Equivalent circuit models for package level discontinuities and chip-package interconnects

Maulin Bhagat

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EQUIVALENT CIRCUIT MODELS FOR PACKAGE LEVEL DISCONTINUITIES AND CHIP-PACKAGE INTERCONNECTS

by

Maulin P. Bhagat

A Thesis submitted in Partial Fulfillment of the Requirements for the Degree of MASTERS OF SCIENCE in Electrical Engineering

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Equivalent circuit models with closed form expressions, in conjunction with a segmentation technique, have been derived to study three different aspects related to chip-package co-design issues. This technique has been successfully applied to package level discontinuities, chip-package interconnects, and the modeling of transients that affect the performance of an integrated microstrip antenna placed at the RF front end.

In each case, circuit models have been developed to include coupling by the use of coupling capacitance and mutual inductance. All circuit elements are defined by closed form expressions. To analyze package level discontinuities various microstrip transmission lines placed in close proximity to each other have been considered. Typical via structures such as the single via connecting signal lines placed on either side of a ground plane, and two-via structures between transmission lines placed on different layers have also been modeled. In each case the S-matrix from the equivalent circuit model compared to that obtained using a full wave simulator shows excellent agreement, thereby establishing the validity of the model.

To address chip-package co-design issues associated with RF front end a test bed consisting of a microstrip antenna with an embedded matching network has been implemented in a Multi-Layered Organic (MLO) material. The impact on the antenna’s input and radiation characteristics due to neighboring circuitry is found to be significant in both the frequency and time domains. Using the equivalent circuit model the coupling has been characterized and compared with that obtained from a full wave simulator.
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5.1 Circuit element values for the equivalent circuit model of the coupled 50 Ω line layout
The current trend in the packaging technology is towards higher operating frequency, smaller overall size, and lower cost. Several different techniques have been adopted to meet these requirements [1-2]. The most common techniques are System on Chip (SOC) and System on Package (SOP).

In recent years, much research has been devoted towards CMOS integration of the complete system (i.e. SOC). But there exists several problems with it that could prevent single-chip integration of complete system in CMOS at higher frequencies. Three major disadvantages of the system on chip (SOC) methodology are: (1) In the single chip integration of the mixed signal system (e.g. RF front end), the signal to noise ratio (SNR) of the analog blocks can degrade due to coupling from nearby digital blocks [3]. (2) The quality factor, Q, of on-chip inductor is very low compared to discrete inductors [2-3], which could severely impact the performance of the RF circuitry. For example the Q of the tank circuit is limited by the component with the lowest Q, which is most often an inductor. (3) Further, in semiconductor technologies for RF applications, typically more than 50% of the expensive chip area is occupied by passive components [1,3]. Therefore, to make it more economical these passive elements are moved to a multi-layered organic
(MLO) package [4] while the active components are retained on the chip, system on package (SOP) technology. The chips are placed on the package using chip-package interconnects like flip chip, QFN (Quad Flat No-lead), UCSP (Ultra Chip Scale Package), TQFP (Thin Quad Flat Package) etc. that help reducing the parasitic components associated with interconnects [2,5].

As opposed to discrete passive devices, integrated passive elements can be fully integrated in a multi-layered substrate during fabrication. Implementing passive components within the package reduces the total fabrication cost considerably. As in this technology the on-chip circuitry is combined with on package passive circuitry using a common multilayered package, it is widely known as Chip-Package Co-design. More information on chip-package co-design challenges and directions are presented in [6-7].

The objective of this work is to study the chip-package co-design issues associated with the following.

1. Package level coupling and discontinuities,
2. Chip-package interconnects, and
3. Impact of transient on microstrip antenna characteristics placed in MLO material [8].

Most common methods to study chip-package co-design issues are using full wave simulators and equivalent circuit models. Even though, results obtained using full wave simulators are accurate they are very time intense and serve merely as an analysis tools rather than design tools. The equivalent circuit modeling overcomes some of these disadvantages. Such models are invaluable tools to predict coupling and to optimize design for physical dimensions such as spacing, dielectric height, width of transmission
lines, interconnect type and dimensions etc. Further, they can also be incorporated into a circuit simulator with active elements to obtain the overall system performance. Such simple and comprehensive models, which include discontinuities and coupling effects both at the package level and interconnect level are not currently available.

Equivalent circuit modeling to predict substrate coupling is given in [9-10]. At present, circuit models for package level coupling [11-13], and chip-package interconnect coupling [14-18] are obtained by three different methods. In one case the S/Y/Z parameters obtained from a full wave simulator is translated to an equivalent circuit transmission line model [15-18]. In another method empirical expression are developed by curve fitting to measured results [12]. The third methodology is to optimize a circuit model using a circuit simulator and match results to full wave solvers [13-14].

In the present work, comprehensive equivalent circuit models are presented for package level coupling and chip-package interconnects with closed form simple expressions for the circuit elements.

1.1 Contributions

Major contributions of this work are as follows,

1. For package level coupling a circuit model has been developed that includes mutual capacitance and inductance to account for the electric and magnetic field coupling between two closely placed transmission lines.

2. In contrast to optimization technique described above, here simple and exact equations are provided to obtain the value of each circuit element.
3. An expression for the coefficient of mutual coupling has been developed, which works well for different microstrip configurations.

4. Simpler circuit models have been presented for via structures that are representative of chip-package interconnect. Two different configurations, a single via passing through ground plane and two-via structure are considered.

5. To study the chip-package co-design issues associated with the RF front end, microstrip antennas [19-27] have been designed on MLO material. Event though, the design of microstrip antennas is well known, in this case, it was a challenge because the MLO material is electrically extremely thin at the frequency of interest.

6. The return loss of the microstrip antenna has been improved by designing an L-section matching network using embedded inductor and capacitor.

7. An equivalent circuit model, to analyze the impact of transients on the microstrip antenna’s input characteristics, has been developed.

1.2 Organization

An equivalent circuit model has been developed analytically in chapter 2 to describe and quantify coupling between transmission lines placed in close proximity of each other. Exact empirical equations are presented for self and mutual inductance and capacitance. The circuit model also includes losses due to conductor and dielectrics. A segmentation technique has been used to cascade the equivalent circuit model for multiple sections of the transmission lines.
An analytical modeling of package level discontinuity and chip package interconnects is given in Chapter 3. Examples such as parallel transmission lines and discontinuous transmission lines have been considered. In each case, the S-parameters using the circuit model have been compared to that obtained by a full wave simulator. Moreover, the segmentation technique is successfully extended to study the coupling at chip-package interconnect level. An example of coupled vias (representing chip-package interconnect) is considered for this purpose and results of circuit model are compared with the results obtained for a full wave solver.

In order to study chip-package co-design issues on integrated microstrip antennas, it is necessary to investigate the feasibility of the design of a microstrip patch antenna in an MLO package, where the dielectric thickness is extremely small. Chapter 4 discusses a variety of designs such as direct feed antennas and electromagnetically coupled square patches and dipole antennas. An L-section matching network using embedded inductor and capacitor was designed to improve the performance of the microstrip antenna. Finally the predicted antenna parameters are compared to measured data.

Chapter 5 illustrates the effects of coupling from neighboring circuitry into the antenna feed line. In order to perform time domain analysis, a transient of five cycles at 2.49 GHz was applied to the adjacent transmission line. The effect on the input and radiation characteristics is studied. The technique is described to reduce such coupling. The equivalent circuit model has been used to obtain the S-matrix and results compare well with that obtained from a full wave solver.

Chapter 6 summarizes the major contributions and outline future work.
CHAPTER TWO

ANALYTICAL DEVELOPMENT OF EQUIVALENT CIRCUIT MODEL

One of the popular methods for the analysis of chip-package co-design issues, package level coupling, and chip-package interconnects is by using a full wave simulator. However, such a method is very time intense and serves merely as an analysis tool rather than a design tool. Another useful method is the development of equivalent circuit models to understand coupling phenomenon. This provides a more efficient and fast method for analysis, which can serve as an intuitive design tool.

Recently, several different models to represent coupling between two parallel transmission lines have been developed and in addition, analytical methods to calculate the values of circuit elements are available [11-18,28-31]. Coupled transmission lines have been modeled as equivalent planar waveguide and coupling between them is represented by a circuit model in [11]. Circuit element values can also be found by empirical equations [12] or optimization tools available in circuit simulators [13-14]. Another popular method to obtain the circuit element values is to obtain S-parameters for a specified frequency range and convert these into either Y or Z parameters, which are then related to the circuit elements [15-18]. Circuit properties of coupled microstrip lines
and a method of finding the modal characteristic impedance are published in [28-29]. Microstrip dispersion has been modeled by coupled TE and TEM transmission lines using a distributed circuit description [30]. From time domain measurement of coupled line structures equivalent circuit models have been extracted [31]. Analytical techniques developed to date, are complicated and do not lead to simple equations to calculate circuit element values.

In order to study the chip-package co-design issues, package level coupling, and chip-package interconnect level coupling effect a technique of segmenting parallel running microstrip transmission lines has been developed. Further, a circuit model for each segment has been presented with closed form equations for each circuit element. To verify the validity of the circuit model, the S-parameters obtained from the circuit model over a frequency range are compared with that obtained from the full wave simulator High Frequency Structure Simulator (HFSS). The segmentation technique for the package level coupling is also extended to study the coupling at chip-package-interconnect level. Examples of single via passing through a ground plane and two via configurations for chip-package interconnects are considered. The circuit model and equations for chip-package interconnects, which differ from the coupled microstrip structure, are discussed.

This chapter discusses the segmentation technique and the analytical development of the circuit model to address package level discontinuities / coupling and chip-package co-design issues related to the microstrip antenna. A similar method has been developed to model chip-package interconnects.
2.1 Segmentation Technique

Two parallel transmission lines shown in Figure 2.1 (a), placed on the substrate shown in Figure 2.1 (b), was the layout of concern. It was required to predict how much amount of energy is getting coupled from one line to another using an equivalent circuit model. In order to obtain such an equivalent circuit model these parallel transmission lines were segmented. One such segment of finite width, \( \Delta z \), is shown with dashed lines in Figure 2.1(a). Thus, each segment consists of two metal planes with ground plane underneath and a gap region through which the coupling occurs. The advantage of using this kind of segmentation is the value of coupling capacitor between two metal planes and the value of capacitor representing the fringing affect near the edges can be found by using the equations available for gap discontinuity. This is discussed in more detail in next section.

\[ \Delta z \]

Microstrip line

Metal Layer

\( \varepsilon_r, \tan \delta \)

Ground Layer

(a)

(b)

Figure 2.1. Segmentation technique for two closely placed transmission lines (a) Layout considered (b) Cross section of the dielectric material used.

2.2 Circuit Model for Package Level Discontinuities

The circuit model for each segment of width \( \Delta z \) should consist of the current flowing through both the metal plates and its and associated losses. It should also include
the affect of the finite area of the metal plane, its close proximity to the ground plane, and the loss associated with the dielectric material between the metal and the ground plane. In addition to these parameters the model should also consider the coupling due to electric and magnetic fields.

The circuit model for each segment can be defined intuitively. The effect of charges on the plate and magnetic flux linkage between the ground and transmission line can be considered by self-inductance and self-capacitance. Losses in the conductor and in the dielectric can be added by series resistor and shunt resistor respectively. The electric and magnetic field coupling can be introduced in the circuit model using a mutual capacitor and a mutual inductor respectively. The circuit model showing all these elements is shown in Figure 2.2.

![Figure 2.2](image-url)

**Figure 2.2.** Equivalent circuit model for a single segment, where L is self-inductance, \(C_a\) is self-capacitance, \(C_b\) is mutual capacitance, \(k\) is coefficient of mutual coupling, \(G\) is conductance of the dielectric material, and \(R\) is loss associated with the metal plate.

In Figure 2.2, \(L\) is self-inductance, \(C_a\) is self-capacitance, \(C_b\) is mutual capacitance, \(k\) is coefficient of mutual coupling, \(G\) is conductance of the dielectric material, and \(R\) is resistor representing loss associated with the metal plate. Cascading the
circuit model for a single segment over the entire length of the coupled section the circuit model for parallel transmission lines can be obtained.

For example, for the coupled microstrip shown in Figure 2.3 the circuit model for the entire layout can be developed using this concept. The length of each segment can be considered equal to the width of each transmission line. Cascading the circuit model of each segment the circuit model for the considered layout can be obtained and is shown in Figure 2.4.

![Circuit Model Diagram](image)

**Figure 2.3.** An example of closely placed transmission lines (a) Layout considered (b) Cross section of the dielectric material used.

![Equivalent Circuit Model Diagram](image)

**Figure 2.4.** Equivalent circuit model for two segments.
Thus, so far in this chapter, the segmentation technique and the circuit model for each segment have been shown. In the next section, equations to calculate each circuit element are presented.

### 2.3 Analytical Expressions for Circuit Elements

As mentioned earlier with the help of segmentation technique, equations for the coupling capacitor and the self / fringing capacitor can be found using a gap discontinuity model. Self-inductance is related to the self-capacitance and the impedance of the line. For the coefficient of the mutual coupling an empirical equation has been developed that works well for package level coupling. Conductor and dielectric losses are obtained using well-known equations [61]. The following sections discuss the equations for circuit elements.

#### 2.3.1 Coupling ($C_b$) and Fringing ($C_a$) Capacitances

The values of $C_a$ and $C_b$ were found using a gap discontinuity model shown in Figure 2.5. Each segment can be seen as a gap discontinuity. Even though the direction of current flow is different in the segment and the gap discontinuity layout shown here the electric field lines are in the same direction in gap region for both the cases. Thus, this model can be used to calculate electric field coupling and fringing effect, which are $C_b$ and $C_a$ respectively. This model cannot be used to find the magnetic field coupling because the direction of magnetic field is different for the segment and the gap discontinuity shown in Figure 2.5.
Chapter 2: Analytical Development of Equivalent Circuit Model

There are several methods available to find the values of $C_a$ and $C_b$. One of these is using well-known variational method published in [32]. Supporting mathematical work was published in [33-34]. The analysis using the variational method to calculate the values of $C_a$ and $C_b$ is very complicated and includes too many iterative equations. Further, these values are very inconsistent with respect to different values of dielectric permittivity ($\varepsilon$), height ($h$), and spacing ($s$).

Another method to calculate these values is to use empirical equations available in [35]. These equations are simple, easy to implement, and do not require iterative calculations. The only disadvantage of this method is that the equations are limited to physical dimension ratios $w/h$ and $s/h$, where $w$ is width of the microstrip line, $s$ is the spacing between two coupled lines and $h$ is the height of the dielectric material used. In this case the values of $C_a$ and $C_b$ are more consistent with changes in $\varepsilon$, $h$, and $s$. It is because of these reasons the empirical equations, even though physical dimensions limit
them, were used to calculate $C_a$ and $C_b$. The empirical equations obtained from [35] are presented here.

The expressions of $C_a$ and $C_b$ are obtained using even and odd mode analysis performed over gap discontinuity shown in Figure 2.5 (a). The capacitances $C_{odd}$ and $C_{even}$ are related to $C_a$ and $C_b$ by,

\[
C_a = \frac{C_{even}}{2} \quad \text{(2.1)}
\]

\[
C_b = \frac{1}{2} [C_{odd} - C_a] \quad \text{(2.2)}
\]

The closed-form expressions for $C_{even}$ and $C_{odd}$ valid for $2.5 \leq \varepsilon_r \leq 15$, and $0.5 \leq w/h \leq 2$, $0.1 \leq s/w \leq 1.0$ are given as follows.

\[
C_{odd} = \left( \frac{\varepsilon_r}{9.6} \right)^{0.8} \left( \frac{s^{m_0}}{w^{m_0-1}} \right) e^{k_0} \quad \text{(2.3)}
\]

\[
C_{even} = \left( \frac{\varepsilon_r}{9.6} \right)^{0.8} \left( \frac{s^{m_r}}{w^{m_r-1}} \right) e^{k_r} \quad \text{(2.4)}
\]

Where,

\[
m_0 = \left( \frac{w}{h} \right) \left[ 0.6191 \log \left( \frac{w}{h} \right) - 0.3853 \right], \quad 0.1 \leq s/w \leq 1.0
\]

\[
k_0 = 4.26 - 1.453 \log \left( \frac{w}{h} \right), \quad 0.1 \leq s/w \leq 1.0
\]
\[ m_e = 0.8675 \]
\[ k_e = 2.043 \left( \frac{w}{h} \right)^{0.12} \quad 0.1 \leq \frac{s}{w} \leq 0.3 \]
\[ m_r = \frac{1.565}{\left( \frac{w}{h} \right)^{0.16}} - 1 \quad 0.3 \leq \frac{s}{w} \leq 1.0 \]
\[ k_r = 1.97 - \frac{0.03}{\left( \frac{w}{h} \right)} \]

In the range over which these expressions are valid, the accuracy is within 7 percent compared to measured values [35].

