N94-20504

ATTENUATION AND ε_{EFF} OF COPLANAR WAVEGUIDE TRANSMISSION LINES ON SILICON SUBSTRATES

Susan R. Taub and Paul G. Young
NASA-Lewis Research Center
Cleveland, OH

Abstract

Attenuation and ε_{eff} of Coplanar Waveguide (CPW) transmission lines have been measured on Silicon substrates with resistivities ranging from 400 to greater than 30,000 ohm-cm, that have a 1000 angstrom coating of SiO_2. Both attenuation and ε_{eff} are given over the frequency range 5 to 40 GHz for various strip and slot widths. These measured values are also compared to the theoretical values.

Introduction

Historically, Silicon has not been the material of choice for microwave applications because of its extremely high loss and its lack of high frequency active devices. Recently, however, high frequency Silicon devices have become available. In addition, Silicon Germanium (SiGe) shows great promise for high frequency Silicon based devices [1].

Theoretical work has been done in calculating the attenuation of microstrip lines on Silicon as a function of resistivity and frequency [2]. There has also been some work on Coplanar Waveguide (CPW) lines on Silicon on Insulator (SOI) [3] as a function of resistivity. This paper presents both measured and theoretical attenuation and ε_{eff} data of CPW lines on Silicon as a function of resistivity, strip and slot width and frequency.

Results

Groups of CPW lines were fabricated on five different Silicon wafers. The wafers were 8 mils thick and had resistivities of 400-720 ohm-cm, 2500-3300 ohm-cm, 5000 ohm-cm, 5000-10,000 ohm-cm, and greater than 30,000 ohm-cm. A 1000 angstrom layer of SiO_2 was deposited on the wafers, followed by a 2.5μm thick Au layer. The CPW lines were fabricated using etch back. Each group of CPW lines consisted of on open, thru, and four delay lines with a different strip width (s) and slot width (w) for each group. The s and w of each of the groups were: s=2 w=1, s=4 w=2, and s=6 w=3. The impedance of these lines was approximately 50 ohms. The CPW lines were on wafer probed using PicoProbes and an HP8510 Automatic Network Analyzer. The data from each group was analyzed using NIST's DEEMBED software to obtain values for the attenuation and ε_{eff}.

Figure 1 shows the measured attenuation of a CPW line with s=4 w=2 on Silicon substrates of varying resistivities. As the resistivity of the material increases, the attenuation decreases. This is attributed to the reduction of dielectric loss. Loss also increases with frequency, due to an increase in the effective length of the line. Note that for substrates of medium resistivity (400-720 ohm-cm), the loss is twice that of those with resistivities greater than 2,500 ohm-cm. However, the improvement in attenuation for substrates with resistivities greater than 2,500 ohm-cm is small.

Figure 2 shows the attenuation of CPW lines on a Silicon wafer with resistivity of 2,500-3,300 ohm-cm as a function of s and w. Measured and theoretical curves are shown.

The theoretical curves were given by Gupta [4]. As was predicted by the theory, the loss of the CPW lines decrease with increasing s and w; however, the measured values of attenuation are slightly lower than the theory predicts.

ε_{eff} is not a function of resistivity but of s and w. Figure 3 shows theoretical and measured values for ε_{eff} of CPW lines with s=6 w=3. The theoretical values were again given by Gupta [4]. The value of ε_{eff} is virtually independent of resistivity and closely matches the theoretical value.

Conclusions

Both measured and theoretical values for attenuation and ε_{eff} for CPW lines on Silicon wafers were shown as a function of resistivity, strip and slot width and frequency. Losses for CPW lines on Silicon can be minimized if the resistivity of the wafer is kept above 2,500 ohm-cm. Thus making Silicon a viable microwave material.

Acknowledgment

The authors would like to thank Dr. Roger Marks of NIST for his help with the software and Dr. Rainee Simons for his support during this project.

References

[1] Patton, G., et al. "75 GHz f_t SiGe-Based Heterojunction Bipolar Transistor", IEEE Electron Device Lett., 1990, EDL-11, (4), pp. 171-173.

[2] Rosen, A., et al. "Silicon as a Millimeter-wave Monolithically Integrated Substrate-a New Look", RCA Rev., 1981, 42, pp. 633-660.

[3] Caviglia, A., et al. "Microwave Performance of SOI n-MOSFET's and Coplanar Waveguides", IEEE Electron Device Lett., 1991, EDL-12, (1), pp. 26- 27.

[4] Gupta, K.C., R. Garg, I.J. Bahl. Microstrip Lines and Slotlines. Artech House, Inc. 1979. pp. 285-287 and 275-276.

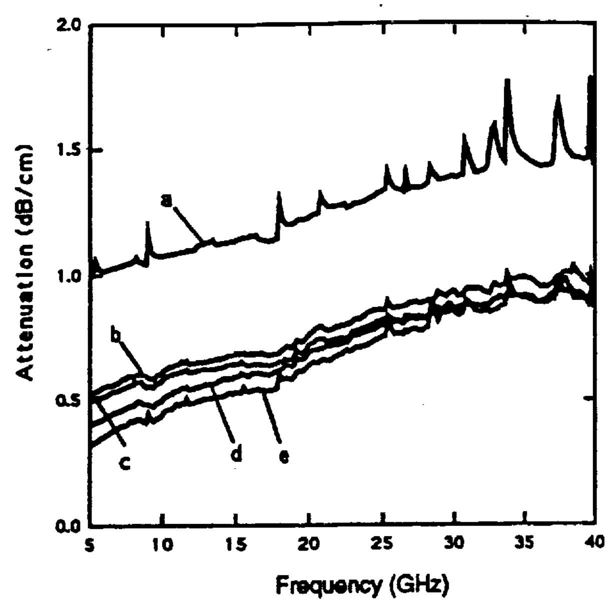

Figure 1: Measured attenuation of CPW lines on Silicon substrates with resistivities: a) 400-720 ohm-cm b) 2500-3300 ohm-cm c) 5000 ohm-cm d) >30,000 ohm-cm

