the ground plane as the return path. Accordingly, the electric and magnetic fields in single-ended interconnects are perpendicular to the direction

![Configurations of electric and magnetic fields](image)

**Fig. 9.** Configuration of the electric and magnetic fields in: a) single-ended stripline and b) ground-coplanar waveguide (G-CPW).
of propagation. Thus, the propagation mode is TEM for perfectly homogeneous lines where the fields travel through only one dielectric media or quasi-TEM mode when effective parameters can be used to represent the phenomena associated with composed dielectric media (e.g., Fig. 9b microstrip lines).

The main characteristic of single-ended interconnects is that few wires are needed to transmit multiple signals [31]. In consequence, the single-ended signaling scheme is less expensive, easy to implement, and compact. However, a disadvantage is that the return current for all the signals is shared within the same ground conductor, causing interference between the signals. Hence, lower data transfer rates per bit are obtained and the interconnects can act as antennas and radiate part of the energy when not properly designed [32].

1.5.2 Edge-coupled interconnects

Edge-coupled interconnects, usually referred to as differential interconnects when used for signaling purposes, are composed of two signal traces and one or two ground planes. Based on the phase of the currents and voltages propagating in the signal traces, the analysis of the interconnect’s properties can be carried out by assuming two different propagation modes. These are the odd and even propagation modes. For the odd-mode propagation case (Fig. 10a), each signal trace carries complementary

![Diagram of edge-coupled interconnects](image)

**Fig. 10.** Configuration of the electric and the magnetic field in edge-coupled striplines presenting: a) odd, and b) even propagation modes.
waves and the electric field is confined between the conductors. On the other hand, for the even mode propagation (Fig. 10b), the signal in both traces presents the same phase and amplitude, whereas the magnetic field surrounds the conductors following Ampere’s Law [26], [27].

Edge-coupled interconnects allow to transmit signals on a couple of metal traces. Typically, higher data rates are propagated in these interconnects when compared to the single-ended counterpart, lower output swing requirements are needed, lower electromagnetic emissions are obtained and mainly, these interconnect present high immunity to common-mode noise. However, some disadvantages such as higher cost, more board space, and more complexity, in general, is observed when compared with the single-ended case [32].

1.6 Purpose of this project

This chapter contextualizes the topic of interest in this project, which is focused on the analysis and modeling of homogeneous planar transmission lines (i.e., single-ended and edge-coupled interconnects). Particular emphasis is paid to effects involving conductor losses in these structures. For this purpose, from a strictly physical perspective, the frequency-dependent current distribution is carefully analyzed when microwave data transmission is desired. Moreover, both IC and PCB technologies are considered and evaluated.
O DESIGN on-chip and PCBs interconnect structures, efficient modeling tools that accurately incorporate the substrate effects, the metal effects, and the geometry of the interconnects are needed [33]. In this regard, the electrical parameters of transmission lines (RLGC parameters) are able to indicate the interaction between the voltages and currents that propagate in the structure and its composite materials. This allows to represent the losses and the delay of a signal when it is propagated through an interconnect.

The RLGC parameters considered in the telegrapher’s model are the resistance ($R$), inductance ($L$), conductance ($G$) and capacitance ($C$), all per-unit length. Basically, these parameters are calculated from the geometry and the properties of the composite materials in a transmission line. Thus, with these parameters other important features of the transmission line such as $\gamma$ and $Z_C$ are determined. An RLGC model can be used to represent either single-ended or edge-coupled transmission lines, in the latter case the model is obtained using a transformation of multiport $S$-parameters to mixed-mode parameters. In general, an accurate representation of the losses and signal delay in interconnects involves a precise representation of the RLGC parameters of the transmission line. In this chapter, a general description of the RLGC parameters and its dependence on the operating frequency, the geometry, and the composite materials is presented.
2. DESCRIPTION OF THE RLGC MODEL

2.1 C and G Parameters

In Fig. 11 the RLGC model of a single-ended transmission line is presented. Particularly, $C$ and $G$ are related to the energy stored in form of electric field, and to the losses in the dielectric material, respectively. These parameters are calculated from the geometry and the properties of the conductor through the following expressions:

$$C = \varepsilon_r K_G$$  \hspace{1cm} (12)

And,

$$G = C \omega \tan\delta$$  \hspace{1cm} (13)

where $K_G$ is a geometry factor of the interconnect and $\tan\delta$ is the loss tangent of the dielectric material ($\tan\delta = \varepsilon_i/\varepsilon_r$), associated with the absorption of energy due to

![RLGC model of a single-ended transmission line.](image)
polarization process.

## 2. DESCRIPTION OF THE RLGC MODEL

### 2.2 R and L Parameters

The series parasitic effects are accounted for by $R$ and $L$. These parameters represent the metal losses and the energy stored as the magnetic field, respectively. Both, $R$ and $L$ depend on the current flow; therefore, they are affected by the current distribution effects, the transmission line geometry, and the operating frequency. In the following paragraphs, an analysis of the effects on $R$ and $L$ for single-ended and edge-coupled interconnects is presented.

#### 2.2.1 Single-ended transmission lines

Several frequency regions can be defined in order to identify the main effects impacting the interconnect. For example, at the lower frequencies the skin effect can be masked and its contribution in $R$ and $L$ is minimum. These frequency regions are defined by considering the effective cross section of the interconnects, which changes within limits established by the dimensions of the signal trace, the ground plane, and the thickness of the dielectric laminate [9], [12]. Fig. 12 shows a qualitative depiction of the current distribution within the cross section of a microstrip line; basically, three regions were identified:

- **a) Low-frequency region.** At low frequencies, the resistance is obtained adding up the resistance of the signal trace ($R_{SO}$) and the ground plane ($R_{GND0}$), as follows [27]:

![Fig. 12](image-url)
2. DESCRIPTION OF THE RLGC MODEL

\[ R_{LF} = R_{SO} + R_{GND0} \]  \hspace{1cm} (14)

Regarding the inductance, the total inductance remains almost constant and it is defined as the sum of the external inductance of the transmission line \((L_{\text{ext}})\), the internal inductance of the signal strip \((L_{SO})\) and the internal inductance of the ground plane \((L_{GND0})\); mathematically [26]:

\[ L_{LF} = L_{SO} + L_{GND0} + L_{\text{ext}} \]  \hspace{1cm} (15)

where \(L_{\text{ext}}\) is the external inductance that is associated with the loops formed by currents flowing at the surface of the conductors, \(L_{SO}\) and \(L_{GND0}\) are the internal inductances related with the loops formed by the currents that propagate through the signal trace and the ground plane, respectively [10]. Thus, when the current distribution within the signal trace and the ground plane varies with frequency, \(L_{SO}\) and \(L_{GND0}\) exhibit this dependency as well. Equations (14) and (15) are valid for the low-frequency region, which is comprised between \(0 < f \leq f_{MF}\), where \(f_{MF}\) is the transition frequency corresponding to the beginning of medium-frequency region.

\textit{b) Medium-frequency region.} As the frequency increases, the proximity effect and the edge effect (highest current concentration in the edges of the trace) takes importance, they yield that the current in the ground plane starts confining under the signal trace [12], [34]. This change in the ground plane impacts the total resistance and inductance of the transmission line. Research work has been oriented to obtain general expressions for \(R\) and \(L\) based on the current distribution when the transmission line is working in the medium-frequency region [9], [12], [35]. The previously proposed methods, however, fail properly accounting for the effects in the ground plane. In this regard, some models will be proposed here based on the technology and the geometry of the interconnect.
The analysis in the medium-frequency region supposes that the current flowing through the signal trace remains almost constant and the ground plane is the one affected by the operating frequency. The resistance in the medium-frequency region can be calculated as follows:

\[ R_{MF} = R_{S0} + R_{GND} = R_{S0} + R_{GND}(f) \]  

(16)

The internal inductance in the ground plane is also frequency dependent and can be related to the resistance through \( L_{GND} = R_{GND}/2\pi f \) \([26]\), whereas \( L_{S0} \) and \( L_{ext} \) can be considered constant. Therefore, the total inductance in the medium-frequency region is obtained from the following expression:

\[ L_{MF} = L_{S0} + L_{ext} + \frac{R_{GND}(f)}{2\pi f} \]  

(17)

The medium-frequency region is defined from \( f_{MF} \) up to \( f_{HF} \), where \( f_{HF} \) is the beginning of high-frequency region.

c) High-frequency region. In this frequency region, the skin effect modifies the current distribution within the cross section of the signal trace. The resistance in the signal trace is calculated by applying \( R_S = K_h k_S \sqrt{f} \), where \( k_S \) is the skin effect constant (similar to (8)). Additionally, the resistance in the ground plane can be calculated by assuming that the skin effect is well established, and it can be obtained by \( R_g \approx K_h/\sigma_{eff} \delta = K_h k_g \sqrt{f} \), where \( k_g \) is a constant. Mathematically, the total resistance in the high-frequency region is given by [9]:
\[ R_{HF} \approx K_H k_S \sqrt{f} + K_H k_g \sqrt{f} = K_H k_T \sqrt{f} \] (18)

where \( k_T = k_S + k_g \) and \( K_H \) is the roughness effect factor, calculated by the Hammerstad and Jenssen formula as follows [26], [36]:

\[ K_H = 1 + \frac{2}{\pi} \text{atan} \left[ 1.4 \left( \frac{r}{\delta} \right)^2 \right] \] (19)

Similarly, the internal inductance is modified by the skin effect in the signal trace. It can be easily calculated from the resistance using: \( L = R/2\pi f \) [26]. The total inductance is obtained accounting the external inductance and the internal inductance. The external inductance is constant. Therefore, the total inductance in the high-frequency region is [9]:

\[ L_{HF} \approx \frac{K_H k_S}{2\pi \sqrt{f}} + \frac{K_H k_g}{2\pi \sqrt{f}} + L_{ext} = \frac{K_H k_T}{2\pi \sqrt{f}} + L_{ext} \] (20)

This frequency region is defined at frequencies at which \( f > f_{HF} \). Actually, in order to calculate \( f_{HF} \), (14) and (16) are simultaneously solved. Again, \( f_{HF} \) is found by solving (18).