### 2.3.2 Self-Inductance

Self-inductance (L) can be found using an equation of inductance per unit length of microstrip line. This equation is given by,

\[ L' = \frac{Z_0 \sqrt{\varepsilon_{re}}}{c} \quad \text{(H/m)} \quad (2.5) \]

Where,

- \( Z_0 \) = characteristic impedance of the transmission line (\( \Omega \))
- \( \varepsilon_{re} \) = effective relative permittivity

The effective relative permittivity (\( \varepsilon_{re} \)) is related to the height of the substrate (h), and the width of the transmission line (w) and is given as follows, [61].
\[ \varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \sqrt{\frac{1}{1+12\frac{h}{w}}} \]  

(2.6)

The inductance \( L \), expressed in Henrys and shown in the equivalent circuit model of Figure 2.4 which corresponds to a segment of length \( w \), is obtained from equation (2.5) by multiplying \( L \) by width \( w \).

### 2.3.3 Losses Associated with Metal

Losses associated with metal plate are considered using a resistor, \( R \). Both DC and AC resistances should be considered to calculate total value of \( R \). DC and AC resistances are represented by \( R_{ac} \) and \( R_{dc} \) respectively. Thus, total resistance is given by,

\[ R = R_{ac} + R_{dc} \]  

(2.7)

\( R_{ac} \) and \( R_{dc} \) are given by,

\[ R_{ac} = \sqrt{\frac{\pi \mu_0}{\sigma_c}} \]  

(2.8)

and

\[ R_{dc} = \frac{1}{\sigma_c t} \]  

(2.9)

Where,

\( f \) = desired frequency of operation (Hz)

\( \mu_0 \) = permeability constant \((4\pi \times 10^{-7} \text{ H} / \text{m})\)

\( \sigma_c \) = conductivity of the metal plane \((\text{S} / \text{m})\)

\( t \) = thickness of the metal plane \((\text{m})\)
A circuit simulator that was used for the simulation (Serenade) does not consider an element value as a variable with respect to some parameter. Here the value of $R_{ac}$ is a function of frequency. Ideally, in the circuit model the value of $R_{ac}$ should change at every frequency. As the variation in the value $R_{ac}$ is not significant in the frequency range of interest (1 to 5 GHz), a center frequency ($f$) is chosen to obtain $R_{ac}$.

Thus, knowing the value of $R_{ac}$ and $R_{dc}$ the value of $R$ can be found using equation (2.7). The unit of $R$ is Ohm ($\Omega$). For a good conductor with finite thickness the value of $R$ is usually very small. For example the value of $R$ is equal to 0.014 $\Omega$ for copper ($\sigma_c = 5.7 \times 10^7$ S / m) with thickness equal to 0.7 mils.

2.3.4 Loss Associated with Dielectric Material

It is usual practice to represent the loss associated with dielectric material as a shunt conductance ($G'$). One end of $G'$ is connected to an inductor and the other end is connected to the ground. In the circuit model this conductance is presented by a resistor, the value of which is simply an inverse of the conductance value. An equation to find the value of conductance ($G'$) is shown bellow [61].

$$G' = \omega C_a \tan \delta$$

(2.10)

Where,

$\omega = 2\pi f$, and $f$ is the frequency of interest

$C_a = \text{self / fringing capacitance value obtained from section 2.3.1}$

$\tan \delta = \text{loss tangent associated with the dielectric material}$
Here, again, the value of $G$ is a function of frequency. For the same reason, as explained in the section 2.3.3, the value of $f$ in the equation (2.10) can be chosen to be equal to middle frequency of the frequency range of interest. Associated value of resistor, represented by $G$ in the circuit model, can be found by just inverting the value obtained from equation (2.10). Thus, the equation for $G$ becomes,

$$G = \frac{1}{\omega C \tan \delta}$$  \hspace{1cm} (2.11)

The value of $G$ is very large, usually in the range on M$\Omega$. For example, for $\tan \delta = 0.02$, $C_a = 0.174$ fF, and frequency of interest $f = 2.45$ GHz, the value of $G$ is equal to 18.7 M$\Omega$.

2.3.5 Mutual Inductance

Mutual inductance represents the coupling between two microstrip lines due to magnetic fields. In the circuit model it is represented by a coefficient of mutual coupling. The relationship between mutual inductor ($M$) and coefficient of mutual coupling ($k$) is given by,

$$M = k \sqrt{L_1 L_2}$$  \hspace{1cm} (2.12)

Where,

$L_1 = $ inductance of one of the microstrip lines, and

$L_2 = $ inductance of the other microstrip line.

The value of $k$ is always between 0 and 1. To date, no equation is reported that calculates the value of $k$ or $M$ for a given coupled line layout. Here, an empirical equation for coefficient of mutual coupling ($k$) has been derived by studying the nature of the
variation in a value of k as a function of spacing. The value of coefficient of mutual coupling reduces as the distance between two coupled line increases. This is because of less amount of magnetic coupling between lines with higher distance between them. The variation of the value of k as a function of spacing between coupled transmission lines for a specified impedance \( Z_0 \) is shown in the Figure 2.6, where \( Z_0 = 50 \, \Omega \).

![Figure 2.6](image)

**Figure 2.6.** Change in the value of coefficient of mutual coupling (k) as a function of distance between the coupled microstrip transmission lines \( (Z_0 = 50 \, \Omega) \)

Noting that the variation of k as a function of spacing is exponential, an equation of this curve can be approximated by an exponential equation of the form shown in equation (2.13).

\[
k = A e^{-\frac{s}{B}}
\]  

(2.13)

Here, \( s \) is the spacing (in \( \mu\text{m} \)) between the coupled microstrip transmission lines and ‘A’ and ‘B’ are constants. The values of A and B are different for different characteristic impedance of the microstrip lines. The values of A and B for 50 \( \Omega \), 70 \( \Omega \),
and 100 Ω lines are tabulated in Table 2.1. The values of A and B, displayed in the table, were obtained for the coupled microstrip lines placed on the MLO material. These values work very well for the range of w/h and s/w in which the equations of self and mutual capacitance are valid.

Table 2.1. Values of A and B for three different characteristic impedances of a parallel microstrip transmission lines

<table>
<thead>
<tr>
<th></th>
<th>50 Ω Lines</th>
<th>70 Ω Lines</th>
<th>100 Ω Lines</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>0.25</td>
<td>0.65</td>
<td>3.24</td>
</tr>
<tr>
<td>B</td>
<td>0.3149</td>
<td>0.1968</td>
<td>0.05827</td>
</tr>
</tbody>
</table>

A matlab code was written that calculates the value of all the circuit elements for a given layout of the closely placed microstrip transmission lines. This code is presented in Appendix B.

### 2.4 Modeling Chip-Package Interconnects

In the chip-package co-design methodology passive elements are built within the package and the active components are built on chip. In order to connect them interconnects are used. There interconnections are commonly termed as “chip-package interconnects”. The commonly used chip-package interconnects are vias, wire bonds, and flip chips. These interconnects give rise to signal integrity issues. Moreover, at higher frequencies the cross talks between closely placed interconnects becomes prominent.

The study of coupling between chip-package interconnect has been concerned of several publications. The Finite Difference Time Domain (FDTD) and Finite Element
Method (FEM) techniques have been used to extract the characteristics of interconnects in [36]. The equivalent circuit models for flip-chip interconnects are presented in [15,37-39], whereas the circuit model presenting the coupling between flip-chips has been presented in [16]. The equivalent circuit model for flipped-chip-mounted monolithic microwave integrated circuits has been presented in [40]. The characteristics of the wire bonds have been analyzed in [41-43]. An equivalent circuit models representing single via / wire bond and coupled vias / wire bonds structure have been proposed in [14], where as the equivalent circuit model of the system consisting of wire bonds have been presented in [44-51].

In this work an equivalent circuit model has been presented for two different configurations of via structure. One is, a single via passing through a ground plane and the other is coupled via structures. The equivalent circuit model for a single via through ground plane published in [14] is complicated, moreover all the circuit element values are obtained using optimization tool available in a circuit simulator. A simpler circuit model, compared to that published in [14], has been proposed here for a single via through ground structure. The circuit model obtained for a via passing through ground plane has been extended to develop the circuit model for coupled via structure. The results of the circuit model for coupled via structure are compared with that obtained from a full wave solver. The following subsection discusses the circuit model for the single via structure and the last subsection is devoted to the equivalent circuit model of a coupled via structure.
2.4.1 Equivalent Circuit Model for a Single Via through a ground plane

A typical single via structure is shown in Figure 2.7, where the via connects two signal lines separated by a ground plane.

![Figure 2.7. Cross section of the single via through ground](image)

The equivalent circuit model for a via should only include the self-inductance and loss associated with the via. Thus, a via can be represented by an inductor and a resistor connected in series with it. As there is a cut in ground plane in the structure shown in Figure 2.7, the circuit model should include a capacitor representing an energy getting stored in a gap between the via and the ground. Such an equivalent circuit model is shown in Figure 2.8. In the figure, inductor, L, represents self inductance associated with via and R represents associated losses. The energy getting stored in the region between the via and the ground plane is represented by capacitor Ca.

![Figure 2.8. Equivalent circuit model for the single via with ground structure](image)
So far in this work, only the equation for self-inductance (L) has been found. The values of loss (R) and the capacitor (Cₐ) were found by optimizing their values using a circuit simulator (Serenade). The equation set to calculate the value of self-inductance for a via is presented in equation (2.14) [52].

\[
L = 2^{-9} \left[ \ln\left(\frac{4l}{d}\right) - 1 + S + 0.5\frac{d}{l} \right]
\]  

(2.14)

Where, \( l \) is the length of the via (in cm), \( d \) is diameter of via (in cm). The skin effect correction (S) is defined as,

\[
S = \frac{7.088}{A}, \text{ For } A > 100
\]

\[
S = 0.25\left(1 + 0.00546A^5\right)^{0.2}, \text{ For } A \leq 100
\]

Where,

\[
A = 0.1071d\sqrt{f}
\]

Where, \( f \) is the frequency at which skin depth is required.

The values of other two circuit elements (R and Cₐ) in the equivalent circuit model can be obtained by using optimization tool available in a circuit simulator. The results, obtained by using the proposed model and equations, are compared to that published in [14] and are shown in the following chapter.

The next section discusses the development of the circuit model for coupled via structure. The circuit model of the single via developed in this section is extended to develop the circuit model for the coupled via structure.
2.4.2 Equivalent Circuit Model for a Coupled Via Structure

In the previous section it was shown that the via can be represented by only its self-inductance and losses associated with it. The capacitor (\(C_a\)) was added in the circuit model to include the effect of energy stored between the ground plane and the via. In this structure a couple via structure as shown in Figure 2.9 is considered.

![Diagram of coupled via structure](image)

Figure 2.9. Coupled via structure. (a) top view (b) side view

As shown in the figure, two transmission lines of width, \(w_1\), are placed on the top metal layer (shown with solid lines) and are running towards each other. Each of these transmission lines is connected through vias to microstrip transmission line of width, \(w_2\), placed on bottom metal layer (shown with dotted lines). Unlike the structure discussed in previous section, this structure is backed by a ground plane, as shown in the cross section of Figure 2.9 (b).

The goal here is to develop the equivalent circuit model for the coupled via structure shown in Figure 2.9. In order to develop the equivalent circuit model, the total length was segmented into two subsections. Such segments are shown in Figure 2.10.
Chapter 2: Analytical Development of Equivalent Circuit Model

The circuit model for the complete structure can be developed by cascading the equivalent circuit model for a single segment. The equivalent circuit model for a single segment is shown in Figure 2.11, where as the circuit model for the complete structure is shown in Figure 2.12. Where L represents self-inductance of the via and R represents losses associated with the via. The electric field and magnetic field coupling between two vias are presented by mutual capacitor (C_b) and coefficient of mutual coupling (k) respectively. It should be noted here that the equivalent circuit model for the single via structure consist of only a self-inductor (L) and a resistor (R).

---

Figure 2.10. The segmentation of the coupled via structure

Figure 2.11. Equivalent circuit model for a segment of coupled vias

Figure 2.12. Complete equivalent circuit model for the coupled via structure
The value of the self-inductance (L) can be obtained from the equation (2.14) presented in the previous section. The equation for mutual capacitor was found out and is shown in equation (2.15).

\[
C_b = \frac{\pi \epsilon_0 \epsilon_r}{\cosh^{-1}\left(\frac{D}{2d}\right)} \quad (F/m)
\]  

(2.15)

Where,

\begin{align*}
D &= \text{distance between the center of two vias (m)} \\
d &= \text{diameter of the via (m)} \\
\epsilon_r &= \text{relative permittivity of the dielectric material between two vias}
\end{align*}

As the unit of \(C_b\) is F/m, the value obtained from the right side of equation (2.15) needs to be multiplied by the length of the segment to calculate the value of \(C_b\) in Farads. The values of the remaining circuit elements (R and k) were found using the optimization tool available in a circuit simulator. The comparison between the results obtained from the circuit model and that obtained from the full wave simulator are compared in next chapter.

In summary, a segmentation technique and the equivalent circuit model have been developed to study package level discontinuity / coupling and coupling between chip-package interconnects. The closed form equations for all the circuit elements are presented to study the coupling at package level. A simpler equivalent circuit model has been presented for a via structure. The circuit model for single a via has been applied to two different configurations, via through ground plane and a coupled via structure.
In the next chapter, the circuit model and the equations presented in this chapter are used to analytically model the package level discontinuities and chip-package interconnects. The results obtained from the equivalent circuit model are compared to that obtained by a full wave solver.
In this chapter, the segmentation technique and the equivalent circuit model, developed in previous chapter, have been used to analytically model the package level coupling / discontinuities and chip-package interconnect level coupling. Two different kinds of problems, one showing the package level coupling and the other displaying the discontinuity, have been analyzed using the circuit model developed for package level coupling. The equivalent circuit model developed for a via structure has been applied to a via passing through a ground plane and coupled via structure. The results obtained from the circuit model are compared with that obtained using a full wave simulator, HFSS.

3.1 Multi-Layered Organic (MLO) Material

In chip-package co-design to ease the interconnections and to implement the passives the multilayered material is commonly used. As the goal of this work is to study the chip-package co-design issues, it makes more sense to use a multi-layered material for the verification of the equivalent circuit model. For this reason, a Multi-Layered
Organic (MLO) material was used as a substrate for all the designs considered. The MLO material was manufactured at NSF Packaging Research Center, Georgia Institute of Technology. This MLO package was developed mainly for digital application, not RF, since the material has poor tangent value (0.02 – 0.05) [8]. In spite of this defect, efforts for adopting MLO package into RF area are going on, since it has advantage over the ceramic-based package such as lower process cost and better thermal and mechanical characteristics. The MLO material used for this work is shown in Figure 2.1, where layers 1A, 1B and 1C are metal layers; 2A, 2B and 2C are dielectric layers; and G serves as the ground plane.