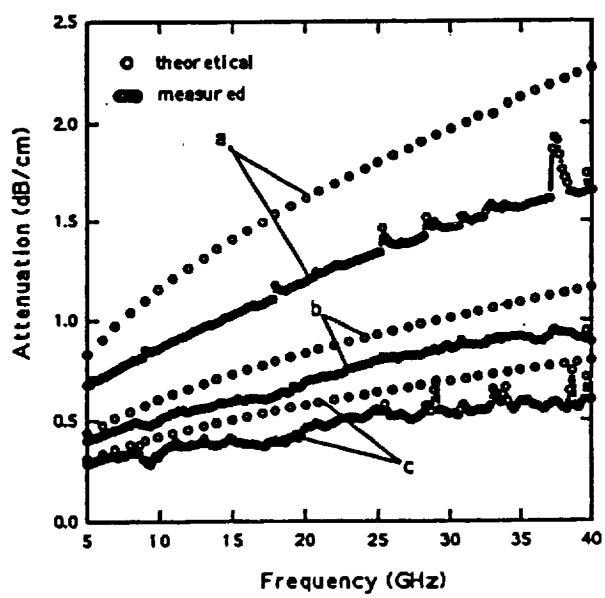

Figure 2: Measured and theoretical attenuation of CPW lines on a Silicon substrate, resistivity= 2500-3300 ohm-cm: a) s=2,w=1 b) s=4, w=2 c) s=6, w=3

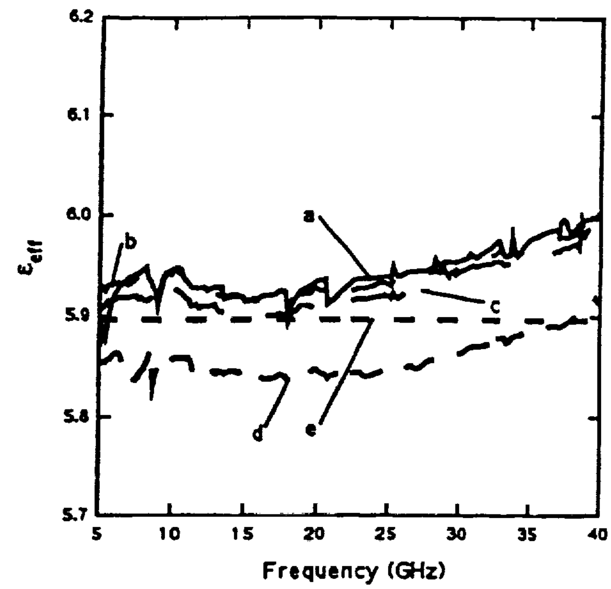

Figure 3: ε_{eff} of CPW lines on Silicon substrates, s=6 w=3, with resistivities: a) 400-720 ohm-cm b) 2500-3300 ohm-cm c) 5000 ohm-cm d) >30,000 ohm-cm e) theory

NOVEL COPLANAR WAVEGUIDE TO SLOTLINE TRANSITION ON HIGH RESISTIVITY SILICON

R. N. Simons, S. R. Taub and P. G. Young

Indexing terms: Coplanar waveguide, High resistivity silicon

Two novel coplanar waveguide (CPW) to slotline transitions have been fabricated and tested on high resistivity silicon. The first transition uses an air bridge to couple RF power from the CPW line to the slotline and has the entire circuit on the top side of the wafer. In the second transition, the grounded CPW (GCPW) line and the slotline are on opposite sides of the wafer and are coupled electromagnetically. The measured average insertion loss and return loss per transition are better than 1.5 and 10dB, respectively, with a bandwidth greater than 30% at C-band frequencies.

Introduction: Recently, $Si_{1-x}Ge_x$-on-silicon devices have emerged as a viable alternative to GaAs- and InP-based heterostructure devices [1]. It appears that this material technology will be suitable for monolithic microwave integrated circuits (MMICs) for a number of reasons, some of which are the following: The technology is compatible with existing silicon technology which is extensively used for digital circuits and the low cost of silicon wafers. The active devices such as transistors and diodes are fabricated on the heterostructure while the interconnections would be laid out over the silicon substrate. By choosing a silicon substrate with sufficiently high resistivity it is possible to make the dielectric attenuation constant of the interconnecting microwave transmission lines approach those of GaAs [2]. For this to be possible, the transmission line interconnects must be characterised on silicon.

This Letter presents two new CPW to slotline transitions fabricated on high resistivity silicon substrates. A transition between a coplanar waveguide (CPW) and a slotline on a high resistivity silicon substrate has several applications. These include: facilitating fast and inexpensive testing of CPW and slotline MMICs using on-wafer RF probes, functioning as a balun in mixer circuits, and providing interconnection between CPW MMIC phase shifters or amplifiers and linearly tapered slot antennas in phased arrays. In the past, several investigators have worked on conventional CPW to slotline transitions on alumina and GaAs substrates [3, 4]. In these transitions the CPW line and the slotlines are orthogonal to each other.

Transition design and fabrication: The first transition presented in this Letter uses an air bridge to couple RF power from the CPW line to the slotline and has the entire circuit on the same side of the wafer. In the second transition, the grounded CPW (GCPW) line and the slotline are on opposite sides of the wafer and are coupled electromagnetically. In both cases