### 2.2.2 Edge-coupled transmission lines

Current distribution effects in coupled interconnects are similar to the ones observed in single-ended interconnects (see Fig. 12); however, an important point to remark here is that the relatively large size of interconnects on PCB technology makes the low and medium frequency regions to become apparent below 1 GHz [9], [11]. In the following paragraphs, the different propagation modes are analyzed.
**Odd mode.** In the same way as for the single-ended case, at low frequencies, the current is homogeneously distributed within the cross-section of the transmission line until certain operating frequency is reached. At this point, the current distribution changes within the ground plane. A further increase in the operating frequency will eventually lead to a change in the current distribution within the signal traces. Fig. 13 shows the current density at three different frequencies obtained through full-wave simulations.

The resistance considering odd-mode propagation ($R_{\text{odd}}$) is dependent on the current flowing within the cross section of the traces and is approximately given by $R_{\text{odd}} = 2R$. On the other hand, the odd inductance ($L_{\text{odd}}$) can be calculated by accounting the external, the internal, and the coupling inductance $M$ [28]. Thus, $L_{\text{odd}}$ is calculated by [26]:

$$L_{\text{odd}} = L_{\text{ext}} + L_S + L_{\text{GND}} - M$$  \hspace{1cm} (21)

where $M$ is obtained from the capacitance in odd ($C_{\text{odd}}$) and even ($C_{\text{even}}$) mode with the air as the propagating medium ($\varepsilon_r = 1$); mathematically, it is calculated by [28]:

$$M = \frac{\mu_0 \varepsilon_0}{2} \left( \frac{1}{C_{\text{even}}} - \frac{1}{C_{\text{odd}}} \right)$$  \hspace{1cm} (22)

**Fig. 13.** Frequency-dependent current distribution in edge-coupled stripline when operated in odd mode.
2. DESCRIPTION OF THE RLGC MODEL

Even mode. Fig. 14 shows a full-wave simulation of the current distribution presented in even mode. Comparing this result with the one obtained in the odd mode (Fig. 13), the current distribution in the high-frequency region differs from the one presented in the odd case. Therefore, modified equations are necessary to model $R_{even}$ and $L_{even}$. In this case, three frequency regions are also observed in edge-coupled interconnects similarly to the ones shown in Fig. 12 and Fig. 13.

The resistance in even-mode ($R_{even}$) can be calculated as $R_{odd} = 0.5R$. The inductance in even-mode ($L_{even}$) tends to increase because the magnetic fields are reinforced. Thus, the total even-mode inductance is [26]:

$$L_{even} = L_{ext} + L_S + L_{GND} + M$$  \hspace{1cm} (23)

2.3 Characteristic impedance and propagation constant

Based on the RLGC parameters, two additional parameters can be obtained: $\gamma$ and $Z_c$ [26], [27]. $Z_c$ is defined as the ratio between the instant voltage and current in a section of the transmission line [37]. In general, $Z_c$ can be calculated by [27]:

\[ \text{Fig. 14. Frequency-dependent current distribution in equally-spaced edge-coupled striplines when operated in even mode.} \]
2. DESCRIPTION OF THE RLGC MODEL

\[ Z_c = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \] (24)

On the other hand, \( \gamma \) is used to define the propagation properties of the signal when it is propagated in a transmission channel [25]. Mathematically, the propagation constant is calculated by:

\[ \gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = \alpha + j\beta \] (25)

where \( \beta \) is the phase delay.

Edge-coupled transmission lines involve two propagation modes: odd and even mode. The corresponding characteristics are defined from the characteristic impedance \( (Z_m) \) and the propagation constant \( (\gamma_m) \), where the suffix \( m \) allows one to distinguish between different propagation modes. Similar to the single-ended case, the propagation constant can be extracted from the RLGC parameters by:

\[ \gamma_m = \sqrt{(G_m + j\omega C_m)(R_m + j\omega L_m)} \] (26)

where the \( R_m, L_m, G_m \) and \( C_m \) are the resistance, inductance, conductance and capacitance per-unit-length, respectively.

2.4 Summary

In this chapter, a description of the parameters involved with the representation of transmission lines using the RLGC equivalence was presented. Particular emphasis was given to the necessary considerations for representing the effects associated with the current distribution within the cross section of single-ended and edge-coupled interconnects. Basically, three frequency regions were identified (a low, medium and
high-frequency region) and expressions for $R$ and $L$ were developed. These expressions show the impact of the current distribution effects in the resistance and the inductance of the transmission line ($R$ and $L$). Finally, $Z_c$ and $\gamma$ were obtained from the $RLGC$ parameters. With this information, a complete analysis of the interconnects can be realized.
Signals in ICs interact with CMOS transistors formed in semiconductor substrates by dielectric materials, polysilicon and impurities. In general, these materials as well as the corresponding structure and dimensions impact the operating frequency (and therefore the performance) of the transistor. So, in actual IC implementations, the signal is propagated through different metal layers, and the dielectric properties of the SiO$_2$, and cooper or aluminum layers determine the propagation properties of the signal.

Particularly, due to the increase in the operating frequency, an accurate model of the parasitic effects in interconnection channels needs to be performed in order to avoid underestimation of the losses and the phase delay in the system [38]. Basically, the interconnect models for IC are different to the ones for PCB technology because the structures in IC technology are smaller than the structures in PCB technology. In IC technology micrometer structures are common [39]; also, it is possible to find structures of hundreds of nanometers, or tens of nanometers such as those employing high-$k$ materials. Therefore, the parasitic effects determine the performance of the structure in different ways [40], [41], [42].

In this chapter, firstly the signal propagation in simple structures presenting high-$k$ materials is analyzed. This analysis allowed to develop a mathematical model to characterize this material through S-parameter measurements. After that, a model for single-ended transmission lines and the pads used to access to the planar structures in CMOS technology is presented. The extraction method and the proposed models account for the current distribution effects in the structures, which is the topic of interest in this project.
3.1 MIM (Metal-Insulator-Metal)

High-\(k\) materials such as hafnium dioxide (HfO\(_2\)), zirconium dioxide (ZrO\(_2\)) and titanium dioxide (TiO\(_2\)) have been used to replace the transistor's silicon dioxide gate. These materials enhance the transistor’s performance because they allow a high carrier charge concentration in the interface metal-dielectric that reduce the gate leakage. In this section, a characterization of high-\(k\) dielectrics is presented, the proposed method involves an analysis of the series parasitic effects using a concentric ring MIM structure (Fig. 15).

The dielectric properties of the films used to form the MIM structure can be obtained measuring its electromagnetic response at high frequency. In Fig. 15, the equivalent circuit is shown, two admittances can be observed: the center pad’s admittance \(Y_1 = G_1 + j\omega C_1\) and the external ring’s admittance \(Y_2 = G_2 + j\omega C_2\); both admittances are dependent on the effective area \(A\) of the corresponding capacitor, which defines the capacitance as \(C_n = \varepsilon_r\varepsilon_0 A_n/h\) and the conductance as \(G_n = C_n\omega\tan\delta\). The subscript \(n\) is used to distinguish between the center and ring pad.

Bear in mind that the experimental S-parameters include the effect of the admittances \(Y_1\) and \(Y_2\), but also the parasitic series impedance \((Z_p = R + j\omega L)\); because of this, the total impedance of the equivalent circuit shown in Fig. 15 is represented by:

![Diagram of MIM structure](image)

**Fig. 15.** Sketch representing three-quarters of the MIM structure used for dielectric characterization; the corresponding equivalent circuit is also shown.
Here, a negligible coupling between the center and ring pads is assumed at the frequency band analyzed. The total impedance in terms of the capacitance can thus be rewritten as:

\[ Z_{\text{DUT}} = \left( \frac{1}{y_1} + \frac{1}{y_2} \right)^{-1} + Z_p \]  \hspace{1cm} (27)

where \( \omega = 2 \pi f \) and \( C_t = C_1 C_2 / (C_1 + C_2) \). Moreover, the total capacitance of the DUT is:

\[ C_t = \frac{\varepsilon_r \varepsilon_0 A_{\text{eq}}}{h} \]  \hspace{1cm} (29)

where the equivalent area of the DUT is \( A_{\text{eq}} = A_1 A_2 / (A_1 + A_2) \), and \( A_1 \) and \( A_2 \) are the areas of the center and ring pads, respectively. Now, at microwave frequencies a further simplification can be done, at these frequencies, the loss tangent of the composite films are negligible (\( \tan \delta^2 \ll 1 \)); thus, (28) simplifies to:

\[ Z_{\text{DUT}} \approx \frac{\tan \delta}{\omega C_t} + R + j \left( \omega L - \frac{1}{\omega C_t} \right) \]  \hspace{1cm} (30)

In order to obtain the dielectric properties of the films, it is necessary to determine the parasitic effects represented by \( Z_p \) in (27). In this case, \( \varepsilon_r \) and \( \tan \delta \) are considered constant at relatively low frequencies (i.e., below 1 GHz). Additionally, since within the frequency range up to 40 GHz the metal layers of the MIM structures are thin when compared with the skin depth of the conductors, \( R \) and \( L \) present negligible
variation with frequency. Thus, \( R \) and \( L \) can be obtained from the real and imaginary parts of (30), respectively. The real part can be arranged as:

\[
\omega \text{Re}[Z_{DUT}]|_{\text{low frequency}} = \frac{\tan \delta}{\varepsilon_r} + \omega R
\]  

(31)

On the other hand, the imaginary part of (30) can be arranged as:

\[
\omega \text{Im}[Z_{DUT}]|_{\text{low frequency}} = \omega^2 L - \frac{1}{\varepsilon_r}
\]  

(32)

Thus, a linear regression of \( \omega \text{Im}[Z_{DUT}] \) versus \( \omega^2 \) is used to obtain \( L \). Afterwards, substituting (29) into (30) and solving for \( \varepsilon_r \) yields:

\[
\varepsilon_r = \frac{h}{\omega \varepsilon_0 A_{eq}} (\text{Im}[(R + j\omega L) - Z_{DUT}])^{-1}
\]  

(33)

Likewise, \( \tan \delta \) can be obtained from:

\[
\tan \delta = \frac{\omega \varepsilon_r \varepsilon_0 A_{eq}}{h} \text{Re}[Z_{DUT} - (R + j\omega L)]
\]  

(34)

b. Prototypes and measurements

Two prototypes were developed, the MIM structures are identical in geometry but the dielectric film is different (either \( \text{Al}_2\text{O}_3 \) or \( \text{HfO}_2 \)). They present a coplanar configuration defined by a center pad and an external ring. The top metallic layer made of aluminum presents a thickness \( t = 0.4 \) \( \mu \text{m} \); the dielectric film can be \( \text{Al}_2\text{O}_3 \) or \( \text{HfO}_2 \) and it has a thickness \( h = 0.04 \) \( \mu \text{m} \), whereas the bottom metal layer is made of
tungsten with a thickness $t_b = 0.2 \, \mu m$. Besides, the radius of the center pad is $R_c = 35 \, \mu m$, the clearance between the center pad and the ring is $R_r = 100 \, \mu m$, and the width of the ring is $w = 20 \, \mu m$.

