GaAs Versus Quartz FGC Lines for MMIC Applications

John Papapolymerou, Jack East, and Linda P. B. Katehi

***Abstract*—The performance of finite-ground coplanar (FGC) lines on GaAs and quartz for high-frequency monolithic-microwave integrated-circuit (MMIC) applications is experimentally investigated in this paper. The FGC lines on GaAs are covered with a thin layer of polyimide for passivation purposes. Permittivity and attenuation characteristics for these lines up to 118 GHz are presented and compared with the corresponding characteristics of FGC lines on a quartz substrate, which are commonly used for millimeter-wave applications. The impact of different characteristic impedance values in attenuation properties for both GaAs and quartz is also addressed. Results indicate that the loss on the lines does not depend on the substrate material, but rather on line geometry. All the lines tested show low-loss and low-dispersion characteristics.**

I. Introduction

The finite-ground coplanar (FGC) transmission line is a modified coplanar waveguide (CPW) structure with improved performance at millimeter-wave frequencies [1]. It consists of three strips, one for the signal and two for the ground, similar to the CPW line, but with ground strips that are narrow. A backside conductor is optional for the FGC line since the characteristics are independent of backside metallization. The main advantage of the FGC lines is that they do not support parallel-plate waveguide modes and, as a result, do not require via holes for ground equalization. These vias introduce parasitics and increase fabrication complexity. The permittivity and attenuation characteristics of FGC lines on GaAs and Si have been extensively investigated by several researchers [2]–[4]. Experimental results show that a nearly TEM mode propagates over a wide frequency range (2–118 GHz) and that loss is mainly ohmic. The use of full-thickness substrates and the elimination of via holes for mode suppression will reduce the cost and complexity of monolithic-microwave integrated-circuit (MMIC) design and fabrication at millimeter-wave frequencies. The purpose of this paper is to discuss the effect of polyimide passivation on these lines and to compare lines with different dimensions on both GaAs and quartz substrates. The results confirm the excellent performance of the FGC lines on GaAs.

High-performance line structures are an important part of MMIC design and fabrication at millimeter-wave frequencies [5], [6]. For a given substrate thickness and cross section, conventional microstrip lines have increasing dielectric loss with frequency. One solution to this problem is to use a low-loss material, such as quartz, as a substrate. However, this requires bonding active circuits onto the quartz, which increases fabrication complexity, especially at higher frequencies, while at the same time, the requirement for thinned substrates for microstrip or via holes for CPW's is still important to satisfy. FGC lines are conductor-loss dominated, thus, the use of semiconductor substrates will have a small effect on the line loss properties. Furthermore, passivation is also an important part of MMIC fabrication. Polyimide is a well-characterized dielectric material that has been extensively used in the past as a passivation layer and substrate for the fabrication of microstrip lines [7]–[9]. In this paper, we investigate the effect of a thin-polyimide coating on top of the GaAs FGC lines and compare their performance with low-loss quartz-based lines. The cross section of the fabricated FGC lines can be seen in Fig. 1. Four configurations have been fabricated and tested:

1) FGC lines on GaAs with $Z_0 = 40$, 50, and $60\ \Omega$;
2) FGC lines on GaAs with a 2-μm-thick polyimide overlay and $Z_0 = 50\ \Omega$;
3) FGC lines on GaAs with a 3-μm-thick polyimide overlay and $Z_0 = 50\ \Omega$;
4) FGC lines on quartz with $Z_0 = 70$, 90, and $100\ \Omega$.

