is small enough for portable radio applications, was confirmed for 1 to 4.5 GHz operation.

IV. Conclusions

A GaAs divide-by-256/258 dual-modulus prescaler has been developed using LSCFL with the new level shift circuits and BP-SAINT FET's. The prescaler operated up to 4.5-GHz input frequency at a supply voltage of 3 V with a power dissipation of 100 mW.

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GaAs on Si as a Substrate for Microwave and Millimeter-Wave Monolithic Integration

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Abstract—**Recent advances in GaAs growth on Si have resulted in high-quality and high-performance GaAs electronic and optoelectronic devices on Si substrates. One therefore must consider this composite structure as a substrate material for microwave and millimeter-wave monolithic integrated circuits. In order for GaAs on Si to be practical for this purpose, the dielectric loss must be small. We have calculated the dielectric losses of GaAs/Si composite in a transmission line configuration and compared them with those of other possible substrates, such as GaAs and Si alone, in the frequency range of 10–100 GHz. Depending upon the thickness, results show that high-resistivity GaAs epitaxial layers on Si substrates having moderate resistivities reduce the dielectric loss.**

I. Introduction

Since Hyltin examined Si as a transmission line medium for monolithic microwave integration [1], there has been a great deal of activity in the search for the most suitable substrate for monolithic microwave and millimeter wave integration [2]–[4]. So far, most of the efforts have been concentrated on Si and GaAs materials because of the availability of three-terminal devices on these substrates. Although microwave and millimeter-wave sources (IMPATT and Gunn diodes) [5] and passive components (transmission lines, Lange couplers, Wilkinson splitters, etc.) [6] have already been achieved monolithically by using either Si or GaAs substrate, both substrates have some disadvantages and advantages when compared with each other. Si is available in large area wafers and has advance processing technology. Si is also mechanically sturdy and has a larger thermal conductivity than GaAs but is not capable of producing optical sources (lasers, LED's) and low-noise millimeter-wave three-terminal devices which are already available on GaAs substrates. If high-quality GaAs layers can be grown on Si substrates, the aforementioned disadvantages of each can be remedied. Moreover, such a process opens the possibility of monolithic integration of GaAs and Si-based devices [7].

GaAs-based three-terminal devices on Si substrates, such as metal semiconductor field effect transistors (MESFET's) [8], modulation doped field effect transistors (MODFET's) [9], and heterojunction bipolar transistors (HBT's) [10], have already been demonstrated with performances nearly identical to those grown on GaAs substrates. Therefore, one of the remaining potential problems is identifying whether or not the dielectric loss of this composite material, GaAs/Si, is small enough to allow transmission line media in the microwave and the millimeter-wave region. We have thus undertaken a theoretical study of shielding microstrip lines on GaAs/Si up to 100 GHz for various parameters in Si and GaAs. The dielectric losses in GaAs, Si, and GaAs/Si materials have been investigated and compared.

II. Theory

The characteristics of a shielded microstrip line with double-layer substrate, Fig. 1, were investigated by using the spectral-domain analysis in which hybrid mode representation is used [11], [12]. Since spectral-domain analysis has been described for shielded multilayer dielectric with arbitrary coplanar conductors [13], [14], it is briefly reviewed here for double-layer substrate.

The total field in the structure can be obtained by a superposition of the TE and TM fields, which can be derived from the scalar electric and magnetic potentials ψ^e and ψ^h, respectively. Here the field components in each layer can be obtained as follows:

$$E_{zi} = j\frac{k_i^2-\beta^2}{\beta}\psi_i^e(x,y)\,e^{-j\beta z} \tag{1}$$

$$H_{zi} = j\frac{k_i^2-\beta^2}{\beta}\psi_i^h(x,y)\,e^{-j\beta z} \tag{2}$$

$$\overline{E}_{ti} = \nabla_t\psi_i^e(x,y)\,e^{-j\beta z} - \left(\frac{\omega\mu}{\beta}\right)\hat{a}_z x \nabla_t\psi_i^h(x,y)\,e^{-j\beta z} \tag{3}$$

$$\overline{H}_{ti} = \left(\frac{\omega\epsilon_i}{\beta}\right)\hat{a}_z x \nabla_t\psi_i^e(x,y)\,e^{-j\beta z} + \nabla_t\psi_i^h(x,y)\,e^{-j\beta z} \tag{4}$$

where the subscript i denotes the regions inside the shield and

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the superscripts e and h stand for the TE and TM fields, respectively.

