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Finite Ground Coplanar (FGC) Waveguide: It's Characteristics and Advantages for use in RF and Wireless Communication Circuits

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Abstract

To solve many of the problems encountered when using conventional coplanar waveguide (CPW) with its semi-infinite ground planes, a new version of coplanar waveguide with electrically narrow ground planes has been developed. This new transmission line which we call Finite Ground Coplanar (FGC) waveguide has several advantages which make it a better transmission line for RF and wireless circuits. Since the ground planes are electrically narrow, spurious resonances created by the CPW ground planes and the metal carrier or package base are eliminated. In addition, lumped and distributed circuit elements may now be integrated into the ground strips in the same way as they traditionally have been integrated into the center conductor to realize novel circuit layouts that are smaller and have less parasitic reactance. Lastly, FGC is shown to have lower coupling between adjacent transmission lines than conventional CPW.

Key words: coplanar waveguide (CPW), finite ground coplanar waveguide, transmission lines, coupling

Introduction

Transmission lines are the most basic microwave circuit element in RF and wireless systems. They are required for interconnecting the individual electrical elements together that comprise a Monolithic Microwave Integrated Circuit (MMIC), for interconnecting MMICs within microwave MultiChip Modules (MCMs), and for interconnecting MCMs together with other microwave components such as antennas to construct RF systems. In addition, filters, couplers, power dividers, tuning stubs, matching networks, and other critical RF system components are all constructed by connecting together transmission lines with different propagation characteristics. While no single transmission line is suitable for this wide variety of tasks, coplanar waveguide (CPW) has been widely used for many of these applications.

Coplanar waveguide consists of a center strip conductor and a pair of semi-infinite ground planes on either side of the center conductor as shown in Figure 1a; this entire structure is on the same surface of an electrically thick dielectric substrate [1]. Therefore, both series and shunt connection of circuit elements is possible without the use of metal filled via holes as required for microstrip and stripline. This reduces the fabrication time, lowers the fabrication costs of MMICs on semiconductor substrates such as GaAs and InP by as much as 30 percent, eliminates the need for wafer thinning and back side polishing which reduces wafer breakage, and eliminates the parasitic inductance associated with via holes. The electrical characteristics of CPW are also good. The frequency variation of the effective permittivity, ε_{eff}, of CPW is lower than for microstrip which simplifies broadband circuit design, and a large variation in characteristic impedance is obtainable by varying the strip and slot widths [2]. Flip-chip bonding of MMICs in MCMs is simplified when both the MMIC and MCM employ CPW because all of the interconnects are on the same plane [3]. Thus, CPW circuits cost less, work better, and are easier to fabricate than circuits based on microstrip and stripline.

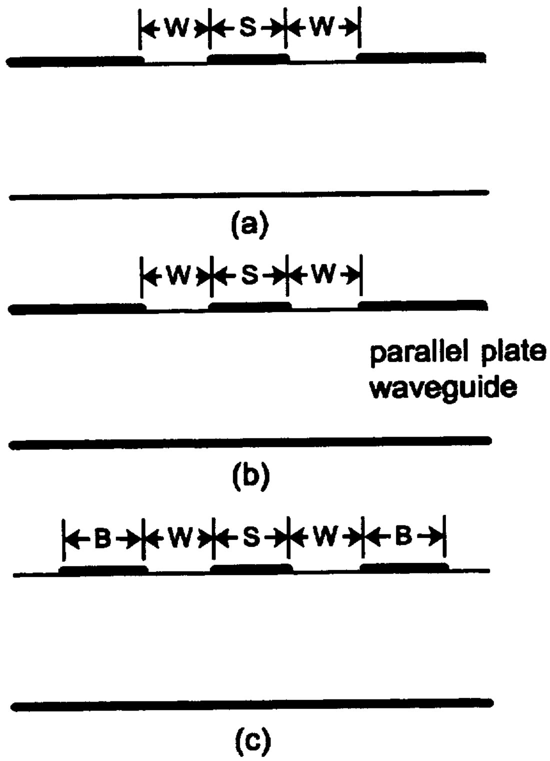


Figure 1: (a) Conventional coplanar waveguide (CPW), (b) Coplanar waveguide with lower ground plane, (c) Finite Ground Coplanar (FGC) waveguide.