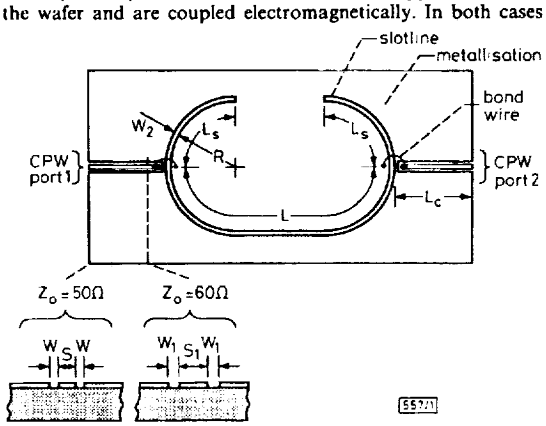

Fig. 1 *Schematic diagram of air-bridge coupled CPW to slotline transition*

$S = S_1 = 0{\cdot}1524$ mm, $W = 0{\cdot}0838$ mm, $W_1 = 0{\cdot}1524$ mm, $L = 19{\cdot}77$ mm, $L_s = 6{\cdot}83$ mm, $L_c = 5{\cdot}08$ mm, $R = 4{\cdot}35$ mm, and $W_2 = 0{\cdot}157$ mm

these circuits, the CPW line and the slotline are collinear. Both transitions are fabricated on a single 5000–10000 Ω cm resistivity silicon wafer. The thickness D of the wafer is 0·381 mm with $\varepsilon_r = 11{\cdot}7$. The thickness T of the gold metallisation is about three times the skin depth at the centre frequency f_0 of 6GHz.

(*a*) *Air-bridge coupled CPW to slotline transition:* An air-bridge coupled CPW to slotline transition is illustrated in Fig. 1. At the input port the characteristic impedance Z_0 is 50Ω for compatibility with on-wafer RF testing. The line transforms to a 60Ω line that terminates in an open circuit. The Z_0 of the slotline is 60Ω. The circular bend at the feed end of the slotline provides a smooth transition. A 1 mil diameter bond wire between the open end of the CPW centre strip conductor and the opposite edge of the slotline functions as an air bridge and couples RF energy. The length L_s is $\sim \lambda_{g(slotline)}/4$ at f_0.

(*b*) *Electromagnetically coupled GCPW to slotline transition:* Fig. 2 shows a transition with electromagnetic coupling between a GCPW and a slotline which are on opposite sides of a wafer. At the input port Z_0 is 50Ω, and in the centre region Z_0 is 60Ω. This is realised by gradually flaring the GCPW slots. At the open end the top ground planes are terminated in two open circuited stubs of length L_g. Owing to the lack of CPW discontinuity models, these stubs are model-

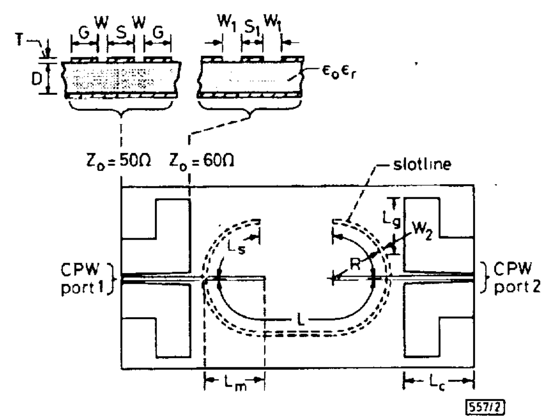

Fig. 2 *Schematic diagram of electromagnetically coupled CPW to slotline transition*

Top side circuit pattern:
$S = S_1 = 0{\cdot}1524$ mm, $W = 0{\cdot}1067$ mm, $W_1 = 0{\cdot}2515$ mm, $G = 2{\cdot}54$ mm, $L_m = 4{\cdot}6$ mm, $L_g = 3{\cdot}98$ mm, $L_c = 5{\cdot}08$ mm
Bottom side circuit pattern:
$L = 18{\cdot}7$ mm, $L_s = 6{\cdot}83$ mm, $R = 4{\cdot}35$ mm and $W_2 = 0{\cdot}157$ mm

led as microstrip lines of very low Z_0 ($\sim 10\Omega$) and length about $\lambda_{g(microstrip)}/4$. They provide a virtual short circuit between the top ground planes of the GCPW and the slotline. In addition, the GCPW has a finite ground plane of width G to suppress the parallel plate waveguide mode [5]. The Z_0 of the slotline is 60Ω. Once again, owing to a lack of CPW discontinuity models, the centre strip conductor of the GCPW which extends beyond the terminated ground planes of the GCPW to form the transition is modelled as a microstrip line. The distances L_s and L_m are $\sim \lambda_{g(slotline)}/4$ and $\lambda_{g(microstrip)}/4$, respectively, at f_0.

Transition performance and discussions: During testing, the circuits were suspended 10mm above the probe station stage. The insertion loss and return loss were measured using Cascade Microtech on-wafer probes. For two back-to-back air-bridge coupled CPW to slotline transitions, with a short length of slotline in between, the measured characteristics are shown in Fig. 3. The insertion loss and return loss per transition are ~1·5dB, and better than 10dB, respectively, over greater than 30% bandwidth centred at f_0. The above insertion loss includes the insertion loss of the two air bridges and the following which were not practical to calibrate out: the 19·8mm length of slotline between the transitions, and two

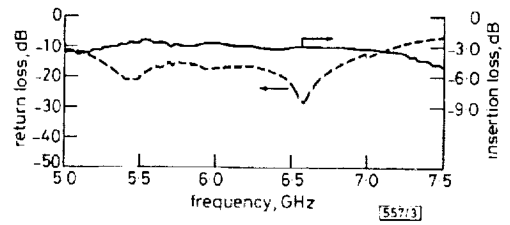

Fig. 3 *Measured insertion loss and return loss of air-bridge coupled CPW to slotline transition*

5·08 mm long CPW lines located at the input and output ports. Fig. 4 shows the measured insertion loss and return loss, respectively, for the electromagnetically coupled GCPW to slotline transitions with a short length of slotline in between. The insertion loss and return loss per transition are ~1·5 dB, and better than 10 dB, respectively, over greater than 40% bandwidth centred at 4·55 GHz. The above insertion loss includes the insertion loss of the two junctions and the following which were not practical to calibrate out: the 18·7 mm length of slotline between the transitions, and two 5·08 mm long GCPW lines at the input and output ports. The return loss of this transition is observed to be better than the previous design because of better impedance match between the strip and the slot. However, the shift in the centre frequency from 6 to 4·55 GHz is caused by discontinuity effects which were not fully accounted for in the design.