Finite Fourier transform in the x direction is applied to the potential functions to obtain the spectral-domain representation of the fields. Satisfying the appropriate boundary conditions at the interfaces, the following equations are obtained:

$$\tilde{E}_{z0}(\alpha_n) = G_{11}(\alpha_n, \beta)\tilde{J}_x(\alpha_n) + G_{12}(\alpha_n, \beta)\tilde{J}_z(\alpha_n) \quad (5)$$

$$\tilde{E}_{x0}(\alpha_n) = G_{21}(\alpha_n, \beta)\tilde{J}_x(\alpha_n) + G_{22}(\alpha_n, \beta)\tilde{J}_z(\alpha_n) \quad (6)$$

where $\tilde{J}_x(\alpha_n)$ and $\tilde{J}_z(\alpha_n)$ are the transforms of the unknown current components on the strip. To solve these equations, Galarkin's method is applied in spectral domain by expanding unknown current components $\tilde{J}_x$ and $\tilde{J}_z$ in terms of known basis functions $\tilde{J}_{xm}(\alpha_n)$ and $\tilde{J}_{zm}(\alpha_n)$ as

$$\tilde{J}_x(\alpha_n) = \sum_{m=1}^{M} a_m \tilde{J}_{xm}(\alpha_n) \quad (7)$$

$$\tilde{J}_z(\alpha_n) = \sum_{m=1}^{N} b_m \tilde{J}_{zm}(\alpha_n) \quad (8)$$

and substituting into (5) and (6). After some mathematical manipulations, one can obtain the following algebraic equations:

$$\sum_{m=1}^{M} a_m K_{im}^{11} + \sum_{m=1}^{N} b_m K_{im}^{12} = 0 \quad (9a)$$

$$\sum_{m=1}^{M} a_m K_{im}^{21} + \sum_{m=1}^{N} b_m K_{im}^{22} = 0. \quad (9b)$$

Simultaneous solution of these two algebraic equations yields the propagation constant β for a specific frequency.

Once the propagation constant β and the field components in each layer are found, (10) and (11) can be used to calculate the characteristic impedance and the dielectric loss coefficient, respectively:

$$Z_0 = \frac{2\mathrm{Re}\int_s \bar{E} \times \bar{H} \cdot \hat{a}_z ds}{I_z^2} \quad (10)$$

$$\alpha_d = \frac{\omega\epsilon \tan\delta \int_{s\mathrm{diel}} |\bar{E}|^2 ds}{2\mathrm{Re}\int_s \bar{E} \times \bar{H} \cdot \hat{a}_z ds}. \quad (11)$$

This dielectric loss calculation assumes that the loss tangents of the dielectric materials being used are sufficiently small.

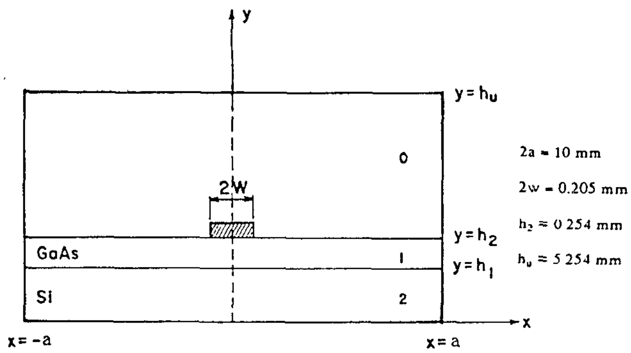

Fig. 1. Shielded microstrip line with double layer substrate.

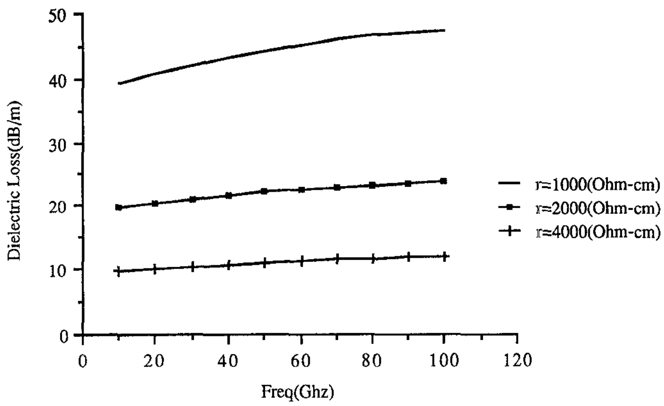

Fig. 2. Dielectric loss versus frequency for Si substrate, $\epsilon_r = 11.8$

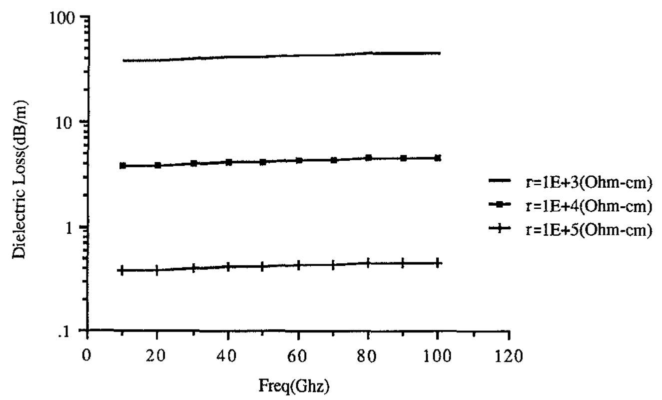

Fig. 3. Dielectric loss versus frequency for GaAs substrate, $\epsilon_r = 12.91$.