However, CPW is not without problems. Typically, MMICs and MCMs incorporate a metal base to facilitate heat removal, provide electromagnetic shielding, and provide mechanical strength. Furthermore, MMIC substrates are thinned to between twenty five and one hundred microns to further aid thermal management. Thus, in practice, the ideal CPW shown in Figure 1a has an additional lower ground plane as shown in Figure 1b which creates a parallel plate waveguide region between the upper and lower ground planes. Since the parasitic parallel plate waveguide mode has a lower phase velocity than the CPW mode over the entire frequency spectrum, power leaks from the CPW mode to the parallel plate waveguide mode [4, 5]. Besides resulting in increased radiation attenuation, energy in the parallel plate waveguide mode creates box type resonances when the dimensions of the MMIC or RF circuit are greater than $\lambda_d/2$ where λ_d is the wavelength in the dielectric medium [6,7].

To solve these problems, several alternatives have been reported. Microwave absorbing materials can be used to attenuate the microwave energy at the edges of the substrate [8], multiple dielectric layers can be used to increase the phase velocity of the Transverse ElectroMagnetic (TEM) parallel plate waveguide mode which reduces coupling between the two modes [4], or metal filled via holes can be used to short the upper and lower ground planes together and thus eliminate the TEM parallel plate waveguide mode [7]. Unfortunately, each of these solutions results in higher fabrication costs and greater system complexity.

An alternative approach has been reported in the literature [9,10] and developed by the authors that uses finite width ground planes so that the total width of the transmission line is electrically small. Therefore, the parallel plate waveguide mode is not established and the problem of spurious resonances is eliminated. This new transmission line is called Finite Ground Coplanar (FGC) waveguide and is illustrated in Figure 1c. In this paper, the methods used to analyze FGC, FGC waveguide propagation characteristics, and how FGC may be used to improve RF and wireless circuits are presented.

Methods of Analysis

The authors have used experimental measurements, a two Dimensional - Finite Difference Time Domain (2D-FDTD) analysis, and conformal mapping to characterize FGC waveguide. For the experimental measurements, FGC circuits were fabricated on double side polished Si wafers with a resistivity of 2500 Ω-cm and a thickness, H, of 411 μm. Two fabrication procedures were used to define the metal lines; a lift-off process consisting of 0.02 μm of Ti and between 1.0 and 1.5 μm of evaporated Au and a Au plating procedure which results in a metal thickness of 3 μm were used. To suppress the parasitic slotline mode that may exist on all coplanar waveguide structures, airbridges spaced 2000 μm apart were used when discontinuities in the FGC waveguide would establish the parasitic mode.

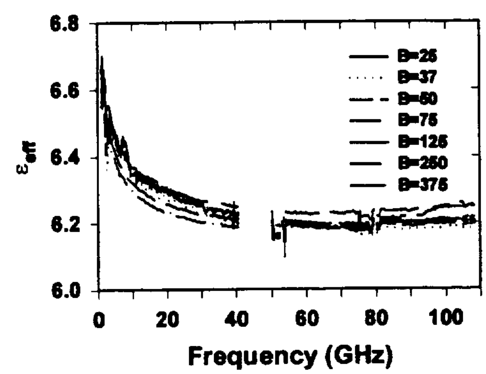
Measurements are made using a vector network analyzer and microwave probes with 150 μm pitch. Calibration of the measurement system is accomplished through a full Through-Reflect-Line (TRL) calibration [11] using calibration standards fabricated on the wafer with the FGC test circuits. This calibration procedure places the reference planes at the center of the through line and establishes a reference impedance equal to the characteristic impedance of the delay lines. The calibration standards consist of a through line, four delay lines, and a short circuit terminated FGC line to cover the measured frequency range. The same probe pad layout is used for the circuits or transmission lines under test so that the mismatch it creates may be accurately removed from the results. Lastly, measurements made for circuits without a backside ground plane are placed on a quartz spacer to isolate the Si substrate from the wafer chuck.