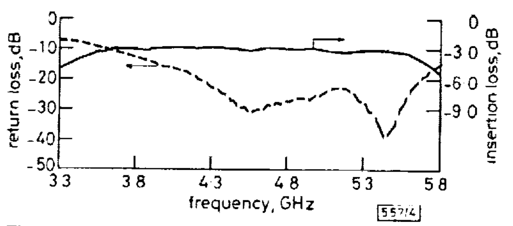

Fig. 4 *Measured insertion loss and return loss of electromagnetically coupled CPW to slotline transition*

Conclusions: Two novel CPW to slotline transitions on high resistivity silicon substrates have been experimentally demonstrated. The transitions are fabricated either on the same side or on opposite sides of a wafer and use an air bridge or electromagnetic coupling, respectively, to couple power. The measurements show that the transitions have good insertion loss, return loss and bandwidth characteristics.

2nd October 1992

R. N. Simons, S. R. Taub and P. G. Young (*Mail Stop 54-5 NASA Lewis Research Center, 21000 Brookpark Road, Cleveland, Ohio 44135, USA*)

References

1 PATTON, G., *et al.*: '75 GHz f_T SiGe-based heterojunction bipolar transistors', *IEEE Electron Devices Lett.*, 1990, **EDL-11**, (4), pp. 171–173

2 ROSEN, A., *et al.*: 'Silicon as a millimeter-wave monolithically integrated substrate—a new look', *RCA Rev.*, 1981, **42**, pp. 633–660

3 HOUDART, M., and AURY, C.: 'Various excitation of coplanar waveguide'. IEEE MTT-S Int. Microwave Symp. Dig., 1979, pp. 116–118

4 HIROTA, T., TARUSAWA, Y., and OGAWA, H.: 'Uniplanar MMIC hybrids—a proposed new MMIC structure', *IEEE Trans.*, 1987, **MTT-35**, (6), pp. 576–581

5 SIMONS, R. N., and LEE, R. Q.: 'Coplanar waveguide aperture coupled patch antenna with ground plane/substrate of finite extent', *Electron. Lett.*, 1992, **28**, (1), pp. 75–76

MICROWAVE CHARACTERIZATION OF SLOTLINE ON HIGH RESISTIVITY SILICON FOR ANTENNA FEED NETWORK

Rainee N. Simons, Susan R. Taub,
Richard Q. Lee, and Paul G. Young
National Aeronautics and Space Administration
Lewis Research Center
Cleveland, Ohio 44135

ABSTRACT

This paper presents, the effective dielectric constant (ε_{eff}) and attenuation constant (α) of a unshielded slotline on a high resistivity (5000 to 10,000 Ω-cm) silicon wafer. The ε_{eff} (DC to 40 GHz) and α (DC to 26.5 GHz) are determined from the measured resonant frequencies and the corresponding insertion loss of a slotline ring resonator. The measurements are carried out at room temperature and without the application of a DC bias. The attenuation for slotline on silicon are compared with microstrip line and coplanar waveguide on other semiconductor substrate materials. Finally, applications of the slotline to antenna feed network are addressed.

I. INTRODUCTION

There are several reasons why silicon is now a viable microwave material. One reason is that silicon MOSFET's with cutoff frequencies as high as 89 GHz have been reported (ref.1). Another, is that silicon MMIC amplifiers, mixers and IMPATT diodes are now commericially available (refs. 2 thru 4). The final reason is that transmission lines, such as, microstrip line (refs. 5 thru 9) and Coplanar Waveguide (CPW) (ref. 10) with low loss have been demonstrated on high resistivity silicon. Silicon has several advantages over GaAs and InP technologies, such as: better thermal conductivity, higher reliability, higher circuit complexity, availability of wafers of very large diameters, better mechanical properties, and lower cost (ref. 9). Furthermore, integration of MMIC's with digital control circuits and radiating elements on a single silicon wafer is possible. This can enhance the reliability, efficiency and lower the cost of phased array antenna systems. However, silicon substrates do have slightly higher dielectric loss than traditional microwave substrates. DC bias and high operating temperatures can increase this loss. Additional loss can be introduced, if during processing, the wafers are exposed to temperatures high enough to significantly lower their resistivity (ref. 9).

Conventional silicon wafers have low resistivity and consequently an unacceptably high value of dielectric attenuation. Therefore, microwave circuits for phased array antenna systems fabricated on these wafers have low efficiency. By choosing a silicon substrate with sufficiently high resistivity it is possible to make the dielectric attenuation of the interconnecting microwave transmission lines approach those of GaAs or InP (refs. 6 and 7).

In order to fabricate microwave circuits on silicon, the transmission lines on this material must be characterized. Recently, the attenuation of microstrip transmission lines on high resistivity, bare and passivated silicon as a function of frequency, temperature and DC bias have been measured (ref. 9). Also, attenuation and ε_{eff} of CPW lines as a function of resistivity, frequency and geometry on silicon substrates has been examined experimentally and theoretically (ref. 10). This paper presents the effective dielectric constant (ε_{eff}) and attenuation constant (α) of an unshielded slotline on a high resistivity (5000 to 10,000 Ω-cm) silicon wafer over the frequency ranges DC to 40 GHz and DC to 26.5 GHz respectively. The measurements are carried out at room temperature and without the application of a DC bias.

II. THEORY

An experimental slotline ring resonator is shown in figure 1. The advantage of a ring resonator over a series gap coupled linear resonator is that the ring resonator is free of end effects. The loaded Q-factor, Q_L, of the resonator is determined from the following equation relating the measured resonance frequency, f_0, and the frequency range, Δf, between the 3-dB points on either side of the resonance:

$$Q_L = f_0 \,/\, \Delta f. \tag{1}$$

The unloaded Q-factor, Q_u, of the resonator is determined from the following equation relating the measured peak insertion loss, L, at resonance and Q_L:

$$L\ (\mathrm{dB}) = 20 \operatorname{Log} \{1 - [Q_L/Q_u]\}. \tag{2}$$

The ε_{eff} is determined from the following equation:

$$\varepsilon_{eff} = (30\ n/f_0\ l)^2 \tag{3}$$

Where n is an integer and denotes the order of resonance. Therefore for n resonances of a particular resonator, n values of ε_{eff} can be obtained. l is the mean circumference of the ring in cm. f_0 is in GHz.