During the fabrication process, e-beam evaporation under ultra-high vacuum conditions was performed for the deposition of the metal materials. In the case of the

![Graph](image)

**Fig. 16.** Regression of experimental $\omega \text{Re}[Z_{DUT}]$ versus $\omega$ data corresponding to $\text{Al}_2\text{O}_3$ to extract the parasitic series resistance $R$.

![Graph](image)

**Fig. 17.** Regression of experimental $\omega \text{Im}[Z_{DUT}]$ versus $\omega^2$ data for $\text{Al}_2\text{O}_3$ to extract the parasitic series inductance $L$. 

37
dielectric materials, atomic layer deposition (ALD) was used to grow them using either tetrakis (dimethyl) amino hafnium (TDMAH) or trimethylaluminum (TMA) as metal precursors for HfO₂ and Al₂O₃, respectively; in both cases, the oxidizing precursor was H₂O. The deposition temperature for these dielectrics was 250°C. Furthermore, in order to evaluate the influence of post-metallization annealing (PMA) on the electrical characteristics of HfO₂-based MIM structures, two different thermal treatments were carried out on them. First, one of the HfO₂ samples was treated with a PMA in pure N₂ at 400°C for 30 min. A second HfO₂ sample had a PMA in an N₂/H₂ atmosphere using the same temperature and time.

Afterwards, one-port S-parameter measurements were performed to the fabricated devices using a ground-signal-ground RF probe with a pitch of 150 μm. For this purpose, a previously calibrated vector network analyzer was used.

c. Experimental results

![Graph](image)

**Fig. 18.** Frequency-dependent extracted parameters for HfO₂. Notice that εᵣ and tanδ are decreasing with the thermal treatment.
Fig. 16 and Fig. 17 show the extraction of $R$ and $L$ for the $\text{Al}_2\text{O}_3$-based MIM structure. Once these parameters are known, $\varepsilon_r$ and $\tan\delta$ are obtained using (31) and (32). The model was also used to characterize $\text{HfO}_2$. Fig. 18, shows the corresponding results illustrating that no exposure to additional thermal treatment originates values of $\varepsilon_r$ and $\tan\delta$ as those reported in [43]. In contrast, when an $\text{N}_2$-based PMA is applied to $\text{HfO}_2$, the value of $\varepsilon_r$ is considerably reduced, suggesting that a low-$k\text{Al}_x\text{O}_y$ interfacial layer (IL) between the top Al electrode and $\text{HfO}_2$ is formed due to a chemical reaction of oxygen species present in the dielectric with the Al; this treatment also impacts $\tan\delta$. Notice in Fig. 18 that the loss tangent tends to decrease due to IL after PMA. In order to validate the results, a comparison of the model with experimental results is presented in Fig. 19. Notice that when $R$ and $L$ in the model, $S$-parameters obtained from experimental data are similar to the proposal.

Fig. 19. $S_{11}$ parameter of $\text{HfO}_2$ without TT, notice that a good agreement between experimental data and model is obtained.
3.2 Single-ended homogeneous transmission lines

Based on the current distribution effects in transmission lines, an accurate analysis of the single-ended structures at microwave frequencies is proposed in this section.

a. Proposed IC Model for \( R \) and \( L \) (involving current distribution effects in the ground plane)

As described in Chapter 2, the response of the transmission line (\( TL \)) changes depending on the frequency region at which it is operating, in this regard, three frequency regions were identified:

**Low-frequency region.** At low frequencies, uniform flow of current within the cross section of interconnects can be assumed. Thus, the resistance and inductance can be calculated by (14) and (15).

**Medium-frequency region.** As the frequency increases, the current in the ground plane starts confining under the signal trace due to the proximity effect. This effect reduces the effective area through which the current is flowing until a minimum width (\( w_{\text{eff}} \)) under the signal trace is reached [26]. At this point, the current distribution is almost constant. Nevertheless, the skin depth (\( \delta \)) is also affected even if the operating frequency is within a range that doesn’t exceed the thickness of the ground plane. In this case, a small change in the current concentration is observed at the edges of the

![Conceptual representation of current concentration within the medium-frequency region](image.png)

**Fig. 20.** Conceptual representation of the current concentration within the medium-frequency region; the parameters used to represent the corresponding frequency dependence of the current distribution are included.
current confining region in the ground plane.

In Fig. 20, a sketch representing these effects is shown. Notice that the cross section of the ground plane is divided into three parallel sections (i.e., ①, ②, and ③), the sections labeled with the number 1 represent the sides of the ground plane where the lateral skin effect occurs. This section is represented by a parameter defined as $\delta_{w1}$. Additionally, the slight depth reduction at the edges of the confining region in the ground plane is represented by the skin effect $\delta_{t1}$; both effects are limited by the minimum values where the current can flow ($w_1$ and $t_1$). The section labeled with the number 2 represents the skin effect in the ground plane where the current is mostly confined, it is represented by $\delta_2$ and it is limited to the width $w_2$ of the ground plane.

In order to determine the resistance of the ground plane in section 1, the general form of the resistance $R_1 = 1/\sigma t'w'$ is used. Notice that the parameters are modified in accordance with the terms in Fig. 20, where $t' = \delta_{t1} + t_1$ and $w' = \delta_{w1} + w_1$. Thus, the resistance of the ground plane in section 1 is $R_1 = 1/\sigma(\delta_{w1} + w_1)(\delta_{t1} + t_1)$. Similarly to the analysis in section 1, the resistance in section 2 is $R_2 = 1/\sigma w_2 \delta$. The total resistance in the ground plane ($R_{GND}$) corresponds to the combined effects observed in the sections 1 and 2 of the interconnect.

The resistance in the signal trace ($R_{S0}$) remains constant because within this frequency region the current is confined within the whole signal trace. Thus, the total resistance in the medium-frequency region can be calculated using (16) as follows:

$$R_{MF} = R_{S0} + R_{GND}(f) = R_{S0} + \frac{f}{k_1f + k_2\sqrt{f} + k_3}$$

$$\tag{35}$$

where $k_1 = 2\sigma t_1 w_1$, $k_2 = \sigma k(w_2 + 2w_1 + 2t_1)$, $k_3 = 2\sigma k^2$, $\delta_{w1} = \delta_{t1} = \delta_2 = k/\sqrt{f}$, $\sigma$ is the electric conductivity, and $k$ is a constant related with the skin effect.

On the other hand, the internal inductance in the ground plane is also frequency dependent and can be related to the resistance through $L_{GND} = R_{GND}/2\pi f[26]$, where
$L_{S0}$ and $L_{ext}$ can be considered constant. Therefore, the inductance in the medium frequency region is obtained from:

$$L_{MF} = L_{S0} + \frac{1}{2\pi (k_1f + k_2 \sqrt{f} + k_3)} + L_{ext}$$  \hspace{1cm} (36)$$

The medium-frequency region is thus defined from:

$$(R_{GND0}k_1 - 1)f_{MF} + R_{GND0}k_2\sqrt{f_{MF}} - R_{GND0}k_3 = 0$$  \hspace{1cm} (37)$$

One of the roots for (37) corresponds to the transition frequency.

**High-frequency region.** In order to calculate the resistance and inductance in this frequency region, (18) and (20) are used. Combining (20) and (18) the frequency defining the high-frequency region ($f_{HF}$) can be obtained from:

$$k_Tk_1f_{HF}^{3/2} + (k_Tk_2 - R_{S0}k_1 - 1)f_{HF} + (k_Tk_3 - R_{S0}k_2)\sqrt{f_{HF}} - R_{S0}k_3 = 0$$  \hspace{1cm} (38)$$

Again, $f_{HF}$ is one of the roots of (38).

**b. Prototypes and experimental set-up**

In order to test the proposed models, measurements over a test chip with transmission lines terminated with ground-signal-ground (GSG) pads so that RF-coplanar probes can be used to perform S-parameter measurements were analyzed.
The microstrip TLs were fabricated previously in a 0.35-μm process technology. An accurate description of the structures is presented in [44]. The signal trace made of aluminum presents widths of $w = w_s = 2$ μm and $w = 4$ μm for the different lines, whereas the height of the SiO$_2$ dielectric is $h = 1$ μm and $h = 3$ μm. For the lines with $h = 3$ μm, the thickness of the ground plane is $t_g = 640$ nm, whereas for the lines with $h =$

**Fig. 21.** a) Plot of $R$ versus $\sqrt{f}$, and b) experimental curve of $L$ used to extract the parameters in the medium-frequency region.
1 μm the metal thickness is $t_g = 665$ nm. In all cases, the thickness of the signal trace is $t = t_s = 925$ nm and lengths $l = 400$ and $1000$ μm were included in the test chip.

The measurements were carried out in a frequency range from 0.01 to 40 GHz using a previously calibrated vector network analyzer setup and RF-coplanar probes presenting a pitch of 150μm. Afterwards, a line to line procedure was applied to remove the parasitic effects associated with the pads. With this procedure, $\gamma$ and $Z_c$

\[ L_{HF} \sqrt{f} = L_{cu} \sqrt{f} + \frac{k_T}{2\pi} \]

Fig. 22. Experimental data used to obtain $L_{ext}$ through a linear regression.

