ANALYSIS OF MICROSTRIP LINE WITH COPLANAR GROUND PLANE

By

TANYA KANWAR 031116

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DEPARTMENT OF ELECTRONICS AND COMMUNICATION
JAYPEE UNIVERSITY OF INFORMATION TECHNOLOGY-WAKNAGHAT

CERTIFICATE

This is to certify that the work entitled, **"ANALYSIS OF MICROSTRIP LINE WITH COPLANAR GROUND PLANE"** submitted by Tanya kanwar in partial fulfillment for the award of degree of Bachelor of Technology Electronics and Communication Engineering Department of Jaypee University of Information Technology has been carried out under my supervision. This work has not been submitted partially or wholly to any other University or Institute for the award of this or any other degree or diploma.

TChakraborty

Dr. Tapas Chakravarthy
PROFFESOR
Electronics & Communication Dept.
JUIT

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CONTENTS

1. INTRODUCTION

1.1 TRANSMISSION LINE 1
1.1.1 Principle 1
1.1.2 Losses ...2
1.2 MICROSTRIP LINE 2
1.2.1 Design Equation 4
1.3 VARIATIONAL ANALYSIS 7
1.3.1 Technique ...7
1.4 CROSSTALK COUPLING 9
1.4.1 RF-IC model and coupling mode ...9

2. CALCULATIONS

2.1 VARAITIONAL ANALYSIS ...10
2.1.1 Program for variational analysis ...11
2.1.2 Code ...12

3. RESULTS

3.1 RESULTS OBTAINED FROM VARAITIONAL ANALYSIS ...15
3.1.1 Graph "Impedance Vs Distance" ...16
3.2 CROSSTALK SIMULATION ...17
3.2.1 S-parameters for 50 ohms ground proximity ...18
3.3 COMPARSION OF ANALYTICAL RESULTS WITH SIMUALTED RESULTS ...19

4. CONCLUSION ...20

5. BIBLIOGRAPHY ...21

ABSTRACT

High-speed multi-chip module interconnects are characterized by planar transmission lines. In this project, we report the analytical expression for line capacitance and characteristic impedance of microstrip transmission line in presence of ground plane aperture using variational analysis. The line capacitance and characteristic impedance in the line are obtained for a range of structure parameters and the dielectric constant. The results obtained from this expressions developed in this work are called variational analysis.

In high speed PCB design, parallel microstrip transmission lines are comman,in which serious crosstalk problem usually arises when dealt with in correctly. so here in order to remove this problem ground plane aperture is introduced in the multi-chip module.

So here the results obtained from variational analysis are compared with those of crosstalk simulation.

List of Figures

1. **Fig 1.1** Transmission line
2. **Fig1.2** A microstrip line showing length and width of strip line
3. **Fig 1.3** Relationship between characteristic impedance and Physical dimensions of microstrip.
4. **Fig 1.4** Geometry of a multilayer microstrip line
5. **Fig 1.5** **(a)** Microstrip and **(b)** Edge coupled microstrip
6. **Fig 1.6** **(a)**-RF-IC model **(b)** coupled pair of microstrip line on Ground substrate
7. **Fig 2.1** Crossection of microstrip line.
8. **Fig 3.1** Values of Impedance for given distances.
9. **Fig 3.2** Relationship between impedance and distance.
10. **Fig 3.3** Stimulated Results for crosstalk simulation.
11. **Fig 3.4** Relationship between Normalized amplitude and frequency.
12. **Fig 3.5** Comparison between analytical and simulated results.

The Introduction

1.1 TRANSMISSION LINE

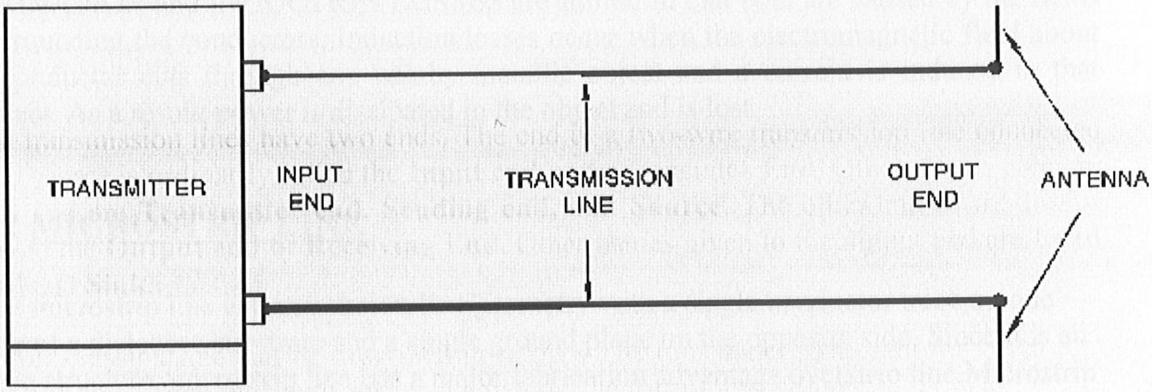
A **transmission line** is a device designed to guide electrical energy from one point to another. It is used, for example, to transfer the output RF energy of a transmitter to an antenna. This energy will not travel through normal electrical wire without great losses. Although the antenna can be connected directly to the transmitter, the antenna is usually located some distance away from the transmitter.

The transmission line has a single purpose for both the transmitter and the antenna. This purpose is to transfer the energy output of the transmitter to the antenna with the least possible power loss.

1.1.1 Principle

All transmission lines have two ends. The end of a two-wire transmission line connected to a source is ordinarily called the **Input end** or the Generator **End**. Other names given to this end are **Transmitter end, Sending end, and Source**. The other end of the line is called the **Output end** or **Receiving End**. Other names given to the output end are **Load end** and **Sink**.

A transmission line can be described in terms of its impedance. The ratio of voltage to current (E_{in}/I_{in}) at the input end is known as the **Input impedance (Z_{in})**. This is the impedance presented to the transmitter by the transmission line and its load, the antenna. The ratio of voltage to current at the output (E_{out}/I_{out}) end is known as the **Output impedance (Z_{out})**.



.Fig 1.1 Transmission line

This is the impedance presented to the load by the transmission line and its source. If an infinitely long transmission line could be used, the ratio of voltage to current at any point on that transmission line would be some particular value of impedance. This impedance is known as the **Characteristic Impendence**.

1.1.2 Losses

"**Line losses**" occur in al lines.Line losses may be any of three types - COPPER, DIELECTRIC, and RADIATION or INDUCTION LOSSES.

