

# Passive Elements Modeling in Microwave/Millimeter-wave Application

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## ABSTRACT

Passive elements (e.g., transmission line, transformer and balun) are widely used in Microwave/Millimeter-wave (MMW) applications. Compared with time-consuming Electromagnetic simulation, equivalent circuit based compact models for these passive elements are more preferable for IC designers. However, most of existing equivalent-circuit models only cover low-frequency characteristic, which is not suitable for MMW applications. In this work, we presented new compact models for passive elements in MMW applications. Experimental results show that the proposed compact models are accurate enough over the entire frequency range up to 67GHz for both transmission line and transformer when the distributive coupling between the metal tracks and substrate effect are emphasized.

**Keywords:** compact model, transmission line, transformer, millimeter-wave.

## 1 INTRODUCTION

Along with the aggressive scaling of CMOS technology, the cut-off frequency ( $F_t$ ) and minimum noise figure ( $NF$ ) of MOSFET are greatly improved to allow high performance millimeter-wave (MMW) applications, such as high speed wireless communication at 60GHz [1], and automobile radar at 24GHz and 77GHz [2]. However, circuit components will suffer from significant parasitic effect at millimeter-wave regime, especially for passive elements (e.g., transmission line, transformer and balun) since the layouts are designed small to achieve high self-resonant frequency. This makes them extremely impressionable on the layout arrangements.

Although electromagnetic (EM) simulation can provide the S-parameters of passive elements with good accuracy, their computational cost is usually very high. Besides, EM based passive components only support small-signal analysis. Direct extension of such method to analysis with large signal operation and noise analysis is difficult. To support SPICE circuit simulation in frequency domain, time domain, and noise analysis etc, frequency-independent equivalent-circuit compact models are preferable for designers.

However, establishing equivalent circuit model at millimeter-wave regime is perplexed by multiple coupling effects because the parasitic components become approaching to the same magnitude of the circuit

components. Therefore, it is extremely difficult to build an equivalent-circuit compact model to predict the MMW performance. For transmission line (TL), analytical model and SPICE models for coplanar waveguides (CPW) and microstrip have been presented in [3][4]. However, it is difficult to maintain high accuracy over a large frequency range with effective extraction methodology. The development of reliable TL models lagged behind inductor models. Meanwhile, although there have been extensive works on constructing SPICE models of on-chip balun and transformers [5]-[7], few researches are done on SPICE-compatible equivalent circuit models at millimeter frequencies. Most of the transformer models from PDK only cover the low frequency characteristics of transformers [8][9]. The electronic specifications, such as quality factor, insertion loss, self-resonant frequency, are not optimized with layout structure. This has hampered the extensive use of integrated transformers in MMW applications.

In this paper, we report recent progresses made by our group about the physics-based lumped-element circuit models of TLs and transformers suitable for millimeter-wave frequency applications.

## 2 MODEL AND VERIFICATION

### 2.1 Transmission Line

Microstrip lines (MS) and coplanar waveguides (CPWs) are the most commonly used TLs in CMOS circuits. Although MS provide shielding from the lossy substrate, the gap between the signal wire and bottom metal is set by the technology, leading to poor flexibility in circuit design. CPWs, on the other hand, have the freedom to adjust both the width and gap, and are easy to be integrated in standard CMOS processes. However, they suffer from losses as high as 1-2dB/mm at 10GHz due to substrate dielectric loss, while grounded CPW (GCPW) are confronted with the similar problem as MS. As a comprise, CPW with slotted shield (SCPW) which realizes low wire losses and high effective dielectric permittivity, as well as chip size [3], is promising candidate for high quality passive components at millimeter-wave frequency and an accurate SPICE-compatible equivalent circuit model is avidly expected.