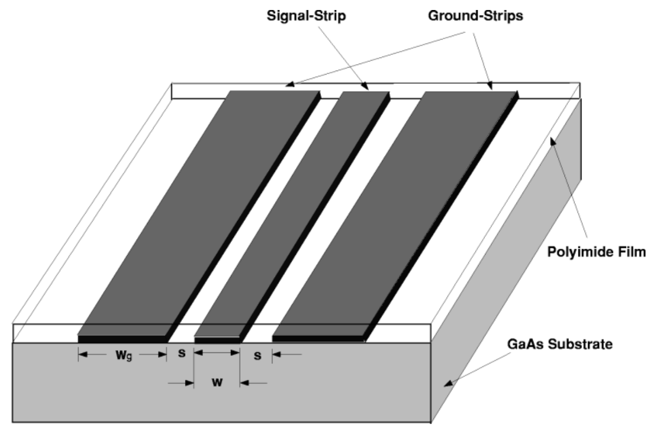

Fig. 1. FGC lines on GaAs with a polyimide overlay.

II. Fabrication

FGC lines have been fabricated on a 525-μm-thick semi-insulating GaAs wafer and a 165-μm-thick quartz wafer. The FGC line signal strip (w), slot (s), and ground strip (w_g) widths are 50, 45, and 160 μm, respectively, and correspond to a 50-Ω line for GaAs substrate and to 90 Ω for quartz substrate. The lines have been created by using a standard liftoff process with a total metal thickness of 1 μm. After the lines are formed, polyimide Pyralin PI2545 is spun at 4- and 2.5-K r/min on top of the lines for two GaAs wafers, in order to get a thickness of 2 and 3 μm, respectively. The precure temperature for the polyimide film is 140 $^\circ$C and the hard-cure 200 $^\circ$C, while the baking time is set to 30 min and 3 h, respectively. At the beginning and end of the FGC lines, the polyimide is chemically etched in order to allow the probe tips of the measurement system to be in electrical contact with the lines. The relative dielectric constant of the polyimide film is 3.5. The test FGC lines consisted of a thru line with a length of 1.0 mm, a short with a length of 0.5 mm, and three delay lines with 1.388-, 4.106-, and 10-mm length, respectively.

III. Results and Discussion

All of the measurements for the FGC line characteristics have been performed with an HP8510 network analyzer and a probe station using a variety of RF probes. Deembeding is achieved by performing a thru-reflect-line (TRL) calibration with the help of multiCal,[1] a measurement program available from NIST. This program also provides the effective dielectric permittivity and attenuation characteristics of the lines-under-test, from the delay-line measurements.

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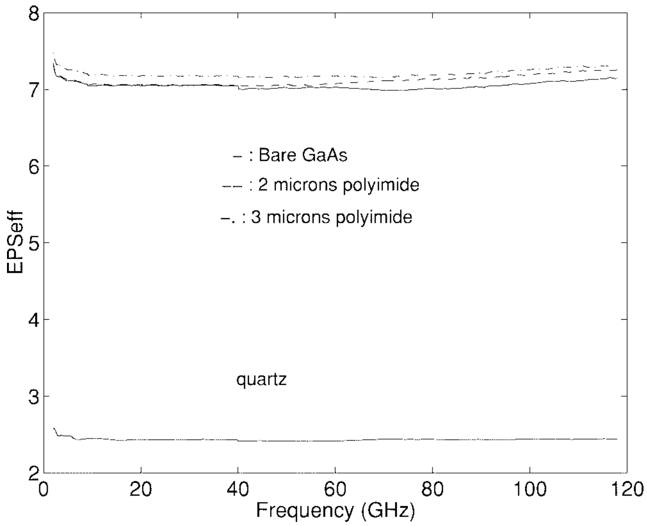

Fig. 2. Effective dielectric constant versus frequency for the various FGC lines.