Copper Losses

One type of COPPER LOSSES is I^2R loss. In RF lines the resistance of the conductors is never equal to zero. Whenever current flows through one of these conductors, some energy is dissipated in the form of heat. This heat loss is a **Power Loss**. With copper braid, which has a resistance higher than solid tubing, this power loss is higher.

Copper losses can be minimized and conductivity increased in an RF line by plating the line with silver.

Dielectric Losses

DIELECTRIC LOSSES result from the heating effect on the dielectric material between the conductors. Power from the source is used in heating the dielectric. The heat produced is dissipated into the surrounding medium. When there is no potential difference between two conductors, the atoms in the dielectric material between them are normal and the orbits of the electrons are circular. When there is a potential difference between two conductors, the orbits of the electrons change. The excessive negative charge on one conductor repels electrons on the dielectric toward the positive conductor and thus distorts the orbits of the electrons. A change in the path of electrons requires more energy, introducing a power loss.

Radiation and Induction Losses

RADIATION and INDUCTION LOSSES are similar in that both are caused by the fields surrounding the conductors. Induction losses occur when the electromagnetic field about a conductor cuts through any nearby metallic object and a current is induced in that object. As a result, power is dissipated in the object and is lost.

1.2 MICROSTRIP LINE

The microstrip line is transmission-line geometry with a single conductor trace on one side of a dielectric substrate and a single ground plane on the opposite side. Since it is an open structure, microstrip line has a major fabrication advantage overstrip line.Microstrip circuit dimensions decrease with increasing substrate dielectric constant. It also features ease of interconnections and adjustments.

In a microstrip line, the wavelength Λ , is given by

$$\Lambda = \lambda / (\varepsilon_{eff})^{0.5}$$

Where: ε_{eff} = the effective dielectric constant, which depends on the dielectric constant of the substrate material and the physical dimensions of the microstrip line, and λ= the free-space wavelength.

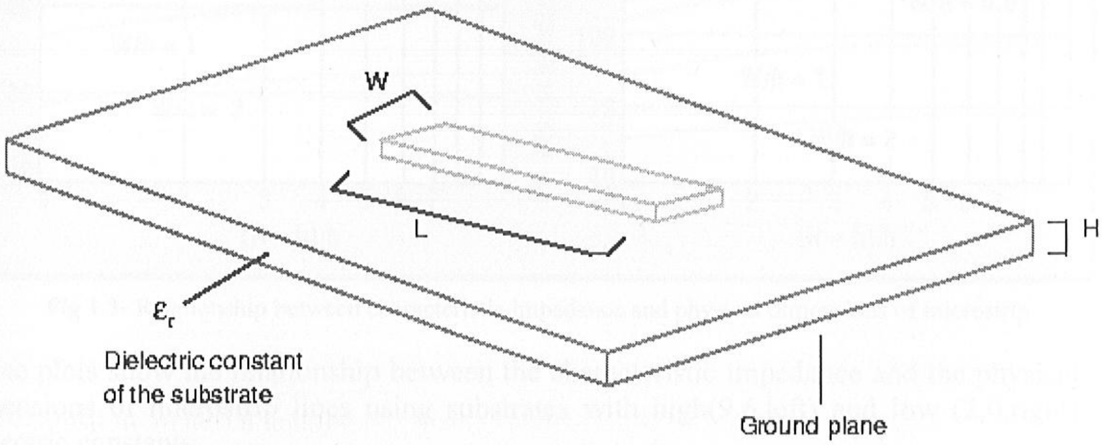


Fig1.2- A microstrip line showing length and width of strip line

L = The length of the microstrip line
W = The width of microstrip line
H = PC board substrate thickness (height of the microstrip line above the ground plane)

In a microstrip line, the electromagnetic (EM) fields exist partly in the air above the dielectric substrate and partly within the substrate itself. The top and side covers essentially redistribute the field of the more theoretical microstrip and understandably have an influence on the effective dielectric constant.
The next figure illustrates the relationships between characteristic impedance and the physical dimensions of shielded microstrip lines for two examples: substrates with low (2) and high (9.6) relative dielectric constants. The top cover tends to reduce the impedance. When the ratio of the distance from the top cover to the dielectric substrate and the substrate thickness [(H – h)/h] is greater than 10, the enclosure effects can be considered negligible.
The characteristic impedance range of a microstrip line is 20 to 120 V. The upper limit is set by production tolerances while the lower limit is set by the appearance of higher-order modes.

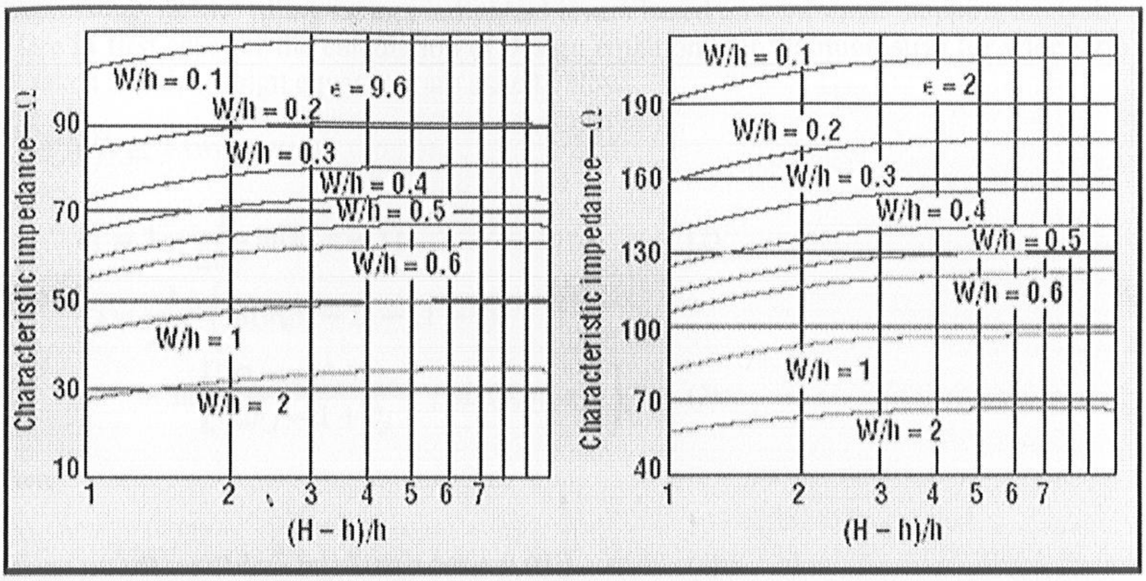


Fig 1.3- Relationship between characteristic impedance and physical dimensions of microstrip

These plots show the relationship between the characteristic impedance and the physical dimensions of microstrip lines using substrates with high(9.6,left) and low (2.0,right) dielectric constants.