Fig. 23. $R$ and $L$ versus frequency curves showing the validation with full-wave simulation.
were obtained from experimental data of each pair of microstrip lines with identical cross section [45]. Once $\gamma$ and $Z_c$ are obtained from the measured data, the experimental series resistance and inductance were determined as $R = \text{Re}(\gamma Z_c)$ and $L = \frac{\text{Im}(\gamma Z_c)}{2\pi f}$ [26].

c. Experimental results

In order to extract the coefficients for the corresponding models, linear regressions are performed. The resistance at low frequencies ($R_{LF}$) remains almost constant and can be obtained from (14). In Fig. 21(a) $R_{LF}$ for a microstrip with $w = 4 \ \mu m$ and $h = 3 \ \mu m$ is presented. On the other hand, the change in the slope of the resistance defines the starting point for the medium frequency region which is upper-limited by the high-frequency region.

Notice that, the upper limit of the medium frequency region is easily identified, it is

![Fig. 24](image_url)  

**Fig. 24.** $R$ and $L$ versus frequency curves showing the confrontation of the proposed model with experimental data.
between the first points that the red line intercepts the experimental data in Fig. 21(a); at this point, the resistance is proportional to the $\sqrt{f}$. However, the actual values ($f_{MF}$ and $f_{HF}$) for the medium-frequency region cannot be obtained until the full model is implemented.

Hence, it is necessary to establish approximate limits for this region in order to extract the medium-frequency parameters in (35) and (36). In this case, Fig. 21(a) illustrates the frequency interval $[f_x, f_y]$ used to extract both $R_{MF}$ and $L_{MF}$. This interval corresponds to the range where a change in the slope of the resistance is observed, at low frequencies it is one of the last frequencies where the green line intercepts the data, and at high frequencies it is one of the first frequencies where the red line intercepts the data.

In Fig. 21(b) the transmission line’s admittance $Y_1$ versus $\sqrt{f}$ is plotted. From (36) $Y_1$ can be written as $Y_1 = 1/(L_{exp} - a) - 2\pi k_1 f$, where $a = L_{S0} + L_{ext}$ and the slope $2\pi k_1 f$ is modified to obtain a linear curve of the experimental data. Afterwards, the value for $k_1$ is obtained. Additionally, the linear regression of data shown in Fig. 21b is used to extract the values of $k_2$ and $k_3$. Correspondingly, the term $R_{S0}$ is obtained by substituting $k_1$, $k_2$, and $k_3$ into (38). Notice that subtracting the frequency-dependent term from the expression for $R_{MF}$, the value of $R_{S0}$ is obtained. Also, once $R_{S0}$ is known, $R_{GND0}$ is obtained using (14).

In order to extract $L_{ext}$, a plot of $L_{exp}\sqrt{f}$ versus $\sqrt{f}$ is used. Fig. 22 shows the extracted value of $L_{ext}$ using a linear regression based on (20). Additionally, the value of $L_{S0}$ is directly determined by using the value of $a$ and $L_{ext}$ obtained before, whereas $L_{GND0}$ is calculated by (15) using the obtained values for $L_{S0}$ and $L_{ext}$. Once the parameters were obtained, the transition frequencies are calculated using (37) and (38).

Finally, $R$ and $L$ can be extracted from measurements. This procedure was repeated for the different TLs. In Fig. 22 a comparison of the extracted $R$ and $L$ with full-wave simulation is presented, notice that extracted parameters present accurate values instead full-wave simulations when are consider the nominal design values (Simulation I) and the error introduced by the manufacturer (Simulation II). Finally, Fig. 23 shows the extracted per-length-unit $R$ and $L$ for each case; both $R$ and $L$ were
obtained for each transmission line following the proposed method. Notice that the method can be used for interconnects with different geometries.

3.3 Pad de-embedding

The pads are large area metal structures which allow to measure the performance of interconnects and other devices. Fig. 25 shows a sketch detailing the test fixture for probing a DUT with RF-probes. An accurate modeling of pads elements allow identifying the dependence of the device under analysis on frequency. Basically, these structures are affected by shunt and series parasitic effects.

In this section, the current distribution effects are analyzed to propose an accurate model of the pads. From this analysis, new expressions for $R$ and $L$ are developed; additionally, a de-embedding methodology is proposed, using an open dummy structure and the geometry of the pad structure.

a. Modeling of $R$ and $L$ in pads

The pad parasitics effects can be represented by means of a shunt admittance ($Y$) and a series impedance ($Z$). In the case of the latter, assuming that the pads are identical and present resistive ($R$) and inductive ($L$) series parasitics, the series impedance $Z = R + j\omega L$, and it is mainly associated with the trace that interconnects the center pad with the DUT, and also with the path formed between the ground shield and the ground pads as it is shown in Fig. 25.

**Fig. 25.** Sketch detailing a test fixture for probing a DUT. Notice the conceptual depiction of the current path in the ground plane at microwave frequencies.
Bear in mind that $R$ and $L$ are strongly dependent on frequency because of the skin and current crowding effects. Thus, in order to develop a broadband model for these parameters so that $Z$ can be determined, the change in the current distribution within the cross section of the structure as the frequency changes must be considered. For this purpose, several equations are necessary to represent $R$ and $L$ as explained hereafter.

**Modeling the series resistance.** The model for $R$ is formulated starting at frequencies considered as very low within the microwave range, these frequencies are in the order of kilohertz. At this starting point, the skin depth is much bigger than the thickness of the ground plane and the total resistance of the pad can be obtained from (14). Notice that the resistance of the signal pad is ignored since the $R_{t0}$ is typically much larger.

Due to the current crowding effect, there is a certain frequency $f_c$ at which the current no longer flows through the entire cross section of the ground plane, and it is confined within a width $w_{eff}$ below the trace (see Fig. 26). In order to model this effect, the cross section of the ground plane is divided into three sections presenting resistances $R_I$, $R_{II}$, and $R_{III}$ respectively. Thus, the parallel connection of these three resistances represents the total resistance of the ground plane.

$R_{II}$ corresponds to the resistance of the central part of the ground plane, which can be assumed to be constant at these frequencies. In fact, from the basic relationship...
between resistance and geometry, it is possible to calculate \( R_{II} \approx l_{t}/\sigma g_{w_{eff}} \), where the effective width of the corresponding section is \( w_{eff} \approx 6h \) when \( w < 6h \) \([26]\). Nevertheless, it is important to remark that \( w_{eff} \approx w \) can be assumed when the width of the signal trace is much larger than the thickness of the SiO\(_2\) dielectric (i.e., \( w > 6h \)). In order to check these relations for \( w_{eff} \), full-wave simulations were performed using different thicknesses for the inter-metal oxide. In Fig. 27, the corresponding results are shown. Notice that for the smallest considered height of the dielectric (i.e., \( h = 3 \) \( \mu \)m), the value for \( w_{eff} \) becomes approximately \( w \). Thus, since in advanced technologies the probing pads are separated from the ground shield by a dielectric layer typically less than 3 \( \mu \)m, \( w_{eff} \approx w \) is assumed here. Bear in mind, however, that for some technologies the width of the signal trace may be very narrow when compared with \( h \). In this case, a good approximation for the effective width was provided in \([26]\) (i.e., \( w_{eff} \approx 6h \)). The resistance \( R_{II} \) can be calculated as \( R_{II} = l_{t}/\sigma g_{w_{t}} \).

\( R_{I} \) and \( R_{III} \) correspond to the outer sections of the region where the current is confined in the ground plane (see Fig. 26). Hence, these resistances tend to rise as frequency increases since the current crowds closer to the section of the ground plane that is under the signal trace, reducing the corresponding effective area. This effect is modeled by the lateral (\( \delta_{t} \)) and vertical (\( \delta_{w} \)) skin effect, which impacts the cross section where the current is flowing in the ground plane. Thus, following the nomenclature defined in Fig. 26, the resistance of these sections is calculated as \( R_{I} = R_{III} = l_{t}/\sigma A_{I-III} \), where \( A_{I-III} \approx \delta_{w}\delta_{t} \) is the area through which the current is flowing in these sections. A good approximation to obtain \( A_{I-III} \) as a function of frequency is involving the metal skin effect (\( \delta \)) assuming a vertical and horizontal reduction of the area. This yields \( A_{I-III} \approx \delta^2 = 1/\pi \mu f \), where \( \mu \) is the permeability of vacuum.

The total resistance of the ground plane can be obtained from the parallel \( R_{I} || R_{II} || R_{III} \). Thus, through simplifications it is possible to write an expression for the resistance of the pad structure for \( f > f_{a} \) as:

\[
R = R_{fa} \approx R_{t0} + \frac{k_{oa}f}{1+k_{a}af} \tag{39}
\]
where $k_{0a} = l\pi\mu/2$, $k_{1a} = l\pi\mu\sigma t_g w/2l_c$, and $R_t$ is the resistance of the signal trace (a constant value).

Beyond certain frequency $f_b > f_a$, it can be considered that most of the current flowing through the ground plane is confined within the width $w$. Nevertheless, at the bottom part of this plane, the current concentration suffers a reduction as frequency increases, which implies that the area where the current is flowing takes a trapezoidal form. This effect is illustrated in Fig. 27 and can be approximately represented by considering a reduction in the area where the current is flowing, which requires the introduction of a frequency-dependent width $a$. From full-wave simulations, $a \approx 0.2w$ is determined, which is verified in the next sections. Using this parameter, the resistance corresponding to the sections with the width $a$ within the ground plane can be obtained as:

$$R_{0.2w} \approx \frac{l_c}{0.2w \sigma \delta} = \frac{l_c k_a}{0.2w} \sqrt{f}$$  \hspace{1cm} (40)

where $k_a = \sqrt{\pi \mu / \sigma}$. On the other hand, the section within the ground plane where the current flow is approximately uniform presents a width $w - 2a$ or $0.6w$. Thus, in this case, the resistance associated with this section can be calculated as:

$$R_{0.6w} \approx \frac{l_c}{0.6w \sigma t_g}$$  \hspace{1cm} (41)

which allows for the calculation of the total resistance of the ground plane as the parallel connection $R_{0.2w} \parallel R_{0.6w} \parallel R_{0.2w}$. Thus, incorporating this resistance into the equation for the total resistance of the pad structure, it is possible to write the following expression for $f > f_b$ after a simplification:
where $k_{0b} = k_0lt / 0.4w$, and $k_{1b} = 1.5k_0\sigma t_g$.