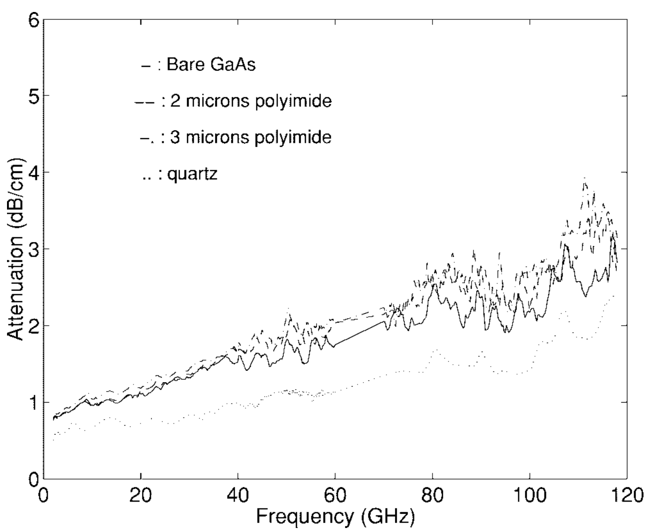

Fig. 3. Attenuation per physical length versus frequency for the various FGC lines.

The effective dielectric constant of the various line configurations is shown in Fig. 2. As can be seen, the nearly constant behavior of $\epsilon_{\rm eff}$ over the entire frequency range indicates the propagation of a nearly pure TEM mode. In addition, we observe that the thin film of polyimide is responsible for a slight increase in $\epsilon_{\rm eff}$ for both the 2-μm polyimide (1.4%) and the 3-μm polyimide (2.8%) when compared to that of bare FGC lines. This increase is more pronounced at higher frequencies.

Fig. 3 shows the attenuation per physical length for the four different cases. The straight line between 60–70 GHz represents a gap in the data. As can be seen, the polyimide increases the attenuation constant with the effect being more pronounced in W-band (12%) and for the thicker polyimide (23% in W-band). The quartz has the smallest attenuation for all the lines. This lower loss when measured in decibels/centimeters is due to the lower effective dielectric constant and higher characteristic impedance of the FGC line on quartz (90 Ω compared to 50 Ω for GaAs). The results of Figs. 2 and 3 are in very good agreement with similar results shown in [2]–[4] and [7]. Since, in most microwave circuits, lengths are expressed in terms of guided wavelengths, the attenuation per guided wavelength for the four different lines has been evaluated and is shown in Fig. 4. The results are comparable for all four cases with GaAs (bare or with polyimide) being slightly better than quartz. This indicates that the loss of FGC lines is ohmic in nature and independent of the substrate material. As a result, FGC lines are very good candidates for high-frequency application circuits. Furthermore, we can conclude that the thin layer of polyimide which covers the FGC lines on GaAs for passivation purposes does not increase the total loss of the lines.

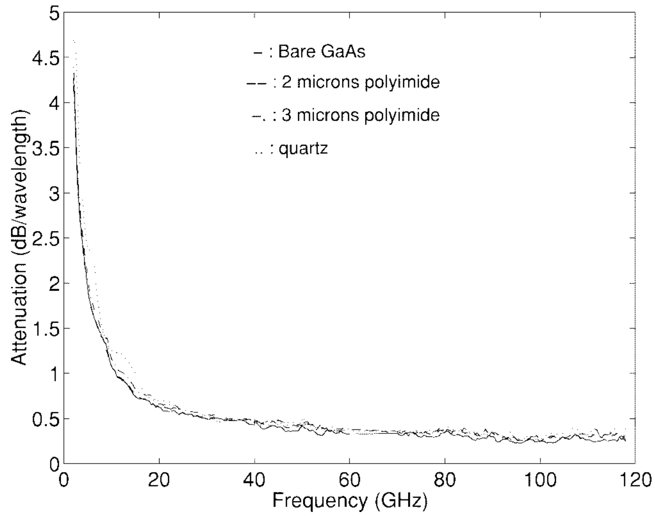

Fig. 4. Attenuation per guided wavelength versus frequency for the various FGC lines.