There are three types of losses that occur in microstrip lines: conductor (or ohmic) losses, dielectric losses, and radiation losses. An idealized microstrip line, being open to a semi-infinite air space, acts similar to an antenna and tends to radiate energy. Substrate materials with low dielectric constants (5 or less) are used when cost reduction is the priority.

1.2.1 Design Equation

To correlate mathematical expressions with physical dimensions a cross-sectional view of a multilayer microstrip line is shown in Fig. below. The width of the signal line, the cumulative thickness of the first layers, and the dielectric constant of the layer are respectively denoted by, and while the thickness of the first layer (substrate) by as in a one-layer microstrip line.

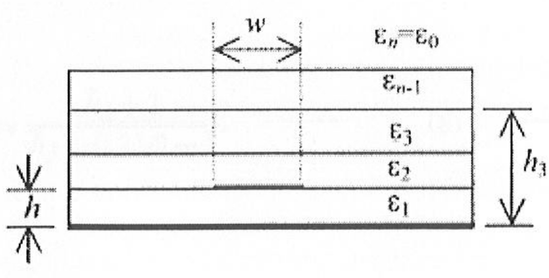


Fig 1.4- Geometry of a multilayer microstrip line

For the multilayer microstrip line in above fig, proposed the following analytical expressions for the filling factors individual layers based on conformal mapping analysis. Here in first case for the calculation of design equations for the microstrip for wide strip is taken and the design equations are stated below.

For $w/h \geq 1$ (wide strip)

$$q_1 = 1 - \frac{1}{2\overline{w}_e}\ln(\overline{w}_e\pi - 1) \quad (2)$$

$$q_j = \frac{1}{2\overline{w}_e}\left\{\ln(\overline{w}_e\pi - 1) - (1 - \overline{v}_j) \times \ln\left[\frac{2\overline{w}_e\cos(0.5\overline{v}_j\pi)}{2\overline{h}_j - 1 + \overline{v}_j} + \sin(0.5\overline{v}_j\pi)\right]\right\} \quad (3)$$

here

$$\overline{w}_e = \overline{w} + \frac{2}{\pi}\ln[17.08(0.5\overline{w} + 0.92)] \quad (4)$$

nd

$$\overline{v}_j = \frac{2}{\pi}\tan^{-1}\left[\frac{2\pi}{\overline{w}_e\pi - 4}(\overline{h}_j - 1)\right]. \quad (5)$$

Now we consider the narrow strip where the ratio of the width of the microstrip to the height is taken to be less than one and the design equations are stated below.

or $w/h < 1$ (narrow strip)

$$q_1 = 0.5 + \frac{0.9}{\pi\ln(0.125\overline{w})} \quad (6)$$

$$q_j = 0.5 - \frac{0.9 + 0.25\pi\ln b_j \cdot \cos^{-1}\left[\frac{\sqrt{b_j}}{\overline{h}_j}(\overline{h}_j - 1 + 0.125\overline{w})\right]}{\pi\ln(0.125\overline{w})} \quad (7)$$

where

$$b_j = \frac{\overline{h}_j + 1}{\overline{h}_j + 0.25\overline{w} - 1}. \quad (8)$$

From (2) to (8), all the bars over their corresponding letters stand for "normalized to h" and $j = 2, 3, \ldots, n-1$. Note that according to [3], other than $1 + \bar{v}_j$ in $1 - \bar{v}_j$ [4] is used in (3). Because of their physical interpretation, filling factors cannot be negative. However, calculations show that for a microstrip line with more than two layers, q_j in (3) or (7) sometimes becomes negative. To avoid this, consider a limiting case with h_j approaching infinity. In this case, the following equation should hold:

$$q_j = 1 - \sum_{i=1}^{j-1} q_i. \tag{9}$$

On the other hand, when $h_j \rightarrow \infty$, (3) and (7) become, respectively,

$$q_j = \frac{1}{2\bar{w}_e} \ln(\bar{w}_e \pi - 1) \tag{10}$$

and

$$q_j = 0.5 - \frac{0.9}{\pi \ln(0.125\bar{w})}. \tag{11}$$

Using (2) and (6), one immediately sees that only if $n = 3$ (two-layer case), (10) and (11) are equivalent to (9). As a result, (3) and (5) should be improved to become, respectively,

$$q_j = 1 - \sum_{i=1}^{j-1} q_i - \frac{1 - \bar{v}_j}{2\bar{w}_e} \times \ln\left[\frac{2\bar{w}_e \cos(0.5\bar{v}_j \pi)}{2\bar{h}_j - 1 + \bar{v}_j} + \sin(0.5\bar{v}_j \pi)\right] \tag{12}$$

and

$$q_j = 1 - \sum_{i=1}^{j-1} q_i - \frac{\ln b_j \cdot \cos^{-1}\left[\frac{\sqrt{b_j}}{b_j}(\bar{h}_j - 1 + 0.125\bar{w})\right]}{4\ln(0.125\bar{w})}. \tag{13}$$

Other expressions in [4] remain unchanged.

As an example, the effective permittivity and characteristic impedance of a three-layer microstrip where the signal line is on the first layer are calculated using previous and improved design equations and compared with the moment method based data It is seen that in the worst case, the effective permittivity and characteristic impedance from the previous equations differ from the MoM-based data by 35.35% and 14.35%. Calculations show that if is used in (3), the case is even worse. In contrast, the corresponding differences between the results from the improved equations and those from the moment method analysis are only 1.40% and 1.14%, respectively.

1.3 VARIATIONAL ANALYSIS

The basic idea of variational analysis is to determine a continuous field approximating the data and exhibiting small spatial variations. In other words, the target of the analysis is defined as the smoothest field that respects the consistency with the observed values over the domain of interest. For analysis of strip and microstrip like transmission, variational method is found to be the simplest. This method requires setting of either the potential function or the green's function for a particular function. These functions are derived by solving a set of algebraic equations obtained by applying the boundary conditions at various boundaries and interfaces.