The phase velocity, v_{ph}, of the electromagnetic wave on the slotline is equal to

$$v_{ph} = 3\mathrm{x}10^8 \,/\sqrt{\varepsilon_{eff}} \quad (\mathrm{mt/sec}). \tag{4}$$

Finally, the attenuation constant α of the slotline is determined from the relation:

$$\alpha = \pi\ f_0 \,/Q_u\ v_{ph} \quad (\mathrm{Np/mt}). \tag{5}$$

III. RESONATOR FABRICATION AND EXPERIMENTAL RESULTS

The slotline ring resonator is fabricated on a silicon wafer which is coated sequentially with 700 Å of silicon dioxide, 200 Å of chromium and 2.5 μm of gold. The thickness, T, of the gold metalization is greater than three times the skin depth at 8.5 GHz and above. The measured resistivity of the silicon dioxide layer is 10^{14} Ω-cm. The ε_{eff} and α are determined by substituting the measured resonant frequencies and the corresponding insertion loss in equations 1 thru 5. Figure 2 presents the ε_{eff} as a function of the frequency. In this figure the slot width W, wafer thickness D, and relative dielectric constant ε_r, are equal to 0.1 mm, 0.381 mm and 11.7, respectively. Also shown in Fig.2 is the computed ε_{eff} which is obtained as described in ref.11. The measured and computed ε_{eff} are in good agreement.

The intrinsic peak insertion loss L of the resonator is corrected for the insertion loss due to the microstrip feed lines and the coaxial connectors of the fixture. This is done by subtracting the feed and connector loss from the overall measured insertion loss. These excess losses are determined from a separate set of measurements using a thru line of length equal to the sum of the feed line lengths in the test fixture. Figure 3 presents the measured attenuation α as a function of the frequency. The W, D and ε_r of the slotline are the same as those in Fig. 2. The attenuation of slotline is compared in Table 1 with the measured results from the open literature for microstrip line and coplanar waveguide on various other semiconductor substrate materials. It is worth mentioning here that the attenuation values quoted in Table 1 depend on the substrate thickness, metalization thickness and also the strip conductor width and/or slot width which are not the same in all cases.

IV. CONCLUSIONS AND DISCUSSIONS

The ε_{eff} and α for a slotline on a high resistivity silicon substrate have been experimentally obtained. The attenuation constant for slotline has been compared with that of microstrip line and coplanar waveguide on other semiconductor substrate materials. The value of attenuation for slotline was found to be comparable to other transmission lines. This, however, can only be a rough comparison because attenuation depends upon the substrate thickness, metalization thickness and strip conductor width and/or slot width which are not the same in all the cases. However our experiments demonstrate the viability of high resistivity silicon for low loss antenna feed network. Application of this information to the feed network will be presented at the symposium.

REFERENCES

1. Yan, R.H., et al.: 89-GHz f_T Room-Temperature Silicon MOSFET's, IEEE Electron Device Letters, Vol. 13, No. 5, pp. 256-258, May 1992.

2. Hanes, M.H., et al.: MICROX™ -An All-Silicon Technology for Monolithic Microwave Integrated Circuits, IEEE Electron Device Letters, Vol. 14, No.5, pp. 219-221, May 1993.

3. Hewlett Packard Communications Components, GaAs and Silicon, Designers Catalogue, Section 8, 1993.

4. Hughes millimeter Wave Products Catalogue.

5. Hyltin, T.M.: Microstrip Transmission on Semiconductor Dielectrics, IEEE Trans. Microwave Theory Tech., Vol.MTT-13, No.6, pp.777-781, Nov. 1965.

6. Young, L. and H. Sobel (Editors), Advances in Microwaves, Vol.8, Academic Press, New York, New York, 1974, pp. 19-25.

7. Rosen, A., et al.: Silicon as a Millimeter-Wave Monolithically Integrated Substrate - A New Look, RCA Rev., Vol. 42, pp. 633-660, 1981.

8. Caviglia, A.L. et al.: Microwave Performance of SOI n-MOSFET's and Coplanar Waveguides, IEEE Electron Device Letters, Vol.12, No.1, pp.26-27, Jan. 1991.

9. Levesque, K., et al.: Microwave Characterization of High Resistivity Silicon, Microwave Hybrid Circuits Conf., Oct. 1992.

10. Taub, S.R. and P.G. Young: Attenuation and ε_{eff} of Coplanar Waveguide Transmission Lines on Silicon Substrates, Eleventh Annual Benjamin Franklin Symposium on Antenna and Microwave Technology in the 1990s Digest, pp.8-11, May 1993.

11. Simons, R,N.: Suspended Slot Line Using Double Layer Dielectric, IEEE Trans. Microwave Theory Tech., Vol. MTT-29, No. 10, pp.1102-1107, Oct.1981.

12. Higgins J.A.: Microwave GaAs FET Monolithic Circuits, 1979 IEEE Inter. Solid-State Circuits Conf. Digest of Tech. Papers, pp. 120-121.

13. Haydl, W.H. et al.: Millimeterwave Coplanar Transmission Lines on Gallium Arsenide, Indium Phosphide and Quartz with Finite Metalization Thickness, 1991 IEEE MTT-S Inter. Microwave Symp. Digest, pp. 691-694, 1991.

14. Haydl, W.H.,: Experimentally Observed Frequency Variation of the Attenuation of Millimeter-Wave Coplanar Transmission Lines with Thin Metalization, IEEE Microwave & Guided Wave Letters, Vol.2, No.8, pp.322-324, Aug. 1992.