Theoretical analysis of FGC waveguide with a 2D-FDTD method [12] is employed as outlined in [13]. In all simulations, the discretization cell has dimensions of 2.5 μm in the horizontal direction and 25 μm in the vertical dimension, see Figure 1c, and the time step takes the value of the Courant limit. The computational domain is terminated to the left, to the right, and to the top (see Figure 1c) by eight cells of Perfectly Matched Layer (PML) absorber yielding a maximum theoretical reflection coefficient, ρ_{max}, of 10^{-5}. Without loss of generality, the CPW mode is excited in the FGC line by applying a horizontal odd electric field across the two gaps between the ground planes and the center conductor. The horizontal, parallel to the dielectric interface, electric field is probed at symmetrical observation points across the FGC waveguide which permits the determination of the CPW and the slotline mode characteristics.

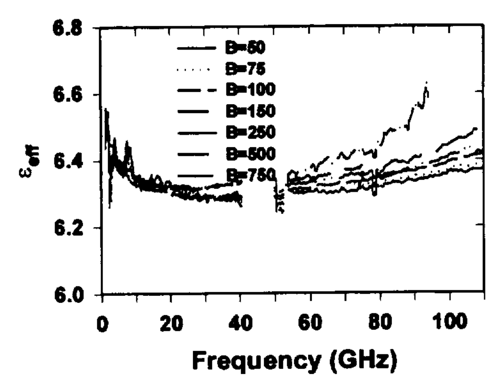
FGC Waveguide Propagation Characteristics

As already stated, the frequency variation of ε_{eff} for CPW is very low. Furthermore, since the characteristic impedance of CPW is dependent on the ratio k=S/(S+2W) [1] and independent of H, wide lines with low conductor loss can be used to realize most characteristic impedances regardless of the substrate parameters. Therefore, before FGC waveguide can be used in place of CPW, it must be proven to have low attenuation and a slowly varying effective dielectric constant as a function of frequency. To investigate the propagation characteristics of FGC lines, two sets of test circuits were fabricated with a center conductor and slot width, S and W respectively, of S=W=25 and S=W=50 μm, a metal thickness of 1.3 μm, and the ground plane width, B, varied between S and 15 S [13], [14].

The measured effective dielectric constant is shown in Figures 2a and 2b for lines with S and W of 25 and 50 μm respectively. For both sets of lines, ε_{eff} varies by less than one percent as the normalized ground width, B`=B/S, varies from one to fifteen for low frequencies which shows that $\varepsilon_{eff}$ is not dependent on B`. Furthermore, while ε_{eff} is nearly frequency independent at low frequencies, the two lines behave differently at higher frequencies. For the line with the narrower strip and slot width, the line is nondispersive until the total line width approaches $\lambda_d/2$, while the lines with wider strip and slot width become dispersive when the total line width approaches $\lambda_d/4$. Generally, dispersion increases when the propagating mode couples to a parasitic mode [15, 16]. Thus, for these two structures, there must be different parasitic modes that are influencing the FGC waveguide. For the wide lines, the parasitic mode is a microstrip mode which is stronger when H/(S+2W) is small [17] while the narrow lines are influenced more by the parallel plate waveguide mode. Further evidence of the parallel plate waveguide mode was seen by the presence of distinct resonances when the total width of the line approaches $\lambda_d/2$. Results for lines above this limit are not presented to maintain clarity in the figures. If the criteria that the total line width must be less than $\lambda_d/4$ is used, ε_{eff} varies by less than five percent over the frequency range of 1 to 110 GHz.