1.3.1 Technique

Here in this technique for finding the varaitinal analysis, a microstip line is placed between two ground planes whose distance is being increased by a fixed amount.

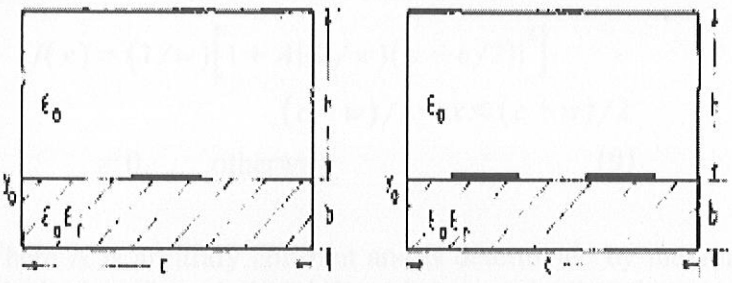


Fig 1.5(a) –Microstrip and **(b)** Edge coupled microstrip

In this case the microstrip line is placed in covered metal shield with ground planes separated by a certain distance which goes on increasing at a constant rate.
Here the characteristic admittance of the line is given as

$\gamma = \beta_n$

Where the following equations are used to determine the capacitance of the microstrip,

$$\beta_n = n\pi/c.$$

$$L_n = \sin(\beta_n w/2)$$

$$M_n = (2/\beta_n w)^3 \left[3\left\{(\beta_n w/2)^2 - 2\right\}\cos(\beta_n w/2) + (\beta_n w/2)\left\{(\beta_n w/2)^2 - 6\right\}\sin(\beta_n w/2) + 6\right]$$

$$P_n = (2/n\pi)(2/\beta_n w)^2$$

$$\gamma = \varepsilon_0\left[\coth\frac{n\pi h}{c} + \varepsilon_r \coth\frac{n\pi b}{c}\right]$$

$$A = -\frac{\sum_{n\,odd} (L_n - 4M_n) L_n P_n / Y}{\sum_{n\,odd} (L_n - 4M_n) M_n P_n / Y}.$$

$$T_n = (L_n + AM_n)^2$$

The capacitance of the structure is obtained by the following variational expression as:-

$$C = \frac{\left[\int_{S_1} f(x)\,dx\right]^2}{\int_{S_1}\int_{S_1} G(x, y_0 / x_0, y_0) f(x) f(x_0)\,dx\,dx_0}$$

Where f(x) is the charge distribution on strip conductor s1 and can be assumed to be of the form

$$f(x) = (1/w)\left[1 + A|(2/w)(x - c/2)|^3\right], \quad (c-w)/2 \leqslant x \leqslant (c+w)/2$$

$$= 0, \quad \text{otherwise} \qquad (9)$$

Where A is arbitrary constant and is determined by maximizing the line capacitance C. substituting (7) and (9) in (8) and then evaluating the integral, we obtain the following expression for calculation of capacitance

$$C = \frac{(1 + 0.25A)^2}{\sum_{n\,odd} (T_n P_n / Y)}$$

Once the capacitance for two different values of relative permittivity is calculated the impedance is calculated by

$$Z_0 = \frac{1}{v\sqrt{CC_0}}$$

Where
υ = 3*10^11

C = capacitance per unit length of microstrip for given relative permittivity of €r=2.2, €r=4.6 and €r=9.9.

Co = capacitance per unit length for same structure but wit €r=1.

1.4 CROSSTALK COUPLING

The ratio of the power in a disturbing circuit to the induced power in the disturbed circuit observed at specified points of the circuits under specified terminal conditions. This is called **crosstalk coupling**. Crosstalk is usually referred as a phenomenon that signal transmitted on one circuit or channel of transmission system creates an undesired effect on other channel or other part of circuit. Crosstalk coupling can cause lots of undesirable results including excessive signal delay, overshoot, undershoot and even reduction in signal delay.

To overcome the effects of crosstalk it is desirable to use differential signals and traces as they are less susceptible to radiated noise and thus radiate less noise and therefore reduce the effect of crosstalk. Crosstalk coupling is usually expressed in dB.

1.4.1 RF-IC model and Coupling mode

The RF-IC model consist pins mounted in plastic model. Where the line of width 0.76mm and spacing of pins is 0.05mm.power transferred by one pin is disturbed among al the pins which cause the crosstalk. In order to reduce the crosstalk a ground plane is made between the two pins.

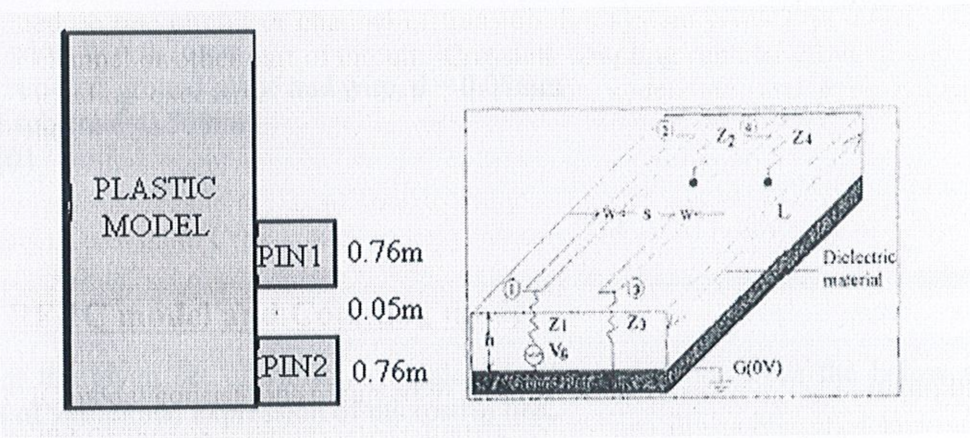


Fig 1.6(a)-RF-IC model **(b)** coupled pair of microstrip line on ground subtrate

In fig (b), we assume that the two microstrips are homogenous and of equal width, w.Here port 1 is excited by a voltage generator and al the four ports are terminated with impedance. The voltages induced at port 3 and port 4 is respectively named as backward and forward crosstalk.

