Resonance Phenomena and Power Loss in Conductor-Backed Coplanar Structures

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***Abstract*—Resonances and associated power loss which occur in conductor backed coplanar transmission lines (CPWs) at selected frequencies have been investigated by experiment and by three-dimensional (3-D) EM simulations from 1 to 230 GHz. It is found that a substantial part of the incident power is lost by reflection and radiation. The resonance frequencies are predictable by patch antenna theory.**

***Index Terms*—Conductor-backed CPW, coplanar, CPW, millimeter-wave, modes.**

I. Introduction

INTEGRATED circuits based on coplanar transmission lines (CPWs) [1], [2] are attractive because there is no need for vias, since the ground planes are on the chip surface. However, when coplanar circuits are packaged, inadvertently more ground planes are added, for instance, when the chip is soldered down to a package for efficient heat sinking. An additional ground plane in close proximity to the fields of the coplanar line, may support unwanted modes. The excitation of modes other than the coplanar mode and their suppression or elimination has been the subject of extensive studies in the past [3]–[12], however generally confined to microwave frequencies. It is the purpose of this study to contribute new experimental and simulation results obtained at frequencies to 230 GHz, a range where increasingly more GaAs- and InP-based coplanar HEMT MMICs find application [13], [14].

II. Results

From Fig. 1, we make the observation that a symmetric coplanar line with two wide ground planes as shown, when placed on metal, represents two parallel plate resonators or patch antennas, coupled capacitively to the center conductor of the coplanar line via the gaps s. The theory of rectangular patch antennas is well documented [15], [16]. If w_g and l_g are the width and the length of the rectangular patch, respectively, the resonant frequencies are

$$f_{mn} = \frac{c}{2\sqrt{\varepsilon_r}}\left[\left(\frac{m}{w_g}\right)^2 + \left(\frac{n}{l_g}\right)^2\right]^{0.5} \quad (1)$$

where c is the speed of light, ε_r is the and relative dielectric constant of the substrate, and m and n are integers.With the patches not shorted at their periphery, the magnitude of the electric field E_z normal to the plane of the patch varies as cosine functions in x and y with peaks in E_z along its periphery, the slot and within the patch, depending on the dimensions of the patch. The number of modes, given by (1), increases rapidly with frequency. Here we compare experimental and simulated data generated over a wide frequency range from 1 to 230 GHz. The experimental as well as the simulated data, for a semi insulating GaAs chip with dimensions $W = 2$ mm, $L = 4$ mm and $h = 0.1$ mm, when placed on metal, is shown in Fig. 2. The individual modes are well resolved in the range 40–100 GHz, corresponding to the group $m = 1$ and varying n. The modes are not resolvable above about 120 GHz. The three groups shown correspond to the integers $m = 1$, 2, and 3. Table I lists the measured, the simulated, and the calculated [(1)] resonant frequencies in the range 1–120 GHz, where we observe excellent agreement. A symmetrical coplanar line with actual dimensions of $w = 17$ μm, $s = 16.5$ μm, $w_g = 0.925$ mm and $l_g = 3.9$ mm was used. For the frequency range 1–120 GHz, commercially available equipment, the HP network analyzer system HP8510XF and CASCADE ACP110C probes, were used. The results in the range 70–230 GHz were obtained with an in-house developed set of two active probes [17], together with our measurement software and an HP 8510C network analyzer. Reasonably good agreement exists over the range 70–120 GHz, the range of overlap of the two systems.

TABLE I
($W = 2$ mm, $L = 4$ mm, $w_g = 0.925$ mm, $l_g = 3.9$ mm, $h = 0.1$ mm, $\varepsilon_r = 12.9$, Frequencies in GHz)

f_{mn}	f_{mn} Equ. 1	f_{mn} measured	f_{mn} simulated
f_{11}	46.4	45.3	43
f_{12}	50.0	49.3	49
f_{13}	55.4	54.3	55
f_{14}	62.2	62.3	62
f_{15}	70.0	69.7	70
f_{16}	78.5	78.6	79
f_{17}	87.5	85	87
f_{18}	96.8	97.3	98
f_{26}	110.8	109.7	111
f_{27}	117.4	117.7	118
f_{28}	124.5	125.7	126
f_{29}	132.1	132.3	132

The three-dimensional (3-D) EM simulation was performed with the commercial software HP HFSS. The sample was treated as an open structure with a radiation surface surrounding it. Two coplanar probes were simulated at the input and output. Lossless materials were used to reduce computation time and to determine the resonance frequencies more

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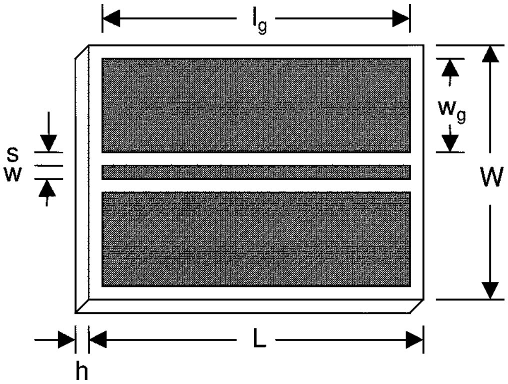

Fig. 1. Coplanar transmission line on a substrate.

precisely. The results of the simulation for lossless and zero thickness metallization and lossless dielectrics is shown in Fig. 2 (bottom).

