Accurate Modelling of Thin-Film Resistor up to 40 GHz

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Abstract

We present an accurate scalable model for Thin-Film Resistors from DC up to 40GHz. The model is based on the basic microstrip theory, and an empirical self-capacitance parameter is introduced. The simulated S-parameter of the model show very good agreement with the measurements.

1. Introduction

Thin-Film Resistors (TFRs) are used in microwave circuits as passive attenuators, terminal loads. To date, TFR model considers the parasitic series inductance and shunt capacitance to be the same as from a lossless microstrip line[2]. However, this model does not provide get good results when the width of resistor is much smaller than substrate thickness. In addition, the self-capacitance has to be taken into account to improve the model accuracy. The self-capacitance of planar resistor was first introduced by Demurie[1]. If a voltage is applied at the terminals of a resistor, a potential different will exist across two arbitrary points *A* and *B* in the resistor. Therefore, a parasitic capacitance exists between *A* and *B*.

In microwave integrated circuit (MIC) technology, a thin-film resistor is realized as a thin strip of a lossy conductor on top of the dielectric substrate. The resistive layer can be a self-passivating Tantalum Nitride (TaN) compound. The sheet resistivity of the process is adjusted by controlling the thickness of the resistive layer. In most processes, a sheet resistivity of 50Ω per square is selected for the convenience it provides to circuit designers. A small area of conductor metal is deposited at the ends of the element as contacts to the resistor. The exposed resistive area defines the resistance of the structure. The fabrication design rules generally require that resistive layer be narrower than the width of the conducting contact by some minimum distance. The requirement arises due to the need to have a good contact between the resistive layer and the conducting layer by process alignment tolerances.

Figure 1 shows layout cross-section of the thin-film resistor. The measurement reference planes at the ends of the conductor contacts are shown in Figure 1(a), so that the conductor contact is removed from the measurement. However, the electrical effect of the step discontinuity will not be removed from the measurement. As a result, the thin-film resistor model should be divided into three sections as shown in Figure 2. The middle section is the intrinsic thin-film resistor, which is modelled as a lossy microstrip transmission line. The other two sections models the step discontinuities at both sides of the intrinsic thin-film resistor.

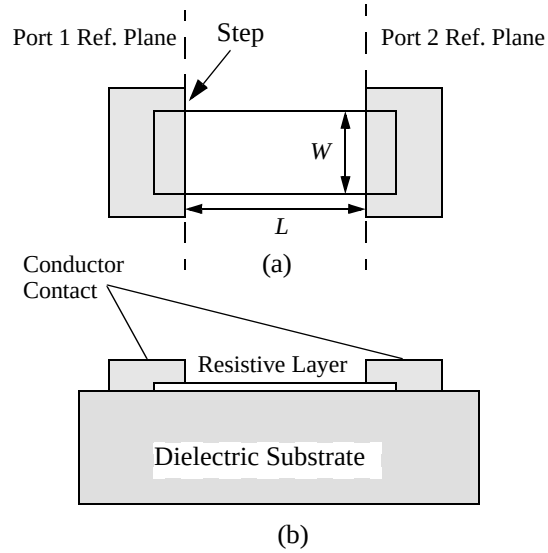


Figure 1. (a) Overhead and (b) side view of thin-film resistor

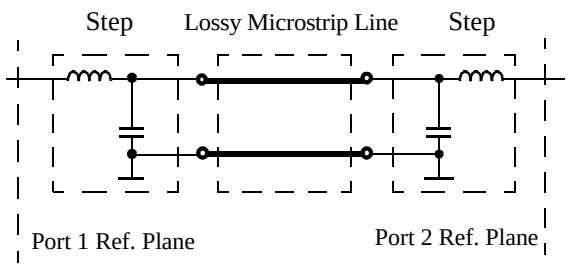


Figure 2. Equivalent circuit of Thin-Film Resistor

2. Step discontinuity in microstrip width

Since the width of the thin-film resistor must be narrower than that of the contacting conduct layer, a step discontinuity exits at both ends of the resistor. The electromagnetic field is discontinuous at the steps because the current density increases from the wider to the narrower conductor and scattered electric fields exist on the front edge of the wider conductor, as shown in Figure 3(a). Figure 3(b) shows the equivalent circuit of the step in conductor width. L_s represents the current compression and C_p represents the electrical scattering fields.