TABLE 1
Comparision of Attenuation Constant of Microwave Transmission Lines on Semiconductor Substrates

TRANSMISSION LINE	SUBSTRATE MATERIAL	DIMENSIONS (Inch)	ATTENUATION @ 10 GHz (dB/cm)	REFERENCE
Microstrip (Z_0 = 50 Ω)	SI GaAs	D = 0.025 W = 0.025 T = 3 μm*	0.105	12
Coplanar Waveguide (CPW) (Z_0 ≈ 50 Ω)	SI GaAs	D = 0.025 S = 0.025 W = 0.0125 T = 3 μm*	0.16	12
Coplanar Waveguide (CPW)[1] (Z_0 ≈ 35 Ω)	SI InP	D = 0.5 mm S = 88 μm W = 16 μm T = 0.25 μm*	4.5	14
Coplanar Waveguide (CPW) (Z_0 = 50 Ω)	SI GaAs	D = 0.5 mm S = 75 μm W = 56 μm T = 3 μm*	0.45	13
Microstrip (Z_0 = 50 Ω)	High Res. silicon (1.5 k Ω-cm)	D = 0.01 W = 0.006$	0.5	5
Microstrip (Z_0 ≈ 50 Ω)	High Res. Silicon (8 k Ω-cm)	D = 0.021 W = 0.016 T ≈ 3 μm	0.16	9
Coplanar Waveguide (CPW)[2] (Z_0 ≈ 50 Ω)	High Res. Silicon (2.5 - 3.3 k Ω-cm)	D = 0.008 S = 0.004 W = 0.002 T = 2.5 μm*	0.62	10
Coplanar Waveguide (CPW)[2] (Z_0 ≈ 60 Ω)	High Res. Silicon (4 k Ω-cm)	D = 400 μm S = 30 μm W = 35 μm T = 1 μm$	3	8
Slotline[2] (Z_0 = 60 Ω)	High Res. Silicon (5 - 10 k Ω-cm)	D = 0.015 W = 0.004 T = 2.5 μm*	0.25	This work

D is substrate thickness and T is metalization thickness
Microstrip: W is strip width
Coplanar Waveguide: S is center strip width and W is slot width
[1]Metal thickness less than one skin depth
[2]A SiO_2 interfacial layer is present
*Gold conductors, $Aluminium conductors

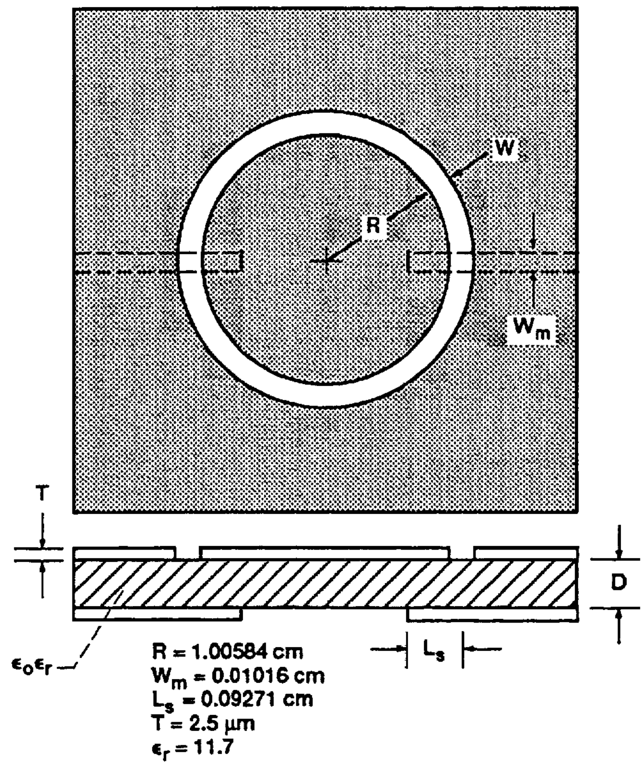

Figure 1.—Slotline ring resonator electromagnetically coupled to microstrip feed lines.

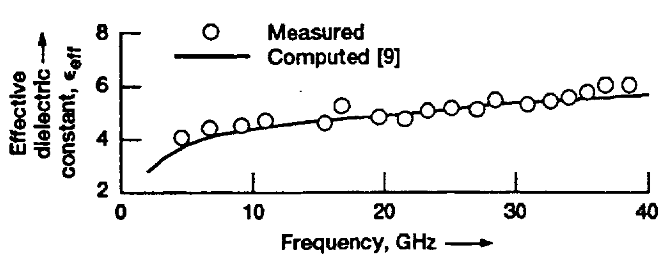

Figure 2.—Effective dielectric constant.

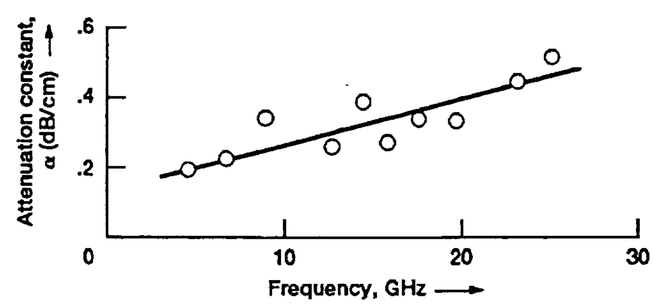

Figure 3.—Attenuation constant.

NEW COPLANAR WAVEGUIDE FEED NETWORK FOR 2 × 2 LINEARLY TAPERED SLOT ANTENNA SUBARRAY

Rainee N. Simons, Thomas D. Perl, and Richard Q. Lee
Mail Stop 54-5
NASA Lewis Research Center
21000 Brookpark Road
Cleveland, Ohio 44135

KEY TERMS

Coplanar waveguide, linearly tapered slot antenna, slot line, array antenna

ABSTRACT

A new technique for exciting 2 × 2 subarray of linearly tapered slot antennas (LTSAs) with coplanar waveguide (CPW) is presented. rf power is coupled to each element through a CPW-to-slotline transition by a coax-to-CPW in-phase four-way radial power divider. The transition and the power divider are coupled by a novel nonplanar CPW right-angle bend. Measured results at 18 GHz show excellent radiation patterns and return-loss characteristics. © 1992 John Wiley & Sons, Inc.

