Open and Short Circuits in Coplanar MMIC's

Klaus Beilenhoff, *Member, IEEE,* Harald Klingbeil, Wolfgang Heinrich, *Member, IEEE,* and Hans L. Hartnagel, *Fellow, IEEE*

***Abstract*— Coplanar MMIC stub configurations are investigated by means of the finite-difference method in the frequency domain. The open end, the short end, and a capacitively loaded short end (MIM-short) are analyzed. Their parasitic effects are described in terms of the effective length extension l_{ext}. The influence of the different line parameters is discussed and simple design rules are given.**

I. Introduction

TODAY, monolithic microwave integrated circuits (MM IC's) rely more and more on coplanar waveguides (CPW's) as the basic transmission line [1]. In contrast to the microstrip case, however, few modeling tools are available so far for simulating CPW structures. This applies also to open and short circuits which are of particular importance regarding phase-sensitive resonator structures and network-analyzer calibration sets. Hence, a reliable description is required that enables one to include their characteristics in the design process. Several papers have treated open and short circuits for CPW's [2]–[6]. Most of them, however, deal with geometries different from the typical MMIC case and, therefore, their results do not apply to the structures considered here.

The *open circuit* in coplanar MMIC's is commonly realized as shown in Fig. 1(a). With regard to the *short end,* two different types have to be considered: either a simple contact between center and ground conductor as illustrated in Fig. 1(b) or, if dc isolation of the center electrode is required, the so-called *MIM short* according to Fig. 1(c).

For the analysis, we employ a finite-difference method in the frequency domain [8]. It is a three-dimensional full-wave approach, which allows one to obtain directly the scattering matrix for nearly arbitrarily shaped waveguide discontinuities. In the following treatment, no losses are taken into account.

The information on the stub parasitics is associated with the phase of the reflection coefficient. Instead of referring to capacitances and inductances caused by the fringing fields at the discontinuity, the parasitic effects are described here in terms of an equivalent length increase of the homogeneous coplanar waveguide, the so-called effective length extension l_{ext}.

Two general comments regarding the modeling of discontinuities for coplanar MMIC's seem to be in order. First, the stubs introduce only relatively small parasitics, i.e., the l_{ext} values range far below the wavelength. That means one does not need an accuracy better than 10–20% since their influence on circuit design is one order of magnitude less than that of the homogeneous line parameters β, α, and Z_c. Second, it is very difficult to derive simplified descriptions by physically based considerations. Due to the full three-dimensional nature of the field at CPW discontinuities, models based on empirical values are mostly the only way to provide an efficient approximation for the electrical behavior of such structures.

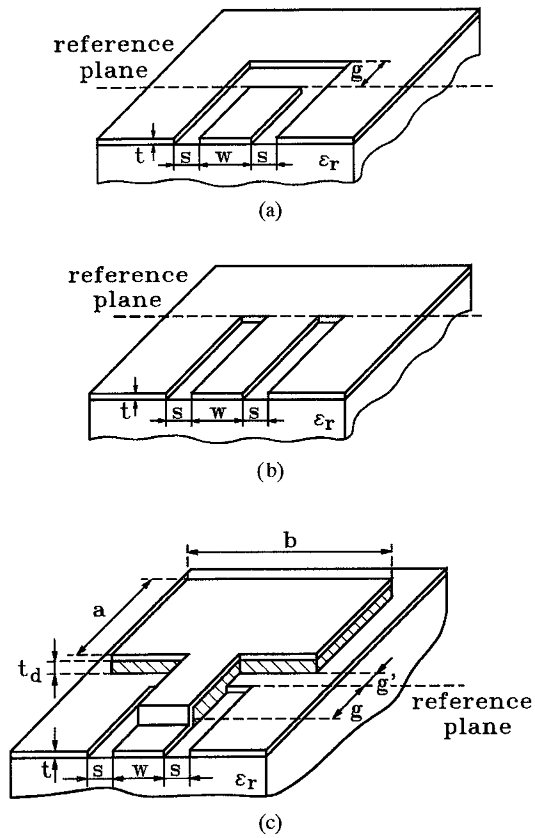

Fig. 1. The different structures considered: (a) open circuit, (b) short circuit, and (c) MIM-short circuit. (Unless otherwise specified the line parameters are chosen as follows: $d = w + 2s = 50\,\mu$m, $t = 3\,\mu$m, $\varepsilon_r = 12.9$, substrate thickness $h_s = 200\,\mu$m, $w = 10, 20$, and $40\,\mu$m for about 65, 50, and 30-Ω impedance, respectively.)