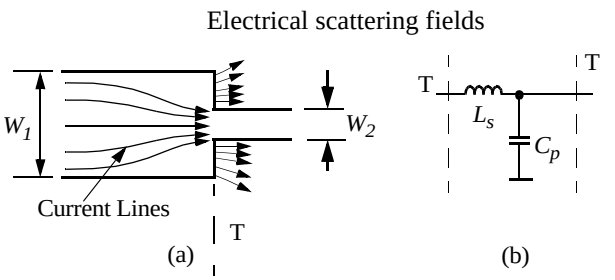


Figure 3. Step discontinuity in microstrip width (a) construction (B) lumped-element equivalent circuit[3]

C_p is approximately given by[3]

$$C_p = C_{f1}(W_1 - W_2). \quad (1)$$

Here, C_{f1} is the fringing capacitance per unit length of wider microstrip, and it is

$$C_{f1} = \frac{1}{2}\left(\frac{\sqrt{\varepsilon_{eff1}}}{c_0 Z_{01}} - \varepsilon_0 \varepsilon_r W_1/h\right) [F/m], \quad (2)$$

where c_0 is the speed of light in free space, Z_{01} and ε_{eff1} are the characteristic impedance and effective dielectric constant of the wider microstrip.

In some cases TFRs may have very small lengths, and the two step discontinuities can be very close to each other. In this case, capacitance C_p at one end of resistor will decrease because the electrical scattering field is constrained by the step at the other end of resistor. To calculate C_p under this condition, the method which derives the even mode fringing capacitance C_f' of two parallel microstrip line is employed, and it can be expressed as[4]

$$C_{f1}' = \frac{C_{f1}}{1 + A(h/l)\tanh(8l/h)} [F/m], \quad (3)$$

where $A = exp[-0.1\,exp(2.33 - 2.53W_1/h)]$ and l is the length of resistor. By putting Eqs.(2) and (3) into (1), C_p can be obtained as

$$C_p = \frac{\frac{1}{2}\left(\frac{\sqrt{\varepsilon_{eff1}}}{c_0 Z_{01}} - \frac{\varepsilon_0 \varepsilon_r W_1}{h}\right)\cdot(W_1 - W_2)}{1 + A(h/l)\tanh(8l/h)} \quad (4)$$

The closed form expression for inductance L_s has been derived by curve fitting the numerical results, and L_s is[6]

$$L_s = [a(\alpha - 1) - b\log((\alpha) + c(\alpha - 1)^2)]h \text{ [nH]} \quad (5)$$

where $a = 40.5$, $b = 75$, $c = 0.2$ and $\alpha = W_1/W_2$.

3. High sheet resistance microstrip model

As a two-port network, a high sheet resistance microstrip line can be represented by an ABCD matrix[5]

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cosh(\gamma l) & Z_0 \sinh(\gamma l) \\ \frac{\sinh(\gamma l)}{Z_0} & \cosh(\gamma l) \end{bmatrix} \quad (6)$$

where Z_0 is characteristic impedance and γ is propagation constant. It is well known that Z_0 and γ can be given by series impedance per unit length Z and shunt admittance per unit length Y of the microstrip line as,

$$Z_0 = \sqrt{\frac{Z}{Y}} \quad (7)$$

$$\text{and } \gamma = \sqrt{Z \cdot Y}. \quad (8)$$

Due to the low value of the losses in alumina substrate, conductance per unit length shunt is neglected. Therefore, Y can be given by per unit length shunt capacitance per unit length C as $Y = j\omega C$. For the low loss microstrip line, Z can be given by series resistance per unit length R, and inductance per unit length L, as $Z = R + j\omega L$. However, for a high sheet resistance microstrip line, the self-capacitance must be taken into account. In the work of Demurie[1], the self-capacitance was derived by numerical calculation. In our work, self-capacitance was derived by using a commercially available full-wave electromagnetic simulator (HP-Momentum).

Two sets of microstrip lines were simulated by HP-Momentum. One set were microstrip lines with high sheet resistance of 50Ω per square, the other set were lossless microstrip line. Electromagnetic simulations were performed from 1GHz to 40GHz in 4GHz steps. The dimensions of the microstrip lines are shown in Table 1.