![Figure 3.1. Multi-Layered Organic (MLO) material](image)

The property of each layer of the package, provided by the manufacturer, is as follows.

Layer 1A: 9 - 18 μm thick (1/4 or ½ oz Copper).

Layer 1B: 10 -15 μm thick copper

Layer 1B: 10 -20 μm thick copper

Layer 2A: 28-40 mil thick, Nelco 4000-13 FR-5 substrate, εᵣ = 3.7, tan δ = 0.01, both at 1 GHz
Layers 2B and 2C: 62.5μm thick, Shipley Dyna Via, εᵣ=3.3, tan δ = 0.02, both at 1 GHz

Different combinations of substrate and metal layers were used for different problems considered.

3.2 Package Level Coupling and Discontinuities

Two kinds of examples were considered here to study package level coupling. These examples are: Transmission lines placed in closed proximity and coupled discontinuous lines.

3.2.1 Transmission Lines Placed in Close Proximity

To verify the validity of the circuit model for package level coupling two different examples are considered in this section. One example studies the coupling between two 70 Ω transmission lines placed in closed proximity and the other considers two 100 Ω lines. In order to compare the results obtained from equivalent circuit model High Frequency Structure Simulator (HFSS) was used as a full wave solver for these examples. HFSS uses a Finite Element method to calculate fields inside the 3-D region of interest.

Example 1: 70 Ω lines

The layout of the nearly placed 70 Ω lines is shown in Figure 3.2. This layout was laid out on the MLO package shown in the Figure 3.3, where the metal layer 1B was removed to combine substrate layers 2C and 2B. The metal layer 1A was considered as a ground plane and the layout shown in Figure 3.2 was placed on metal layer 1C. The substrate thickness is extremely thin (t = 4.92 mils) in the frequency range of interest.
(1 to 5 GHz). On this substrate the width of the microstrip line (w), corresponding to 70Ω characteristic impedance, is equal to 6.48 mils. The region of interest, that is the coupled region, has a length (l) equal to 25.92 mils. In order to simplify the port definition in HFSS, all four ports were directed away from each other. The length of each additional line (l) is equal to 25.92 mils. The spacing between the two lines was set to 4 mils.

![Diagram](image)

Figure 3.2. 70 Ω lines placed in close proximity. \( l = 25.92 \) mils, \( w = 6.48 \) mils, \( s = 4 \) mils.

![Diagram](image)

Figure 3.3. The cross section of the MLO material used to study the coupling between transmission line places in close proximity.

To analyze a structure in HFSS several things needs to be taken care of. In HFSS every single element in the structure, for example ground, dielectric, and metal layer, needs to be drawn. Moreover, as the structure like microstrip has electromagnetic field
both in dielectric and air above it, the structure in HFSS has to be modified in order to include field outside the dielectric. This is the reason why an air box having height equal to at least five times the substrate thickness has to be added to the structure. Such an air box was drawn for this example, as shown in the Figure 3.4. In order to assign port a rectangular face placed symmetrically around the microstrip line was drawn. The width of the port-face should be at least five times the width of the microstrip line. For this example the width of the port was chosen to be seven times the thickness of the microstrip line. Moreover, the port-face should also cover the area above the dielectric, that is air, because to assign a microstrip port the field distribution in the air needs to be considered. Such a port-face with width equal to seven times the thickness of the microstrip line and the height equal to five times the substrate thickness is shown in Figure 3.4

Figure 3.4. The layout of the 70 Ω transmission lines placed in close proximity in HFSS
Once the drawing is complete the material properties can be assigned in 'setup material' option. The material can be assigned to a 3D element only. A 2D element has to be assigned a boundary condition. The microstrip lines, having thickness of 0.7 mils, were assigned the properties of copper. The dielectric box was assigned the properties of the MLO material ($\varepsilon_r = 3.3$ and $\tan \delta = 0.02$) and the air-box was assigned the properties of an air. A perfect-E boundary condition was applied to ground layer. All four port-faces were considered as wave-ports.

After defining ports and assigning appropriate material / boundary conditions to every element in the structure, the structure can be analyzed for a frequency range of interest. The frequency range of interest is 1 to 5 GHz. In order to make sure the mesh supports the maximum frequency of interest, the mesh was adapted at 5 GHz with maximum delta S set equal to 0.01. An interpolating sweep was then performed over 1 to 5 GHz with maximum number of solution equal to 50. The structure was then solved. The simulation took 5 minutes and 13 seconds on P-III processor (866 MHz) with 256 MB RAM. The S-parameters obtained from HFSS were used as a reference to verify the validity of the equivalent circuit model.

In order to develop equivalent circuit model the coupled region was divided into four segments as shown in Figure 3.2. This way, the length of each segment becomes 6.48 mils. The complete circuit model for coupled region is shown in Figure 3.5. The circuit element values of the equivalent circuit model can be found by using equation set (2.1) to (2.13).
By using equations (2.1) to (2.4), with line width, \( w = 6.48 \) mils; substrate height, \( h = 4.92 \) mils; relative permittivity of the substrate material, \( \varepsilon_r = 3.3 \); and separation, \( s = 4 \) mils, the values of \( C_a \) and \( C_b \) come out to be 0.174 fF and 2.455 fF respectively. For \( \varepsilon_r = 3.3 \), the value of effective relative permittivity can be found from equation (2.6). The value of \( \varepsilon_{re} \) is equal to 2.5117. With this value of effective relative permittivity, the value of self-inductance can be found from equation (2.5). As the equation (2.5) gives the value of inductance per unit length of microstrip line, the value of self inductance (L) for the length of segment equal to 6.48 mils becomes 0.0609 nH. The loss associated with the metal can be calculated from equations (2.7) to (2.9). Equations (2.8) calculates the AC resistance of the metal at certain frequency. The value of \( R_{ac} \) at \( f = 2.45 \) GHz, and conductivity \( \sigma_c = 5.7 \times 10^7 \) S/m is equal to 0.013 \( \Omega \). The value of DC resistance, obtained from equation (2.9), is equal to 0.9867 m\( \Omega \). Here, metal thickness, \( t = 0.7 \) mils; and conductivity of the metal (copper) \( \sigma_c = 5.7 \times 10^7 \) S/m were considered. Thus the total loss associated with metal (R), from equation (2.7) becomes 0.014 \( \Omega \). The value of loss associated with dielectric was calculated using equation (2.11). For frequency (f) equal to
2.45 GHz, and \( \tan \delta = 0.02 \) the value of \( G \) is equal to 18.72 M\( \Omega \). Finally, the value of coefficient of mutual coupling can be calculated from equation (2.13) and Table 2.1. For 70 \( \Omega \) lines the table reads the value of \( A \) and \( B \) equal to 0.65 and 0.1968 respectively. Using these values of \( A \) and \( B \), \( k \) can be obtained from equation (2.13). The value of \( k \) is equal to 0.292. All these values of circuit elements are tabulated in Table 3.1.

<table>
<thead>
<tr>
<th>Circuit Elements</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. ( C_s )</td>
<td>0.174 fF</td>
</tr>
<tr>
<td>2. ( C_b )</td>
<td>2.455 fF</td>
</tr>
<tr>
<td>3. ( L )</td>
<td>0.0609 nH</td>
</tr>
<tr>
<td>4. ( R )</td>
<td>0.014 ( \Omega )</td>
</tr>
<tr>
<td>5. ( G )</td>
<td>18.72 M( \Omega )</td>
</tr>
<tr>
<td>6. ( k )</td>
<td>0.292</td>
</tr>
</tbody>
</table>

The circuit model shown in Figure 3.5 with circuit element values mentioned in Table 3.1 was simulated in a circuit simulator (Serenade). The S-parameter obtained from the circuit simulator were compared with that obtained from HFSS. These results are compared in Figure 3.6. In the figure solid lines corresponds to results obtained from equivalent circuit model and dashed lines corresponds to results obtained from HFSS. The Figure 3.6 (a) compares return loss \((S_{11})\), Figure 3.6 (b) compares near end coupling \((S_{12})\), Figure 3.6 (c) compares far end coupling \((S_{13})\), and Figure 3.6 (d) compares the through transmission. All results are plotted in dB and for easy comparison they are all plotted on a same scale. The results obtained from equivalent circuit model compares well with that obtained from HFSS.
Figure 3.6. Comparison of the S-parameters for the 70 Ω transmission lines placed in close proximity
(a) S_{11} (b) S_{12} (c) S_{13} and (d) S_{14}

The S_{11} parameters obtained from the circuit model differ slightly from that obtained from HFSS but follows the same curve. Whereas, S_{12} and S_{13} parameters obtained from equivalent circuit model compares really well with that obtained from HFSS. Almost all the energy input at port 1 is transmitted to port 4 because they are connected to each other by a microstrip transmission line. This is the reason why the S_{14} parameters (Figure 3.6 (d)) are very close to zero through out the frequency range of
interest. The $S_{14}$ parameters obtained from equivalent circuit model are so close to that obtained from HFSS that they are hard to identify.

Thus, the circuit model predicts both near and far end coupling, through transmission, and reflection very close to that predicted by HFSS. The circuit model has been verified for spacing between the microstrip lines from 2 mils to 7 mils. In conclusion, the circuit model can be trusted to predict through transmission, return loss, near end coupling, and far end coupling for coupled 70 $\Omega$ line layouts in the frequency range from 1 to 5 GHz.

**Example 2: 100 $\Omega$ lines**

A similar kind of analysis was performed for 100 $\Omega$ transmission lines placed in close proximity. Figure 3.7 shows the layout that was considered. It is similar to that shown for 70 $\Omega$ line except that the dimensions are different. For this structure the dielectric material shown in Figure 3.3 was considered, where the transmission line structure was placed on metal layer 1C and the metal layer 1A was considered as a ground plane. The width of the microstrip line is 2.99 mils (corresponding to 100 $\Omega$ line) and the length is equal to 11.96 mils. In order to simplify the segmentation, the length of the coupled region was set equal to four times the thickness of the microstrip line. To simplify the port definition in HFSS all four ports were directed away from each other.

To make sure the mesh is good enough for the highest frequency of interest (5 GHz), the mesh was adapted at 5 GHz with maximum delta $S$ set equal to 0.01. The interpolating frequency sweep was then performed over a frequency range of 1 to 5 GHz. The S-parameters obtained this way were used as a reference to verify the circuit model.
Chapter 3: Addressing Package Level Discontinuities & Chip-Package Interconnect

Figure 3.7. 100 Ω lines placed in close proximity. l = 11.96 mils, w = 2.99 mils, s = 2 mils.

The circuit model for 100 Ω lines placed in close proximity will be the same as that shown in Figure 5.8. The only difference would be the values of circuit elements. The circuit element values corresponding to microstrip line thickness; w = 2.99 mils, substrate height; h = 4.92 mils, substrate relative permittivity ε_r = 3.3, loss tangent; tan δ = 0.02; spacing between the microstrip lines, s = 2 mils; metal thickness, t = 0.7 mils; metal conductivity, σ_c = 5.7 x 10^7 S/m; frequency of interest, f = 2.45 GHz; and microstrip line characteristic impedance, Z_0 = 100 Ω are shown in Table 3.2.

Table 3.2. Circuit element values for the equivalent circuit model of the 100 Ω lines placed in close proximity

<table>
<thead>
<tr>
<th>Circuit Elements</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. C_a</td>
<td>0.075 fF</td>
</tr>
<tr>
<td>2. C_b</td>
<td>1.741 fF</td>
</tr>
<tr>
<td>3. L</td>
<td>0.0392 nH</td>
</tr>
<tr>
<td>4. R</td>
<td>0.014 Ω</td>
</tr>
<tr>
<td>5. G</td>
<td>42.33 MΩ</td>
</tr>
<tr>
<td>6. k</td>
<td>0.8388</td>
</tr>
</tbody>
</table>
In the figure solid lines corresponds to results obtained from equivalent circuit model and dashed lines corresponds to results obtained from HFSS. Figures 3.8 (a) to (d) compare return loss ($S_{11}$), near end coupling ($S_{12}$), far end coupling ($S_{13}$), and the through transmission respectively. All results are plotted in dB and for easy comparison they are all plotted on a same scale. It can be seen from the graphs presented that the equivalent
circuit model predicts all four S-parameters very well compare to that obtained from HFSS.

Thus, so far, the equivalent circuit model has been successfully applied to 70 Ω lines and 100 Ω lines placed in close proximity. In the next section the equivalent circuit model has been applied to a discontinuous lines that are placed in close proximity.

3.2.2 Package Level Discontinuities

In the previous section the equivalent circuit model has been verified with different characteristic impedance transmission lines placed nearby. In this section a distinct example has been used to verify the validity of the circuit model. The Layout is shown in Figure 3.9. As shown in the figure, discontinuous lines are considered as opposed to what was considered so far. A 70 Ω line is going into a 50 Ω line and thus giving rise to a discontinuity. Such two transmission lines are placed parallel to each other. Just like other examples, the layout shown in Figure 3.9 was also laid out on the MLO material shown in Figure 3.3. The goal is to use the circuit model and segmentation technique developed so far to predict through transmission, return loss, near end coupling, and far end couplings.

In the figure, \( W_1 = 11.58 \) mils (corresponding to a 50 Ω line), \( W_2 = 6.48 \) mils (corresponding to a 70 Ω line), \( l_1 = 26.16 \) mils, \( l_2 = 12.97 \) mils, \( S_1 = 2 \) mils, and \( S_2 = 7.1 \) mils. Thus the coupled region of interest has the total length equal to \((l_1 + l_2)\) 39.13 mils. Such a four-port structure was simulated in HFSS with 18 GHz as adaptive frequency and maximum delta S equal to 0.01. The mesh of the structure gets adapted at this frequency. The interpolating frequency sweep was then performed over 1 to 20 GHz.
In order to obtain the equivalent circuit model for the layout shown in Figure 3.9, each line (50 Ω and 70 Ω) was divided into two segments. This way, the total number of segments is equal to four. The equivalent circuit model for each segment was found and cascaded together. The complete equivalent circuit model is shown in Figure 3.10. The circuit element values for both coupled 50 Ω lines and 70 Ω lines can be found using equations (2.1) to (2.13). The circuit element values for coupled discontinuous lines were obtained by considering microstrip line thickness; \( w_1 = 11.58 \) mils (50 Ω), \( w_2 = 6.48 \) mils (70 Ω), substrate height; \( h = 4.92 \) mils, substrate relative permittivity \( \varepsilon_r = 3.3 \), loss tangent; \( \tan \delta = 0.02 \); spacing between the microstrip lines, \( s_1 = 2 \) mils (50 Ω), \( s_2 = 7.1 \) mils (70 Ω); metal thickness, \( t = 0.7 \) mils; metal conductivity, \( \sigma_c = 5.7 \times 10^7 \) S/m; frequency of interest, \( f = 10 \) GHz (because the frequency range of interest is 1 to 20 GHz).
Chapter 3: Addressing Package Level Discontinuities & Chip-Package Interconnect

GHz); and microstrip line characteristic impedance, $Z_0 = 50 \, \Omega$ (for 50 \, \Omega line) and $Z_0 = 70 \, \Omega$ (for 70 \, \Omega line). The values of circuit elements are displayed in Table 3.3.