A new unified model for CPW, GCPW, and SCPW is proposed by our group and corresponding direct extraction methodology is provided to acquire the parameters of the model. Based on the quasi-TEM assumption, the CPWs could be fully characterized by the Telegrapher's equation with RLGC parameters. In this work,  $R_x$ ,  $L_x$ ,  $G_x$ , and  $C_x$  are

adopted to describe frequency dependent per-unit-length RLGC parameters. The CPWs are modeled by multiple cascaded blocks  $B_i$ , in which equivalent circuits are adopted to capture the behaviors of  $R_x$ ,  $L_x$ ,  $G_x$ , and  $C_x$ , as shown in Figure 1.

The series branch is consisted of an inductor  $L_{hf}$  in series with a R/L ladder composed of  $R_{hf}$ ,  $R_1$  and  $L_1$  to characterize the skin effect at high frequencies. The low resistivity of silicon substrate causes serious signal loss, and the increasing attenuation makes the implementation of a CMOS millimeter-wave system difficult. So in the TL models, how to describe the substrate coupling accurately is the most difficult. Modification of parallel branches is the emphasis of the modeling. As shown in Figure 1, the parallel branch is comprised of three parts to describe three different coupling effects: (1)  $C_{sg1}$  characterizes the pure capacitance between the signal wire and ground for various types of CPWs, as shown in Figure 2(a). (2) C-R-C network is incorporated here to describe the EM coupling to the lossy substrate, as shown in Figure 2(b). (3) The C-L-R series path describes the EM coupling to the lower metal, especially for the CPWs with large lower ground or shield, e.g. GCPW and SCPW, as shown in Figure 2(c). The direct extraction method is shown in Figure 3.

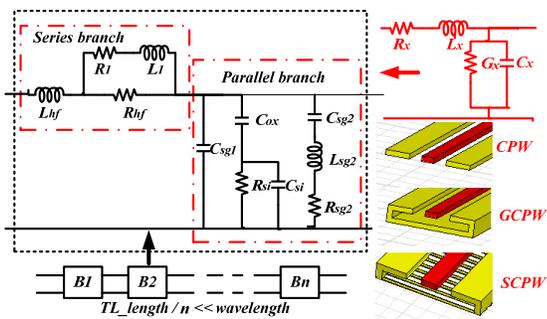


Figure 1: Proposed unified circuit model for CPWs.

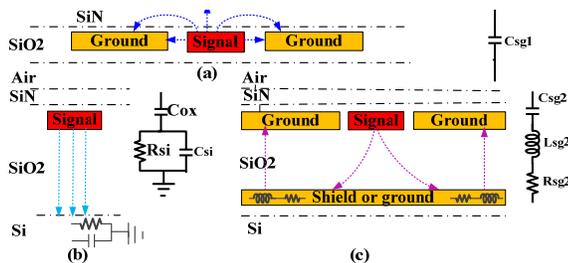


Figure 2: Parallel branch of the CPWs model, (a)  $C_{sg1}$  (b) C-R-C network, (c) C-L-R series path.

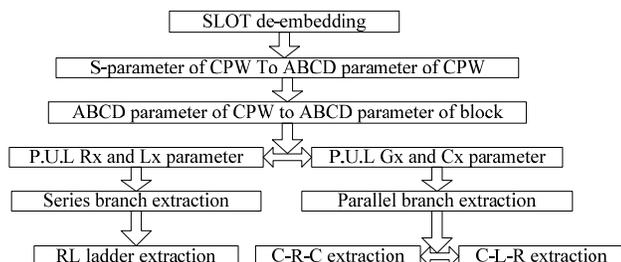


Figure 3: Proposed extraction flow of the CPWs model.

In order to verify the proposed unified CPWs model and direct extraction methodologies, various CPWs structures, including CPW and SCPW with lengths of  $400\mu\text{m}$  are designed and fabricated in 90nm IBM CMOS process. S-parameters are measured up to 67GHz. Short-Load-Open-Through (SLOT) technique [8] has been applied for parasitic de-embedding to acquire the intrinsic CPWs data. The measured per-unit-length  $R_x$ ,  $L_x$ ,  $G_x$ ,  $C_x$  parameters of different structures are compared with the proposed model in Figure 4. Excellent agreement between measured and simulated result is obtained with maximal  $C_x$  and  $L_x$  errors less than 5% over the entire frequency range up to 67GHz.