2

CALCULATIONS

2.1 Variational analysis

The variational analyis of microstrip line were done by set of formulas, where This method requires setting of either the potential function or the green's function for a particular function. These functions are derived by solving a set of algebraic equations obtained by applying the boundary conditions at various boundaries and interfaces. The given set of formulas used for the calculation where the given set of values are:-

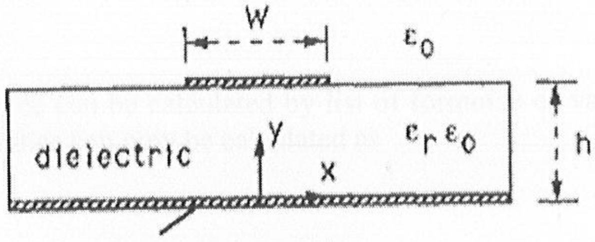


Fig 2.1- Crossection of microstrip line.

Width of wire, W = 0.76mm
Distance between ground plane and wire, d = 0.05mm
Height of substrate=0.508mm
n=0 to 1001
€r=1
€r=2.2
€r=4.6
€r=9.9

Now from the given set of values, we calculate the impedance from the below set of equations of variational expression of microstrip line.

$$\gamma = \beta_n$$

Where,

$$\beta_n = n\pi/c.$$

$$L_n = \sin(\beta_n w/2)$$

$$M_n = (2/\beta_n w)^3\left[3\{(\beta_n w/2)^2 - 2\}\cos(\beta_n w/2) + (\beta_n w/2)\{(\beta_n w/2)^2 - 6\}\sin(\beta_n w/2) + 6\right]$$

$$P_n = (2/n\pi)(2/\beta_n w)^2$$

$$Y = \epsilon_0 \left[\coth \frac{n\pi h}{c} + \epsilon_r \coth \frac{n\pi h}{c} \right]$$

$$A = -\frac{\sum_{n\,\mathrm{odd}} (L_n - 4M_n) L_n P_n / Y}{\sum_{n\,\mathrm{odd}} (L_n - 4M_n) M_n P_n / Y}.$$

$$T_n = (L_n + AM_n)^2$$

Where

$$C = \frac{(1 + 0.25A)^2}{\sum_{n\,\mathrm{odd}} (T_n P_n / Y)}$$

That's how the capacitance can be calculated by list of formulae of variational analysis. The characteristic impedance can now be calculated as

$$Z_0 = \frac{1}{\upsilon\sqrt{CC_0}}$$

Where

υ = 3*10^11

C = capacitance per unit length of microstrip for given relative permittivity of €r=2.2, €r=4.6 and €r=9.9.

Co = capacitance per unit length for same structure but wit €r=1.

2.1.1 <u>Program for variational analysis of a miscrostrip line</u>

The following is the code written for variational analysis of microstrip line using the formulas stated above. This code calculates the capacitance for following values of relative permittivity:-

€r=2.2
€r=4.6
€r=9.9

Where n runs from 0 to 1001.After getting the capacitance of each relative permittivity, it multiplies it with relative permittivity of one. And then impendence is calculated correspondingly for each permittivity for a given value of n.

2.1.2 Code

```
Clear;
num1 = 0;
den1 = 0;
num2 = 0;
den2 = 0;
num3 = 0;
den3 = 0;
num4 = 0;
den4 = 0;

er1 = 1;
er2 = 2.2;
er3 = 4.5;
er4 = 9.9;
for i = 1:24
for j = 1:501

  d = i/20;
  n = (2*j)-1;
  a = (n*3.14*0.38)/ ((2*d) + 0.76) ;
  l = sin (a);

 m = power ((1/a),3)*(3*(power(a,2)-2)*cos(a) + a*(power(a,2)-6)*sin(a)+6);
 p = (2/ (n*3.14))*power ((1/a), 2);

  y1   =      8.854*10^-15*[coth ((n*3.14*200)/ ((2*d) + 0.76)) +
              (er1*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))] ;

  y2   =      8.854*10^-15*[coth ((n*3.14*200)/ ((2*d) + 0.76)) +
              (er2*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))] ;

  y3   =      8.854*10^-15*[coth ((n*3.14*200)/((2*d) 0.76))+
              (er3*coth ((n*3.14*0.508)/((2*d) + 0.76)))];

  y4   =      8.854*10^-15*[coth ((n*3.14*200)/((2*d) + 0.76))+
              (er4*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))];

  num1 = num1 + (((l-(4*m))*l*p)/y1);
  den1 = den1 + (((l-(4*m))*m*p)/y1);
  num2 = num2 + (((l-(4*m))*l*p)/y2);
  den2 = den2 + (((l-(4*m))*m*p)/y2);
  num3 = num3 + (((l-(4*m))*l*p)/y3);
  den3 = den3 + (((l-(4*m))*m*p)/y3);
```

```
  num4 = num4 + (((l-(4*m))*l*p)/y4);
  den4 = den4 + (((l-(4*m))*m*p)/y4);
  end

a11 = -1*num1/den1;
a12 = -1*num2/den2;
a13 = -1*num3/den3;
a14 = -1*num4/den4;


denn1 = 0;
denn2 = 0;
denn3 = 0;
denn4 = 0;

for j = 1:501
  n = (2*j)-1;
  d = i/20;
  a = (n*3.14*0.38)/((2*d) + 0.76);
  l = sin (a);

  m = power ((1/a), 3) * (3 * (power(a,2)-2)*cos(a) + a*(power(a,2)-6)*sin(a)+6);
  t1 = power ((l + (a11*m)), 2);
  t2 = power ((l+(a12*m)), 2);
  t3 = power ((l+ (a13*m)), 2);
  t4 = power ((l+ (a14*m)), 2);
  p = (2/ (n*3.14))*power ((1/a), 2);


  y1    =       8.854*10^-15*[coth ((n*3.14*200)/ ((2*d) + 0.76)) +
               (er1*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))] ;

  y2    =       8.854*10^-15*[coth ((n*3.14*200)/ ((2*d) + 0.76)) +
               (er2*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))] ;

  y3    =       8.854*10^-15*[coth ((n*3.14*200)/((2*d) 0.76))+
               (er3*coth ((n*3.14*0.508)/((2*d) + 0.76)))];

  y4    =       8.854*10^-15*[coth ((n*3.14*200)/((2*d) + 0.76))+
               (er4*coth ((n*3.14*0.508)/ ((2*d) + 0.76)))];

  denn1 = denn1 + ((t1*p)/y1);
  denn2 = denn2 + ((t2*p)/y2);
  denn3 = denn3 + ((t3*p)/y3);
  denn4 = denn4 + ((t4*p)/y4);

end
```

```
c1 = power((1+ (0.25*a11)),2)/denn1;
c2 = power ((1+(0.25*a12)),2)/denn2;
c3 = power ((1+(0.25*a13)),2)/denn3;
c4 = power ((1+(0.25*a14)),2)/denn4;
d

z1 = 1/ (3*10^11*sqrt (c1*c2))
z2 = 1/ (3*10^11*sqrt (c1*c3))
z3 = 1/ (3*10^11*sqrt (c1*c4))
end
```

The above code runs for n varying from 0 to 1001 and for initial separation between the ground planes and the wire taken as 0.05 which is increasing at 0.05% for every increase in distance, d between the two. It calculates the characteristic impedance for every value of d.The expressions used are general and can be used to obtain data for a class of stripline and microstrip structures with single strip conductor by simply substituting the appropriate expression for Y at charge plane.