![Equivalent circuit model of the coupled discontinuous lines](image)

**Figure 3.10. Equivalent circuit model of the coupled discontinuous lines**

<table>
<thead>
<tr>
<th>Circuit Elements</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. $C_{a70}$</td>
<td>0.231 fF</td>
</tr>
<tr>
<td>2. $C_{b70}$</td>
<td>1.89 fF</td>
</tr>
<tr>
<td>3. $L_{70}$</td>
<td>0.061 nH</td>
</tr>
<tr>
<td>4. $R_{70}$</td>
<td>0.027 , \Omega</td>
</tr>
<tr>
<td>5. $G_{70}$</td>
<td>3.45 M,\Omega</td>
</tr>
<tr>
<td>6. $k_{70}$</td>
<td>0.1571</td>
</tr>
<tr>
<td>7. $C_{a50}$</td>
<td>0.21 fF</td>
</tr>
<tr>
<td>8. $C_{b50}$</td>
<td>4.799 fF</td>
</tr>
<tr>
<td>9. $L_{50}$</td>
<td>0.0793 nH</td>
</tr>
<tr>
<td>10. $R_{50}$</td>
<td>0.027 , \Omega</td>
</tr>
<tr>
<td>11. $G_{50}$</td>
<td>3.79 M,\Omega</td>
</tr>
<tr>
<td>12. $k_{50}$</td>
<td>0.1947</td>
</tr>
</tbody>
</table>

Table 3.3. Circuit element values for the equivalent circuit model of the discontinuous lines placed in close proximity

In the circuit model first two segments corresponds to a coupled 70 \, \Omega lines and last two segments corresponds to a coupled 50 \, \Omega lines. The circuit model shown in Figure 3.10 with the elements values listed in Table 3.3 was simulated in a circuit simulator (Serenade). Results obtained from the circuit simulator and the full wave
simulator were compared. Figure 3.11 shows this comparison. Figures 3.11 (a) to (d) compare return loss ($S_{11}$), near end coupling ($S_{12}$), far end coupling ($S_{13}$), and the through transmission respectively. All results are plotted in dB and for easy comparison they are shown on a same scale. In the figure the solid lines correspond to results obtained from the equivalent circuit model and the dashed lines corresponds to results obtained from HFSS.

![Figure 3.11](image)

Figure 3.11. Comparison of the S-parameters for the discontinuous lines placed in close proximity (a) $S_{11}$ (b) $S_{12}$ (c) $S_{13}$ and (d) $S_{14}$
The comparison shows that the $S_{11}$ obtained from the circuit simulator follows the same curve as that by $S_{11}$ obtained from HFSS up to about 10 GHz. The curve then deviates slightly from that predicted by HFSS. The $S_{12}$ parameter obtained from the equivalent circuit model and that obtained from HFSS follow the same curve through out the frequency range of interest. The $S_{13}$ parameter does not follow the same curve as that obtained by HFSS but the values predicted by the equivalent circuit model are close to that predicted by HFSS. These two curves meet at about 6.5 GHz. Just like other other examples the $S_{14}$ parameter is close to zero decibel for both equivalent circuit model and HFSS through the frequency range of interest. The deviation in the results obtained from the circuit model and HFSS could be because of the fact that the circuit model does not consider the effect of charge getting stored at the transition from 70 Ω lines to 50 Ω lines.

Thus, in this section the validity of the segmentation technique and the circuit model have been verified using several different examples. The circuit model has been used to predict coupling between 50 Ω lines and 70 Ω lines placed in close proximity. Finally the circuit model was used to study near and far end coupling between discontinuous lines placed very close to each other. The results obtained from the circuit model and the full wave solver (HFSS) matched very well. This finishes the verification of the segmentation technique and the equivalent circuit model for package level coupling and discontinuity.

The segmentation technique has also been extended to study the coupling at chip-package interconnect level. The next section verifies validity the circuit model obtained
for a single via passing through ground plane and for couple via structure by comparing the circuit model results with that obtained from a full wave solver.

3.3 Chip-Package Interconnects

In section 2.4 the simpler circuit model for a single via passing through ground structure was proposed. In addition, the circuit model for single via was also extended to study the coupling between the coupled via structure. In this section, proposed circuit models are verified by comparing the results obtained from the circuit model to that obtained from a full wave solver (HFSS) and published results. The first section discusses the circuit model for the via passing through a ground plane and compares the results of the circuit model to that published in [14]. The second section compares the results of the circuit model for the coupled via structure with that obtained from the HFSS.

3.3.1 Single Via Passing Through Ground Plane

A single via passing through a ground plane is shown in Figure 2.7. An example is considered to compare with published results [14]. The dimensions of the structure are: $D_{\text{thruhole}} = 1.6$ mm, $D_{\text{via}} = 0.8$ mm, $D_{\text{air}} = 0.6$ mm, and $H = 0.5$ mm. The equivalent circuit model for the single via structure is shown in Figure 2.8.

Using equation (2.14), with $D_{\text{via}} = 0.08$ cm, length of the via $l = 0.1$ cm, and frequency of interest $f = 40$ GHz, the value of inductance is equal to 0.2027 nH. The values of loss associated with via (R) and capacitance $C_a$, which represents the storage of energy between the ground plane and the via were obtained using the optimization tool available in the circuit simulator (Serenade). The optimized values of R and $C_a$ and the
value of $L$ are tabulated in Table 3.4. The circuit model of Figure 2.8 with the circuit element values presented in Table 3.4 was simulated in Serenade. Figure 3.12 shows the comparison of the S-parameters obtained from the circuit model and that published in [14].

Table 3.4. The circuit element values for single via passing through ground plane structure

<table>
<thead>
<tr>
<th>Circuit Element</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. $L$</td>
<td>0.2027 nH</td>
</tr>
<tr>
<td>2. $R$</td>
<td>2.2 $\Omega$</td>
</tr>
<tr>
<td>3. $C_a$</td>
<td>0.1765 pF</td>
</tr>
</tbody>
</table>

![Figure 3.12](image)

Figure 3.12. Comparison of the S-parameters for the discontinuous lines placed in close proximity (a) $S_{11}$ (b) $S_{12}$

The dashed lines in the figures correspond the results published in [14] and the solid lines corresponds to that obtained from a simpler circuit model proposed in this work. The results obtained from the equivalent circuit model follows very well to that published. The return loss ($S_{11}$) shows a resonance at 34 GHz. This is probably because of the energy getting stored between the via and the ground plane. The equivalent circuit
model presented in this thesis also predicts this resonance. Both the results match very well up to 40 GHz. With these results, the equivalent circuit model for a single via can be trusted. The circuit model for a single via can, now, be used further to verify the validity of the in the equivalent circuit model for the coupled-via structure.

### 3.3.2 Coupled Via Structure

The equivalent circuit model for the coupled via structure shown in Figure 2.9 is considered in this section. First, the coupled via structure was simulated in HFSS to obtain S-parameters. These results were then used as a reference to verify the validity of the equivalent circuit model.

As shown in the Figure 2.9, two transmission lines of width \( w_1 = 11.58 \) mils (corresponding to 50 \( \Omega \)); placed on dielectric thickness, \( h + h = 2.46 \) mils; are running towards each other. These transmission lines are connected to microstrip transmission lines of width \( w_2 = 5.79 \) mils (corresponding to 50 \( \Omega \)); placed on dielectric thickness, \( h = 1.43 \) mils; through vias. Both transmission line of width \( w_1 = 11.58 \) mils and \( w_2 = 5.79 \) mils corresponds to 50 \( \Omega \) because they are placed at different elevation and so the substrate thickness used by them is different. As all four lines are moving away from each other, the only possible way the coupling between ports 1–3 and ports 2–3 can occur is through vias. The goal here was to predict such coupling using equivalent circuit model.

A 3-D structure drawn in HFSS looked like the one shown in Figure 3.13. In this Figure the via transition from one metal plane to another can be seen easily. As shown in the figure, two air boxes, one having thickness equal to about six times the substrate
thickness and the another with three times the thickness of the substrate, were drawn to help HFSS building smoother mesh in the air region. The microstrip lines and ground planes were considered to be infinitely thin and assigned a perfect E boundary (assigning perfect E boundary to infinitely thin element is equivalent to assigning a perfect conductor to them). The vias were assigned the properties of a copper material. The dielectric material has relative permittivity, $\varepsilon_r = 3.3$ and loss tangent, $\tan \delta = 0.02$. In order to assign ports, four different rectangular sheets were drawn. These rectangular port-faces had height equal to total height of the structure and width equal to six times the thickness of the corresponding microstrip line. All ports were considered as wave ports.

![Diagram](image)

**Figure 3.13.** A 3-D view of the coupled via structure drawn in HFSS

In order to obtain solution up to 40 GHz, the mesh of the structure was adapted at 38 GHz with maximum delta S set to 0.02. An interpolating frequency sweep was then
performed over 1 to 40 GHz. The solution obtained from HFSS was considered as a reference for the equivalent circuit model.

The equivalent circuit model for the coupled via structure was developed using a segmentation technique, shown in Figure 2.10 and shown below for convenience. The coupled via was segmented into two segments. The equivalent circuit model for each segment, shown in Figure 2.11, was cascaded in order to obtain the circuit model for the whole coupled via structure. Such a circuit model is shown in Figure 2.12.

For each segment of the coupled via structure, the length of via, \( l = 1.23 \text{ mils} = 3.124 \times 10^{-3} \text{ cm} \), the diameter of the via, \( d_{\text{via}} = 2 \text{ mils} = 50.8 \times 10^{-6} \text{ cm} \), distance between the center of vias, and \( D = 7.79 \text{ mils} = 19.78 \times 10^{-3} \text{ cm} \). With this dimensions the value of self-inductance, \( L \), and mutual capacitance, \( C_b \), can be found using equations (2.14) and (15). The value of loss associated with the via and the coefficient of mutual coupling were optimized using the optimization tool available in the circuit simulator. The values of \( L \) and \( C_b \), along with optimized values of resistor, \( R \), and coefficient of mutual coupling, \( k \), are displayed in Table 3.5.

<table>
<thead>
<tr>
<th>Circuit elements</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. ( L )</td>
<td>4.86 fH</td>
</tr>
<tr>
<td>2. ( C_b )</td>
<td>2.2298 fF</td>
</tr>
<tr>
<td>3. ( k )</td>
<td>0.4</td>
</tr>
<tr>
<td>4. ( R )</td>
<td>3.3 ( \Omega )</td>
</tr>
</tbody>
</table>

The circuit model, shown in Figure 2.12 with circuit element values tabulated in Table 3.5, was simulated in Serenade. The results obtained from the circuit simulator
were compared with that obtained from the full wave simulator (HFSS). Figure 3.14 compares these two results.

![Comparison of S-parameters](image)

Figure 3.14. Comparison of the S-parameters for the discontinuous lines placed in close proximity (a) S\(_{11}\) (b) S\(_{12}\) (c) S\(_{13}\) and (d) S\(_{14}\)

Just like other figures, in this figure too the solid lines corresponds to results obtained from the equivalent circuit model and the dashed lines corresponds to the results obtained from HFSS. Figure 3.14 (a) shows that the S\(_{11}\)-parameter obtained from the equivalent circuit model deviates from that obtained from HFSS after 20 GHz. Whereas,
the near end coupling ($S_{12}$) and far end coupling ($S_{13}$) obtained from the equivalent circuit model compare very well with that obtained from HFSS. Such is the case with $S_{14}$ parameter too. The graphs of $S_{12}$, $S_{13}$, and $S_{14}$ obtained from both equivalent circuit model and that HFSS almost follow each other throughout the frequency range of interest. This verifies the validity of the segmentation technique and the equivalent circuit model.

In summary, in this chapter the segmentation technique and the equivalent circuit model for both package level discontinuity / coupling and chip-package interconnect level coupling was verified. The results obtained from the equivalent circuit model were compared with that obtained from a full wave solver or that published in papers. The results compared very well.

To study the chip-package co-design issues as related to integrated antennas, a test bed consisting of a microstrip antenna and associated embedded matching network was implemented. The next chapter is devoted to the design of different test bed microstrip antennas and associated matching network. Chapter 5 presents the chip-package co-design issues associated with the integrated antennas in both frequency and time domain.
In order to study the chip-package co-design issues associated with the integrated antenna, it is at first necessary to design microstrip patch antenna on the multilayered organic material (MLO). The goal is to design a microstrip antenna within the multilayered organic material (MLO), which is electrically very thin at 2.45 GHz, which therefore makes the task extremely challenging. The performance of the antenna can be improved using a matching network. In the present work the matching network is built using embedded components. Even with matching network the performance of the microstrip antenna on an electrically thin substrate was in question and so other antennas, including electromagnetically coupled patch and dipole, on thicker substrate have been built as alternative antennas. Total four different kinds of microstrip antennas are designed and implemented on MLO. All these antennas are designed to work at 2.45 GHz. The next step after designing the antennas is to study chip-package co-design issues. This is the study of how sensitive the microstrip antenna is to other nearby
components both in digital and analog domain. Therefore this study is also termed as sensitivity analysis. Out of all designed microstrip antennas, the one built on very thin substrate has been chosen to perform sensitivity analysis. This is because the antenna on thinner substrate is the desired one. This chapter is devoted to the design of four different microstrip antennas built for a RF front end. This chapter also includes the design of integrated matching network for the antenna built on extremely thin substrate. The study of chip-package co-design issues is left to chapter 5.

4.1 The Integrated Antenna

The design procedure for an integrated microstrip antenna is well known. A survey of microstrip antenna elements, with emphasis on theoretical and practical design techniques, is presented in [19]; and the progress in microstrip antennas as applied to different kinds of applications and the reliability of such systems is reviewed in [20-22]. The design of Electro-Magnetically Coupled (EMC) microstrip patch antenna [23-26] and EMC dipole antennas [27] for bandwidth enhancement are discussed in great depth.

Microstrip antennas are low profile, comfortable to planar and nonplanar surface, simple and inexpensive to manufacture using modern printed-circuit technology, mechanically robust when mounted on rigid surfaces, compatible with MMIC designs, lightweight antennas most suitable for aerospace, missile, satellite and mobile applications [53]. Moreover, when the particular patch shape and mode are selected they are very versatile in terms of resonant frequency, polarization, pattern and impedance.
Because of their low power handling capability these antennas could be used with low power transmitters and receiver applications.

Major operational disadvantages of microstrip antennas are their low efficiency, high Q (sometimes in excess of 100), poor polarization purity, poor scan performance, spurious feed radiation and very narrow frequency bandwidth.

Microstrip antennas consist of a patch of metalization on a grounded substrate. There are numerous substrates that can be used for the design of microstrip antennas, and their dielectric constants are usually in the range of $2.2 \leq \varepsilon_r \leq 12$. The dielectric substrate that are most desirable for antenna performance are thick substrates whose dielectric constant is in the lower end of the range because they provide better efficiency, larger bandwidth, loosely bound fields for radiation into space, but at the expense of larger element size. Thin substrates with larger dielectric constants are desirable for microwave circuitry because they require tightly bound fields to minimize undesired radiation and coupling, and lead smaller element sizes.

A variety of passive microstrip antenna configurations, including Electro-Magnetically Coupled (EMC) square patch and dipole, were analytically modeled using a full wave analysis provided by the Ensemble design tool. Ensemble was chosen for designing the microstrip antennas compare to HFSS because Ensemble is supposed to be very good for microstrip antenna simulations. Ensemble solves for far field, near field, $S/Z/Y$ parameters, input impedance, and current distribution of a given 2D layout using Method of Moment. As the antenna configuration modeled on very thin substrate is desired one, only that antenna was chosen to perform sensitivity analysis out of all
designs. These antennas were designed on a multilayered organic material, the property of which is discussed in Section 3.1. These kinds of multilayered packages are mainly used for implementing passives for system on package (SOP) designs.

Four kinds of microstrip antennas were designed using various combinations of metal, ground, and substrate layers. These antennas are:

1. Microstrip square patch antenna on an electrically thin substrate (antenna and feed system on 1C, 1A as ground, G and 1B were removed, and 2A served as supporting material).

2. Microstrip square patch antenna on an electrically thick substrate (antenna and feed system on 1C, G as ground, 1A and 1B were removed).

3. Electro-Magnetically Coupled (EMC) square patch antenna (antenna on 1C, feed system on 1A, G as ground, 1B was removed), [23-26].