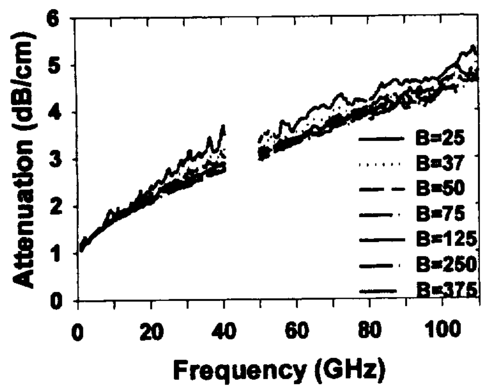


(a)

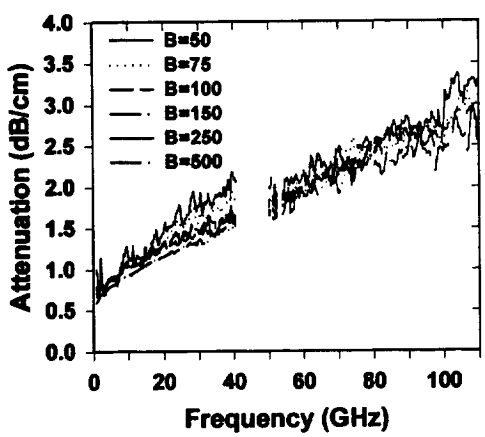


(b)

Figure 2: Measured effective dielectric constant of FGC waveguide as a function of frequency and ground plane width for (a) S=W=25 μm and (b) S=W=50 μm.



(a)



(b)

Figure 3: Measured attenuation of FGC waveguide as a function of frequency and ground plane width for (a) S=W=25 μm (b) S=W=50 μm.

Figures 3a and 3b show the measured attenuation in dB/cm for the two lines. The attenuation decreases as the normalized ground width increases until it exceeds two, at which point the attenuation as a function of B` saturates. This is shown by plotting the attenuation of finite ground lines normalized to the attenuation of conventional CPW with the same strip and slot width as shown in Figure 4. Notice that the attenuation is high for very narrow ground plane strips but it quickly becomes comparable to the attenuation of CPW as B` increases. Furthermore, it is found that at low frequencies, there is no measurable difference in attenuation as a function of the ground plane width. As a basis of comparison, Figure 4 also shows the ratio of attenuation between FGC and CPW predicted using conformal mapping [13] and closed form equations for conductor loss [18]. The data in Figure 3 also shows that the attenuation has a $f^{0.5}$ dependence when B'<2, and when B'>2, the frequency dependence increases to $f^{0.68}$. This implies that conductor loss dominates for narrow ground plane lines and that radiation loss increases as the ground plane width increases.

Based on these results, FGC waveguide has low attenuation, low dispersion, and no parasitic resonances if the ground plane width is at least twice the center conductor width and the total line width is less than $\lambda_d/4$. Since the width of the center conductor of CPW is typically less than $\lambda_d/10$ to maintain a quasi-TEM mode [5], it follows that the ground planes for FGC are also electrically narrow if B=2S. Thus, the ground planes of FGC may be thought of as two strips in parallel on either side of the center strip of the CPW, or FGC waveguide is a variation of coplanar stripline. With this view in mind, the ground strips may be used to implement circuit elements in the same way as they are currently implemented in the center conductor of CPW.

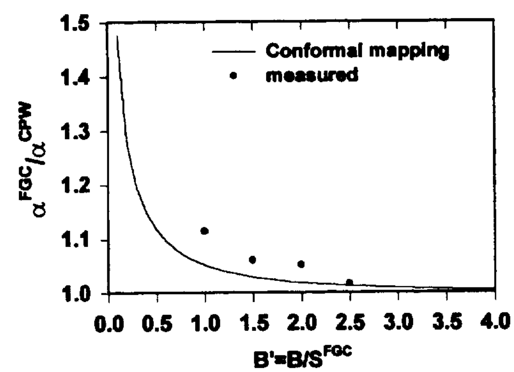


Figure 4: Attenuation of FGC waveguide with S=W=25 μm normalized to the attenuation of conventional CPW with S=W=25 μm as a function of the normalized ground plane width.