TABLE I
GEOMETRICAL CHARACTERISTICS FOR FABRICATED LINES

substrate	w (μm)	s (μm)	Z_0 (Ω)
GaAs	50	15.45	40
GaAs	50	64.5	60
Quartz	50	18.61	70
Quartz	50	64.5	100

Since the characteristic impedances for the lines investigated were different for GaAs and quartz, additional lines with varying dimensions and impedances have been fabricated. The impedance range for the two substrates corresponds to a convenient range for line fabrication. The line dimensions and the corresponding Z_0 can be seen in Table I. From the measured attenuation per physical length, the attenuation per guided wavelength has been evaluated and can be seen in Fig. 5 for GaAs and Fig. 6 for quartz. From Fig. 5, we observe that the 50- and 60-Ω lines have practically the same attenuation, while for the 40-Ω line, there is an increase of about 100% at 60 GHz, which is the center of the entire frequency range. Similarly, from Fig. 6, we observe that the attenuation for the 90- and 100-Ω lines is almost the same, and that the 70-Ω line exhibits a 60% increase from the other two lines at 60 GHz. We should note here that the small ripple observed in the 90- and 100-Ω lines is due to ripple in the mismatched measurement system.

In order to better understand the behavior of the FGC lines versus impedance, the measured attenuation per physical length data have been curve fitted to a $a + b\sqrt{f}$ function, and the extracted functions have been used in order to evaluate the attenuation per guided wavelength for three different frequency points in the center of each measured band. The final results can be seen in Fig. 7 for both quartz and GaAs. From these figures, we observe that the attenuation decreases as the impedance increases in a nonlinear way, as expected [10]. In terms of loss in dB/λ_g a 50- or 60-Ω line on GaAs is

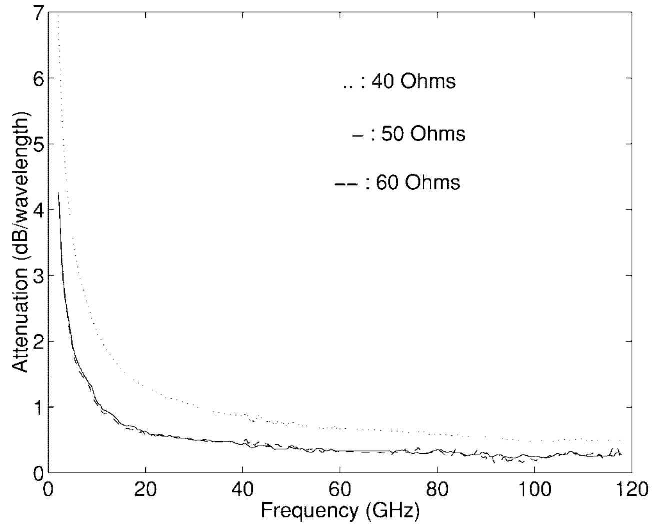

Fig. 5. Attenuation per guided wavelength for lines on GaAs with different Z_0.

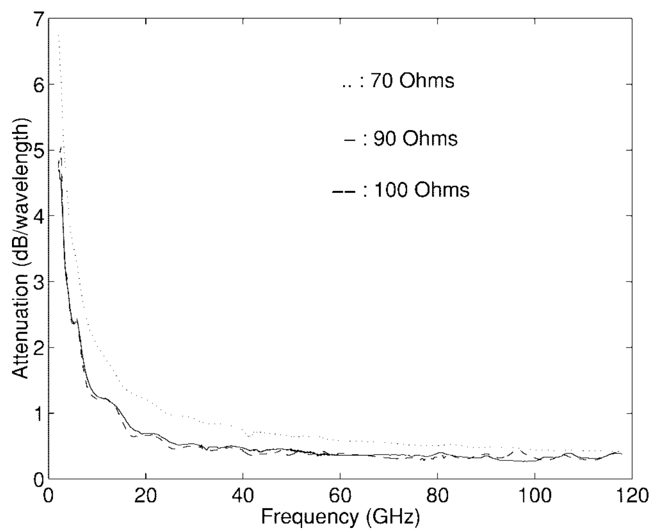

Fig. 6. Attenuation per guided wavelength for lines on quartz with different Z_0.

equivalent to a 90-Ω line on quartz. However, we should note that the 50- and 60-Ω GaAs lines have the same geometrical dimensions with the 90- and 100-Ω lines on quartz, respectively, indicating a strong dependence of the total line loss on the geometrical characteristics rather than the substrate material and thickness. This feature makes FGC lines ideal for high-frequency MMIC's.