III. Results and Discussions

The effective dielectric constant, characteristic impedance, and dielectric loss of the shielded microstrip line whose geometry is shown in Fig. 1, were obtained for three different substrates, namely GaAs, Si, and GaAs/Si. Dielectric loss versus frequency of GaAs and Si substrate are shown in Figs. 2 and 3. The same parameters for GaAs/Si substrate were plotted for two different resistivities of the GaAs overlay material, each at three different GaAs overlayer thicknesses as shown in Figs. 4 and 5.

Since Si-based millimeter-wave high-power IMPATT diodes, GaAs-based high-frequency MESFET's, MODFET's, HBT's, and several millimeter-wave passive circuit elements have already been designed, they may be monolithically integrated on the GaAs/Si substrate. The dielectric loss of this composite substrate is not higher than that of a Si substrate alone, provided that the resistivity of the GaAs overlayer is higher than the resistivity of the Si substrate, Fig. 5.

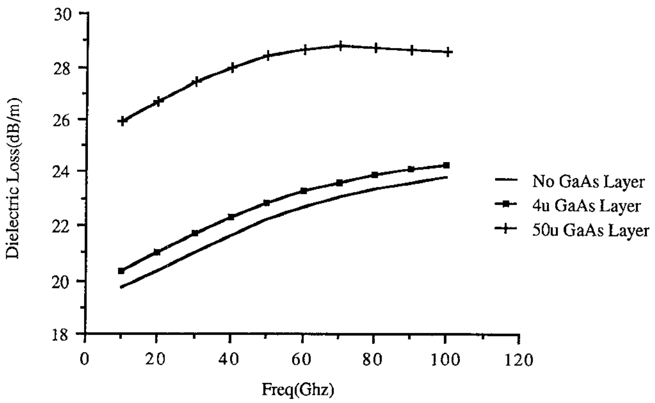

Fig. 4. Dielectric loss versus frequency for $\rho_{Si} = 2000\ \Omega \cdot cm$, $\rho_{GaAs} = 1000\ \Omega \cdot cm$.

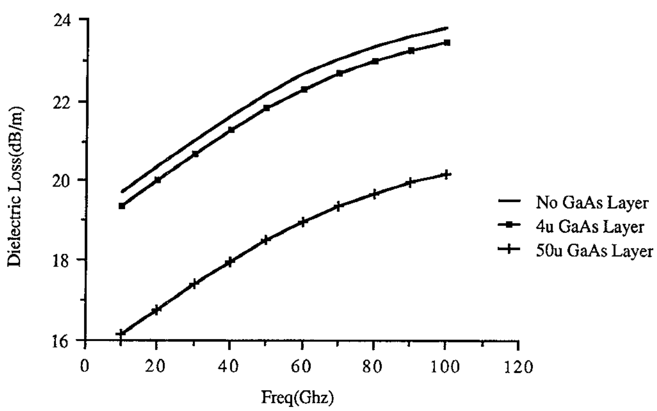

Fig. 5. Dielectric loss versus frequency for $\rho_{Si} = 2000\ \Omega \cdot cm$, $\rho_{GaAs} = 4000\ \Omega \cdot cm$

From Figs. 4 and 5 and one can conclude that a) the resistivity of the substrate material will not be degraded considerably if the resistivity of the GaAs overlay material is less than that of the Si substrate, provided that the thickness of the overlay material is relatively small, and b) depending upon the GaAs overlayer thickness, the equivalent resistivity of the entire structure can be higher than that of the Si substrate when a high-resistivity GaAs overlay material is used. Table I includes data to show the effect of GaAs overlayer, quantitatively. If the GaAs resistivity cannot be made sufficiently large, the transmission line can be fabricated on dielectric materials, such as SiO_2 deposited on Si substrates.

IV. Conclusions

The dielectric loss in GaAs/Si material has been investigated as substrate material for 50-Ω shielded microstrip line, and it has been found that GaAs overlay material modifies the resistivity of the entire structure depending upon its thickness and resistivity relative to those of Si material. As far as the dielectric loss is concerned, by using high-resistivity GaAs overlayer on Si, the dielectric loss can be reduced. Since some microwave and millimeter-wave passive components, active components, and sources have already been established monolithically either on GaAs substrate or Si substrate, they might be well established on GaAs/Si substrate to yield a monolithically integrated system. It can then be stated that this combination offers exciting possibilities for monolithic microwave and millimeter-wave integration.

TABLE I
Dielectric Loss P_{dL}(dB/m) at $f = 100$ GHz

	w/o GaAs	4 μm GaAs	50 μm GaAs
$\rho_{Si} = 2000\ \Omega$−cm $\rho_{GaAs} = 1000\ \Omega$−cm	23.8	24.2	28.6
$\rho_{Si} = 2000\ \Omega$−cm $\rho_{GaAs} = 4000\ \Omega$−cm	23.8	23.5	20.2
$\rho_{Si} = 2000\ \Omega$−cm $\rho_{GaAs} = 1 \times 10^4\ \Omega$−cm	23.8	23.3	18.5
$\rho_{Si} = 2000\ \Omega$−cm $\rho_{GaAs} = 1 \times 10^6\ \Omega$−cm	23.8	23.2	17.4

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