II. The Open

The MMIC open-circuit is presented in Fig. 1(a). Under the assumption that the length extension remains small compared to the wavelength on the CPW, l_{ext} can be described by the ratio between the end capacitance C_0 due to the fringing electric fields and the capacitance per unit length C' of the

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K. Beilenhoff, H. Klingbeil, and H.L. Hartnagel are with the Institut für Höchfrequenztechnik, Technische Hochschule Darmstadt, 64283 Darmstadt, Germany. W. Heinrich is with the Ferdinand-Braun-Institut für Höchstfrequenztechnik, 12489 Berlin, Germany.
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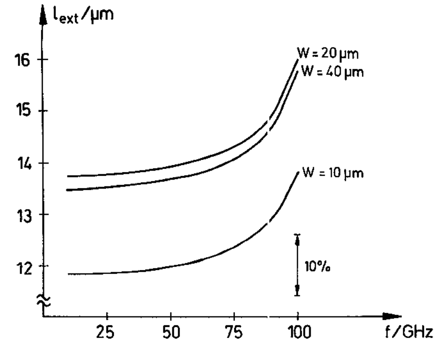

Fig. 2. Effective length extension l_{ext} of the open circuit (Fig. 1(a)) against frequency for $g = 15\,\mu$m.

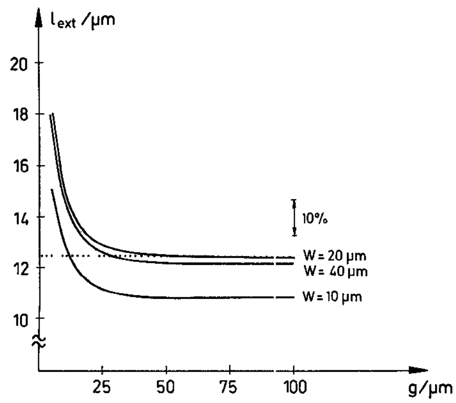

Fig. 3. Effective length extension l_{ext} of the open circuit against gap width g for a frequency of $f = 10$ GHz. (Dotted curve denotes value obtained by the design rule in (2).)

homogeneous waveguide:

$$l_{ext} = \frac{C_0}{C'} \,. \tag{1}$$

Fig. 2 shows the frequency dependence of l_{ext} for three different line geometries. Because the dimensions of the structure are small compared to the wavelength, dispersion is negligible. Also the metallization thickness t exerts only a small influence on l_{ext} as pointed out in [9]. Varying t from 0 to 5 μm, for example, leads to about 1% deviation only.

In Fig. 3 the influence of the spacing g is studied. The behavior can be explained by subdividing the end capacitance C_0 into two parts [3]. The capacitance due to the fringing field across the gap decreases with growing g while the capacitance due to the fringing field across the slot s is nearly independent from g. As can be seen from Fig. 3, the curves saturate for g values larger than d. Hence, it is sufficient to choose $g = d$ in order to minimize the parasitic length extension.

Summarizing the results for the open circuit, one can state that the l_{ext} values are in the 10-μm range. Metallization thickness t and gap width g (for $g \geq d$) do not influence l_{ext} significantly. Also, the dependence on frequency and on w/d is relatively weak. Roughly speaking, therefore, one has a length extension that only scales with d. This result is very interesting, because it enables one to provide a simple design rule, namely

$$l_{ext} \approx \frac{1}{4} \cdot d \,. \tag{2}$$

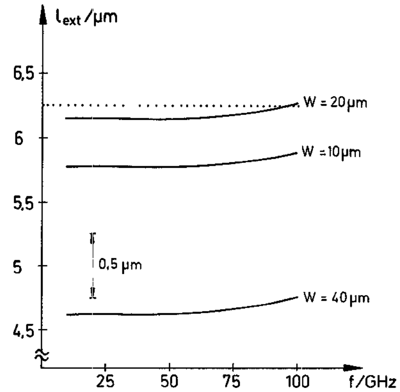

Fig. 4. Effective length extension l_{ext} of the short circuit (Fig. 1(b)) against frequency. (Dotted curve refers to the design rule in (4).)