Having found that, in terms of loss, GaAs and quartz are equivalent for the same FGC geometry, the choice of material for a substrate depends on other design criteria. If low cost is a major issue, then quartz can be chosen with the active devices being flip-chip bonded to it. On the other hand, if the active devices must be monolithically integrated with the rest of the circuitry, then GaAs is more appropriate with a thin overlay of polyimide for passivation. GaAs is also more suitable for applications above 120 GHz where the flip-chip bonding process increases fabrication complexity considerably.

IV. Conclusions

FGC lines on GaAs with a thin overlay of polyimide have been fabricated and tested. Experimental results show a negligible increase in the effective dielectric constant and a small increase in the attenuation per physical length when compared with bare FGC lines

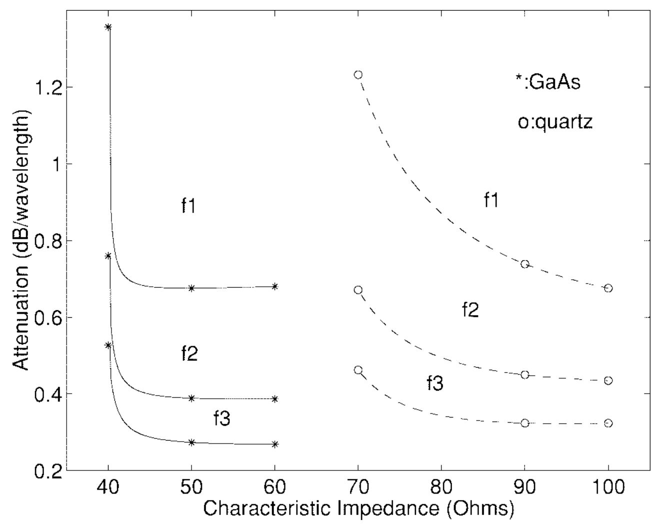

Fig. 7. Attenuation per guided wavelength versus characteristic impedance Z_0 for lines on GaAs and quartz at three different frequencies: $f_1 = 19.1$ GHz, $f_2 = 50$ GHz, and $f_3 = 94$ GHz.

on GaAs. The attenuation per physical length of FGC lines on GaAs with or without polyimide is higher than that of FGC lines on quartz. However, the attenuation per guided wavelength is almost the same for all types of lines investigated in this paper, indicating that the total loss of FGC lines with the same geometry is independent of the substrate material. This allows for the use of a thin layer of polyimide over FGC lines on GaAs without increasing the total loss in actual circuits, while providing passivation at the same time. In addition, the attenuation of the lines decreases in a nonlinear fashion versus characteristic impedance. Finally, FGC lines with a polyimide overlay can be used in millimeter-wave receivers and transmitters fabricated on GaAs, where the active devices are monolithically integrated with the other circuitry and do not need to be flip-chip bonded, as in the case of quartz.

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An Alternative Method for End-Effect Characterization in Shorted Slotlines

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***Abstract*—In this paper, we propose an alternative method to characterize the end effects in shorted slotlines. Using a commercial software to model the transition, a simple mathematical expression of the end impedance, valid for a given substrate family and in a wide frequency domain, was achieved. Several dielectric substrates were considered and the obtained results were compared with experimental ones. A correct agreement was observed. This approach can also be useful in more complex simulations relative to multilayer structures.**

***Index Terms*—End effect, slotline.**

I. Introduction

The development of microwave and millimeter-wave integrated circuits (MIC's) has increased and plays an important role in more recent mobile and satellite communication systems. Thus, the use of planar structures in interconnects is of greatest interest, especially microstrip lines because of their simple structure and their well-developed characterization. As a consequence, more sophisticated multilayer circuits must be considered in order to reduce the dimensions. Hence, the study of transitions between different planar transmission lines becomes necessary. For many years, several computational methods have been developed to solve electromagnetism equations in complex structures.