I. INTRODUCTION

Linear tapered slot antennas (LTSAs) are useful in applications which require high gain, narrow beamwidth and wide bandwidth [1]. The most common approach of exciting a single-element LTSA is with a microstrip/slotline transition. Recently, the use of coplanar waveguide feeding a single-element LTSA has been demonstrated [2]. For LTSA arrays, finline-to-waveguide [1] and slotline-to-microstrip [3] feed structures have been reported. The former feeding approach is bulky in size, while the latter could produce spurious radiation.

This article proposes a new technique for exciting a 2 × 2 LTSA subarray using a CPW-to-slotline transition in conjunction with a coax-to-CPW in-phase four-way radial power divider. The transition and the power divider are coupled by a novel nonplanar CPW right-angle bend. This compact feed design is easy to integrate, and using CPW as the transmission medium produces less spurious radiation.

II. CPW FEED NETWORK AND SUBARRAY DESIGN

Figure 1 illustrates the construction of the 2 × 2 LTSA subarray. The feed network for this subarray consists of a CPW-to-slotline transition and a coax-to-CPW in-phase four-way, radial power divider [4] on separate dielectric substrates coupled by a novel nonplanar CPW right-angle bend.

a. LTSA and CPW-to-Slotline Transition. The LTSA and the CPW feed are etched on opposite sides of the circuit board and electromagnetically coupled, as shown in Figure 1. The LTSA is formed by gradually flaring the width of the slotline by an angle 2α. In general, a symmetric beam is required to illuminate a reflector for maximum aperture efficiency, this is achieved by choosing 2α equal to 10.6 degrees [1]. Similarly, to optimize the radiation efficiency of the LTSA, H is chosen to be 0.75 λ_0, where λ_0 is the free-space wavelength at the center frequency f_0 of 18 GHz. The length L of the antenna as determined by α and H is 4.1 λ_0.

To couple power to the antenna, the center strip conductor of the CPW is extended to form a CPW-to-slotline transition with the LTSA. The distances from the short-circuit termination of the slotline and the open termination of the extended center strip conductor, to the CPW-to-slotline junction is approximately a quarter of a wavelength at f_0. To provide a smooth transition, the slotline at the feed end of the LTSA has a circular bend instead of a right-angle bend. The radius of curvature of the bend is approximately $\lambda_{g(slot)}/6$. The finite ground planes of the CPW lines are connected to the antenna ground plane via holes to ensure odd-mode excitation. Further, the CPW ground planes are tapered to provide good impedance match.

The coax-to-CPW in-phase four-way radial power divider circuit board is also shown in Figure 1. Design details for the radial power divider are given in [4].

b. Nonplanar CPW Right-Angle Bend. The nonplanar CPW right-angle bend is illustrated in Figure 2. In this bend, the center strip conductor of the CPW line is enlarged, forming a circular island that facilitates drilling a hole for a pin con-