Passive Circuit Elements in FGC

To demonstrate the utility of using the ground strips of FGC waveguide, we consider in this section the integration of a NiCr thin film resistor and a Si_3N_4 Metal-Insulator-Metal (MIM) capacitor [19]. As discussed in the prior section, these circuit elements may be placed in either the center conductor or the ground strips as shown in Figures 5 and 6. Notice that symmetry is maintained when elements are placed in the ground strips to avoid exciting the parasitic slotline mode. Both of the elements were characterized over the frequency band of 1 to 40 GHz using a vector network analyzer and microwave probes, and from the measured data, the equivalent circuits shown in Figures 7 and 8 are developed.

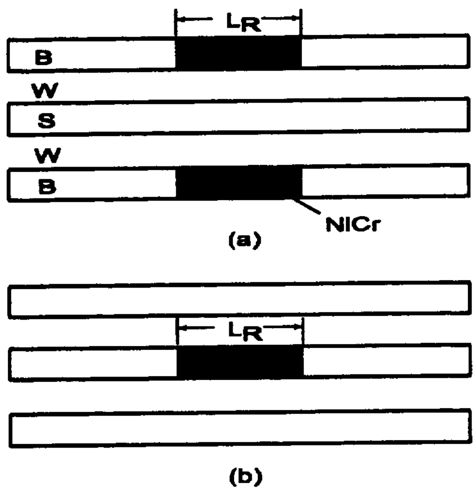


Figure 5: Schematic of thin film resistor in the (a) ground planes and (b) center conductor of FGC waveguide.

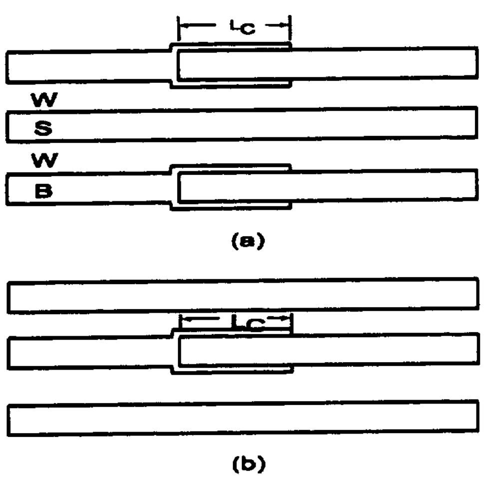


Figure 6: Schematic of MIM capacitor in the (a) ground planes and (b) center conductor of FGC waveguide.

To model the thin film resistors, the parasitic reactance is modeled by a pair of equivalent shunt capacitors to ground and a series inductance as shown in Figure 7. The equivalent circuit resistance as a function of the resistor length is shown in Figure 9 where it is seen that the DC measured resistance of the resistor placed in the center strip is twice as large as the same length resistor placed in the ground planes, while the RF determined resistance of the center strip resistor is approximately three times larger than the ground resistor. Furthermore, it is found that the parasitic reactances are independent of the placement of the resistor which is interesting since the associated inductance is expected to vary the same as the resistance values.

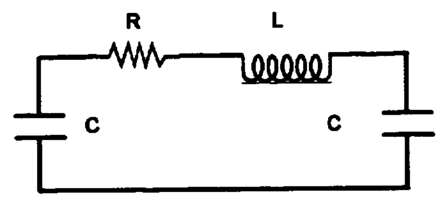


Figure 7: Equivalent circuit model of thin film resistor in FGC waveguide.

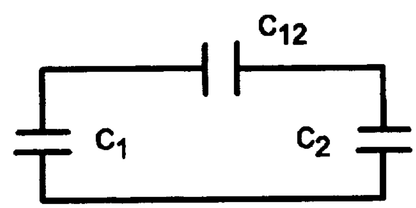


Figure 8: Equivalent circuit model of MIM capacitor in FGC waveguide.