Two main approaches are investigated. One, generally based on semianalytical formulations, uses the method of moments and the spectral-domain approach [1]–[9], the second one considers essentially numerical algorithms like the finite-difference time-domain (FDTD) or the transmission-line matrix (TLM) method [10]–[17]. The respective benefits of these two approaches have already been discussed in the literature [3], [8], [14], [15].

In this paper, we consider the slot–microstrip transition. A rigorous study of such a structure needs to take into account the end effects which appear on these transmission lines and depend on the working frequency and linewidth. These end effects have already been characterized as capacitive or inductive according to the type of line. For microstrip lines, simple calculated models have been proposed [20]–[22]. They give a good evaluation of capacitive effects at the line termination. For the slotlines, the existing calculations are somewhat heavy and they are not so easy to use for circuit design. Furthermore, many experimental measurements have been made for different substrates and different linewidths at given frequencies in order to obtain families of charts. However, all these proposed measurement techniques [23], [24] need a large number of experimental circuits, although the results are necessarily limited because of the time needed for experiment. Using a relatively more simple method based on MDS Momentum simulation [18], we achieved and propose an efficient analytical model of end effects in a slot line, which can be simply used for many applications and more general computations.

Our results are then compared to those obtained both by experimental approaches [23], [24] and by sophisticated numerical computations [25].

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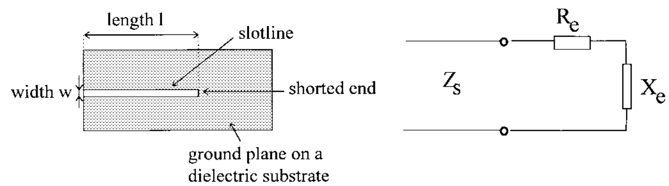

Fig. 1. Shorted slotline and its end equivalent circuit.

II. Theoretical Approach

As has been shown in the literature [25], slotline termination can be considered as a simple equivalent circuit such as a resistance and a serial inductance (see Fig. 1). Hence the expression of Z_e, the slotline end-effect impedance, is $Z_e = R_e + jX_e$ where R_e represents the radiation losses and X_e represents the inductive phenomena at the line termination.

Considering Z_s the slotline characteristic impedance, due to the non-TEM nature of the transmission mode in slot lines, Z_s cannot be defined uniquely [19], but in the present approach, we simply use the results given by the simulator. Let Z be the line impedance at a given point and $z = Z/Z_s$ be the normalized impedance at this point. z can also be written $z = (1 + \Gamma/1 - \Gamma)$, where Γ represents the corresponding reflection coefficient. With our simulator, it is possible to set a port at a given point and find Γ as S_{11}. In the case of a single-port measurement, the literal expression of the input impedance is then given by

$$z_{\text{in}} = \frac{(1 + S_{11})}{(1 - S_{11})}.$$

Let us consider a line of length l. For a given frequency, linewidth, and type of substrate, two simulation steps are necessary. The first one determines the characteristic impedance Z_s and the slot guided wavelength λ_s. In the second one, the MDS Momentum optimization module gives the S_{11} maximum value (corresponding to the maximum VSWR) in accordance to a determined length $l = \delta + (\lambda_s/2)$ where $\delta \le (\lambda_s/2)$. Therefore, the impedance at the input port (z_{in}) is calculated and the end-effect impedance of the shorted slotline z_e is determined as

$$z_e = \frac{z_{\text{in}} - jtg\left(\dfrac{2\pi\delta}{\lambda_s}\right)}{1 - jz_{\text{in}}tg\left(\dfrac{2\pi\delta}{\lambda_s}\right)}.$$

Normalized resistances r_e and reactances x_e are plotted versus frequency for different width w and different substrates. Each curve can be simply fitted with a linear equation. Therefore, for a given substrate with a thickness h and a dielectric constant ε_r, we obtain a single expression of end reactance, which can be simply used to accurately take into account end effects in the slot for mixed microstrip–slotline structures.