It can be seen that the per-unit-length inductance of SCPW is almost identical to that of CPW, while the per-unit-length capacitance is larger than that of CPW. Meanwhile, the per-unit-length conductance of CPW is larger than that of SCPW, which implies larger loss of the CPW than that of SCPW. These conclusions are in agreement with the results in [3].

S-parameter comparisons between measurements and simulated results of  $400\mu\text{m}$  CPWs with 10 cascaded blocks are shown in Figure 5. Excellent data agreement of the three structures is achieved over the entire frequency range up to 67GHz.

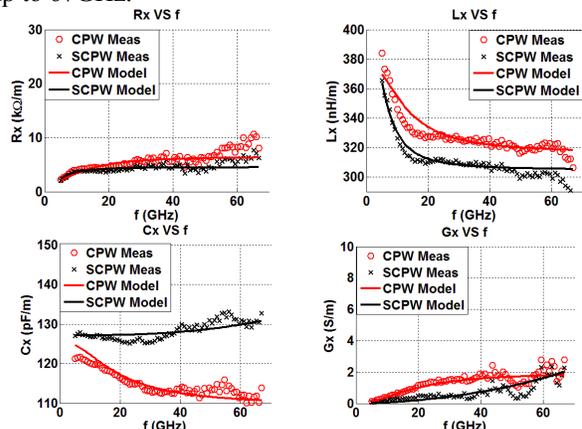


Figure 4: Comparisons between the experimental data (dot) and the modeled results (line) of per-unit-length resistance, inductance, capacitance, and conductance.

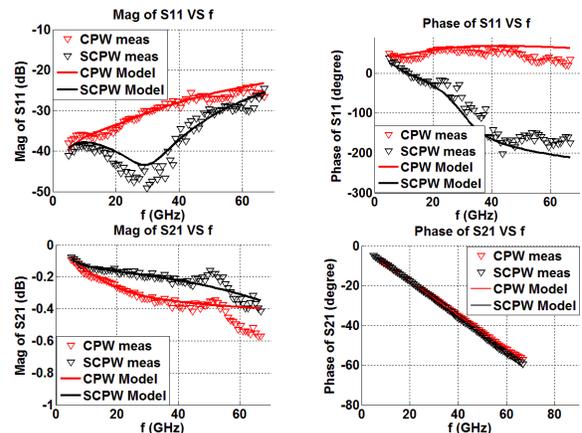


Figure 5: S-parameter comparisons between the measured data (dot) and the modeled results (line).

## 2.2 Transformer

The proposed model for transformer in this work is shown in Figure 6. Each metal spiral is considered as an inductor and can be modeled by a well developed  $2\text{-}\pi$  inductor model [10]. The DC inductances and resistances of the primary and secondary coils are modeled by  $L_{ij}$  and  $R_{ij}$  ( $i=p,s; j=1,2$ ) respectively.  $L_{ijsk}$  and  $R_{ijsk}$  ( $i=p,s; j=1,2$ ) are used to describe the skin effect and approximate effect. The inductive coupling between coils is represented by mutual inductances  $k_{p1s1}, k_{p1s2}, \dots, k_{p2s2}$ .

For MMW applications, parasitic distributive effects and substrate coupling become more and more significant. These effects should be described properly, which is the key to ensure the accuracy of the proposed model.