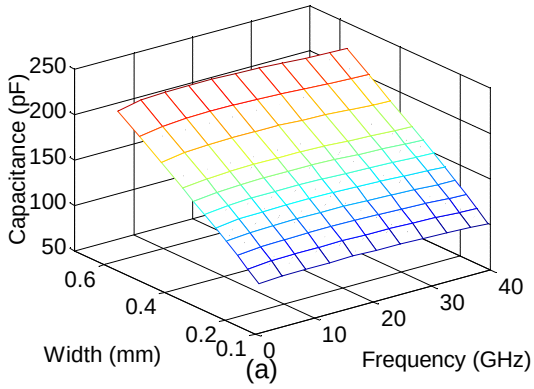
Table 1: Parameters of simulated microstrips

Input Parameter	Minimum Value	Maximum Value
Frequency (GHz)	1	40
Width (μm)	100	500
Length (μm)	200	500

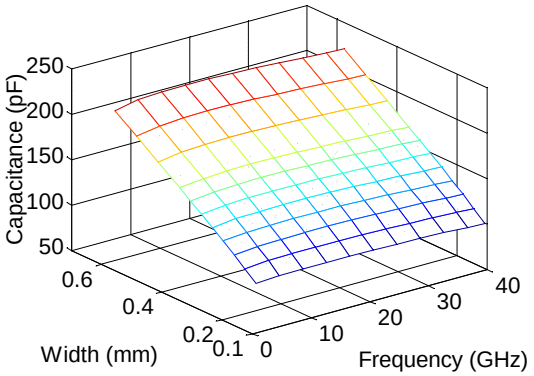
The series resistance, inductance and shunt capacitance per unit microstrip length were extracted from the simulated S-parameter. The extraction procedure is:

1. Transfer S-parameters to ABCD parameters.

2. Determine the propagation constant γ as $\gamma = \operatorname{acosh}(A)/l$ and the characteristic impedance as $Z_0 = B/[\sinh(\operatorname{acosh}(A))]$.
3. Determine series impedance Z and shunt admittance Y of the microstrip line $Z = Z_0 \cdot \gamma$ and $Y = \gamma/Z_0$ respectively.
4. Obtain series resistance per unit length as equal to the real part of Z, series inductance per unit length as $L = Im(Z)/\omega$, and shunt capacitance per unit length as $C = Im(Y)/\omega$.



(a)



(b)

Figure 4. Shunt capacitance per unit length of (a) resistive line (50Ω/□) and (b) lossless line.

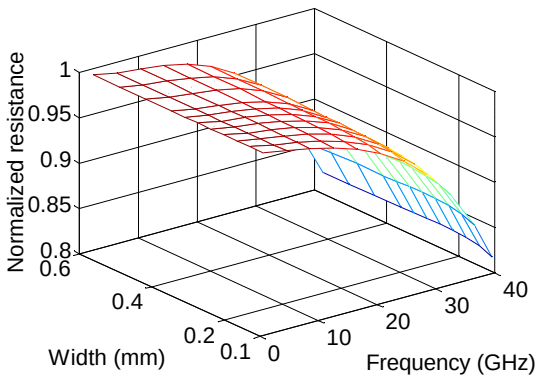
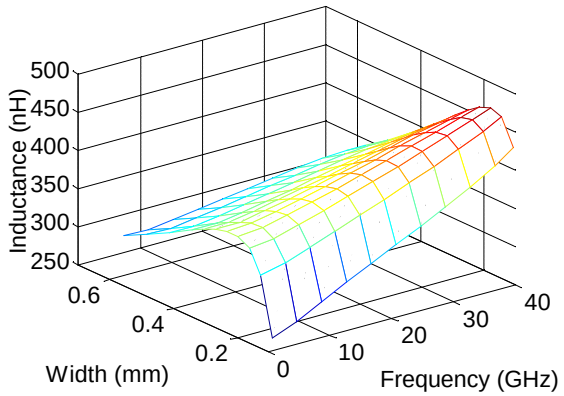
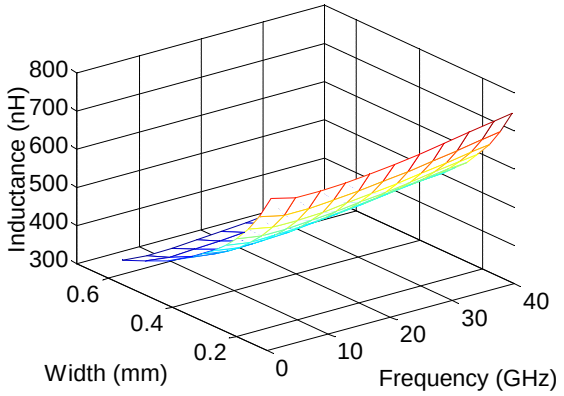


Figure 5. Frequency dependence series resistance per unit length normalized to *DC* series resistance per unit length of high sheet resistance (50Ω/□) line



(a)



(b)

Figure 6. Series inductance per unit length of (a) resistive line (50Ω/□) and (b) lossless line.