As can be seen from the dotted curve in Fig. 3, (2) involves less than 1% error for a 50-Ω line considering a gap of $g \geq d$. For other line geometries, a deviation up to 20% is observed. Such an accuracy appears to be sufficient since the l_{ext} values themselves are small and, consequently, they can be classified as second-order parameters in circuit design.

Equation (2) is limited to $0.2 \leq w/d \leq 0.8$. It holds also for semiconductor substrates other than GaAs because the substrate permittivity affects the end capacitance C_0 and the line capacitance C' in the same way.

III. The Short

Fig. 1(b) illustrates the geometry of the short stub. Its length extension is defined by the end inductance L_0 related to the distributed line inductance L' of the homogeneous CPW:

$$l_{ext} = \frac{L_0}{L'} \,. \tag{3}$$

L_0 is caused by the current distribution in the discontinuity region. In Fig. 4, the behavior of the length extension is plotted against frequency. As observed already for the open circuit, dispersion is negligible due to the miniaturized line geometries. Quantitatively, the values for l_{ext} are about one half of those obtained for the open circuit.

As demonstrated in [9], the short exhibits a significant dependence on metallization thickness t. Increasing t from 0 to 3 μm causes about 15% deviation for the line geometries considered here. Due to the fact that the length extension again is small compared to the wavelength, however, the influence of t on l_{ext} can be neglected for a first-order modeling approach.

Hence, for typical MMIC dimensions a simple design rule can be extracted from the results presented here and those of [9]. It reads

$$l_{ext} \approx \frac{1}{8} \cdot d \,. \tag{4}$$

It is clear that this approximation (see also Fig. 4) only holds in the frequency range where no dispersion is observed and if the metallization thickness does not become too large ($t < s/3$). The permittivity ε_r of the substrate does not need to be included in this formula because under quasi-TEM conditions the magnetic field that dominates at the short is not affected by this parameter.

IV. The MIM-Short

In practical MMIC design, the simple short configuration of Fig. 1(b) may cause problems since it does not allow for different bias potentials on center and ground metallization. This restriction can be circumvented loading the short by a large metal–insulator–metal (MIM) capacitance (see Fig. 1(c)) that blocks dc and low-frequency signals and approximates an ideal short circuit at microwave frequencies.

Commonly, the MIM capacitance consists of an approximately 0.2-μm-thick siliconnitride dielectric layer covered with a metallization that is connected to the center strip by an air-bridge technique. The lateral dimensions of the capacitance depend on the frequency range of operation. In the following, we assume a top metallization area of 100 μm × 100 μm, which leads to a value of $C_{\text{MIM}} = 2.6$ pF for a permittivity of $\varepsilon_r = 6$. Originally, the structure considered was designed for a frequency of $f = 60$ GHz.

The electrical behavior of the MIM short can be described by an end inductance L_0 connected with the MIM capacitance C_{MIM} in series. Together with the line inductance L', the corresponding effective length extension of the MIM short reads

$$l_{ext} = \frac{L_0}{L'} - \frac{1}{\omega^2 L' C_{\text{MIM}}}. \tag{5}$$

In contrast to the other two stubs discussed before, l_{ext} here can also reach negative values due to the capacitive loading. Hence, a zero length extension is possible when adjusting C_{MIM} appropriately.

Fig. 5 presents the frequency dependence of l_{ext} obtained from the three-dimensional field calculations. At frequencies lower than $f < 40$ GHz the assumption that the MIM capacitance forms a very low impedance no longer holds. The capacitance more and more dominates the behavior shifting l_{ext} to negative values. Increasing the frequency above 80 GHz, on the other hand, the length extension becomes independent of the MIM capacitance. Evaluating the results of Fig. 5, one finds an l_{ext} value similar to that of the simple short stub described in Section III. Hence, in a first approximation the term L_0/L' in (5) can be substituted by the value given in (4) ($L_0/L' \approx d/8$). The MIM capacitance can be calculated with good accuracy by the common parallel-plate formula. Thus, one has

$$l_{ext} \approx \frac{d}{8} - \frac{1}{\omega^2 L' C_{\text{MIM}}}$$

with

$$C_{\text{MIM}} = \varepsilon_{r\text{MIM}} \cdot \frac{ab}{t_d}. \tag{6}$$

On the one hand, the value of C_{MIM} should be as small as possible in order to keep the circuit size low. On the other hand, C_{MIM} must be large enough because otherwise the short-circuit characteristics deteriorate. More precisely, C_{MIM} should fulfill the condition

$$\frac{1}{\omega C_{\text{MIM}}} \leq \frac{1}{50} \cdot Z_c \tag{7}$$

where Z_c denotes the CPW characteristic line impedance.