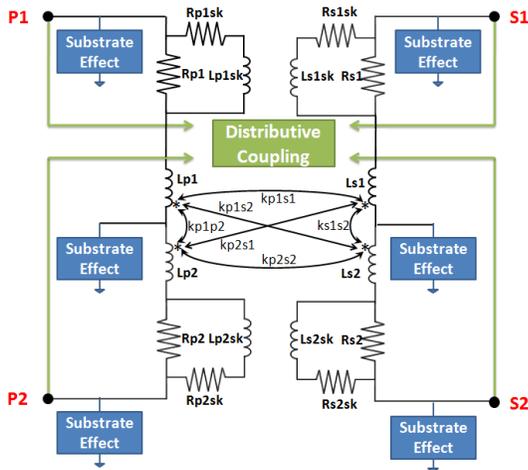


Figure 6: The basic topology of the proposed model.

To illustrate the distributed nature of the coupling capacitance between the two coils, a schematic view of inverting transformer is shown in Figure 7 (a). The gray segments are the metal spirals and the blue ones are the under-pass for the secondary spiral. Substrate parasitic effects and the mutual inductances are left out for simplicity. Coupling capacitance distributes along the metal tracks of two spirals. The cascaded multi-section equivalent circuit along the metal tracks of two spirals is adopted to approximate the distributed effects in Figure 7 (b). The coupling capacitors are highlighted in red and the possible signal coupling paths from primary port to secondary one are indicated by the blue lines with arrows.

In conventional model the capacitive effects are handled by only a single capacitor for simplicity as shown in Figure 7 (c). When frequency is not high, impedance of serial inductors and resistors is small compared to the coupling capacitors, all the distributed capacitors are in effect shorted together and can be modeled by a single lumped capacitor. Nevertheless, at millimeter band the impedance of serial inductance and resistance is comparable or even larger than that of capacitors. Furthermore, such capacitive coupling may allow the signal to flow directly from one port to the other without passing through the conductor coil. So the

distributive coupling between coils, which includes the parasitic coupling between different ports within the same coil and the distributed capacitive coupling across two coils, has to be reevaluated due to intensive distributed effects. At millimeter region, a simple but effective solution is to add an extra effective coupling inductance  $L_c$  and a resistance  $R_c$  in series with the effective coupling  $C_c$  to illustrate the impedance in distributed paths, as shown in Figure 7 (c). The improved equivalent circuit model will be reduced to the conventional one at low frequency and it also describes the impedance seen by signals more accurately at high frequency band.

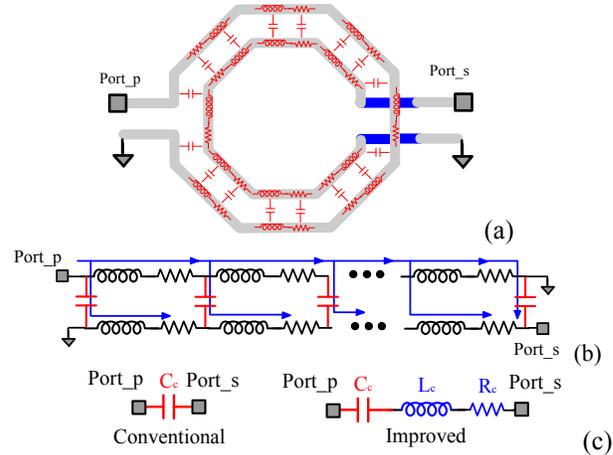


Figure 7: Schematic illustration of distributed effects in an inverting transformer (a) and its multi-section model (b) and compact equivalent circuits model (c).

Traditionally, the electromagnetic field coupling to the lossy Si substrate can be properly modeled by the circuit topology shown in Figure 8 (a). A patterned ground shield (PGS) which can be realized in standard silicon technologies is usually inserted between an on-chip metal coil and silicon substrate to improve isolation. However, with PGS, the impacts on inductance, parasitic resistance and capacitance have to be reevaluated. Here, another  $R\text{-}L\text{-}C$  network series are added to describe the electromagnetic field coupling to the lower ground metal, especially for the PGS network, as shown in Figure 8 (b).