3

RESULTS

3.1 Results obtained from variational analysis

The following are the characteristic impedance obtained for varying values of distance. As the distance between the wire and ground increases, the characteristic impedance almost stabilizes to a constant value.

D	z1(Ωs)	z2(Ωs)	z3(Ωs)
0.05	39.4409	30.0202	21.2934
0.1	47.0333	35.7458	25.3289
0.15	52.5163	39.8457	28.2017
0.2	56.7133	42.9522	30.3632
0.25	60.0169	45.3695	32.0321
0.3	62.6622	47.2814	33.3407
0.35	64.8068	48.8115	34.3789
0.4	66.5634	50.0484	35.2106
0.45	68.0153	51.0575	35.8831
0.5	69.2255	51.8879	36.4317
0.55	70.2422	52.577	36.8833
0.6	71.1028	53.1535	37.2583
0.65	71.8363	53.6397	37.5723
0.7	72.4659	54.0528	37.8375
0.75	73.0096	54.4064	38.0632
0.8	73.4821	54.7112	38.2569
0.85	73.895	54.9757	38.4243
0.9	74.2578	55.2066	38.5701
0.95	74.5782	55.4094	38.6978
1	74.8625	55.5886	38.8103
1.05	75.1159	55.7476	38.9101
1.1	75.3427	55.8895	38.9991
1.15	75.5465	56.0166	39.0787
1.2	75.7304	56.1311	39.1504

Fig 3.1 Values of Impedance for given distances.

Where Z1gives the impedance (Ωs) with €r=2.2 and Z2 gives the impedance (Ωs) with €r=4.6 and Z3 gives the impedance (Ωs) with €r=9.9

3.1.1 Graph "impedance Vs distance "

The plot of "impedance Vs distance "given below shows that with increasing value of distance between the wire and the ground plane, the characteristic impedance almost stabilizes to constant value for different relative permittivities varying from 2.2 to9.9.

Z1 shows that for relative permittivity of 2.2, intinally the impedance is 39.44 Ω and with increase in distance it increases to 75 Ω with stabilization. Similarly for Z2 and Z3 with relative Permittivities of 4.6 and 9.9.

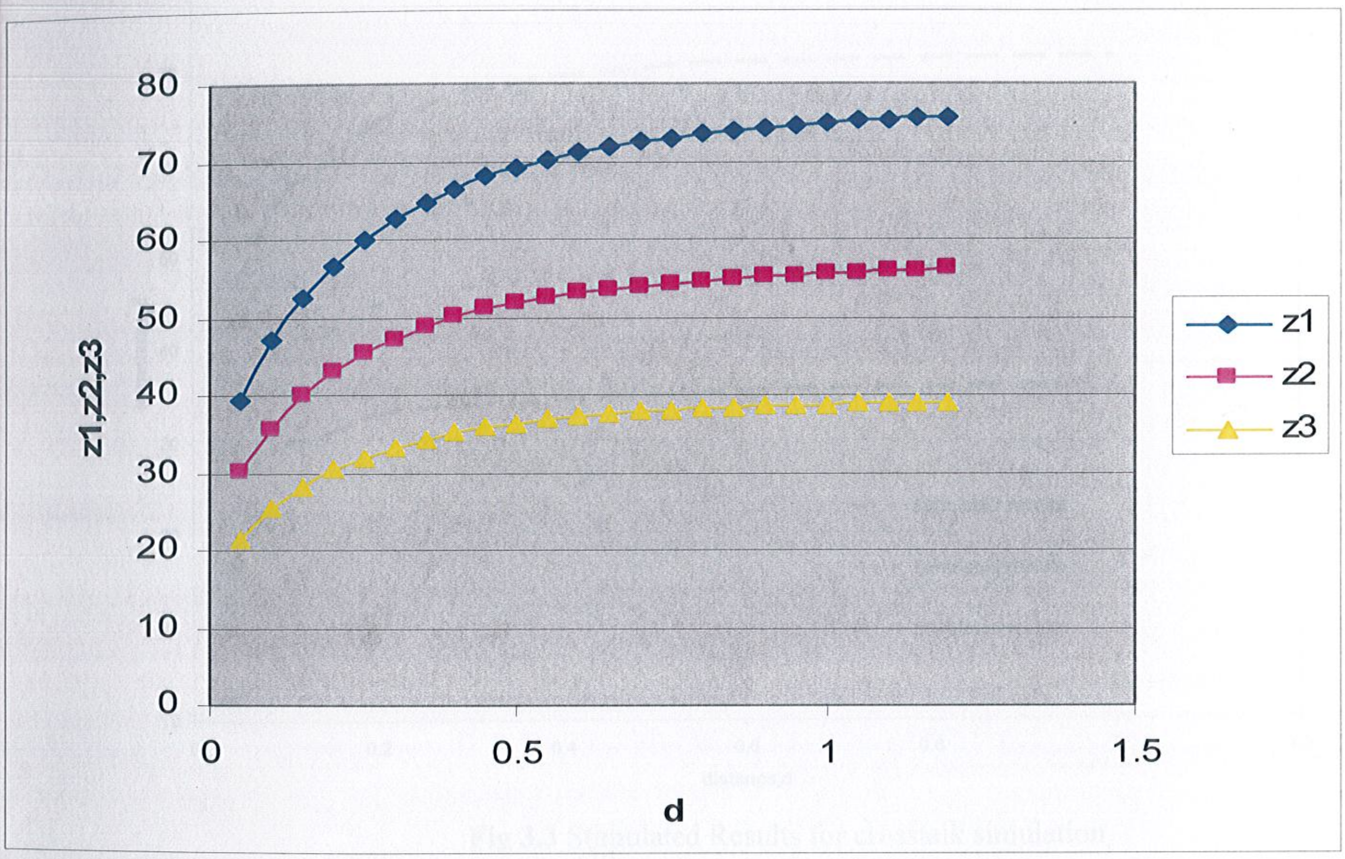


Fig 3.2 Relationship between impedance and distance

3.2 CROSSTALK SIMULATION

In this **crosstalk simulation technique**, a ground plane separates two wires initially by a distance of 0.05mm and then this distance goes on increasing by a constant of 0.05mm.With the introduction of ground plane coupling effect is reduced. And the graph stabilizes to a constant value.