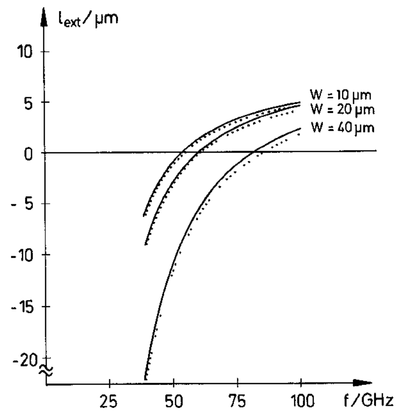

Fig. 5. Effective length extension l_{ext} of the MIM-short circuit (Fig. 1(c)) against frequency. (Dotted curve: design rule of (6).)

V. Conclusions

Typical MMIC stubs have been analyzed systematically by means of a full-wave three-dimensional finite-difference method in the frequency domain. The investigations point out that the parasitics associated with these circuit elements can be approximated by simple design rules. The main results can be summarized as follows.

• Generally, all stubs show length extensions below 20 μm which is small compared to the wavelength. Therefore, the resulting influence on circuit performance is limited, roughly speaking they contribute only second-order effects.

• Due to the miniaturized dimensions, dispersion effects remain small as well. They cause deviations in the 15% range up to 100 GHz.

• For the *open circuit,* one has a length extension of approximately $l_{ext} \approx d/4$ with d being the ground-to-ground spacing of the line. All other geometrical parameters can be neglected for a gap g larger than d.

• The *short circuit* seems to be the better choice because it causes smaller l_{ext} values of about $l_{ext} \approx d/8$. Here, l_{ext} depends on metallization thickness t to a larger extent than in the open-circuit case. If bias isolation is required, the *MIM short* is to be used. Designing the MIM capacitance properly, the resulting electrical behavior closely resembles that of the simple short.

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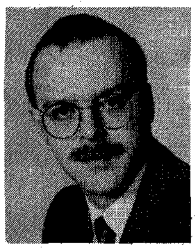


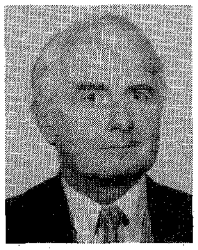
Klaus Beilenhoff (M'90), was born in Soest, Germany, in 1963. He received the Dipl.-Ing. degree in electrical engineering from the Technical University of Darmstadt, Germany, in 1989.

Since then he has been working at the Institut für Hochfrequenztechnik of the same university, where he is engaged in numerical computation of electromagnetic fields applied to MMIC structures.

Harald Klingbeil, was born in Frankfurt, Germany, in 1968. He received the Dipl.-Ing. degree in electrical engineering from the Technical University of Darmstadt, Germany, in 1992.

Wolfgang Heinrich (M'84), was born in Frankfurt, Germany, in 1958. He received the Dipl.-Ing., Dr.-Ing. and habilitation degrees in 1982, 1987 and 1992, respectively, all from the Technical University of Darmstadt, Germany.

In 1983, he joined the staff of the Institut für Hochfrequenztechnik of the same university, where his primary interests included field-theoretical analysis and simulation of planar transmission lines and FET noise-modeling. In April 1993, he joined the Ferdinand-Braun-Institut at Berlin, Germany, as head of the department for computer-aided design.



Hans L. Hartnagel (SM'72–F'92), was born in Germany, studied for the "Diplom-Ingenieur" at the Technical University of Aachen, Germany, and received the Ph.D. and Dr. Eng. from the University of Sheffield in England.

He worked in Germany, France, England, Japan and the USA, partly in industry, partly at universities or government Laboratories. From January 1971 to September 1978 he was a Professor of Electronic Engineering at the University of Newcastle upon Tyne in England and since then has been Professor of High-Frequency Electronics at the Technical University in Darmstadt, Germany. His speciality is compound semiconductor technology, devices and their circuits and he held many consulting positions.