4. EMC dipole (antenna on 1C, feed system on 1A, G as ground, 1B was removed), [27].

Out of all these antennas, the antenna that was built a thin substrate is desired because of its low profile. The antenna (1) has been built on such thin substrate, and so is chosen to perform sensitivity analysis. The following subsections discuss the layout and the MLO package used for each of above-mentioned antennas.
4.2 Microstrip Square Patch Antenna on an Electrically Thin Substrate

The initial challenge in the antenna design is provided by the MLO package where the thickness of the dielectric layer is (4.92 mils) is electrically extremely small at 2.45 GHz. The square microstrip patch antenna has been designed mainly to reduce the size of the antenna. The difficulty with using such a thin substrate is to obtain a desired 50 Ω input impedance point in order to match it with next RF front end block, band pass filter. It is because of this reason a matching network, using integrated (embedded) passives, has been designed to match antenna’s input impedance to that of bandpass filter. The MLO material used for this antenna configuration is shown in Figure 4.1 (b). The antenna layout, shown in Figure 4.1 (a), is placed on metal layer 1C. The metal layer 1B is removed in order to have total substrate thickness equal to 4.92 mils. The metal layer 1A is used as a ground plane, and the ground plane G is removed. The substrate layer, 2A, having thickness of 30 mils, is added in the substrate just to add mechanical support for easy handling.

In the microstrip patch antenna, the size of the patch, that is ‘a’ and ‘b’, determines the resonance frequency of the antenna. The inset (s) determines the input impedance looking in to the antenna. The gap (g), in the design, avoids cross talk between feed and nearby edges of the antenna and does not have much impact on overall performance of the antenna. For a given substrate, the characteristic impedance of the feed line of the antenna is determined by the line width, w. The width ‘w’ of the feed line was chosen so as to have the characteristic impedance equal to 50 Ω. The Co-Planar
Wave guide (CPW) shown in the Figure 4.1(a) is only for measurement purpose. The layout of the CPW used for this antenna configuration is shown in Figure 4.2. The CPW, laid out on metal layer 1C, has an input impedance of 50 Ω. As the impedances of both the feed line of the antenna and that of CPW, they can be connected directly to each other.

Figure 4.1. Microstrip square patch antenna-built on thin substrate- layout and MLO cross-section. (a) Layout of the Microstrip Square Patch Antenna (a = 1360 mils, b = 1360 mils, s = 100 mils, l = 200 mils, w = 11.96 mils (corresponding to 50 Ω), g = 36 mils) (b) The cross-section of the MLO material used (thickness of 2C + 2B = 4.92 mils, thickness of 2A = 30 mils).

Figure 4.2. Co-planar wave guide layout used for measurement; where, a = 36 mils, b = 40 mils, l = 20 mils, w = 12 mils, g = 2 mils.
In Figure 4.2, the rectangular patches having size of ‘a x b’ are connected to ground plane (G) through circular via and so they work as ground planes for CPW arrangement. The transmission line of length ‘l’ needs to be connected to the signal line (feed line) of the antenna. As the thickness of the signal line of the CPW (12 mils) is very close to that of feed line of the microstrip antenna (11.96 mils) they are connected directly to each other. The only discontinuity that exists here is the transaction of ground plane from side ways to bottom. This can be taken care of by using some kind of calibration process during measurement. More rigorous analysis and different ways of CPW to microstrip transition has been discussed in [54-56].

4.2.1 Predicted Results

The antenna shown in Figure 4.1, with \( a = 1360 \) mils, \( b = 1360 \) mils, \( s = 100 \) mils, \( l = 200 \) mils, \( w = 11.96 \) mils (corresponding to 50 \( \Omega \)), \( g = 36 \) mils, thickness of \( 2C + 2B = 4.92 \) mils, thickness of \( 2A = 30 \) mils has been analytically modeled in Ensemble. An adaptive analysis is performed at 2.45 GHz with “Target Maximum Delta Norm” set to 0.02. The mesh is first adapted at 2.45 GHz. After making sure that the mesh is good enough at adaptive frequency, the solution can then be obtained over a frequency range of interest, that is 2 GHz to 3 GHz. The solution has been obtained using fast frequency sweep. The solution took 56 seconds on P-III processor (866 MHz) with 256 MB RAM. With mentioned set of values Ensemble predicted the resonant frequency at 2.49 GHz, which is close to the desired resonant frequency. At the resonance the Return Loss of the antenna has been found to be equal to \(-5.35\) dB, which corresponds to a VSWR of 3.36.
The variation of Return Loss and the VSWR over a frequency range of interest are shown in Figure 4.3 (a) and (b) respectively.

A poor value of the Return Loss and so the VSWR is expected, because of very thin substrate (4.92 mils) used. In order to improve the Return Loss a matching circuit using embedded components has been designed. This matching network with embedded components improved the Return Loss from −5.35 dB to −27.93 dB. The design of the matching network is explained in section 4.3

Another useful characteristic of antenna is far field pattern, which gives an idea of how well the antenna is radiating. The antenna far field pattern is shown in Figure 4.4. The plot is in dB and is normalized to the maximum radiation value at theta = 0 degree. As shown in the figure the antenna is radiating well in the direction right angle to the plane of the patch. Moreover the directivity of the antenna is poor.
Chapter 4: Integrated Antenna System on MLO Material for RF Front End

For the configuration being discussed in this section so far, three different kinds of antennas with different insets (s) have been analytically modeled using Ensemble. The Return Loss and VSWR, at resonant frequency, for different inset are tabulated in Table 4.1. The table shows change in resonant frequency \( f_r \), input impedance \( Z_{in} \), VSWR, and Return Loss with respect to change in inset (s).

Table 4.1. Change in antenna parameters with respect to inset for square patch antenna built on thin substrate

<table>
<thead>
<tr>
<th>No.</th>
<th>a = b (mil)</th>
<th>s (mil)</th>
<th>Thickness (mil)</th>
<th>( f_r ) (GHz)</th>
<th>( Z_{in} ) (( \Omega ))</th>
<th>VSWR</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1360</td>
<td>90</td>
<td>4.92</td>
<td>2.49</td>
<td>17.3 – j 17.9</td>
<td>3.31</td>
<td>-5.42</td>
</tr>
<tr>
<td>2</td>
<td>1360</td>
<td>100</td>
<td>4.92</td>
<td>2.49</td>
<td>18.1 – j 22.1</td>
<td>3.36</td>
<td>-5.35</td>
</tr>
<tr>
<td>3</td>
<td>1360</td>
<td>120</td>
<td>4.92</td>
<td>2.49</td>
<td>19.1 – j 27.6</td>
<td>3.51</td>
<td>-5.08</td>
</tr>
</tbody>
</table>
4.2.2 Measured Results

The predicted layouts and results of the microstrip square patch antenna and other antennas like EMC patch and dipole antennas, discussed in following sections, were sent for fabrication to NSF Packaging Center, Georgia Institute of Technology. Microstrip layouts usually take two to three fabrication runs before the design is finalized, because the layout performance is usually predicted using full wave solvers considering several ideal conditions. Moreover, fabrication process cannot be error proof. The fabrication for discussed antennas was done by students, and not by professionals. In this section the predicted and measured results are compared.

The layout of the microstrip square patch antenna- built on a thin substrate looked similar to the one shown in Figure 4.5 (a). The figure shows the metal layer 1C, where the black region is dielectric material and shaded region is metal (copper). The CPW region is expanded and shown in Figure 4.5 (b). More than one vias are used in order to ground each of the two metal planes. This is done just to make sure that the planes are perfectly grounded. The metal layer 1B is removed for this particular configuration of the antenna and metal layer 1A is all covered with copper material acting as a ground plane. The vias shown in Figure 4.5 (b) connect metal layer 1C to metal layer 1A.

As there was a problem with the deposition of the materials to build the MLO material, the MLO package contained many air bubbles in it. Moreover, there were difficulties in laying out the CPW next to signal line because they are placed only 2 mils apart. These reasons avoided the successful implementation of most of the microstrip antennas.
The predicted and measured return losses for the antenna shown in Figure 4.5 are compared in Figure 4.6. The figure shows reasonable agreement between predicted and measured data. These data are separated by less than 1 dB. The predicted resonance frequency is slightly less than that measured. The resonant frequency is dependent on the patch size. The required resonance frequency can be obtained by slightly tweaking the patch size. Thus, Ensemble predicts the input characteristics of the antenna very well.
Unfortunately, the implementation of other antenna configurations was not very successful. It is because of this reason measured results for the microstrip antenna built on electrically thick substrate, EMC patch, and EMC dipole are not discussed in this thesis.

4.3 Matching Circuitry with Embedded Passive Components

As mentioned in the previous section, there was difficulty in obtaining a good return loss for the microstrip patch antenna. The best value for return loss that could be achieved, with inset equal to 100 mils, was $-5.35$ dB. Hence it is necessary to design a matching network. For the same reasons as in the case of microstrip radiators, embedded passives (inductors and capacitors) are vary attractive for their low profile, ease of
fabrication and low cost. Therefore the matching network has been designed using embedded components. For the layout of the microstrip antenna, shown in previous section, the input impedance at resonance, obtained from Ensemble, is $(18.8 + j \ 28.1) \ \Omega$. The matching circuit matches this impedance to $50\Omega$. In order to obtain the value of elements in matching network, it is first designed using lumped elements. Designing matching network using lumped elements is easier and faster using circuit simulator. Once the circuit element values are obtained, integrated passives can be designed having the same value as that of lumped element.

4.3.1 Lumped Matching Network

At first, the lumped matching network is designed. The value of each element can be found using a method using smith chart, which is very well explained in [57]. Using these circuit element values a lumped matching network has been simulated in a circuit simulator, Serenade. With the help of Serenade the simulation becomes straightforward and less time consuming. Moreover such simulation makes sure that the value of elements used in the circuit is correct. The values of inductor and capacitor have been kept in the range, $L < 7 \ \text{nH}$ and $C < 5 \ \text{pF}$, in order to have high value of $Q$. Working with these constraints an L-matching network using lumped elements is designed. The L-matching network is desirable since it has reduced number of components and can be made as a part of the band pass filter that follows in the RF front-end circuitry.

The lumped matching network, simulated in Serenade, is shown in Figure 4.7. The matching network consists of a shunt inductor ($L = 2.56 \ \text{nH} < 7 \ \text{nH}$) and a series capacitor ($C = 1.29 \ \text{pF} < 5 \ \text{pF}$). This combination of inductor and capacitor is connected
to the S-matrix box of the microstrip patch antenna. Figure 4.8 shows dramatic improvement in the Return Loss. Using lumped matching network the Return Loss improves from $-5.35$ dB to $-27.93$ dB corresponding to a VSWR of 1.08. The similar kind of response is predicted for the embedded matching network.

Figure 4.7. T-section matching network with lumped elements, where, $L = 2.56$ nH and $C = 1.29$ pF

Figure 4.8. The improvement in Return Loss with matching network.
4.3.2 Embedded Matching Network

At present the methodology for inductor design is to obtain an inductance value for a chosen layout as opposed to obtaining a layout for a given value of inductance. Due to this reason obtaining the layout of an embedded inductor that has inductance value equal to 2.56 nH is challenging. Several simple and accurate expressions for the inductance of various shapes are mentioned in [58]. Circuit modeling approach to analyze on chip inductors is shown in [59-60]. For this work it is required to develop the layout of the planar inductor that has a required inductance value. As, at present, no such equations are available, the layout has been obtained using optimization method. The layout of embedded inductor is optimized in Ensemble with initial values obtained from equations available in [58]. These equations are based on Current Sheet Approximation, Data Fitting Technique, and Wheeler Formula. All these equations are for finding out the value of inductance for a given layout of the embedded inductor. A matlab program has been written that keeps some of the parameters constant and iterates other [Appendix-A]. This way, this program gives a layout of the inductor that has desired value of inductance. As these expressions are not accurate enough, the layout obtained with the program is also not accurate. These values are used as initial values for the optimization done in Ensemble. Similar kind of optimization has been performed, in Ensemble, for the layout of the capacitor. For this optimization the initial values are obtained from parallel plate capacitor equation.

The layout of the embedded matching network is displayed in Figure 4.9, and corresponding cross-section is shown in Figure 4.10. The inductor is placed on the top
layer with the center connected to the ground, metal layer 1A. The dimensions of the inductor are tabulated in Table 4.2. The integrated inductor is a two-turn \( (n = 2) \) circular spiral with the outer diameter of 106.6 mils, line width of 9.97 mils and spacing 4.72 mils. The other end of the inductor is connected to metal layer 1B through a via on which is placed a microstrip transmission line that leads to one of the plates of the capacitor. The second plate of the capacitor is placed on the top layer (Layer 1C). The plate size is 65.5 mils square with a separation of 2.46 mils. A 50Ω microstrip transmission line of length 360 mils leads to the antenna, which is 1360 mils square with an inset feed of 100 mils and substrate thickness of 4.92 mils.

![Figure 4.9. The layout of the matching network with embedded inductor and capacitor](image)

Table 4.2. Dimension of the embedded inductor

<table>
<thead>
<tr>
<th>Integrated Inductor Dimensions</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameters</td>
<td></td>
</tr>
<tr>
<td>Outer diameter</td>
<td>106.6 mils</td>
</tr>
<tr>
<td>Width</td>
<td>9.97 mils</td>
</tr>
<tr>
<td>Spacing</td>
<td>4.72 mils</td>
</tr>
<tr>
<td>Number of turns</td>
<td>2</td>
</tr>
</tbody>
</table>
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Figure 4.10. Cross Sectional View of Matching Network

The response of the entire circuit with the antenna has been simulated in Ensemble. The return loss of the distributed circuit compares well with that of the lumped circuit where the antenna is modeled as a black box containing S-parameters. These results are compared in Figure 4.11. Table 4.3 compares the results for the antenna without matching network simulated in Ensemble, antenna with lumped matching network simulated in Serenade, and the antenna with integrated matching network simulated in Ensemble. As can be seen from the table the results of lumped and embedded matching network are comparing well.

Figure 4.11. Comparison of Return Loss with lumped matching network and with embedded matching network.
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Table 4.3. Comparison of the microstrip antenna response without matching network, with lumped matching network and with embedded matching network.

<table>
<thead>
<tr>
<th></th>
<th>Without Matching Network (Ensemble)</th>
<th>With Lumped Matching Network (Serenade)</th>
<th>With Embedded Matching Network (Ensemble)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$F_r$ (GHz)</td>
<td>2.49</td>
<td>2.49</td>
<td>2.37</td>
</tr>
<tr>
<td>RL (dB)</td>
<td>-5.35</td>
<td>-27.93</td>
<td>-25.57</td>
</tr>
<tr>
<td>VSWR</td>
<td>3.36</td>
<td>1.08</td>
<td>1.11</td>
</tr>
</tbody>
</table>

As can be seen from Figure 4.11 and Table 4.3, very good Return Loss ($-25.57$) and VSWR (1.11) have been obtained using embedded matching network. However, there is slight shift in resonance frequency using embedded components as compare to lumped matching network. This is probably because of the fact that the embedded components were optimized individually and not for the combined circuitry, which would account for the loading effect of the vias. Changing the value of inductor and capacitor a little can easily shift the resonance frequency.

This finishes design of the microstrip antenna built on a thin substrate. The next section discusses the design of rest three kinds of microstrip antennas.

4.4 Alternative Antenna Designs

As mentioned in previous section the microstrip antenna built on extremely thin substrate is not expected to work very well and so other kinds of microstrip antennas were analytically modeled in Ensemble. Three different antennas are modeled as alternatives. These antennas are microstrip square patch antenna built on electrically thick substrate, EMC square patch, and EMC dipole. This section discusses the design and the predicted results for each of these antennas.
4.4.1 Microstrip Square Patch Antenna on an Electrically Thick Substrate

In order to increase the total substrate thickness, the substrate layer 2A of the MLO material, shown in Figure 3.1, is merged into substrate layer 2B + 2C. The microstrip antenna is placed on metal layer 1C. The layout of the antenna and cross section of the MLO is shown in Figure 4.12, whereas the actual fabricated layout is shown in Figure 4.13. As shown in figure 4.12 (b); the metal layers 1A and 1B are removed and the substrate layer 2B + 2C, having $\varepsilon_r = 3.3$, $\tan \delta = 0.02$, and thickness = 4.92 mils, is stacked to substrate layer 1A having $\varepsilon_r = 3.7$, $\tan \delta = 0.01$, and thickness = 30 mils. This increases the total substrate thickness from 4.92 mils to 34.92 mils. The metal layer G serves as a ground plane. As the total substrate thickness has been changed the thickness of the 50 $\Omega$ feed line is 8215 mils, which is different from the previous antenna configuration (11.96 mils).