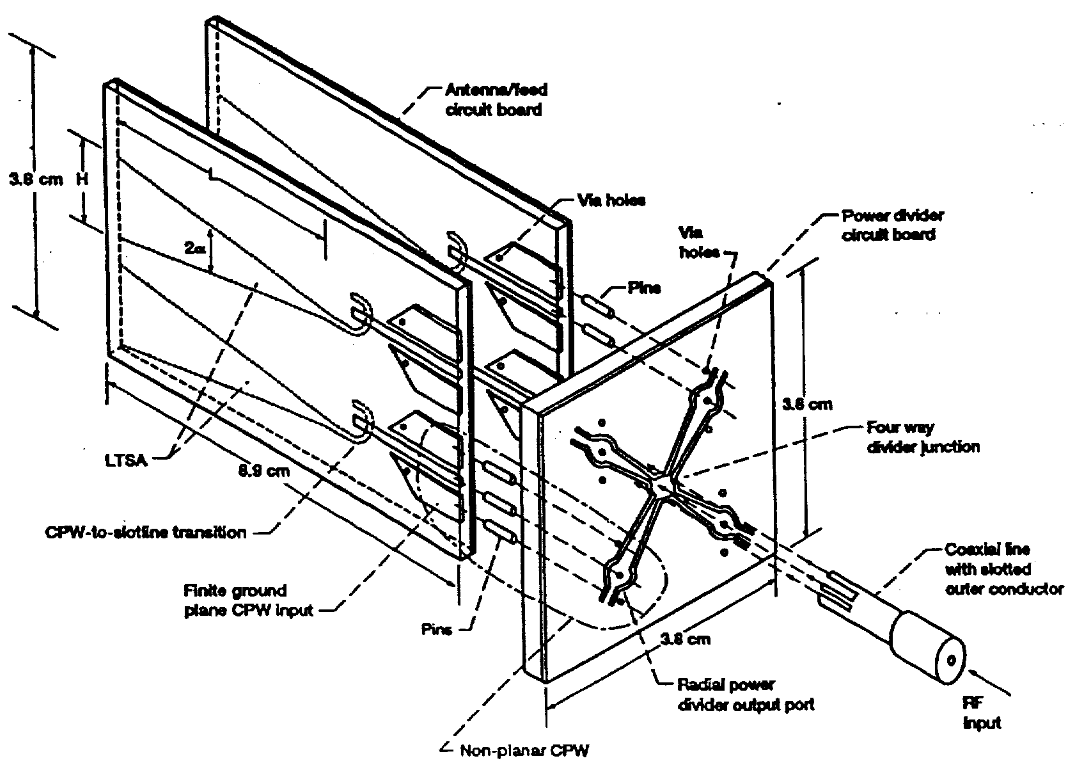

Figure 1 Schematic diagram illustrating the construction of 2 × 2 LTSA subarray

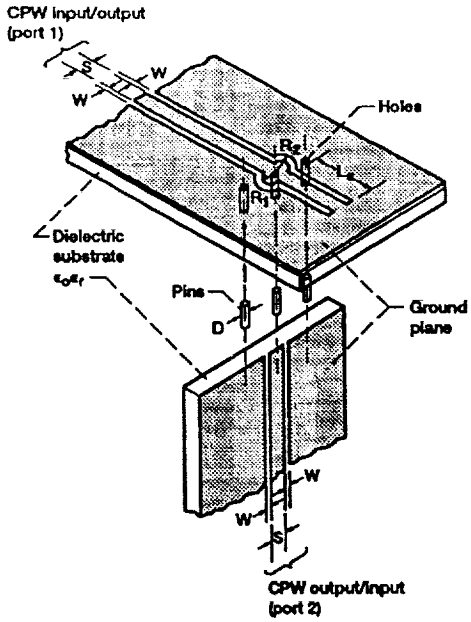

Figure 2 Schematic diagram illustrating nonplanar CPW right-angle bend

nection. Further, the CPW line beyond the island is terminated in a short-circuited stub for impedance matching. The radius of the island R_1, the surrounding slot region R_2, the diameter of the pin D, and the length of the CPW short-circuited stub L_s were experimentally optimized to obtain the best insertion loss and return loss characteristics. R_1, R_2, D, and L_s are approximately 0.043 λ_g, 0.063 λ_g, 0.024 cm, and 0.2 λ_g, respectively, where λ_g is the CPW guide wavelength at f_0. The power divider network as well as the subarray are fabricated on 0.0508-cm-thick RT/Duroid 5880 ($\epsilon_r = 2.2$) circuit board.

III. CPW RIGHT ANGLE BEND AND SUBARRAY PERFORMANCE

The measured insertion loss (S_{21}) of the nonplanar CPW right-angle bend is shown in Figure 3. The measured insertion loss

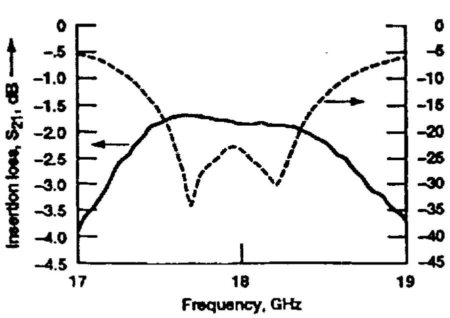

Figure 3 Measured insertion loss (S_{21}) and return loss (S_{11}) of the nonplanar CPW right-angle bend

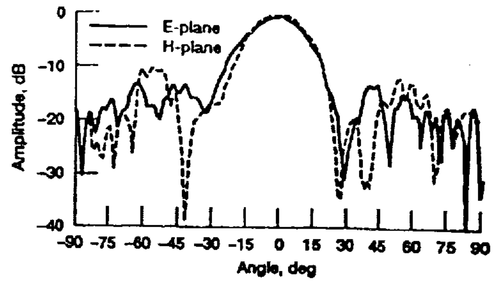

Figure 4 Measured radiation pattern of 2 × 2 LTSA subarray. (a) E plane, (b) H plane

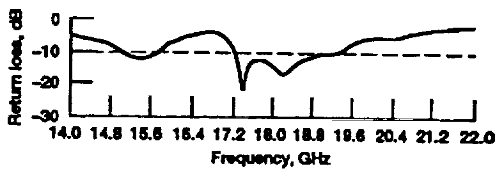

Figure 5 Measured return loss at coaxial input port of the 2 × 2 LTSA subarray

includes the losses occurring in the bend, the attenuation of 2.54-cm length of CPW line on either side of the bend, and that of the two coaxial connectors used at the measurement ports. Also superimposed on Figure 3 is the return loss (S_{11}), which is better than −10 dB and has a bandwidth of 1.45 GHz centered at f_0.

The measured E- and H-plane radiation patterns of the 2 × 2 LTSA subarray at f_0 are shown in Figure 4. The 3-dB beamwidth is approximately 23 degrees in the principal planes, and the measured cross-polarization is less than −16 dB. The measured gain of the single-element LTSA is approximately 11 dB, and hence the gain of the 2 × 2 LTSA subarray is estimated to be 17 dB. Last, as shown in Figure 5, the measured return loss at the coaxial input port of the subarray is better than −10 dB, and the 2:1 VSWR bandwidth is the same as the CPW right-angle bend.

IV. CONCLUSIONS AND DISCUSSIONS

A new feeding technique for a 2 × 2 LTSA subarray using a CPW-to-slotline transition and a coax-to-CPW in-phase four-way radial power divider have been demonstrated. The characteristics of a novel nonplanar CPW right-angle bend have also been presented. The subarray has excellent radiation patterns and symmetric beamwidth. Because of its compactness, this subarray module is suitable as a feed for a reflector antenna or as building blocks for large arrays.

REFERENCES

1. K. S. Yngvesson et al., "The Tapered Slot Antenna—A New Integrated Element for Millimeter-Wave Applications," *IEEE Trans. Microwave Theory Tech.*, Vol. MTT-37, No. 2, Feb. 1989, pp. 365–374.
2. A. Nesic, "Endfire Slot Line Antennas Excited by a Coplanar Waveguide," *IEEE AP-S Int. Symp. Digest*, 1991, pp. 700–702.
3. D. Rascoe et al., "K_a-Band MMIC Beam Steered Planar Array Feed," *1990 IEEE MTT-S Int. Microwave Symp. Digest*, pp. 809–812.
4. R. N. Simons and G. E. Ponchak, "Coax-to-Channelised Coplanar Waveguide In-Phase N-Way, Radial Power Divider," *Electron. Lett.*, Vol. 26, No. 11, May 1990, pp. 754–755.

Received 2-17-92