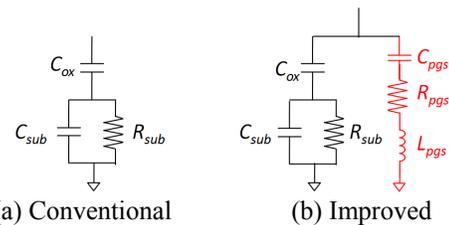


Figure 8. The lumped-circuit topology to describe the substrate effect.

The proposed compact model is validated by a 2:2 interleaved transformer with metal width = 8  $\mu\text{m}$ , metal spacing = 2  $\mu\text{m}$  and inner radius = 50  $\mu\text{m}$ , which is

fabricated using SMIC 65nm RF CMOS process.  $S$ -parameters are measured by Agilent N5247A VNA up to 67GHz. The de-embedded  $S$ -parameters are transformed to  $Y$ -parameter for model parameter extraction by the scalable characteristic ensured parameter extraction technique [11].

Figure 9 shows the measured results and the corresponding model extracted self-inductance, quality factor of the primary port and mutual-inductance between coils. It can be seen that the predicted inductance and quality factor are accurate for whole sampling frequency compared with measured value. The proposed model can cover multi-resonant frequency up to 67GHz accurately.

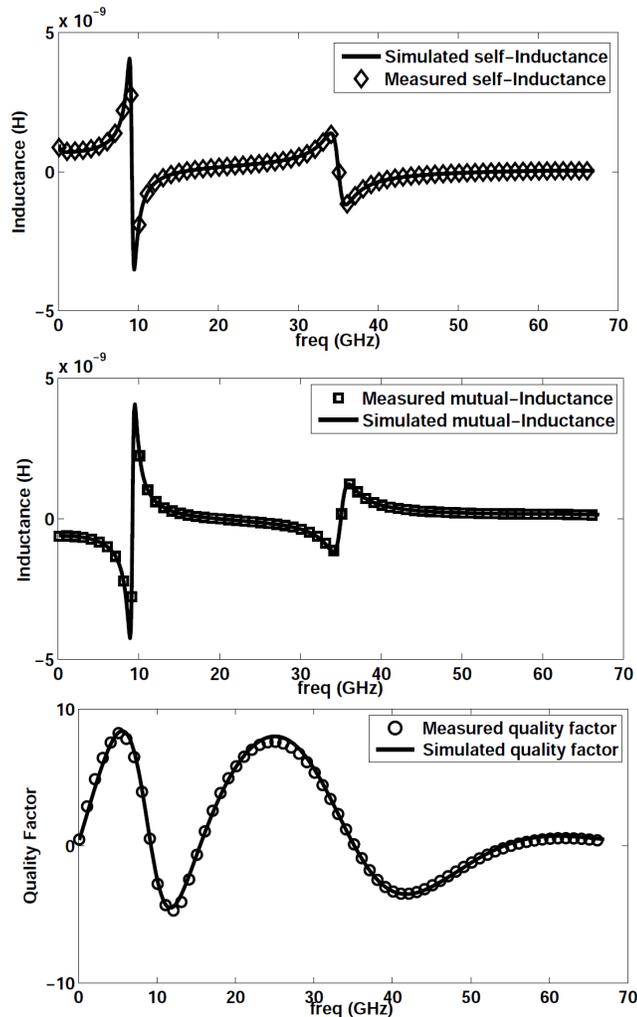


Figure 9. Comparison of measured and simulated self-inductance, mutual inductance and quality factor by the proposed model.

### 3 CONCLUSION

In this paper, high frequency compact models which can cover MMW frequency for passive elements are presented. For TL, as the parasitic effects are already taken into

account in the line model, we add a new C-L-R series path together with conventional branches to describe the EM coupling for CPWs with large lower ground and substrate. For transformer, both distributed effect and substrate coupling need to be modified.  $R-L-C$  series network is used to characterize the distributive and substrate coupling. Experimental results show the excellent agreement of the proposed compact model with measurement data up to 67GHz accurately.

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