![Figure 4.12](image)

Figure 4.12. Microstrip Square Patch Antenna, (a) Layout of the Microstrip Square Patch Antenna ($a = 1260$ mils, $b = 1260$ mils, $s = 300$ mils, $l = 600$ mils, $w = 82.15$ mils (corresponding to 50 $\Omega$), $g = 240$ mils (b) The cross section of the MLO material (thickness of $2C + 2B = 4.92$ mils, thickness of $2A = 30$ mils).
The CPW is used for the measurement purpose. The CPW layout that has input impedance equal to 50 Ω is shown in Figure 4.14. This CPW has a signal line having a
thickness of 12 mils. This signal line needs to be connected to the feed line of the antenna, which is 82.15 mils thick. In order to connect two lines of different width a smooth transition is used between them. This smooth transition and corresponding dimensions are shown in Figure 4.14.

The antenna layout shown in Figure 4.12, with $a = 1260$ mils, $b = 1260$ mils, $s = 300$ mils, $l = 600$ mils, $w = 82.15$ mils (corresponding to $50 \, \Omega$), $g = 240$ mils, has been simulated in Ensemble. An adaptive analysis has been performed at 2.45 GHz with “Target Maximum Delta Norm” set to 0.02. The mesh was first adapted at 2.45 GHz and then the solution has been obtained over a frequency range 2 GHz to 3 GHz. The Return Loss and corresponding VSWR as a function is shown in Figure 4.15. The antenna is resonating at 2.5 GHz. At resonance the Return Loss is $-23.44 \, \text{dB}$, which corresponds to a VSWR of 1.14.

![Figure 4.15. Frequency response for the microstrip antenna built on thick substrate (a) Return loss over a range of frequencies (b) VSWR over a range of frequencies](image-url)
Similar to what was done for the previous antenna configuration, for this antenna configuration three different antennas with different values of inset have been implemented. The change in resonant frequency, input impedance, VSWR and Return Loss with respect to change in inset (s) is shown in Table 4.4.

Table 4.4. Change in antenna parameters with respect to inset for square patch antenna built on thick substrate

| No. | a  = b (mil) | s (mil) | Thickness (mil) | $f_r$ (GHz) | $Z_{in}$ (Ω) | VSWR | Return Loss (dB) |
|-----|--------------|--------|----------------|-------------|--------------|------|----------------|}
| 1   | 1260         | 250    | 34.92          | 2.51        | 32.52 – j 7.1 | 1.58 | -12.85         |
| 2   | 1260         | 300    | 34.92          | 2.5         | 43.76 + j 0.9 | 1.14 | -23.44         |
| 3   | 1260         | 350    | 34.92          | 2.5         | 67.6 – j 6.2  | 1.37 | -15.98         |

**4.4.2 Electromagnetically Coupled (EMC) Square Patch Antenna**

In EMC antenna the feed line and antenna patch are not connected directly to each other, rather they are coupled electromagnetically. The guidelines on designing such antennas are given in [26], where as the analysis techniques are published in [24-25].

The layout of the antenna and corresponding cross section of the MLO material are shown in Figure 4.16. Figure 4.17 depicts the fabricated layout of EMC patch. As shown in figure, the feed line, shown dotted, is placed on metal layer 1A and the antenna, shown with solid line, is placed on metal layer 1C. Metal layer G served as a ground plane. The signal transmitted through the feed line gets electromagnetically coupled to the patch. The input impedance of the antenna can be changed by varying inset (s). For
the measurement the CPW of Figure 2.3 is used. The CPW has to be placed on top metal layer (1C) where as the feed line is on the metal layer 1B. Therefore, a via is used to connect the feed line of the antenna and the signal line of the CPW, as shown in Figure 4.16 (a).

Figure 4.16. Electromagnetically coupled microstrip square patch antenna (a) Layout of the Electromagnetically Coupled Microstrip Patch Antenna (a = 1260 mils, b = 1260 mils, s = 250 mils, l = 500 mils, w = 65.16 mils (corresponding to 50 Ω)) (b) Cross section of the MLO material used (thickness of 2C + 2B = 4.92 mils, thickness of 2A = 30 mils).

Figure 4.17. Actual layout of the EMC microstrip square patch antenna
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The layout of Figure 4.16, with \( a = 1260 \) mils, \( b = 1260 \) mils, \( s = 250 \) mils, \( l = 500 \) mils, \( w = 65.16 \) mils (corresponding to 50 \( \Omega \)), has been simulated in Ensemble. The Return Loss and VSWR, as a function of frequency, predicted by Ensemble are shown in Figure 4.18 (a) and (b) respectively. At resonant (2.44 GHz) the Return Loss is \(-36.42\) dB, which corresponds to VSWR of 1.1. The performance of this antenna configuration is far much better than other configurations discussed so far.

![Figure 4.18. Frequency response for electromagnetically coupled microstrip square patch antenna (a) Return Loss as a function of frequency (b) VSWR as a function of frequency](image)

Similar to other antenna configurations, for this antenna the change in antenna parameters with respect to change in inset (s) is shown in Table 4.5.

Table 4.5. Change in antenna parameters with respect to inset for EMC microstrip square patch antenna

<table>
<thead>
<tr>
<th>No.</th>
<th>( a = b ) (mil)</th>
<th>( s ) (mil)</th>
<th>( f ) (GHz)</th>
<th>( Z_{in} ) (( \Omega ))</th>
<th>VSWR</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1260</td>
<td>230</td>
<td>2.45</td>
<td>41.5 - j 8.5</td>
<td>1.3</td>
<td>-17.65</td>
</tr>
<tr>
<td>2</td>
<td>1260</td>
<td>250</td>
<td>2.44</td>
<td>50.0 + j 0</td>
<td>1.1</td>
<td>-36.42</td>
</tr>
<tr>
<td>3</td>
<td>1260</td>
<td>300</td>
<td>2.44</td>
<td>70 - j 5</td>
<td>1.4</td>
<td>-15.46</td>
</tr>
</tbody>
</table>
4.4.3 Electromagnetically Coupled (EMC) Microstrip Dipole Antenna

Electromagnetically coupled dipole antenna is another configuration that was designed. The design and analysis of such antenna are discussed in [27]. In dipole antenna instead of using a square / rectangular patch a microstrip line of finite thickness is used as a dipole antenna. Just like in EMC patch antenna the feed line is not directly connected to the patch of the antenna, rather they are electromagnetically coupled to each other.

![Diagram of Electromagnetically Coupled Microstrip Dipole Antenna](image)

Figure 4.19. Electromagnetically coupled microstrip dipole antenna (a) Layout of the Electromagnetically Coupled dipole Antenna ($w_1 = 100$ mils, $w_2 = 65.16$ mils, $l = 1380$ mils, $s = 335$ mils) (b) Cross section of the MLO material (thickness of $2C + 2B = 4.92$ mils, thickness of $2A = 30$ mils)

The layout of the dipole antenna and corresponding cross section of the MLO is shown in Figure 4.19 (a) and (b) respectively. Figure 4.20 shows the layout after fabrication. As shown in the figure, like EMC square patch, the feed line is placed on metal layer 1A and the antenna is placed on metal layer 1C. It is to be noted here that there is an offset between the antenna and feed. The feed line is not placed exactly below the antenna patch. For the examples discussed here the offset is set equal to zero. More information on offset and its effect on antenna performance can be found in [17]. For the
measurement, just like other antenna configurations, the CPW arrangement is used. The feed line, placed on metal layer 1A, is accessed through a via similar to EMC patch configuration.

![Actual layout of EMC microstrip dipole antenna](image)

Figure 4.20. Actual layout of EMC microstrip dipole antenna

The layout of Figure 4.19, with $w_1 = 100$ mils, $w_2 = 65.16$ mils, $l = 1380$ mils, $s = 335$ mils, thickness of $2C + 2B = 4.92$ mils, thickness of $2A = 30$ mils, is simulated in Ensemble. The predicted results for this configuration, obtained from Ensemble, are displayed in Figure 4.21. The Return Loss and VSWR as a function of frequency are displayed in Figure 4.21 (a) and (b) respectively. At resonance ($f_r = 2.456$ GHz) the Return Loss is equal to $-27.61$ dB, which corresponds to a VSWR of 1.08. Thus the antenna is radiating very well at resonance.

Just like other antenna configurations, three different dipole antenna design for different insets (s) are designed. The change in antenna resonance frequency, input impedance, VSWR, and return loss with respect to change in inset (s) are tabulate in Table 4.6.
Figure 4.21. Frequency response for electromagnetically coupled microstrip dipole antenna
(a) Return Loss as a function of frequency (b) VSWR as a function of frequency

Table 4.6. Change in antenna parameters with respect to inset for EMC microstrip dipole antenna

<table>
<thead>
<tr>
<th>No.</th>
<th>l</th>
<th>s (mil)</th>
<th>f₀ (GHz)</th>
<th>Z₀₉ (Ω)</th>
<th>VSWR</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1380</td>
<td>335</td>
<td>2.456</td>
<td>50.5 + j 4.145</td>
<td>1.08</td>
<td>-27.61</td>
</tr>
<tr>
<td>2</td>
<td>1380</td>
<td>355</td>
<td>2.446</td>
<td>42.65 + j 1.9</td>
<td>1.17</td>
<td>-21.76</td>
</tr>
<tr>
<td>3</td>
<td>1380</td>
<td>375</td>
<td>2.45</td>
<td>43.265 – j 3.6</td>
<td>1.17</td>
<td>-21.74</td>
</tr>
</tbody>
</table>

This finishes the design of all four kinds of microstrip antennas that have been analytically modeled using Ensemble. Even though, the performance of all alternate antennas is really well, for sensitivity analysis (i.e. to study chip-package issues) the antenna built on thin substrate is used. This is because it was built within the MLO material only. The next chapter is devoted to the study of chip-package issues in both frequency and time domain for individual components and for the system as a whole.
In the chip-package co-design of the RF front end both digital and analog circuitry share the same substrate as the microstrip patch antenna and associated integrated circuitry. The microstrip patch antenna and its associated matching network in the package can suffer a great deal of degradation from the presence of circuitry in the digital or analog domains, which share the same package. It is required to understand how exactly neighboring, active / passive, components would affect the performance of RF front end. In this chapter, instead of considering the whole RF front end only the microstrip antenna and its associated matching network are considered to study the chip-package co-design issues. To study such issues the sensitivity analysis needs to be performed on the antenna.

The analysis, both in frequency and time domain, of the field penetration through planes in multilayered packages is shown in [9]. Circuit modeling approach to study the substrate coupling is discussed in [10]. The work done for this thesis includes the study of substrate coupling using full wave simulators as well as using circuit-modeling approach.

First, this chapter discusses a study of chip-package co-design issues, using a full
wave simulator, performed on the test bed in both frequency and time domain. Then, the equivalent circuit model showing the impact on the antenna performance is presented. The first section in this chapter discusses the affect on microstrip antenna performance in frequency domain due to nearby transmission lines of different characteristic impedances, which is followed by a study of transient coupling in time domain. One of the possible techniques of reducing the coupling of noise into the system is presented in the third section of this chapter. It has been shown that by placing sections of grounded transmission lines the effect of the transient can be reduced. In the last section, the S-parameters similar to that obtained using the full wave simulator, showing the coupling into the antenna feed system due to neighboring circuitry, are obtained using an equivalent circuit model and finally the results are compared.

5.1 Sensitivity Analysis in Frequency Domain

The sensitivity analysis, a study of how sensitive the antenna is to nearby components, is performed in both frequency and time domain. This section is dedicated to the study of how near by components impact the frequency domain characteristics of the microstrip antenna. A frequency domain sensitivity analysis is performed for the system with antenna only (that is without matching network) as well as on the antenna with matching network. As, the full wave solver, Ensemble, creates smaller mesh and large number of segments in the area consisting of passives; it takes considerably long time to simulate a structure that has a microstrip antenna and matching network with it. This is the reason why more rigorous analysis was performed on the structure consisting
only the microstrip patch antenna. One frequency domain result showing the impact of nearby component on the microstrip antenna with matching network is displayed.

5.1.1 Effect of Transient Coupling on Antenna Input & Radiation Characteristics

As mentioned in previous chapter, the microstrip antenna on electrically very thin substrate is the desired antenna. This is because the antenna uses only MLO, and not the supporting material, as a substrate. It is because of this reason; the sensitivity analysis is performed over the antenna built on thin substrate.

Figure 5.1. (a) Layout of the Microstrip Square Patch Antenna, where, \( a = 1360 \) mils, \( b = 1360 \) mils, \( s = 90 \) mils, \( l_1 = 200 \) mils, \( l_2 = 150 \) mils, \( w_1 = 11.96 \) mils (corresponding to 50 \( \Omega \)), \( w_2 = 11.96 \) mils (variable, also 6.36 mils, and 24.14 mils), \( d = 4.027 \) mils (variable), \( g = 36 \) mils, thickness of \( 2C + 2B = 4.92 \) mils, thickness of \( 2A = 30 \) mils, (b) The MLO material used for the design shown in (a).

The objective is to obtain the spacing and the type of metalization that would impact the antenna performance the most. To achieve this goal, microstrip lines of different characteristic impedances and spacing from the antenna feed are placed near to
the feed line of the antenna as shown in Figure 5.1. The arrangement shown in Figure 5.1 was simulated in Ensemble.

The microstrip square patch antenna depicted in Figure 5.1(a); having dimensions: \(a = 1360 \text{ mils}, b = 1360 \text{ mils}, s = 90 \text{ mils}, l_1 = 200 \text{ mils}, w_1 = 11.96 \text{ mils}\) (corresponding to 50 \(\Omega\)); is placed on a substrate shown in Figure 5.1 (b). The microstrip line, representing nearby component, having length, \(l_2 = 150 \text{ mils}\), and width, \(w_2\), is placed the distance, \(d\), apart from the feed line of the antenna. The width (\(w_2\)) and the spacing (\(d\)) are changed in order to see the impact of these changes on antenna performance. Three different widths of microstrip lines that were selected are: 6.36 mils corresponding to 70.71 \(\Omega\) line, 11.96 mils corresponding to 50 \(\Omega\) line, and 24.14 mils corresponding to 30 \(\Omega\) line. The distance between the microstrip line and the feed line of the antenna is varied from 4 mils to 45 mils and the change in VSWR of the antenna system was plotted. Such a plot of change in VSWR, as a function of spacing between transmission line and feed line of the antenna and characteristic impedance of the microstrip lines, is shown in Figure 5.2.

Figure 5.2 shows that for every line the impact on VSWR of the microstrip antenna reduce as the distance of it from the feed line increases. It is interesting to note that the impact of different characteristic impedance lines is different on antenna performance. The transmission line having higher characteristic impedance affects more to the VSWR compare to the transmission line with lower characteristic impedance. This is most likely due to the increased fringing affects around narrower strips transmission lines. Moreover, as the transmission lines are placed away from the feed line, the VSWR
of the antenna moves closer to what it had without anything placed nearby. That confirms that the transmission line placed in close proximity affects the antenna performance. The results shown in Figure 5.2 can also help RF circuit designer in order to choose the minimum distance between antenna system and other active or passive components in order to avoid undesired coupling.