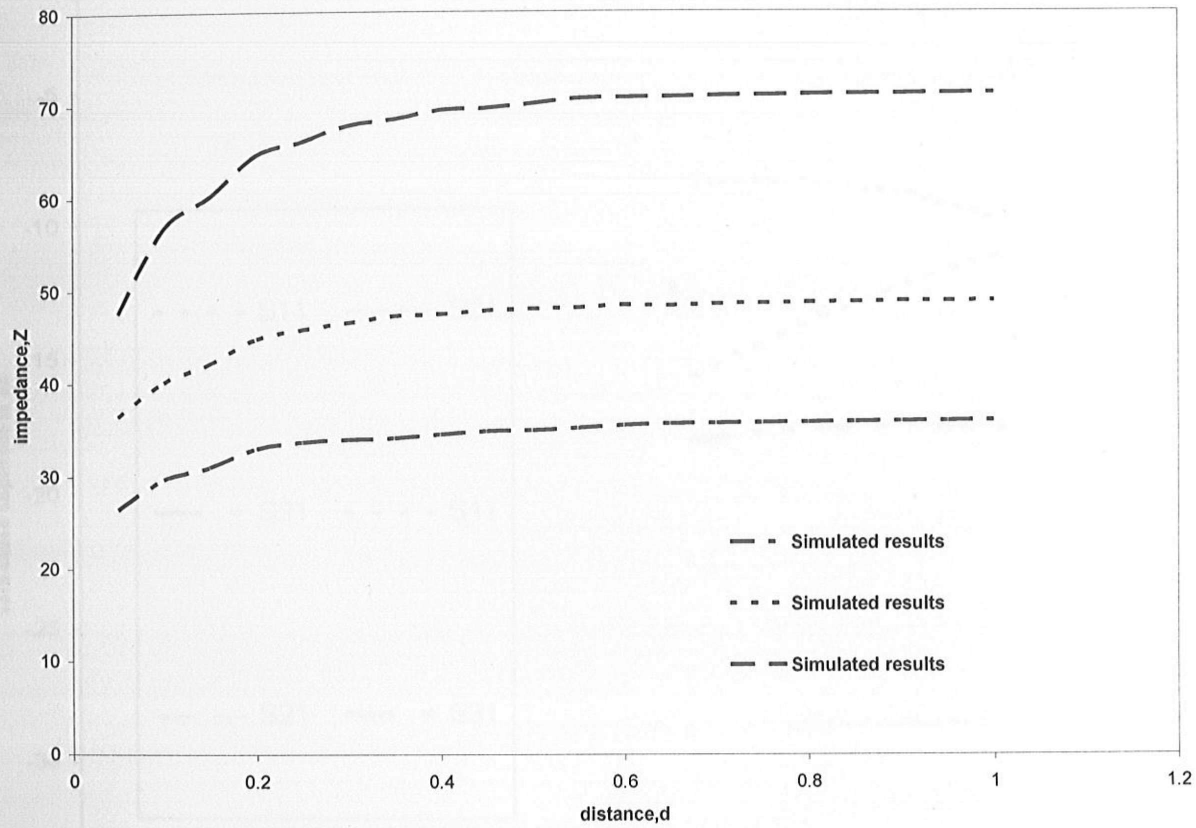


Fig 3.3 Stimulated Results for crosstalk simulation

3.2.1 S-Parameters for 50 ohms ground proximity

This plot shows the simulated S-parameters for coupled microstrip line with ground plane placed between the two lines. It is seen from the graph that S_{31} decreases significantly with placement of ground in between. However it is worth noting that S_{11} also improves with ground proximity. Thus this design methodology can be used in practical systems