The simulated electric field (magnitude) in the center of the substrate (50 μm below the ground plane) for $f_{15} = 70$ GHz is shown in Fig. 3, where arrows indicate its direction. The signal enters the chip at the lower end via small coplanar probes. The electric field distribution for each ground plane represents the $m = 1, n = 5$ mode of (1). Excitation of the parallel plate resonance occurs by the electric field across the slot s of the travelling coplanar wave and thus at several points along the ground plane. The electric field below the ground planes is stationary, in contrast to the field along the coplanar line which is a travelling field. This parallel plate resonance or patch antenna mode differs from the microstrip like mode which is characterized by travelling fields.

Shorting the ground planes of the chip to the metal plane at the bottom of the chip along its perifery of the chip, does not suppress the resonances. Instead, a new field pattern is generated with the new boundary condition $E_z = 0$ along the periphery. The new electric field distribution now varies as a sine function in x and y. The result is merely a change in the resonance frequencies.

Conducting via holes, especially if placed at the points of maximum electric field, are an effective method of mode suppression. From Fig. 3, they would suppress the resonance at 70 GHz if placed along the outer and inner edges of the ground planes. We have found by simulation that resonances are not excited if vias are placed sufficiently close only along the inner edges of the ground planes along the coplanar line, consistent with observations made in [8]. Vias placed 500 μm apart (center to center) will suppress resonances to 150 GHz in a quartz substrate and to 90 GHz in a GaAs substrate. However the use of via holes greatly eliminates the simplicity and attractiveness of the CPW technology.

An analysis of the impedance of the CPW under study is illustrated in Fig. 4. The simulation results of reflected ($|S_{11}|^2$) and transmitted ($|S_{21}|^2$) power are shown, together with the real and imaginary input impedance Z_{11} which assumes low values at the three selected resonant frequencies f_{14}, f_{15}, and f_{16}. The term $(1 - |S_{11}|^2 - |S_{21}|^2)$ represents the power lost. The radiated power with associated radiation pattern at the frequency $f_{15} = 70$ GHz is shown in Fig. 5. About 40% of the incident

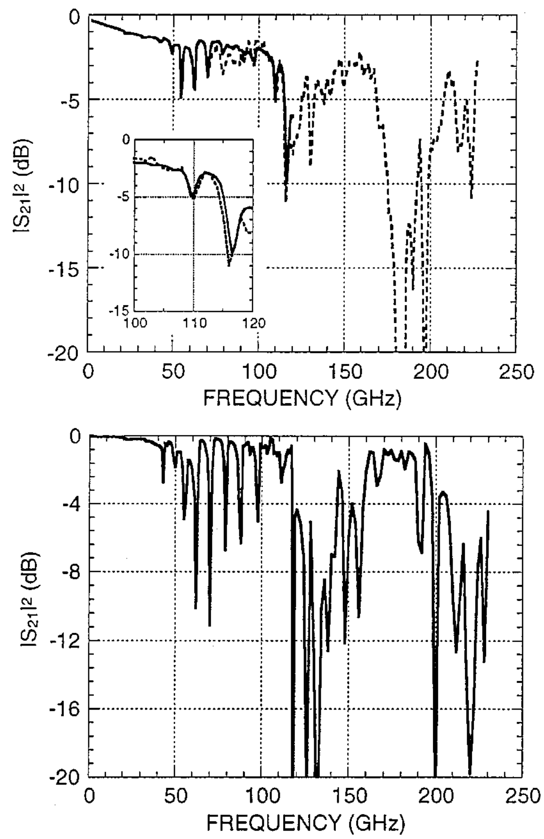

Fig. 2. Measured (top) and simulated (bottom) transmission $|S_{21}|^2$ of a $2 \times 4 \times 0.1\ \mathrm{mm}^3$ conductor-backed CPW on GaAs, measured over the two frequency ranges 1–120 GHz (solid) and 70–230 GHz (dotted).