![Figure 5.2. Change in VSWR as a function of both, distance from antenna feed and transmission line characteristic impedance](image)

The neighboring components not only affect the VSWR and so the Return Loss parameter but also affect the radiation characteristics of the antenna. Figure 5.3 shows the change in antenna radiation characteristic because of the transmission lines placed 4 mils apart from the feed line of the microstrip antenna. The radiation characteristic is plotted in dB and is normalized with respect to highest value of radiation characteristic of antenna by itself. Figure shows that the radiation degrades because of the sizable coupling from the nearby transmission line.
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5.1.2 Sensitivity Analysis on the Microstrip Antenna with Matching Network

The analysis similar to that shown in previous section is also performed for the whole system that includes microstrip antenna and the matching network. The goal here is to see the impact of transmission line placed in close proximity of a matching network performance. The 50 Ω microstrip line is placed 4 mils apart from the matching network as shown in Figure 5.4. The figure shows only the matching network section for ease of visualization. The whole layout, including microstrip antenna, matching network and adjacent microstrip line, is simulated in Ensemble. The simulation took about 8 hours to run on P-III processor (866 MHz) with 256 MB RAM. The simulation took long mainly because Ensemble creates very narrow mesh in inductor and capacitor region.
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Figure 5.4. The layout simulated in Ensemble to study the affect of neighboring transmission line to the performance of the antenna system

The impact of adjacent transmission line on the antenna system was measured in terms of Return Loss. The change in Return Loss of the antenna system with and without adjacent transmission is shown in Figure 5.5. It can be seen from this figure that the Return Loss becomes very poor at the resonance frequency. The antenna becomes very reflective at the frequency of interest. This is probably because of the change in inductor and capacitor values due to the transmission line running nearby. As can be seen from the figure, the affect on the values of inductance and capacitance changes the resonance frequency considerably.

The frequency domain results shown in this section prove that there is a sizable amount of coupling going on between the antenna system and the closely placed components. The amount of coupling depends on the type of metalization (characteristic impedance) and the distance between the elements. Moreover, due to such coupling antenna parameters like VSWR, Return Loss and radiation are sizably affected. In the
next section such coupling is shown in time domain and the waveforms of the excitation signal and the coupled signal are shown.

![Figure 5.5. Comparison of Return Loss of the antenna system with and without adjacent transmission line](image)

**5.2 Sensitivity Analysis in Time Domain**

As shown in previous section, in mixed signal applications where both digital and analog circuitries are placed on the same substrate, the impact on the microstrip antenna performance due to a closely placed component is obvious. To study the affect of such adjacent circuitries on the microstrip antenna in time domain the transient analysis has been performed. In order to perform time domain simulation Ansoft’s Schematic Capture (version 5) is used. This software uses the frequency domain data (S-parameters) to generate time domain plots. It considers the frequency domain data of the structure as a black box with number of ports corresponding to the size of the S-matrix. For example, if it were required to see the reflected waves of the antenna at the antenna port in time domain, the $S_{11}$ parameter over some frequency range can be imported in Schematic
Capture. The Schematic Capture would then consider it as a one-port black box because the size of the S-matrix is 1x1. It also allows user to connect additional lumped elements (like resistors, inductor, and capacitors) and sources (voltage source and current source) to that black box. Voltage and/or current probes need to be placed at required nodes in order to measure time domain signals.

The layout of the Figure 5.1 (with, $a = 1360$ mils, $b = 1360$ mils, $s = 100$ mils, $l_1 = 200$ mils, $l_2 = 150$ mils, $w_1 = w_2 = 11.96$ mils (corresponding to 50 $\Omega$ line), $d = 4.027$ mils, $g = 36$ mils, thickness of $2C + 2B = 4.92$ mils, thickness of $2A = 30$ mils) is considered to obtained time domain results.

![Figure 5.6. Circuit diagram used in Schematic Capture ver. 5, where, $R = 50$ $\Omega$ and $V_{\text{in}} = \text{sinusoidal voltage source operating at 2.46 GHz}$](image)

The S-matrix obtained from the Ensemble for the frequency range of 1 to 10 GHz (mesh adapted at 5 GHz) was exported to Schematic Capture. As shown in the Figure 5.6 the S-matrix, obtained from Ensemble, is represented by a black box with number of port
equal to three (As the S-matrix obtained from Ensemble was 3x3). The port 1 of the black box represents the antenna feed, whereas port 2 and port 3 represent the ports of adjacent running transmission line. Port 2 is connected to a voltage source having impedance of 50 Ω to represent the signal flowing through the adjacent running transmission line. Where as, other two ports (port 1 and port 3) are connected to a matched load (that is, 50 Ω). Three different voltage probes were connected to the ports in order to measure time domain signal at that node.

To represent digital or analog noise on the package, a transient as shown in Figure 5.7, is applied at port 2. The transient is in the form of a sine wave, 1 Volt peak with a frequency of 2.49 GHz truncated after 5 cycles. The signal is applied at 2.49 GHz because the antenna is resonating at that frequency. Moreover, it is terminated after five cycles in order to see the affect of abrupt termination on coupling. With this setup the, SPICE simulation was performed. The time domain signals at port 3 and port 1 have been observed and are shown in Figure 5.8 (a) and (b) respectively.

The source impedance is considered to be equal to 50 Ω and so the actual amplitude of the voltage signal, applied at port 2, can be calculated using a voltage divider rule. As the impedance of both the source and the port 2 are set equal to 50 ohm, the amplitude of the signal at port 2 is 0.5 volts peak. The signal coupled at port 3 because of the signal applied at port 2 is shown in Figure 5.8 (a). As port 2 and port 3 are connected directly through a microstrip transmission line, most of the energy from port 2 gets coupled to port 3. This is the reason; the amplitude of the signal getting coupled to port 3 is close to 0.5 volts.
Chapter 5: Chip-Package Co-design Issues Related RF Front End

Figure 5.7. Transient signal applied at port 2

Figure 5.8. Coupled signals because of the transient applied (a) Signal getting coupled at port 3 (b) Signal getting coupled at port 1
Figure 5.8 (b) shows the signal observed at port 1. This signal is the signal getting coupled in to the antenna feed line due to transient applied at nearby transmission line. Sizeable coupling of about 6 % percent of the excitation can be seen in the presence of the transient. The amplitude of the signal is decaying, probably, because the signal is still getting settled down. Moreover, even after the transient is terminated after five cycles, remnant amount of noise is present at the feed of the antenna, which is probably because of the resonance nature of the antenna at that frequency. The coupled signal dies down slowly because of the loss associated with the dielectric.

The magnitude of the coupled signal depends on the characteristic impedance of the adjacent transmission line and the distance of it from the feed line. The shown results are for a 50 Ω line placed 4 mils apart from the feed. The parameter that controls the amount of coupling is $S_{12}$. As the distance between the feed line and the adjacent line increases the coupling parameter ($S_{12}$) reduces as shown in Figure 5.9. With increase in
the distance between transmission line and the feed line of the antenna the amplitude of
the coupled signal will decrease.

Thus, it has been shown in this section that sizable amount of noise gets coupled in to the RF front end because of the signal traveling in neighboring component. The effect of which continues even after the signal in the neighboring component is terminated. This effect dies with time because of the lossy nature of the dielectric material. The next section discusses a technique with the help of which such a transient coupling can be reduced.

5.3 Decoupling Circuitry

It is clear from the results shown in this chapter so far that noise coupled from the adjacent components into the antenna system affects its performance. In order to reduce such transient coupling of noise to the antenna port, a grounded microstrip line (Decoupling Transmission Line) is placed next to the radiating edge of the antenna at some distance apart as shown in Figure 5.10. The placement of such grounded microstrip line is a random guess, which worked out very well. In the figure, the antenna layout is same as that discussed so far in this chapter, and the adjacent microstrip line was placed 4 mils apart from the feed line of the antenna.

The grounded microstrip line (on the right) having a length of 500 mils is placed next to the radiating edge of the antenna at a distance of 4 mils. The grounded microstrip line placed near to the radiating edge of the antenna might affect the polarization of the antenna. In order to avoid the impact on antenna polarization another grounded
microstrip line is placed on the left side of the antenna. The arrangement shown in Figure 5.10 has been simulated in Ensemble and S-parameters are obtained. These S-parameters were, then, exported to Schematic Capture in order to obtain time domain results. A circuit diagram similar to the one shown in Figure 5.6, with black box containing S-parameters obtained from the layout of Figure 5.10, is drawn. In order to compare the results, the same transient sine wave having 1 Volt peak at a frequency of 2.49 GHz and truncated after 5 cycles, has been applied at port 2 of the layout shown in Figure 5.10. The coupled signal at port 1 can then be observed.

Figure 5.10. The layout simulated in Ensemble in order to reduce coupling of transient noise into the antenna system

Figure 5.11 compares the time domain signal coupled at antenna feed (port-1) with and without grounded line. The dashed line is the signal showing coupling without the decoupling line, whereas the solid line shows the signal getting coupled in the presence of grounded microstrip line. There is negligible amount of coupling to the
antenna port in the presence of the grounded decoupling line. The reason, probably, is the fact that the grounded microstrip line attracts the waves traveling in the dielectric.

![Graph showing voltage signal comparison](image)

**Figure 5.11.** Comparison of voltage signal getting coupled with and without decoupling line

It is observed that the decoupling circuitry improves the radiation characteristic significantly. It has been shown in section 5.1.1, that the coupling from adjacent component degrades the radiation characteristics. The grounded microstrip line seems to improve the radiation characteristic. Figure 5.12 compares the radiation characteristic of antenna by itself, antenna with adjacent microstrip line and antenna with grounded microstrip line. The graph is plotted in dB and is normalized with respect to the highest value of radiation characteristic of antenna by itself. The solid line is the radiation of the antenna by itself. The dashed line shows the degradation in the radiation characteristic due to noise coupling, whereas, the dotted line shows the significant improvement in radiation characteristic compare to other two cases. Thus, the grounded microstrip line
placed next to the radiating edge of the antenna not only reduces the transient coupling but also improves the radiation characteristic of the antenna.

![Graph showing radiation characteristic comparison](image)

**Figure 5.12.** Comparison of radiation characteristic of antenna by itself, antenna with adjacent microstrip line and antenna with grounded microstrip line

So far in this chapter the impact on antenna system has been studied using a full wave solver. The change in antenna input and radiation characteristics are observed. In the next section the impact on the antenna feed system due to a transmission line placed near by is studied using an equivalent circuit model approach. The results obtained from the circuit model are then compared with that obtained from the full wave simulator.
5.4 Equivalent Circuit Model Approach to Study the Impact on the Integrated Antenna

In this section an example that is being discussed so far in this chapter is considered. The layout similar to that shown in Figure 5.1 is considered. The layout is redrawn in Figure 5.13 in such a way that helps developing the equivalent circuit model for the coupling region. As shown in the figure a 50 Ω line is placed in close proximity of a feed line of the microstrip patch antenna. The goal, here, is to see the near end and far end coupling from the closely placed microstrip line into antenna system with the help of equivalent circuit model. The results obtained from Ensemble are considered as a reference for this purpose.

Figure 5.13. Microstrip antenna connected to a coupled 50 Ω lines. a = 1360 mils, b = 1360 mils, g = 36 mils, s = 100 mils, L₁ = 200 mils, L₂ = 150 mils, L₃ = 47.92 mils
The antenna used here is of size: \(a = 1360 \text{ mils}, \ b = 1360 \text{ mils}, \ g = 36 \text{ mils}, \ s = 100 \text{ mils}\). The feed line is 200 mils long \((L_1)\). A microstrip line of length 150 mils \((L_2)\) is placed 4 mils away from the antenna setup. The region of interest is the ‘coupled region’ shown in the figure. The whole setup shown in Figure 5.1 was laid out on the MLO package shown in Figure 5.1 (b).

The equivalent circuit model for the ‘coupled region’ was developed. The coupled region with its segments is drawn separately in Figure 5.2. The coupled region is divided into 4 segments. The length of each segment is kept equal to the width of each transmission lines, that is 11.58 mils. This way, in a segment each metal plate has a square layout. The goal was to obtain the return loss, near end coupling \((S_{12})\), and far end coupling \((S_{13})\) from an equivalent circuit model and compare it with that obtained from Ensemble. The whole layout shown in Figure 5.13 has been simulated in Ensemble and S-parameters are obtained.

![Figure 5.14. The coupled region of interest and its segments. \(l = 47.92 \text{ mils}, \ s = 4.027 \text{ mils}, \ w = 11.58 \text{ mils}\) (corresponding to 50 \(\Omega\)).](image)

For a parallel running line layout considered here the values of circuit elements \((C_a, C_b, L, R, G, \text{ and } k)\) can be obtained using equations (2.1) to (2.13). By using
equations (2.1) to (2.4), with line width, \( w = 11.58 \) mils; substrate height, \( h = 4.92 \) mils; relative permittivity of the substrate material, \( \varepsilon_r = 3.3 \); and separation, \( s = 4.027 \) mils, the values of \( C_a \) and \( C_b \) come out to be 0.271 fF and 3.663 fF respectively. For \( \varepsilon_r = 3.3 \), the value of effective relative permittivity can be found from equation (2.6). The value of \( \varepsilon_{re} \) is equal to 2.6157. With this value of effective relative permittivity, the value of self-inductance can be found from equation (2.5). As the equation (2.5) gives the value of inductance per unit length of microstrip line, the value of self-inductance (L) for the length of segment equal to 11.58 mils becomes 0.0792 nH. The loss associated with the metal can be calculated from equations (2.7) to (2.9). Equation (2.8) calculates the AC resistance of the metal at certain frequency. The value of \( R_{ac} \) at \( f = 2.45 \) GHz, and conductivity \( \sigma_c = 5.7 \times 10^7 \) S/m is equal to 0.013 \( \Omega \). The value of DC resistance, obtained from equation (2.9), is equal to 0.9867 \( \Omega \). Here, metal thickness, \( t = 0.7 \) mils; and conductivity of the metal (copper) \( \sigma_c = 5.7 \times 10^7 \) S/m were considered. Thus the total loss associated with metal (R), from equation (2.7) becomes 0.014 \( \Omega \). The value of loss associated with dielectric is calculated using equation (2.11). For frequency (f) equal to 2.45 GHz, and \( \tan \delta = 0.02 \) the value of G is equal to 11.98 M\( \Omega \). Finally, the value of coefficient of mutual coupling can be calculate from equation (2.13) and Table 2.1. For 50 \( \Omega \) lines the table reads the value of A and B equal to 0.25 and 0.3149 respectively. Using these values of A and B, k can be obtained from equation (2.13). The value of k is equal to 0.1511. All these values of circuit elements are tabulated in Table 5.1.

The circuit model for the whole coupled region is shown in Figure 5.15. In the figure, the circuit model for each segment is cascaded four times. This is because the
coupled region was divided into four segments. The antenna in the Figure 5.13 can be represented by its S-parameter in the equivalent circuit model. The black box, in the Figure 5.15, represents the S-parameter (S_{11} to be specific) of the microstrip patch antenna. These S-parameters are obtained from Ensemble by simulating the ‘antenna region’ separately.

Table 5.1. Circuit element values for the equivalent circuit model of the coupled 50 Ω line layout

<table>
<thead>
<tr>
<th>No.</th>
<th>Circuit Elements</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>C_a</td>
<td>0.271 fF</td>
</tr>
<tr>
<td>2.</td>
<td>C_b</td>
<td>3.663 fF</td>
</tr>
<tr>
<td>3.</td>
<td>L</td>
<td>0.0791 nH</td>
</tr>
<tr>
<td>4.</td>
<td>R</td>
<td>0.014 Ω</td>
</tr>
<tr>
<td>5.</td>
<td>G</td>
<td>11.98 MΩ</td>
</tr>
<tr>
<td>6.</td>
<td>k</td>
<td>0.1511</td>
</tr>
</tbody>
</table>

Figure 5.15. Equivalent circuit model to study the impact on antenna input characteristics due to adjacent transmission line. Where, C_a = 0.271 fF, C_b = 3.663 fF, L = 0.0791 nH, R = 0.014 Ω, G = 11.98 MΩ, and k = 0.1511.