As the frequency of operation continues increasing, a certain frequency $f_c > f_b$ is reached at which $\delta = t_g$. At this point, the current flow through the ground plane tends to confine at the top of this metal layer. In this case, a single vertical variation of the area through which the current is flowing in the ground plane can be considered. Therefore, a simple square-root-of-frequency function describing the dependence on the skin depth of the resistance associated with the ground plane can be used. This allows expressing the total resistance of the pad structure for $f > f_c$ as:

$$R = R_{fc} \approx R_{t0} + k_c\sqrt{f}$$  \hspace{1cm} (43)
where $k_c = l_k a / w$.

The final frequency range considered here is starting at $f = f_d$ which is the frequency at which the skin effect also changes the current concentration within the cross section of the signal trace (i.e., $\delta = t$ at $f = f_d$). In this case, the cross-sectional area through which the current is flowing within the ground plane and the signal trace simultaneously become frequency-dependent. An expression similar to (43) can be written for this case, but considering that the resistance of the signal trace is also dependent on frequency; this is:

$$R = R_{f_d} \approx k_t \sqrt{f} + k_c \sqrt{f} = k_d \sqrt{f} \quad (44)$$

where $k_t = l_k a / w$ and $k_d = k_t + k_c$.

**Modeling the series inductance.** The total inductance of the pad can be obtained from (15). As a matter of fact, $L_t$ for frequencies below $f_d$ remains constant since the current concentration within the signal trace suffers no change. Thus, for $f < f_d$ the value for $L_t$ can be calculated evaluating $k_t \sqrt{f}$ at the frequency at which this inductance becomes frequency dependent [26]; mathematically:

$$L_t = L_{t0} = \left. \frac{k_t}{2\pi \sqrt{f}} \right|_{f=f_d} \quad (45)$$

On the other hand, for the case of interest, there are two inductances contributing to $L_{ext}$: the one associated with the coplanar array formed by the ground-signal-ground pads, and the one associated with the loop formed by the current flowing through the signal trace and the ground plane. Since the signal trace is much narrower and longer than the signal pad, $L_{ext}$ can be obtained considering only the latter contribution without significantly penalizing accuracy. Hence, the expression for calculating the inductance of a microstrip line can be applied using the dimensions of the signal trace.
to obtain the external inductance as in [46]. Once the internal and external inductances have been defined, the expressions to obtain the total inductance of the pad structure can be derived. In this formulation, the same frequency ranges defined for the different variable regions for $R$ are considered.

Firstly, the low-frequency inductance ($L_{LF}$) is analyzed. In this frequency range, $\delta$ is much larger than any dimension of the pad structure and the current can be assumed to uniformly flow through the cross section of the signal trace and the ground plane. In order to simplify the analysis, bear in mind that $L = L_{LF}$ can be assumed provided that $f < f_a$. Moreover, following the same nomenclature used for $R$, when $f > f_a$ the inductance is obtained as $L = L_{fa}$. Thus, taking advantage of the continuity of the inductance versus frequency curve, the following expression allows for the calculation of the low frequency (i.e., $f < f_a$) inductance:

$$L = L_{LF} = L_{fa} \bigg|_{f=f_a}$$  \hspace{1cm} (46)

In this regard, the expression for $L_{fa}$ is obtained by adding $L_{ext}$ to the internal inductance when $f > f_a$. This can be mathematically expressed by

$$L = L_{fa} \approx L_{ext} + L_{t0} + \frac{k_{oa}}{2\pi (1+k_{oa}f)}$$ \hspace{1cm} (47)

where $L_{t0}$ and $L_{ext}$ are calculated from (44) and (61), respectively. Notice that the last term in (47) corresponds to the frequency-dependent internal inductance associated with the ground plane, which is obtained by substituting the frequency-dependent term in the second member of (39).

For $f > f_b$, $f > f_c$, and $f > f_d$, the expressions for the total inductance can be obtained in the same way as for the case of (47). In these cases, however, the frequency
dependent terms in (42), (43) and (44) were used to obtain the frequency-dependent inductance. The corresponding expressions are:

\[ L = L_{fb} = L_e + L_{t0} + \frac{1}{2\pi \sqrt{f}} \frac{k_{ob}}{1+k_{1b} \sqrt{f}} \]  
\[ L = L_{fc} = L_e + L_{t0} + \frac{k_c}{2\pi \sqrt{f}} \]  
\[ L = L_{fa} = L_e + \frac{k_d}{2\pi \sqrt{f}} \]

Notice that the proposed equations allow for the modeling of microstrip transmission lines. However, when the interconnect presents coplanar configuration (e.g., CPW and GCPW), the current flow within the conductors is impacted by the lateral coupling of the signal trace and the ground traces. Nonetheless, a modified version of the proposed equations can be applied by accounting for the additional resistances and inductances introduced by the edge-coupled ground lines.

**Frequency ranges of operation.** In order to apply the models developed for \( R \) and \( L \), the transition frequencies have also to be analytically determined by considering the continuity for the resistance model. For this purpose, expressions corresponding to adjacent ranges are equated and then solved for the frequency to obtain \( f_a, f_b, f_c \), and \( f_d \). The resulting equations are:

\[ f_a = \frac{R_{y0}}{k_{0a} - k_{1a} R_{y0}} \]  
\[ f_b = \left[ \frac{1}{k_{0a} k_{1b} - k_{0b} k_{1a}} \left( \frac{k_{0a}}{2} + \sqrt{\frac{k_{0a}^2}{4} + k_{0a} k_{0b} k_{1b} - k_{0b}^2 k_{1a}} \right) \right]^2 \]
3. ANALYSIS ON IC

\[ f_c = \left( \frac{k_{ab} - k_c}{k_{1b} k_c} \right)^2 \] (53)

\[ f_d = \left( \frac{1}{\sigma_{t_v} \alpha} \right)^2 \] (54)

b. Prototypes and experimental set-up

Two different devices fabricated in a 65 nm RF-CMOS technology were used to verify the usefulness of the proposed model: an inductor and a transmission line. At this point, it is important to remark the fact that a new de-embedding technique to characterize this type of devices is proposed in this work. In this case, only ‘open’ dummies are needed to perform the de-embedding of the collected measurements. Nonetheless, ‘short’ dummies were also included in the prototype so that a comparison with the conventional open-short de-embedding can be carried out. Two-port S-parameter measurements were performed on the fabricated devices up to 40 GHz using the VNA.

Case 1: Inductor

The inductor consists of three turns in parallel formed with traces with a width \( w = 20 \) μm and a maximum radius \( r_v = 200 \) μm. The turns are made of copper with a thickness \( t = 0.55 \) μm and separated from the substrate by a SiO\(_2\) layer with a thickness \( h = 3 \) μm. Following the definitions given in Fig. 25, the length of each ground pad is \( l_p = 130 \) μm, the length of the signal pad \( l_s = 100 \) μm, the distance between pads is \( s = 30 \) μm, the width of the pads is \( w_p = 90 \) μm, and the length of the trace connecting the DUT with the signal pad is \( l_t = 240 \) μm. The pads are shielded from the substrate by means of a ground plane with a width \( w_g = 400 \) μm. The thickness of the ground plane is \( t_g = 0.5 \) μm. Moreover, each ground pad is connected to this plane using an array of vias presenting an average diameter of \( d_v = 0.6 \) μm.

Case 2: Transmission line

The TL presents a length \( l_{TL} = 2450 \) μm and a width \( w = 9 \) μm. Following the definition described above, \( t = 0.55 \) μm, and \( h = 1.8 \)μm. The geometry of the pads corresponding
to Fig. 25 are \( l_p = l_s = w_p = s = 80 \, \mu\text{m} \), and \( l_t = 160 \, \mu\text{m} \). Likewise the inductor case, the pads are shielded from the substrate using a ground plane with the same characteristics.

**Case 3: 3D Simulated Model**

In order to provide an additional assessment of the proposal, full-wave simulations were performed in Ansys HFSS. In this regard, the 3D model was firstly implemented for the open dummy and strictly considering the geometry of the pad array. Afterward, the simulation was performed. A good correlation of the model and the experimental data was observed. Once the electromagnetic model was validated, the short dummy was simulated and the corresponding data was employed to perform the open-short de-embedding. This procedure was performed for comparison purposes for both the inductor and transmission line considered in this project.

**c. Experimental results**

Once the usefulness of the model to represent the electrical behavior of the ‘short’ dummy structure is demonstrated, the \( S \)-parameter measurements performed on the inductor are corrected from the pad parasitics using the conventional open-short de-embedding procedure.

In Fig. 28, the equivalent circuit for the DUT including the series \( (Z_1 \text{ and } Z_2) \) and shunt \( (Y_1 \text{ and } Y_2) \) parasitic effects from the pads is shown. Notice that the corrected two-port network parameters for the DUT can be determined by applying:
3. ANALYSIS ON IC

where $Z_M$ is the $Z$-parameter matrix associated with the uncorrected data (i.e., DUT plus pad parasitic effects), $Y_{DUT}$ is the $Y$-parameter matrix associated with the DUT, and:

$$Y_{DUT} = (Z_M - Z_{series})^{-1} - Y_{shunt}$$

(55)

**Fig. 29.** $S$-parameters and intrinsic inductance for the prototyped inductor processed with different de-embedding methods.
Fig. 30. Complex propagation constant data and characteristic impedance for the fabricated transmission line obtained after processing with different de-embedding methods.
In these equations, it is clear that $Y_1$ and $Y_2$, and $Z_1$ and $Z_2$ can be determined from measurements performed to ‘open’ and ‘short’ dummies structures, respectively. However, the de-embedding method can be simplified using only the measurements corresponding to the ‘open’ dummy in combination with the closed form formulas proposed above.