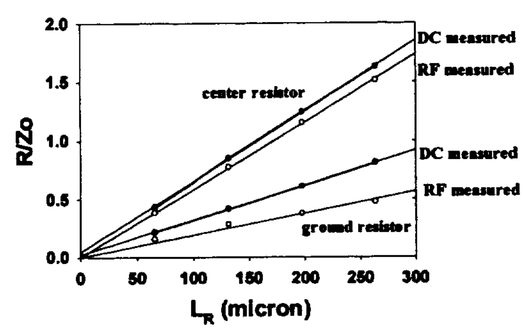


Figure 9: Measured resistance values of thin film resistor in FGC waveguide as a function of the resistor length.

The model of the MIM capacitor is shown in Figure 8 and its equivalent circuit element values as a function of the length L_C are shown in Figures 10 and 11. The capacitance C_{12} is approximately two and a half times larger when the capacitor is placed in the ground planes, while the parasitic shunt capacitances are independent of the capacitor placement as shown in Figure 11. In Figure 10, the self resonant frequency of the capacitors is also plotted where it is seen that for the same value of capacitance, the resonant frequency is higher when the capacitor is placed in the ground plane. If it is assumed that the resonance is due to a series inductance, it is found that the value of this parasitic inductance is dependent only on the length of the capacitor and not on its placement.

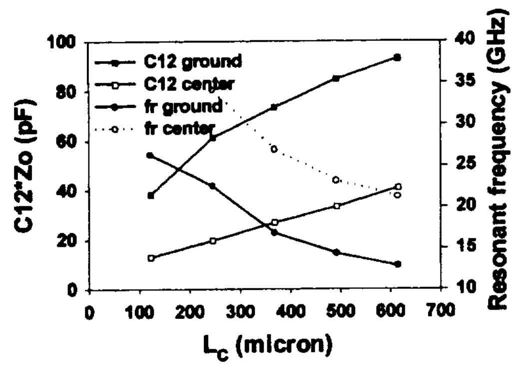


Figure 10: Measured series capacitance and self resonant frequency of MIM capacitors in FGC waveguide as a function of capacitor area.

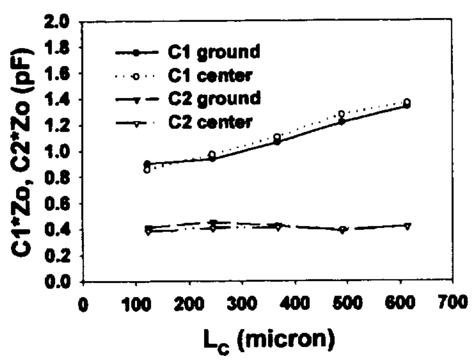


Figure 11: Measured shunt capacitance values of MIM capacitors in FGC waveguide as a function of the capacitor length.

While we have only reported here the characteristics of thin film resistors and MIM capacitors, the results are typical of those found for the integration of spiral inductors and series connected open and short circuit terminated stubs [20]. In all cases, the characteristics indicate that circuit elements may be placed in the ground strips of FGC waveguide and that they behave to a first order as two elements connected in parallel. Furthermore, the parasitic reactances do not appear to be dependent on the placement of the elements. Thus, it is possible to obtain better electrical characteristics for some circuit elements when they are placed in the ground strips. In particular, capacitive elements are best placed in the ground strips since it is possible to obtain twice the capacitance per unit length and a higher self resonant frequency compared to capacitors placed in the center strip.

A word of caution must be given though. We have also observed that large discontinuities such as series gaps in the ground strips do not act the same as if the element were placed in the center conductor. Rather, it has been observed that they excite higher order modes that severely degrade circuit performance [21].