The circuit model shown in Figure 5.15 with the circuit parameters tabulated in Table 5.1 has been simulated in Serenade. The model is simulated for the frequency range from 1 to 5 GHz. The S-parameters, S_{11}, S_{12}, and S_{13}, obtained from circuit
simulator are compared with that obtained from Ensemble, a full wave solver. The comparison of the results is shown in Figure 5.16.

![Figure 5.16](image-url)

**Figure 5.16.** Comparison of the S-parameters for the 50 Ω line placed in close proximity of an antenna feed (a) $S_{11}$ (b) $S_{12}$ (c) $S_{13}$ and (d) $S_{14}$

In the figures, the solid lines correspond to the results obtained from the circuit simulator and the dashed lines correspond to results obtained from Ensemble. All the graphs are plotted on the same scale for easy comparison. The Figure 5.16 (a) shows the
comparison of return loss. The return loss obtained from equivalent circuit model and that obtained from Ensemble are so close to each other that they are hard to identify of the graph. Figure 5.16 (b) and (c) display near end coupling and far end coupling into antenna feed because of the microstrip transmission line placed in close proximity. The $S_{12}$ parameter obtained from the circuit model matches very well with that obtained from Ensemble up to 4 GHz. Where as $S_{13}$ parameter obtained from the equivalent circuit model shows good agreement with Ensemble-result up to about 3 GHz.

The equivalent circuit model has also been verified for spacing other than the one discussed in this example ($s = 4.027$ mils). The circuit model has been verified for spacing from 2 mils to 19 mils between the transmission line and the feed line of the microstrip antenna. In conclusion, it can be said that the equivalent circuit model predicts return loss, near end coupling, and far end coupling considerably well within the frequency range of interest. Thus the equivalent circuit model has been successfully applied to study the chip-package co-design issues related to the microstrip antenna.

Thus, in this chapter the change in antenna parameters like radiation, VSWR and Return Loss because of the adjacent components are discussed. By applying a transient at nearby transmission line, the time domain signal coupled into antenna feed is displayed. A technique to reduce the amount of coupling has also been shown. The decoupling line reduces the amount of transient coupling in the antenna feed line and improves the radiation characteristics. Finally, with the help of equivalent circuit model the impact on antenna input characteristic is studied. The next section summarizes the work done in this thesis and suggests the future work.
CHAPTER

SIX

CONCLUSION

This thesis has investigated the issues associated with chip-package co-design methodology. This chapter summarizes the major contributions of this work and identifies areas that merit future study.

6.1 Contributions of Present Work

- To study the chip-package co-design issues at the package level a novel segmentation technique was introduced. An equivalent circuit model for each such segment was also presented. It was shown that the circuit model for a finite length coupled lines can be developed by just cascading the equivalent circuit model for a single segment.

- As opposed to commonly used procedure of optimizing the equivalent circuit model using circuit simulator, closed from equation for every circuit element was found. An empirical equation to calculate the value of coefficient of mutual coupling, $k$, was developed.

- To study the coupling at chip-package interconnect level a simpler equivalent circuit models of a single via structure and coupled via structure were presented. The
segmentation technique was successfully applied to study the coupling between two coupled vias; representing chip-package interconnects.

- The validity of the circuit models for both package level and chip-package interconnect level was verified by considering various package level coupling / discontinuities and chip-package interconnect examples. The results obtained by simulating equivalent circuit model into circuit simulator were compared with that obtained from a full wave simulator. Very good agreements between these results were obtained.

- In order to study the chip-package co-design issues related to RF front end, a test bed was designed using a typical MLO material. The test bed consisted of the microstrip patch antenna built on extremely thin substrate. It was shown that the return loss of the antenna could be improved by using a L-section matching network built with integrated components. EMC patch and dipole antennas were also designed as alternative designs.

- The impact on the antenna input and radiation characteristics was studied by placing a transmission line adjacent to the antenna feed. The impact on the antenna input characteristics was shown as a function of both characteristic impedance of the nearly placed transmission line and spacing.

- The coupling into the antenna feed was studied in time domain by injecting transient into the neighboring circuitry and the signal coupled to the antenna feed was presented.
Chapter 6: Conclusion

- It was also shown that by strategically placing the grounded metal strip the coupling into antenna feed could be dramatically reduced.
- The impact on antenna input characteristics due to a transmission line placed in close proximity of the antenna feed line was studied using the equivalent circuit model.

6.2 Future Work

This section identifies topics covered in this thesis which merit more detailed study.

Chapter 2 develops the segmentation technique and the equivalent circuit model to study the package level coupling. It also presents the equations to calculate the value of each element. Equations presented in this work are only applied to typical MLO material and not to any other kind of substrate. More work is needed to develop more general-purpose equations that can be applied to different kind of dielectric material with different thickness.

In chapter 2, the equivalent circuit model for single and coupled vias was developed. In this work the equations for some of the circuit elements are presented. It is required to find equations for remaining circuit elements to complete the model.

The circuit model has been applied to only one kind of chip-package interconnects. There are couple of other commonly used chip-package interconnects. These are solder bump and wire bond. The circuit model for the via structure can be easily extended to study the coupling between solder bumps and wire bonds. However,
more work is needed to obtain a simple and accurate expression to calculate values of equivalent circuit elements.

6.3 Summary

This thesis has contributed to a better understanding of chip-package co-design issues with the help of equivalent circuit modeling technique. This work has provided simple and accurate expressions to calculate circuit element values. This work has also shown a possible way to study the coupling between chip-package interconnects.
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APPENDIX - A

CALCULATION OF APPROXIMATE LAYOUT FOR PLANAR INDUCTOR

So far the research has been done to calculate the value of inductance for a given layout of the planar inductor. Researches are still going on to go other way round, that is to calculate the layout of the planar inductor for a given inductance value and substrate parameters.

Figure a.1. A typical spiral inductor layout with three turns
For this work, it was required to find out the layout of the planar inductor that gives a certain value of inductance. As there are no such equations available for this purpose, available equations for calculating inductance value were used to find approximate layout. For this purpose the expression based current sheet approximation [1] was used. For the layout shown in figure (a.1), the value of inductance can be found using equation (a.1). In figure (a.1), \( d_{out} \) is outer diameter, \( d_{in} \) is inner diameter, \( s \) is spacing between the lines, and \( w \) is the width of the line.

\[
L = \frac{\mu n^2 d_{avg} c_1}{2} \left[ \ln \frac{c_2}{\rho} + c_3 \rho + c_4 \rho^2 \right]
\]  

(a.1)

Where,

\( \mu = \) permeability constant

\( n = \) number of turns

\[
d_{avg} = \frac{d_{out} + d_{in}}{2}
\]

\[
\rho = \frac{d_{out} - d_{in}}{d_{out} + d_{in}}
\]

\( c_n = \) constants presented in table a.I

<table>
<thead>
<tr>
<th>Layout</th>
<th>( C_1 )</th>
<th>( C_2 )</th>
<th>( C_3 )</th>
<th>( C_4 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Square</td>
<td>1.27</td>
<td>2.07</td>
<td>0.18</td>
<td>0.13</td>
</tr>
<tr>
<td>Hexagonal</td>
<td>1.09</td>
<td>2.23</td>
<td>0.00</td>
<td>0.17</td>
</tr>
<tr>
<td>Octagonal</td>
<td>1.07</td>
<td>2.29</td>
<td>0.00</td>
<td>0.19</td>
</tr>
<tr>
<td>Circle</td>
<td>1.00</td>
<td>2.46</td>
<td>0.00</td>
<td>0.20</td>
</tr>
</tbody>
</table>

Table a.I. Coefficients for current sheet expression
A matlab code is developed that considers inner diameter \( (d_{in}) \), number of turns \( (n) \) and spacing \( (s) \) as a constant and iterates the value of outer diameter \( d_{out} \) and calculates the right hand side of equation \( (a.1) \). The iteration continues until the right hand side of the equation gives the required value of the inductance. The layout obtained this way is not very accurate because the expression itself is not that accurate. The layout obtained from such code can be used as a initial layout for further optimization. The matlab code that iteratively calculates the layout is presented here.
% This matlab code generates approximate layout of the embedded inductor
% for a given value of inductance

% This code uses Current Sheet Approximation method to calculate approximate
% layout of the planar inductor

% The current sheet approximate method is to calculate the value of inductance
% for a given layout dimensions. In order to obtain the layout of the planar
% inductor for a given inductance value, this code considers inner diameter of the
% coil, number of turns, and spacing between the lines as constant and iterates
% the value of outer dimeter untill the required value of inductance is reached

% Knowing the value of outer dimeter (dout), inner diameter (din), spacing (s),
% and number of turns (n), the value of width of the line (w) was calculated

% This code does not calculate accurate layout of the planar inductor. This is
% because the equation used for this code is not accurate. The user needs to
% optimize the layout in a full wave solver in order to have layout that gives
% correct value of inductor. The values obtained from this code can be a good
% initial values for the optimization.

clear;
clc;

% Input parameters
L = input('Input required inductace value in nH : '); % Converting the unit of L to Henry
L = L*1e-9;

din = input('Input inner diameter of the planer inductor (micro meters) : ');
n = input('Input the number of turns : ');
s = input('Input the spacing between lines (micro meters) : ');
Appendix - A

Converting the unit of input parameters into meters

din = din * 1e-6;
n = 2;
s = s * 1e-6;

% Defining permiability constant
u = pi*4e-7;

% Defining constants associated with a circular geometry

c1 = 1;
c2 = 2.46;
c3 = 0;
c4 = 0.2;

% Iterating the value of outer diameter (dout) to obtain desired value of inductance

% Specifying initial and final value of outer dimeter (dout)
doutstart = din + 1e-6;
doutstop = 1000*din;

% Defining the flag for stopping criteria
flag = 1;

% main loop
for d = doutstart : 1e-6 : doutstop,
    if (flag == 1)
        davg = (d + din)/2;
        fill = (d - din)/(d + din);
        L1 = 0.5 * (u * n^2 * davg * c1)*( log(c2/fill) + c3 * fill + c4 * fill^2 );
    end
end
if (abs(L - L1) < 0.01e-9)
    flag = 0;
    dout = d;
    Lmeas = L1;
else continue;
end

end
end

% Calculating the width of the embedded inductor

w = (dout - din - 2*(n-1)*s)/(2*n);

% Displaying the results

% The value of outer diameter and width of the lines are displayed in micrometers

sprintf('Lmeas = %2.12f nH', Lmeas/1e-9)
sprintf('dout = %f micrometers', dout/1e-6)
sprintf('w = %f micrometers', w/1e-6)
APPENDIX - B

CALCULATION OF CIRCUIT ELEMENT VALUES

This section presents the matlab code that has been written to calculate the values of circuit elements of the equivalent circuit model.

% This matlab code calculates values of equivalent circuit elements for a given parallel running transmission line geometry

% This code takes width of the transmission line (w), conductivity of the metal plane (sigma), thickness of the metal plane (t),
% spacing between two transmission lines (s), height of the substrate (h),
% relative permittivity of the substrate (eR), loss tangent of the substrate material
% (tan_delta), characteristic impedance of the transmission line (Z0),
% and frequency of interest (f) as input parameters and calculates
% self capacitance (Ca), mutual capacitance (Cb), self inductance (L),
% loss associated with metal (R), Dielectric loss (G), and
% coefficient of mutual coupling

clear;
clc;

% Input parameters

w = input('Input the width of the transmission line (mils) : ');
sigma = input('Input the conductivity of the metal plane (S/m) : ');
t = input('Input the thickness of the metal plane (mils) : ');
s = input('Input the spacing between the two transmission lines (mils) : ');
h = input('Input the height of the substrate (mils) : ');
eR = input('Input relative permittivity of the substrate material : ');
tan_delta = input('Input the loss tangent of the substrate material : ');
Z0 = input('Input the characteristic impedance of the transmission line (Ohm) : ');
f = input('Input the frequency of interest (Hz) : ');

% Converting the unit of input parameters into meters
\[
w = w * 25.4e-6;
t = t * 25.4e-6;
s = s * 25.4e-6;
h = h * 25.4e-6;
\]

% Defining permiability constant
\[
m = \pi * 4e-7;
\]

% Calculating the value of Self Capacitance (Ca) and Mutual Capacitance (Cb)
% Defining the constant required to calculate Ca and Cb
\[
m0 = (w/h) * (0.619*\log_{10}(w/h) - 0.3853);
k0 = 4.26 - 1.453*\log_{10}(w/h);
\]

if (0.1 \leq (s/w) \leq 0.3)
   me = 0.8675;
   ke = 2.043 * (w/h)^0.12;
else
   me = (1.565 / (w/h)^0.16) \cdot 1;
   ke = 1.97 - 0.03/(w/h);
end

% Calculating Ceven and Codd using equation (3) and (4)
\[
Codd_96 = w * (s/w)^m0 * \exp(k0);
Ceven_96 = w * (s/w)^me * \exp(ke);
\]

\[
Codd_eR = Codd_96 * (eR/9.6)^{0.8};
Ceven_eR = Ceven_96 * (eR/9.6)^{0.9};
\]

% Calculating Ca and Cb using equation (1) and (2)
\[
Ca = Ceven_eR / 2;
Cb = (Codd_eR - Ca)/2;
\]

% Converting units into Farad
\[
Ca = Ca * 1e-12;
Cb = Cb * 1e-12;
\]
eRe = (eR + 1)/2 + (eR - 1)/2 * ( 1/ sqrt(1 + 12*h/w));

% Calculating value of self inductance using equation (5)
L = ( Z0 * sqrt(eRe)/ 3e8 );

% As the equation (6) gives the value of self inductance (L) per unit length, 
% the value L should be multiplied with the appropriate length of the transmission line 
% In all the cases the length of the subsegment is considered to be equal to 
% the width of the transmission line. And so the value of L needs to be multiplied with 
% width w

L = L * w;

% Calculating the value of loss associated metal (R)
% Calculating DC resistance (Rdc) using equation (9)
if (t == 0)
    Rdc = 0;
else
    Rdc = 1/(sigma*t);
end

% Calculating AS resistance (Rac) using equation (8)
Rac = sqrt( (pi * f * m)/(sigma ) );

% Calculating total resistance (R) associated with metal using equation (7)
R = Rac + Rdc;

% Calculating loss associated with dielectric material (G) using equation (11)
if (tan_delta == 0)
    sprintf('G can not be defined (Choose some high value of G)');
else
    G = 1 / (2*pi*f * Ca * tan_delta);
end

% Calculating the value of coefficient of mutual coupling (k) using equation (13)
% So far the values of A and B for parallel running 50 Ohm, 70 Ohm and 100 Ohm lines
% have been found out
% Converting unit of the spacing (s) into mils
s = s / 25.4e-6;

% Assigning the values to constants A and B using table I

% So far the values of A and B for parallel running 50 Ohm, 70 Ohm and 100 Ohm lines
% have been found out and so this code does not calculate the value of k for any
% other values of characteristic impedance of the transmission line

if (Z0 == 50)
    A = 0.25;
    B = 8.00;
elseif (Z0 == 70)
    A = 0.65;
    B = 5.00;
else
    A = 3.24;
    B = 1.48;
end

% Calculating the value of coefficient of mutual inductance using equation (13)

k = A * exp(-s/B);

% Displaying results

% Here the value of self inductance (L) is displayed in nH, value of loss associated
% with metal (R) in Ohm, value of loss associated with substrate in MOhm, and
% value of self and mutual capacitance (Ca and Cb) are displayed in fF

sprintf('L = %fnH R = %f OHM G = %f OHM \nk = %f \nk Ca = %f fF\nk Cb = %f fF,L*1e9,R*1e-6,k,Ca*1e15,Cb*1e15)
PUBLICATIONS