**Inductor.** In order to remark the importance of removing the parasitic series elements from the measurements to obtain realistic results for the DUT, in Fig. 29 the results of applying a simple open de-embedding are also shown. Observe that a significant deviation is observed between the curves corresponding to this method and the methods that account for the series elements. Moreover, even when using full-wave simulations to generate data corresponding to the ‘short’ structure, discrepancies in the de-embedded data are observed, especially at higher frequencies when the effect of the vias used to short the end of the pad structure to represent this dummy are not properly considered. This is the reason why in Fig. 29 the de-embedding involving full-wave simulations differs from the data obtained after using the physical ‘open’ and ‘short’ structures.

**Transmission lines.** As it is well known, a uniform transmission line can be completely characterized by its characteristic impedance ($Z_c$) and propagation constant ($\gamma = \alpha + j\beta$). Hence, these parameters were calculated, after processing the raw (i.e., uncorrected) data with the open de-embedding, the de-embedding based on full-wave simulations, the open-short de-embedding, and the proposed de-embedding procedure. Fig. 30 shows a comparison between the curves corresponding to $\gamma$ and $Z_c$. 

$$Z_{\text{series}} = \begin{bmatrix} Z_1 & 0 \\ 0 & Z_2 \end{bmatrix}$$ (56)

and

$$Y_{\text{shunt}} = \begin{bmatrix} Y_1 & 0 \\ 0 & Y_2 \end{bmatrix}$$ (57)
Notice that neglecting the series inductance introduced by the pads (as assumed in the open de-embedding) introduces a resonance at around 30 GHz, which may lead to an overestimation of the attenuation. Fortunately, when the series parasitic effects are removed, this resonance vanishes. This is achieved when applying either the open-short de-embedding procedure or the methodology proposed in this document.

3.4 Summary

In this chapter, a model for the series parasitic effects in IC technology was proposed. Basically, the models are based on the current distribution effects in the structures and their impact on the total resistance and inductance of the interconnects. In the first section, a MIM structure was used to characterize dielectric films, good results were obtained. In the second section the importance of the current distribution effects in the ground plane was discussed. Finally a more elaborated model was proposed using mathematical expressions that can be applied to replace full-wave simulations of a text fixture for de-embedding purposes. This latter contribution allows for the simplification of the measurement de-embedding process by using only one dummy structure (open TL) instead of the traditional open and short structures; this strategy reduces the total area reserved to characterization on chips.
ANALYSIS ON PCB

AFTER THE signal leaves the die and the package, it propagates through the PCB. The package supports the ICs and other components and allows the electrical communication between these devices. Typically, the package and the PCB use similar dielectric and conductor materials; however, the cross section of the interconnects for PCB laminates is bigger than the cross section of the package substrates [47]. On the other hand, the length of the interconnects in PCB laminates is higher than the observed in package substrates [48]. So, the current distribution in this technology is expected to be different to the one observed in IC technology.

PCBs typically use a lamination process that consists of stacked materials. So, to reach mechanical stability, apply pressure and heat is applied to obtain a multilayer board, the metallic layers are usually made of copper whereas the dielectric layers vary depending on the losses that the application of interest tolerates. Thus, there are many choices such as FR4, polyimide, and teflon [11]. Modern lamination processes are able to produce minimum resolutions, obtaining trace widths and gaps between 4 and 100 μm [49]. In these structures, the skin, proximity effects, and roughness impact the response of the structure within the microwave range [9] [26].

In this chapter, the current distribution in single-ended and edge-coupled transmission lines in PCBs are analyzed. These analyzes involve the geometry, the conductor roughness, and the current distribution effects. Afterwards, a model for $R$ and $L$ for the TLs is extrapolated to analyze the current distribution effects in edge-coupled interconnects when the even and odd modes occur. Additionally, these
models are used to propose different methods to extract the dielectric properties of the laminates in ultra-low-loss, low-loss, and standard laminates.

4.1 Single-ended homogeneous transmission lines

Based on the properties of the homogeneous interconnects, two different methods to characterize PCB materials are presented. The methods allow characterizing single-ended transmission lines.

4.1.1 Standard and low-loss laminates

The most common laminates are standard and low-loss laminates. These laminates present tanδ between 0.001 and 0.02. For a detailed analysis of the transmission line response, the effect of the surface roughness and the frequency dependency of the resistance and the inductance needs to be included. In the following sub-sections, the models for R and L for single and edge-coupled interconnects are presented.

a. Dielectric characterization using RF-Measurements

In Fig. 31, a sketch of single-ended microstrip and striplines is presented. Notice the roughness of the different layers. For a single-ended interconnect implemented on PCB, R and L can be calculated from the dimensions of the cross section.

Considering only the high-frequency region, R and L can be calculated from (18) and (20), and the conductance from (13). With this parameter, and applying the line-line method, it is possible to obtain the experimental γ [45]. Thus, using this information, it is possible to characterize the dielectric material because from (25) and simplifying the real and imaginary parts of γ², an equation dependent on the propagation constant and the geometry of the transmission line is obtained as follows:

\[ F = \frac{kK_{H}(1+\tan\delta)+2\pi L_{d}\tan\delta\sqrt{f}}{kK_{H}(\tan\delta-1)-2\pi L_{d}\tan\delta\sqrt{f}} \]  \hspace{1cm} (58)

where \( F = \frac{Im(\gamma^2)}{Re(\gamma^2)} \). Rewriting (58), the frequency-dependent loss tangent is obtained as:

62
4. ANALYSIS ON PCB

\[ \tan \delta = \frac{kK_H[1+F]+2\pi L_eF\sqrt{f}}{kK_H[F-1]-2\pi L_e\sqrt{f}} \]  \hspace{1cm} (59)

Notice in (59), that the loss tangent can be calculated from the propagation constant, the roughness and the geometry of the structure. On the other hand, \( \varepsilon_r \) is obtained by \( \beta \) using:

\[ \varepsilon_r \approx \left( \frac{c\beta}{2\pi f} \right)^2 \]  \hspace{1cm} (60)

where \( c \) is the speed of light at vacuum. In order to obtain the external inductance the relation proposed in [46] was used:

\[ L_{ext} \approx \frac{\mu}{2\pi} \left[ \ln \left( \frac{2h}{w_s+t_s} + \frac{\pi h}{w_{eff}} + 1 \right) + \frac{3}{2} \right] \]  \hspace{1cm} (61)

b. Prototypes and experimental set-up

The microstrip transmission lines (Fig. 31a) were prototyped in two different substrates: FR4 (with \( \varepsilon_r = 4.0 \), \( \tan \delta = 0.02 \) and \( h = 178 \mu m \)), and teflon (\( \varepsilon_r = 2.0 \), \( \tan \delta = 0.001 \) @ 4 GHz and \( h = 190 \mu m \)). For the transmission lines over FR4, the signal strip width is \( w_s = 127 \mu m \), and the metal thickness is \( t_s = t_g = 38 \mu m \). On the other hand, for the microstrip over teflon \( w_s = 152 \mu m \), \( t_s = t_g = 36 \mu m \), \( r = 2.5 \mu m \), and \( h = 190 \mu m \).

The single-ended stripline (Fig. 31b) was fabricated embedded between dielectric layers with \( \varepsilon_r = 3.3 \) and \( \tan \delta = 0.005 \) @ 4 GHz. The signal strip width is \( w_s = 200 \mu m \), the metal thickness is \( t_s = t_g = 36 \mu m \), \( r = 2.5 \mu m \), and \( h_1 = h_2 = 260 \mu m \).
The $S$-parameter measurements were performed from 0.01 to 50 GHz using a 2-port Vector Network Analyzer (VNA), which was previously calibrated by applying a Through-Reflect-Line (TRL) calibration method and an impedance standard substrate.

**c. Experimental results**

As described before, the first step is extracting $\gamma$ from the measured data as in [45]; then, the $R$ and $L$ parameters are calculated from (18) and (19) based on the geometry of the single-ended structures. Finally, the $\tan \delta$ and $\varepsilon_r$ parameters can be calculated using (59) and (60), respectively. Fig. 32 shows the effective permittivity and the loss tangent in microstrips lines. Notice that these parameters change with frequency and the dielectric permittivity and loss tangent exhibit similar values than the ones reported by the vendor at 4 GHz. At higher frequencies, however, the extracted loss tangent presents higher values than the proposal; particularly above 50 GHz.

Fig. 33 shows the extracted data for the stripline. Observe that $\varepsilon_r$ and $\tan \delta$ at 4 GHz, are close to the values reported by the vendor. However, at higher frequencies, the dielectric losses increases up to 0.003. Notwithstanding, this is expected because the losses produced by polarization are no longer negligible.

**4.1.2 Ultra-low-loss laminates**

Ultra-low loss laminates exhibit exceptional electrical properties that are stable over a broad frequency and temperature range. These materials were developed to support high-frequency applications and present low dissipation factors ($\tan \delta < 0.001$), whereas the copper foils employed to form the PCB exhibit a surface profile with relatively small rms roughness ($r \approx 2 \mu m$).
4. ANALYSIS ON PCB

a. Dielectric characterization using RF-Measurements

Ultra-low loss laminates minimize the transmission losses by improving the fabrication processes and the composite materials of the laminate. Moreover, in this
case, $\varepsilon_r$, $\tan\delta$ and the coefficient associated with the surface roughness $K_H$ are considered constant.

The $RLGC$ model for the transmission line is simplified, the capacitance $C$ can be considered constant, $G$ is obtained from (13) and the analysis only takes into consideration the current distribution effects on $R$ and $L$ in the high-frequency region. The latter parameters are obtained from (18) and (20), respectively.