The values of capacitance are nearly identical in Figure 4(a) and (b), which verifies that the shunt capacitance per unit length of high sheet resistance microstrip line can be simply derived from the case which is lossless line. Figure 5 shows that the series resistance per unit length of high sheet resistance line decreases with the frequency, because of the self-capacitance effect. Figure 6(a) and (b) are totally different. This means that series inductance per unit length of high sheet resistance microstrip line cannot be simply obtained from the case of lossless line. Further, we find that series impedance per unit length cannot be represented just by a resistor in series with an inductor. An appropriate first order model for the series impedance Z is shown in Figure 7 and Eq.(9).

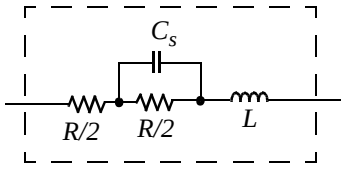


Figure 7. Equivalent circuit of per unit length series impedance

$$Z = R\left[1 - \frac{(\omega R C_s/2)^2}{1 + (\omega R C_s/2)^2}\right] + j\omega\left[L - \frac{R^2 C_s/4}{1 + (\omega R C_s/2)^2}\right] \quad (9)$$

where L is equal to series inductance per unit length in the case of lossless line, R is series resistance per unit length at DC, C_s represents the self-capacitance, which has been derived by curve fitting of the numerical Momentum simulation results.

$$C_s = 1.58 \times 10^{-18}\left(\frac{W}{h} + 1.26\right) \quad (10)$$

where W is the width of resistor and h is the substrate height which is 15 *mil* in our samples.

In the model given by Eq.(9), the real part of Z is the series resistance per unit length, which decreases with frequency in agreement with the trend in Figure 5. The equivalent per unit length inductance is a function of the DC resistance per unit length R, self-capacitance C_s and series inductance per unit length L in the case of lossless line, and it is smaller than that of lossless line.

The series inductance per unit length L and shunt capacitance per unit length C of lossless line can be directly calculated from the empirical formula for Z_c and ε_{eff} of a lossless transmission line, which is under quasi-TEM assumption[3]. The resistance per unit length R is determined by the technology. C_s is given by Eq.(9). Thus, impedance per unit length Z and admittance per unit length Y of high sheet resistance microstrip line can be calculated from L, R, C_s and C. By using Eqs.(7) and (8), the characteristic impedance Z_0 and propagation constant γ can be calculated. Finally, the model can be implemented as matrix form by using ABCD-matrix shown in Eq.(6).

4. Experimental Verification

Nine Thin-Film Resistors were fabricated and measured. They are different in width and length. The planar physical dimensions of the TFR test structures were measured using a microscope. The DC resistance of TFRs were measured using a digital multimeter. Based on the measured resistance and the physical dimensions of a resistor, the sheet resistance of the resistive layer was determined. S-parameters were measured with a Wiltron 37396A Vector Network Analyzer. The S-parameters of two TFRs are shown in Figure 8. The physical dimensions of these two TFRs are listed in Table 2. As it can be seen in Figure 8, good agreement between model simulations and measurements has been obtained.

Table 2: Measured physical parameters of TFRs

dc Resistance	Width	Length
52.4 Ω	270 μm	265 μm
104.2 Ω	270 μm	530 μm

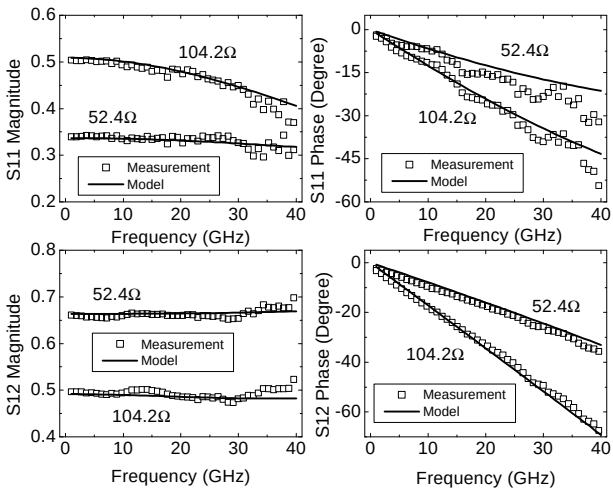


Figure 8. Comparison between measured and simulated S-parameters for two TFRs(53.4Ω and 104.2Ω)

5. Conclusions

We proposed a scalable model for thin-film resistor. Details for the inductance and capacitance parasitics in the TFR have been discussed. A self-capacitance is used to get better fitting of the model. Good agreement between simulations with the proposed model and measurements has been obtained up to 40 GHz.

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6. References

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