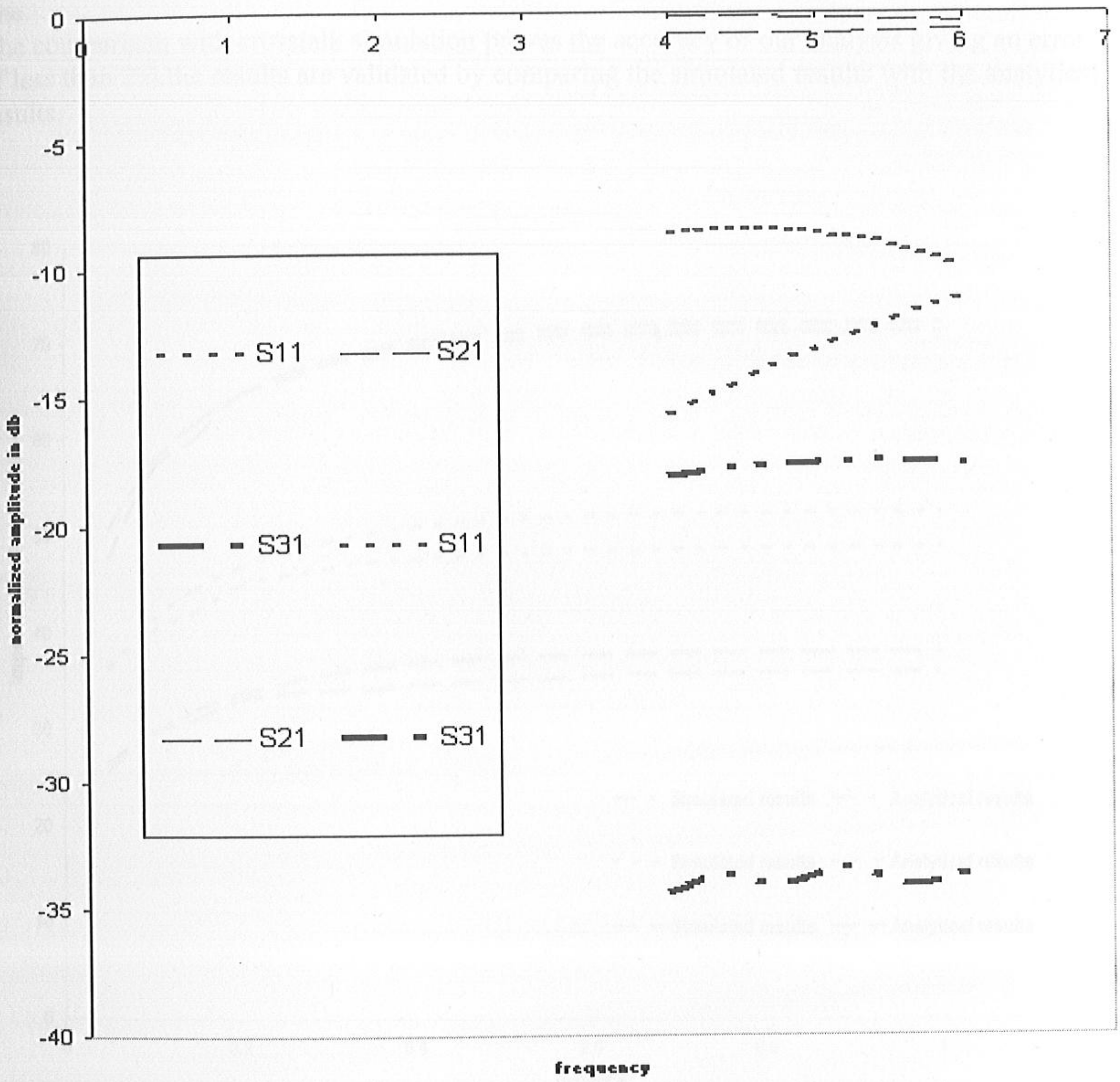


Fig 3.4 Relationship between Normalized amplitude and frequency.

Here for the above graph of ground proximity
S11=reflection coefficient
S21=Transmission coefficient
S31=coupling coefficient
S41=isolation coefficient

3.3 Comparison of analytical results and simulated results

The comparative results shown below are obtained from for different substrates with dielectric constants ranging from 2.2 to 9.9.the choice of these materials is governed by the fact that commonly occurring PCB substrates are often in this range. The results are obtained with electric wall separation far more than the ground plane aperture width. it is found that beyond a particular high value wall separation has a marginal effect on characteristic impedance of the line.

The comparison with crosstalk simulation proves the accuracy of our analysis giving an error of less than 2%.the results are validated by comparing the simulated results with the analytical results.

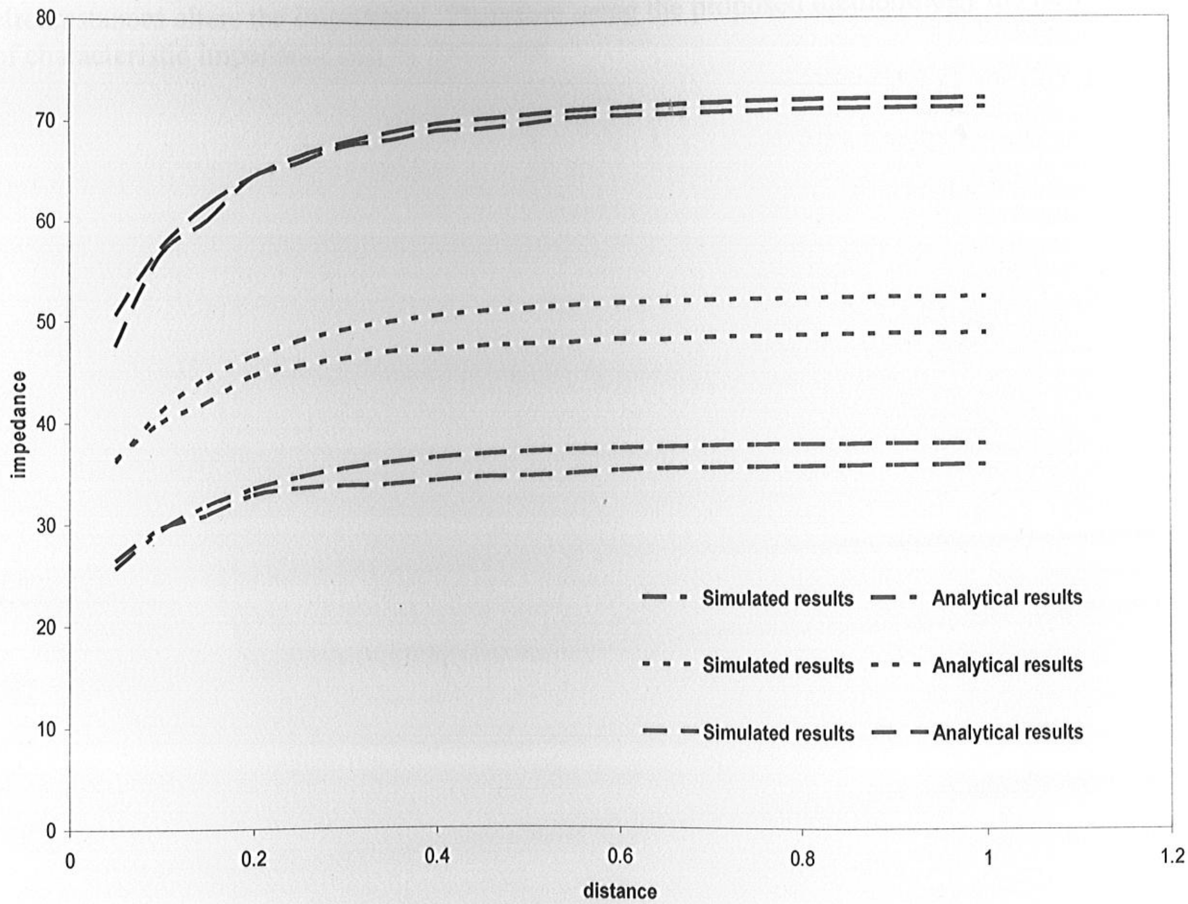


Fig 3.5 Comparison between analytical and simulated results

4

CONCLUSION

In this project work it is demonstrated that when coplanar ground is brought near the microstrip line the characteristic impedance of the line is affected. The characteristic impedance of the line reduces significantly. It is also seen that when the ground plane is separated at least one line-width away, the characteristic impedance value stabilizes to the normal value without the ground plane. Using cross-talk simulation it is shown that in real life multi-chip module applications ground plane can placed in between two signal carrying lines thereby reducing cross-talk. It is worth noting that by placing ground plane in close vicinity under such circumstances alters the impedance. Therefore using the proposed methodology the new value of characteristic impedance can be found out.

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