Thus, based on $Z_0$, $R$, $L$ and $G$, it is possible to get a frequency-dependent attenuation. Firstly, notice that $\alpha$ depends on $R$ and $G$ as shown in (3) and (4). Replacing these terms in (2), the total attenuation can be calculated by:

\[
\alpha = \frac{k\sqrt{f}}{2Z_c} + \pi Z_c C_0 \tan\delta f
\]  

(62)

Rewriting (62),

\[
\frac{\alpha}{\sqrt{f}} = \frac{k}{2Z_c} + \pi Z_c C_0 \tan\delta \sqrt{f}
\]  

(63)

In (63), a linear function can represent the $\alpha/\sqrt{f}$ vs $\sqrt{f}$ plot. Notice in (63) that the slope associated with $\sqrt{f}$ is used to determine $\tan\delta$, and the intercept with the $\alpha/\sqrt{f}$ in low frequencies is used to determine $k$ (i.e., the skin effect constant). Based on this equation, it is possible to calculate the parameters of the conductor and the dielectric, from a linear regression involving attenuation data.

**b. Prototypes and experimental set-up**

Microstrip lines (width $w = 1300$ μm, and length $l = 2.54$ cm) fabricated on an ultra-low loss laminate were used to verify the usefulness of the proposal. The PCB stack-up consists of two metallic layers (copper foils with a thickness $t_s = t_g = 25$ μm) separated
by a 400 μm PTFE sheet. This sheet was mechanically polished and the roughness of the surface is about 0.02 μm.

The microstrip samples were measured using a universal test fixture and a VNA, the S-parameters measurements were taken in the frequency interval from 0.01 to 30 GHz. The test fixture was calibrated employing the SOLT calibration method and an impedance standard substrate.

$Z_c$ and $\gamma$ were extracted from the S-parameters using the method proposed in [50]. The characteristic impedance can be extracted directly from the S-parameters by:

$$Z_c^2 = \frac{Z_{REF}^2 (1+S_{11})^2 - S_{21}^2}{(1-S_{11})^2 - S_{21}^2}$$  \hspace{1cm} (64)

where $Z_{REF}$ is the reference impedance of the measurement equipment and $S_{11}$ ($S_{22}$) and $S_{21}$ ($S_{12}$) are the S-parameters of the transmission line. And, the propagation constant $\gamma$ is mathematically defined by:

$$e^{-\gamma l} = \left[ \frac{1-S_{11}^2 + S_{21}^2}{2S_{21}} \pm \sqrt{\left( \frac{S_{11}^2 - S_{21}^2 + 1}{2S_{21}} \right)^2 - \left( \frac{2S_{11}}{S_{21}} \right)^2} \right]^{-1}$$  \hspace{1cm} (65)

On the other hand, the $RLGC$ parameters can be extracted from measurements using $R = Re\{\gamma Z_c\}$, $L = Im\{\gamma Z_c\}/2\pi f$, $G = Re\{\gamma/Z_c\}$ and $C = Im\{\gamma/Z_c\}/2\pi f$ [51].

c. Experimental results

In Fig. 34, the experimental results for the capacitance and inductance of the transmission line ($L$ and $C$) are shown. Notice the resonances becoming apparent due to the long length of the interconnect. The resonances are inherent to a real transmission line and make difficult to experimentally determine a stable and
accurate circuit model; however, the proposal can be used even if the resonances are not previously filtered from the experimental data.

In order to obtain \( C_0 \), the mean value of \( C \) is extracted from Fig. 34 (\( C_0 \approx 0.16 \text{nF/m} \)); also, \( L_e \) is obtained at frequencies higher than \( 400f_\delta \approx 11 \text{ GHz} \) where \( f_\delta \approx 27 \text{ MHz} \). Accordingly, the extracted value in Fig. 34 is \( L_e \approx 396 \mu\text{H/m} \), where \( Z_c \approx 50 \Omega \).

**Fig. 34.** \( L \) and \( C \) extracted from experimental data. Notice the resonances impacting experimental data.

**Fig. 35.** Plot of \( \alpha/\sqrt{f} \) vs square root of \( f \); linear curve fitting is used in order to extract \( k \) and \( \tan\delta \) using (63).
4. ANALYSIS ON PCB

In Fig. 35, a plot of $\sqrt{f}$ versus $\sqrt{f}$ of single-ended transmission lines is shown. Notice that a linear regression is used in order to obtain $k$ and $\tan\delta$. Therefore, $k \approx 6 \times 10^{-4}$ $\Omega$-m$^{-1}$Hz$^{-0.5}$ and $\tan\delta \approx 0.0034$ are obtained. Finally, the $RLGC$ parameters using the proposal are $R = 0.6\sqrt{f}$, $L = 0.396 + 95.4/\sqrt{f}$, $G = 3.42f$, and $C_0 = 0.16$.

Once the $RLGC$ parameters are obtained, the $S$-parameters are reconstructed and compared with the experimental data (see Fig. 34). Notice the good agreement between the $S$-parameters obtained with the proposal and the experimental data. Furthermore, if $\tan\delta$ is neglected, the losses at high frequencies (mainly due to the dielectric material) are underestimated and $|S_{21}|$ exhibits discrepancy with the measurements.

4.2 Differential homogeneous lines

In coupled transmission lines, the current distribution effects show dependence on the interaction between both traces. In particular, $R$ and $L$ parameters are impacted by the separation of the conductors in the structure. In this section, an analysis at high frequencies considering this fact is presented.
a. Modeling $R$ and $L$ based on current distribution effects

**Differential mode.** Consider Fig. 37 for the analysis of the frequency-dependent resistance and inductance in the high-frequency region.

The cross section of one of the signal traces is shown at the left-hand side of this figure. Notice that the current is considered to be flowing through two sections within each signal trace. In section-1, it is assumed that the current approximately flows on top and bottom of the trace, whereas in section-2 and section-3 the current flows at the edges of the trace. Thus, any change in the trace spacing will modify the current distribution in the latter sections in the following fashion.

As the traces get closer, the current in section-2 is confined closer to the metal surface and the current density in section-3 is diminished, which increases $R$. Moreover, in applications where the trace thickness and width are comparable, any modification in the current density occurring in sections 2 and 3 will impact the resistance associated with sections 1 and 2. Nevertheless, effective dimensions can be considered for these sections, they represent the corresponding contribution to the total resistance of the interconnect. Furthermore, since the lines analyzed here are based in symmetrical striplines, the current distribution on the top and bottom metal surfaces can be considered to be the same. Therefore, the resistive contribution of the top and bottom sides of the metal traces can be calculated as:

**Fig. 37.** A modeling approach to representing the current distribution and inductance loops for the case of a differentially operated interconnect.
4. ANALYSIS ON PCB

\[ R_1 = q_1 K_H \sqrt{f} \] \hspace{1cm} (66)

On the other hand, for the contributions associated with sections 2 and 3, \( K_H \approx 1 \) can be assumed without significantly penalizing accuracy. The equations to carry out the corresponding calculation of these resistances are:

\[ R_2 = q_2 K_H \sqrt{f} \] \hspace{1cm} (67)

and

\[ R_3 = q_3 \sqrt{f} \] \hspace{1cm} (68)

Therefore, considering that all the resistive contributions present a parallel connection, the effective resistance can be calculated as:

\[ R_D = 2 \left[ \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} \right]^{-1} = \frac{d_1 K_H \sqrt{f}}{d_2 K_H + 1} \] \hspace{1cm} (69)

where \( d_1 \) and \( d_2 \) depend on the material properties and the interconnect’s dimensions.

Regarding the inductance \( L \), notice in Fig. 37 that three current loops can be considered for each signal trace: two vertical loops representing the current flow between the center on the signal trace and the physical ground, two horizontal loops representing the current flow between the signal trace and the virtual ground and two external loops representing the current flow between the edges of the signal trace and the physical ground.
The horizontal loops are the ones more strongly affected by the spacing between traces. As the trace spacing increases, the external inductance associated with these loops increases, but the internal inductance will be reduced due to the reduction in the current concentration within section-2. However, it is reasonable to consider, that at high frequencies the current is confined close to the metal surface and then the change in the internal inductance would not be significant to the total inductance.

A similar reasoning can be applied to the other loops. Thus, since the external inductance of these loops is not substantially modified as the space between traces changes, the variation in the interconnect’s total inductance as the spacing changes is mainly due to changes in the horizontal loops.

An equation for predicting the interconnect’s total inductance, based on (69) can be derived in a similar way as in the case of the resistance; this is:

\[
L_D = \frac{d_3 [d_s K_H + 1] \sqrt{T} + d_s K_H}{d_s f + [d_s K_H + 1] \sqrt{T}}
\]

(Common mode). In the common mode, the current in both traces presents the same phase and amplitude. So, as can be seen in Fig. 38, the current distribution differs from the one presented in the differential case.
However, it is possible to employ the same approach than in the differential mode, the cross section of the interconnect is divided into section and the current distribution is analyzed on each of them. Thus, again the current flow is divided into three frequency regions as shown in Fig. 38. Since the current distribution is considerably smaller at the edges of the traces than at the top and bottom, the resulting equations are simpler than in the differential case; as follows:

\[ R_C = c_1 K_H \sqrt{f} \]  \hspace{1cm} (71)

where \( c_1 \) is a constant derived from the parallel contribution of the two regions where the current is concentrated in the even mode, and

\[ L_C = \frac{K_H c_2 \sqrt{f} + c_3 K_H^2}{c_4 f + c_5 K_H \sqrt{f}} \]  \hspace{1cm} (72)

where \( c_{2-5} \) are constants.

**b. Prototypes and experimental set-up**

Edge-coupled striplines were fabricated on a PCB using ultra-low loss laminates that present \( \varepsilon_r = 3 \) and \( \tan\delta = 0.003 \). The metal layers have a surface roughness \( r = 0.4 \) \( \mu \)m. The width of the traces is \( w = 165 \) \( \mu \)m, the thickness of the signal trace is \( t = t_s = 66 \) \( \mu \)m, the thickness of the dielectric is \( h_1 = h_2 = 76 \) \( \mu \)m, and the thickness of the ground plane is \( t_g = 15 \) \( \mu \)m. In summary, four cases of the study were considered on each case the cross section of each signal trace is the same (i.e., \( t_s, w_s, w_g, t_g \) and \( h \)) but the space between the traces \( s = 100, 200, 300 \) and \( 500 \) \( \mu \)m. Furthermore, pairs of striplines (i.e., 10 cm and 20 cm length) for each case were fabricated.