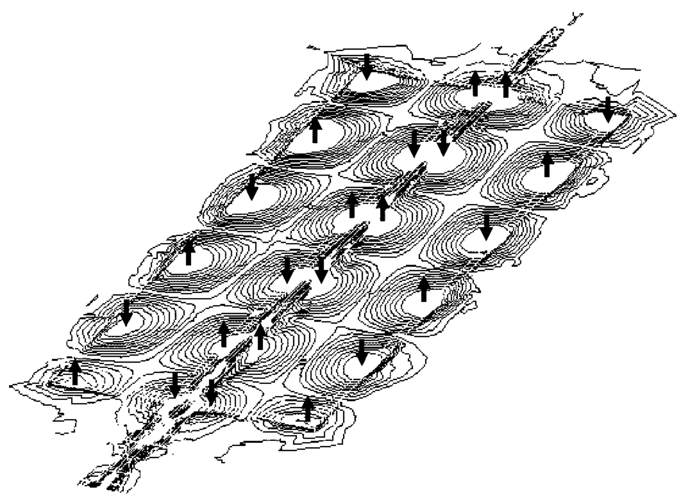

Fig. 3. Simulated electric field (magnitude) pattern, 50 μm below the surface of the chip. The pattern under each ground plane corresponds to the $m = 1$, $n = 5$ mode.

power is radiated, about 50% is reflected, and about 10% is transmitted.

An extensive study of line and substrate parameters was undertaken to determine the effect on the resonances and associated power loss. These simulations clearly show that the resonances are determined solely by the dimension of the resonator formed by a ground plane and the backside metallization and are not related to the dimensions of the dielectric substrate.

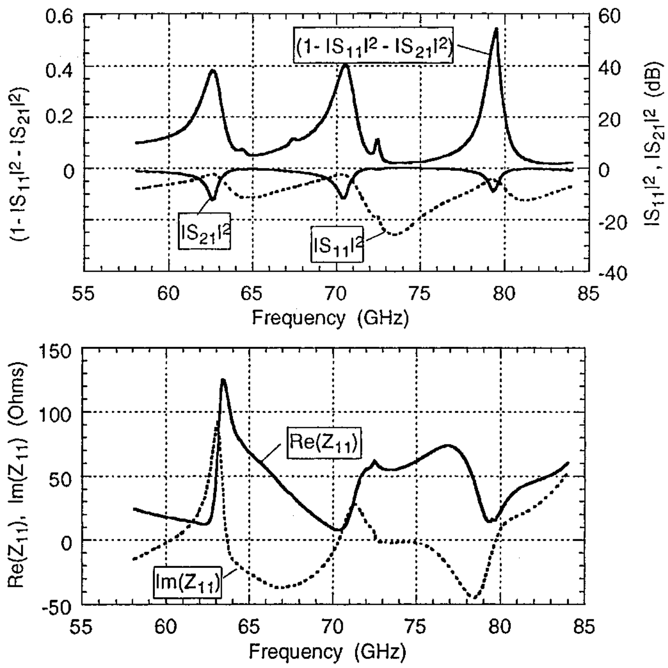

Fig. 4. Simulated power lost, transmitted and reflected power (top), and simulated input impedance (bottom), over the range 58–84 GHz.

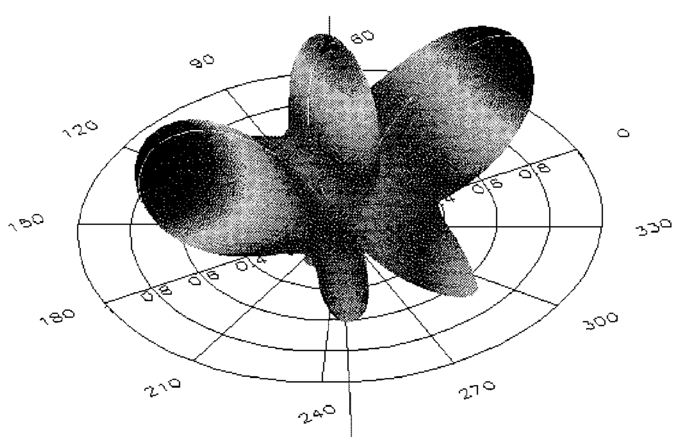

Fig. 5. Simulated radiation pattern at 70.5 GHz. Total power radiated is 40% of the incident power.

Small ground dimensions lead to the well known "finite ground" coplanar line [18], [19], where resonant frequencies generally lie above the operating frequency, resulting from the small dimension of w_g, as predicted by (1).

III. Summary

The ground planes in conductor backed CPWs behave like parallel plate resonators, capacitively coupled to the CPW. Their resonance frequencies f_{mn} are determined by their dimension according to patch antenna theory. At resonance, transmission is reduced by reflection and radiation, and a feedback is provided within the substrate which may cause amplifiers to be unstable [20]. The resonances may be eliminated by reducing at least one of the ground plane dimensions, such as w_g, causing f_{mn} to lie above the operating frequency. Other methods for the reduction of resonances, such as the use of conducting via holes or the total or partial elimination of the backside ground plane, have been simulated.

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