Coupling Between Adjacent FGC Waveguides

We have shown that FGC waveguide has electrical properties similar to those of CPW while being electrically narrow and that circuit elements may be implemented in novel ways to reduce the circuit size. However, these advantages are lost if coupling between adjacent transmission lines forces the circuit designer to separate the lines further than they would be for conventional CPW. Therefore, we investigated the coupling between adjacent FGC lines as shown in Figure 12 and compared it to the coupling between CPW lines [22], [23]. Two methods for measuring the coupling are used: direct measurement using a vector network analyzer and theoretically through a 2-Dimensional Finite Difference Time Domain (2D-FDTD).

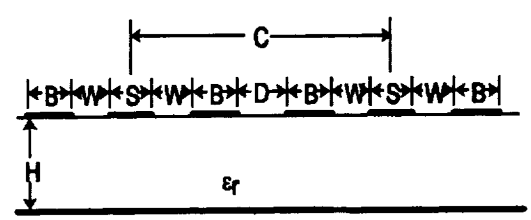


Figure 12: Coupled FGC waveguides.

The results are summarized in Figures 13 and 14 which show the coupling as a function of the ground plane width and the center to center line spacing, C, determined theoretically and experimentally respectively. It is seen that for a given spacing between the center lines of two FGC lines, the coupling is lower when B is smaller. Therefore, to minimize coupling between FGC lines, it is advantageous to have a narrower ground plane width and larger D for a specified center to center spacing. Also shown in Figure 13 is the coupling for FGC lines with a continuous ground plane between the lines, D=0. Note that the ground plane dimension, B, given for this data is only for the ground plane on the outside of the coupled lines. The results show that FGC lines with a continuous ground plane have greater coupling than the conventional FGC lines; coupling is reduced by as much as 15 dB by using finite ground planes between the coupled lines, and even a small value of D greatly reduces the coupling.

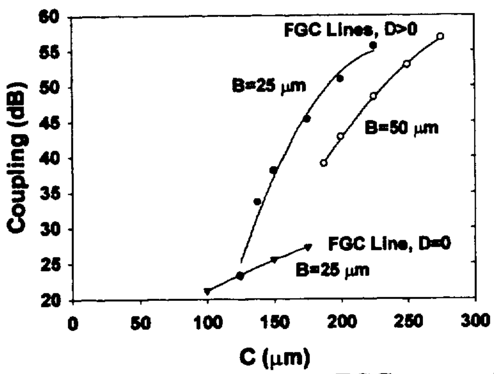


Figure 13: Coupling between FGC waveguides with D>0 and between FGC waveguides with a continuous ground plane (D=0) as a function of B and C determined by a 2D-FDTD analysis (S=W=25 μm).

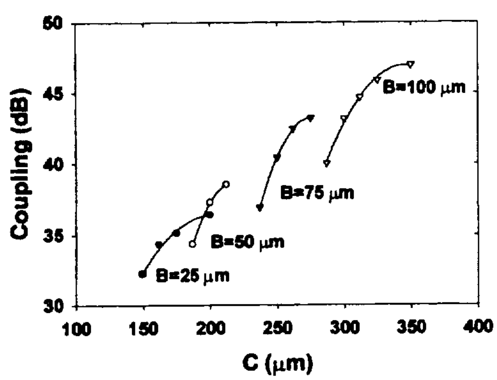


Figure 14: Measured coupling between FGC waveguides as a function of B and C (S=W=25 μm).

The nature of the coupling is understood by examining the horizontal electric field magnitude for an isolated FGC line as shown in Figure 15 and two coupled FGC lines as shown in Figures 16 and 17. The FGC line geometry is the same for all three lines. First, the electric field distribution of the isolated FGC line shows the same field distribution in the slot region as CPW, but it also shows a strong electric field component in the plane of the substrate on both sides of the FGC line. This electric field envelops adjacent FGC lines as shown in Figures 16 and 17 and gives rise to coupling between lines. Further examination of the field plots shows a strong asymmetry in the electric field distribution indicating a strong slotline mode. In fact, the slotline mode is 10 dB stronger than the CPW mode in the coupled lines with D>0, while it is 5 dB stronger when D=0. Thus, airbridges are required to suppress the slotline mode when FGC lines are adjacent to each other.