The mixed-mode parameters were extracted from the \( S \)-parameter measurements using the methodology from [11]. These data allow for the determination of \( R \) and \( L \) in both differential and common modes by respectively applying:
Fig. 39. Experimentally determined a) $R$ and b) $L$ for coupled stripline pairs when operated in differential mode.

Table 1. Extraction of the model parameters for $R$ and $L$ in differential mode.
4. ANALYSIS ON PCB

**Fig. 40.** Experimentally determined a) $R$ and b) $L$ for coupled stripline pairs when operated in common mode.

**Table 2.** Extracted $R$ and $L$ data in common mode.
4. ANALYSIS ON PCB

\[ R_m = \text{Re}(Y_m Z_m) \quad (73) \]

And

\[ L_m = \frac{\text{Im}(Y_m Z_m)}{2\pi f} \quad (74) \]

where the subscript \( m \) corresponds to either differential or common mode. Once \( R \) and \( L \) are known, the next step is to determine the parameters of the model in equations (69)-(72) by means of a parameter optimization tool.

c. Experimental results

The Fig. 39 and Fig. 40 show the resistance and inductance of the transmission lines, this parameter was extracted from the measurements following the process described above.

For the differential mode, the resistance \( R \) in Fig. 39a decreases as the spacing between the traces increases. This is expected since the electric field is inversely proportional to the distance between the driven traces, thus when the trace signals are close a higher electric field induces a current crowding at the inner edge of the striplines reducing the effective area and increasing the resistance. In contrast, the same scenario (Fig. 39b) makes the inductance to increase as the spacing is larger. In this case, the current loop between the signal traces is bigger and so \( L \).

For the common-mode case, weakening the coupling between traces by increasing the spacing reduces the current crowding effect, which in turn lowers the resistance (Fig. 40a). Nevertheless, in contrast to what occurs in the differential-mode case, when the trace spacing increases the inductance is reduced (Fig. 40b). This is due to the fact that the mutual inductance between traces, which adds up to the self-inductance of the individual traces to the total inductance of a stripline pair driven in common mode, is diminished.
4. ANALYSIS ON PCB

d. Model Scalability

In order to verify the scalability of the models for $R$ and $L$, the corresponding parameter extraction was performed. In Table 1, the extracted model’s parameters for the different stripline pairs operating in the differential mode are presented. In Table 2, the model’s parameters for common mode are presented. In both cases, the parameters showed excellent linearity allowing the easy interpolation and extrapolation for additional geometries.

4.3 Summary

In this chapter, different models to represent frequency-dependent resistance and inductance in single-ended and edge-coupled transmission lines in PCB laminates were presented. The results showed that the current distribution effects are impacting in a different way the frequency-dependent resistance and inductance, and it can be predicted with the geometry, the configuration of the conductors, and the properties of the dielectric.
CHAPTER 5

CONCLUSIONS

In this thesis, the impact of the distribution of the current within the cross section of different structures for IC and PCB technologies was analyzed. From this analysis, it was found that the frequency-dependent behavior of the resistance and inductance of high-speed interconnects requires dividing the bandwidth of microwave operation into several frequency regions. Thus, within each one of these frequency regions, there are different dominant effects that yield different response of the structure. The mathematical expressions developed here for these operation regions showed accuracy when were confronted with experimental data and full-wave simulations. In fact, the proposed models were successfully applied to MIM test fixtures, CPW-configured pads, microstrip, and striplines. Thus, in order to summarize the main contributions derived from this work, each one of the topics where the state-of-the-art was advanced is mentioned in the following paragraphs.

5.1 Modeling of microwave transmission channels

Modeling interconnects based on physical models has proven to be a powerful tool to accurately quantify the impact on resistance and inductance by current distribution effects in a microwave transmission channel. Basically, three frequency regions were identified in this work. I) In the low-frequency region the current is flowing within the entire cross section of an interconnection; so, $R$ and $L$ remain approximately constant with frequency and depend on the geometry and the electric properties of the structure. II) For the medium-
frequency region, the current crowding and skin effects occurring in the ground plane become noticeable and $R$ and $L$ increase with frequency. III) In the high-frequency region, the skin effect modifies the current distribution within the cross section of the signal trace and also the roughness of the metal surface takes importance.

Bear in mind that IC technology and PCB technology differ respecting the size of the structures. So, since the current distribution effects are directly determined by the geometry and the electrical properties of the dielectric and metal structures, even if there are still three frequency regions, for the IC technology. Thus, the combination of the current crowding and skin effects makes the equations for $R$ and $L$ to become more complex. For PCB technology, the skin effect is impacting the signal in MHz frequencies while that in IC technology the skin effect arises in GHz frequencies. In addition, roughness of the metal surface plays an important role in the high frequency region where $R$ and $L$ are strongly affected by the current distribution near the metal-to-dielectric interface.

The different proposed models differ from reported works in the consideration of the manufacturing technology and the associated physical effects associated with each case. Moreover, these new formulations remain the advantage of being implementable in SPICE-like simulators, representing the total losses and the phase delay introduced by the structure with accuracy. Below there is a small description of each of them:

- **IC technology model:** A model to represent $R$ and $L$ for single-ended transmission lines implemented in CMOS technology was presented. Details about the association of the corresponding equations with physical effects are given as well as a step-by-step explanation of the parameter extraction strategy to perform the model implementation.

- **PCB technology models for single-ended interconnects:** Different models involving series parasitic effects for single-ended transmission lines on PCB laminates were presented. The models were classified by the properties of the laminates as standard, low-loss and ultra-low-loss materials. These models allow to understand the impact of the geometry and the materials with the electrical properties of the structure.

- **PCB technology models for edge-coupled interconnects:** A model to represent the $R$ and $L$ parameters associated with different propagation modes are proposed for edge-coupled transmission lines. It was found that the spacing in edge-coupled interconnects is an important parameter in the analysis of interconnects. As it was presented, the variation of the resistance remains almost constant with the spacing, but the change in
the trace’s external inductance is the main modified effect. For the case of the resistance, variations are more evident as the frequency increases due to the confinement of the current in sections where the electric field is more intense. On the other hand, the inductance is dependent on the mutual inductance and the area formed between the conductors. This information helps to quantify the frequency-dependent effect in the design of high-speed interfaces, understanding the total losses by the configuration on the conductors.

Bear in mind that the models were obtained by physical modeling, allowing an accurate representation of the current distribution effects in order to simplify to the designer the calculation of $Z_C$ and $\gamma$ by basic mathematic tools. In this regard, these models result in a good alternative to avoid expensive full-wave simulators.

5.2 Electrical characterization of materials

Determining the electrical properties of the materials used on PCB and IC technology is an important task in the package and board design, when a new design is developed, it often includes dummy structures to calculate these parameters or they are assumed to be constant. Nevertheless, as the operating frequency increases, these techniques would need to be replaced with more accurate methods. In this regard, this thesis proposes three methods. They are listed as follows:

- **High-k dielectrics**: A model accounting for the series parasitic effects of an MIM test fixture was proposed. The model is used to extract $\varepsilon_r$ and $\tan\delta$ for high-$k$ dielectrics such as $\text{Al}_2\text{O}_3$ and $\text{HfO}_2$ (high-$k$ materials). Thus, it was demonstrated that the effect of the parasitic inductance and resistance should be appropriately represented at microwave frequencies to avoid errors in the device characterization.

- **Ultra-low-loss laminates**: The proposed method involves measurements in the high frequency region and also considers the electrical parameters constant; with this assumption, further simplifications are done, and then the process is reduced to make a linear regression to find the parameters. The method was validated with microstrip prototypes over a PTFE substrate.

- **Low-loss laminates**: A method to characterize low-loss dielectrics were proposed, it simplifies the analysis by taking measurements on the high-frequency region, where the
5. CONCLUSIONS

effects are plenty identified. The method was applied on microstrips and striplines; the results were compared with experimental data and good correlation was found.

- **Medium-loss laminates**: A method to characterize differential transmission lines was proposed; the method allows the implementation of a scalable model and characterization process to determine the series parasitics in edge-coupled interconnects.

- **Standard laminates**: The model used to extract the electrical properties in low-loss materials was also applied to standard materials. This kind of materials is widely used in the industry because of their thermal properties and their relatively low cost. In this case, a simplified method based on closed-form formulas was proposed.

Notice that the characterization of the structures includes the current distribution effects. Thus, it is possible to extract the electrical properties of the conductor and the dielectric in microwave frequencies at PCB and IC technologies in a broadband. In fact, notice that similar results are obtained when these values are compared with the proposed data from the manufacturer.

### 5.3 Pad de-embedding

A two-step de-embedding for $S$-parameter measurements performed on on-wafer devices fabricated in IC technology was developed using a single ‘open’ dummy structure in combination with a model for the pad series parasitic. For this purpose, a model for $R$ and $L$ was developed and verified. The model implementation is achieved by means of a parameter determination based on the geometry of the pad used to probe the devices. This represents an attractive alternative to save die space when designing and fabricating test chips including structures to be measured at microwave frequencies and to determine the parasitic effects during the pad design in IC technology. Consequently, the proposed model allows saving money, avoiding unnecessary design area and, simplifying also the de-embedding process. All these advantages can be obtained by the use of one single dummy structure on the IC avoiding additional simulations.
PUBLICATION LIST

6.1 Journal

“Semi-empirical Model for IC Interconnects Considering the Coupling between the Signal Trace and the Ground Plane,” in *Circuits, Systems & Signal Processing (CSSP)*, Accepted, pending (minor) revision, 2017.


6.2 Conference Publications


“Caracterización Eléctrica de Líneas de Transmisión de Baja Pérdida con Resonancias en el Rango de las Microondas,” in LVIII Congreso Nacional De Física, Yucatán (MX), 2015.

“Modelado de Resistencia e Inductancia Dependientes de Frecuencia en Líneas de Transmisión Diferenciales,” in LVII Congreso Nacional de Fisica, Sinaloa (MX), 2014.
BIBLIOGRAPHY


Wirel., vol. 22, no. 2, pp. 100–102, February 2012.


[36] E. O. Hammerstad and Ø. Jensen, "Accurate models of computer aided microstrip