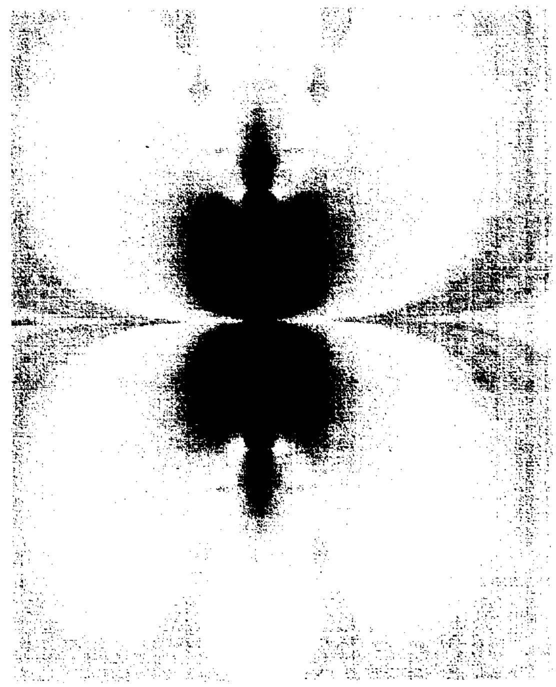


Figure 15: Horizontal electric field component of FGC waveguide (S=W=25 μm, B=50 μm).

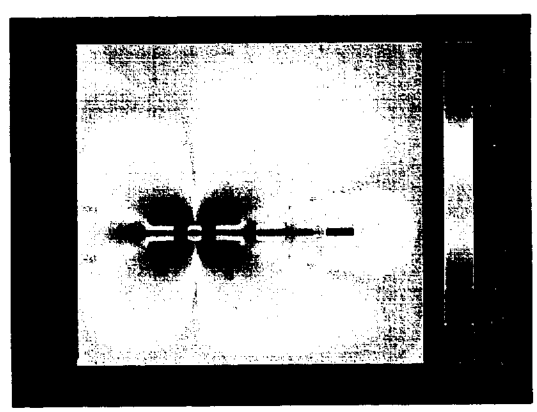


Figure 16: Horizontal electric field component of coupled FGC waveguides (S=W=25 μm, B=50 μm, D=25 μm).

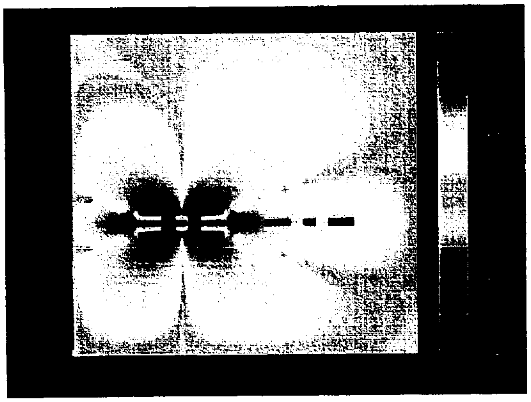


Figure 17: Horizontal electric field component of coupled FGC waveguides (S=W=25 μm, B=50 μm, D=75 μm).

Conclusions

In this paper, the authors have reviewed the advantages of Finite Ground Coplanar (FGC) waveguide for RF and wireless communication circuits. It has been shown that FGC line is better than CPW for MMICs and MCMs. In particular, it has been shown that FGC waveguide has an effective dielectric constant and attenuation comparable to CPW, does not give rise to spurious resonances, permits novel circuit layouts that enables the layout of smaller circuits with improved performance, and has lower coupling between adjacent lines than CPW if certain criteria are followed. Specifically, the ground planes should be twice the center conductor width and the total line width should be less than $\lambda_d/4$.

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