New Uniplanar Transitions for Circuit and Antenna Applications

Nihad I. Dib, Rainee N. Simons, *Senior Member, IEEE*, and Linda P. B. Katehi, *Fellow, IEEE*

Abstract—**New uniplanar microstrip-to-slotline, microstrip-to-coplanar strips (CPS) and microstrip-to-coplanar waveguide (CPW) transitions for MIC/MMIC and slotline antennas for phased array applications are described. Such transitions are compact and suitable to be used in an open environment or inside a package or a multichip module. The transitions share the concept of using a balun which consists of two microstrip lines connected to a slotline through a pair of coupled microstrips. In this paper, the transitions are studied theoretically using the Finite Difference Time Domain (FDTD) technique and measured experimentally using an HP8510C Network Analyzer. For a back-to-back microstrip-to-slotline transition, an insertion loss of less than 1.3 dB per transition is achieved over a 49% bandwidth with a minimum of 0.6 dB around the design frequency.**

I. Introduction

In most slotline antenna applications, a transition is needed to couple the slotline to another planar line, e.g., microstrip line. Although there are techniques to couple the RF energy from a microstrip line to a slot-line, both sides of the substrate are needed to achieve a successful transition [1]–[3]. Such a configuration requires accurate alignment during fabrication due to the demonstrated sensitivity of the transition to the relative position between the microstrip, which is printed on the top side of the substrate, to the slotline printed on the opposite side. Furthermore, such a transition requires a multi-wafer/substrate configuration and presents many practical problems when there is a need for packaging.

In 1983, a single sided microstrip to slot-line transition was proposed and measured [4], [5]. It used a balun in the form of an open ring consisting of two microstrip lines connected to a slot-line through a pair of coupled microstrip. Recently, the same transition has been studied by the authors both theoretically and experimentally [6], [7]. It has been found that the insertion loss and return loss for a single transition are better than 1.5 and 10 dB, respectively, over 8% bandwidth centered at 9 GHz (the design frequency was 10 GHz) [7]. Moreover, a linearly tapered slot antenna (LTSA) excited by this microstrip-to-slotline transition has been experimentally investigated [7]. A major limitation in this approach when used in slotline antennas is the off-axis feeding mechanism and the resulting large space requirement for such a feed.

Recently, a new approach for designing a uniplanar microstrip-to-slotline transition has been proposed in [8]. In this transition, a slot-line radial stub was utilized to create resonance conditions at the junction of the two transmission lines so that maximum power transfer was achieved. It was reported that the bandwidth of the tested transition was 30% at a design frequency of 3.8 GHz [8].

In this paper, new microstrip-to-slotline uniplanar topologies are proposed which provide broadband transition characteristics, require much less real estate and give an on-axis input/output geometry. In addition, these transitions operate without the need for via-holes in contrast to the transition in [8] where grounding pins were used to ensure a good short at the end of the slotline radial stub. Because of these properties, these new transitions can easily be used as feeds for slotline antennas in phased array applications. As an extension of these microstrip-to-slotline transitions, new microstrip-to-coplanar strips (CPS) and microstrip-to-coplanar waveguide (CPW) transitions are demonstrated. The advantage of using the microstrip-to-CPW transition is that there is no need to use via-holes and/or air-bridges to connect the two transmission lines as the case with the transitions reported in [9]–[11]. All of these transitions are studied both theoretically, using the finite difference time domain (FDTD) method, and experimentally.

II. Design Description

Fig. 1(a) shows a back-to-back uniplanar microstrip to slot-line transition geometry. It consists of a 50 Ω microstrip line which branches into two orthogonal paths. The characteristic impedance of each microstrip path is chosen as 70 Ω for easy fabrication. Ideally, the two microstrip paths should be made 0.75 and 0.25 $\lambda_{g(\mathrm{microstrip})}$ long at a specific design frequency so that the fields at the locations (a) and (b) are 180° out of phase. This is necessary to excite the odd mode on the intermediate coupled microstrip line ($Z_0 = 48\,\Omega$) which transitions easily into a slot-line [7]. The configuration shown in Fig. 1(a) has the two transitions on the same side of the slotline and will be called an unbalanced configuration in contrast to the balanced configuration shown in Fig. 1(b) where the two transitions are located on opposite sides of the slotline. The latter configuration has the advantage that surface waves excited by the discontinuities associated with each transition may cancel each other. In practice, right angle bends parasitics contribute an additional phase change which

Manuscript received March 1, 1995; revised July 10, 1995. This work was supported by the Army Research Office.

N. I. Dib was with the Electrical Engineering and Computer Sciences Department, University of Michigan, Ann Arbor, MI USA. He is now with the Jordan University of Science and Technology, Jordan.

L. P. B. Katehi is with the Radiation Laboratory, University of Michigan, Ann Arbor, MI 48109–2122 USA.

R. N. Simons is with the NASA Lewis Research Center, OH 44135 USA.

IEEE Log Number 9415468.

0018–9480/95$04.00

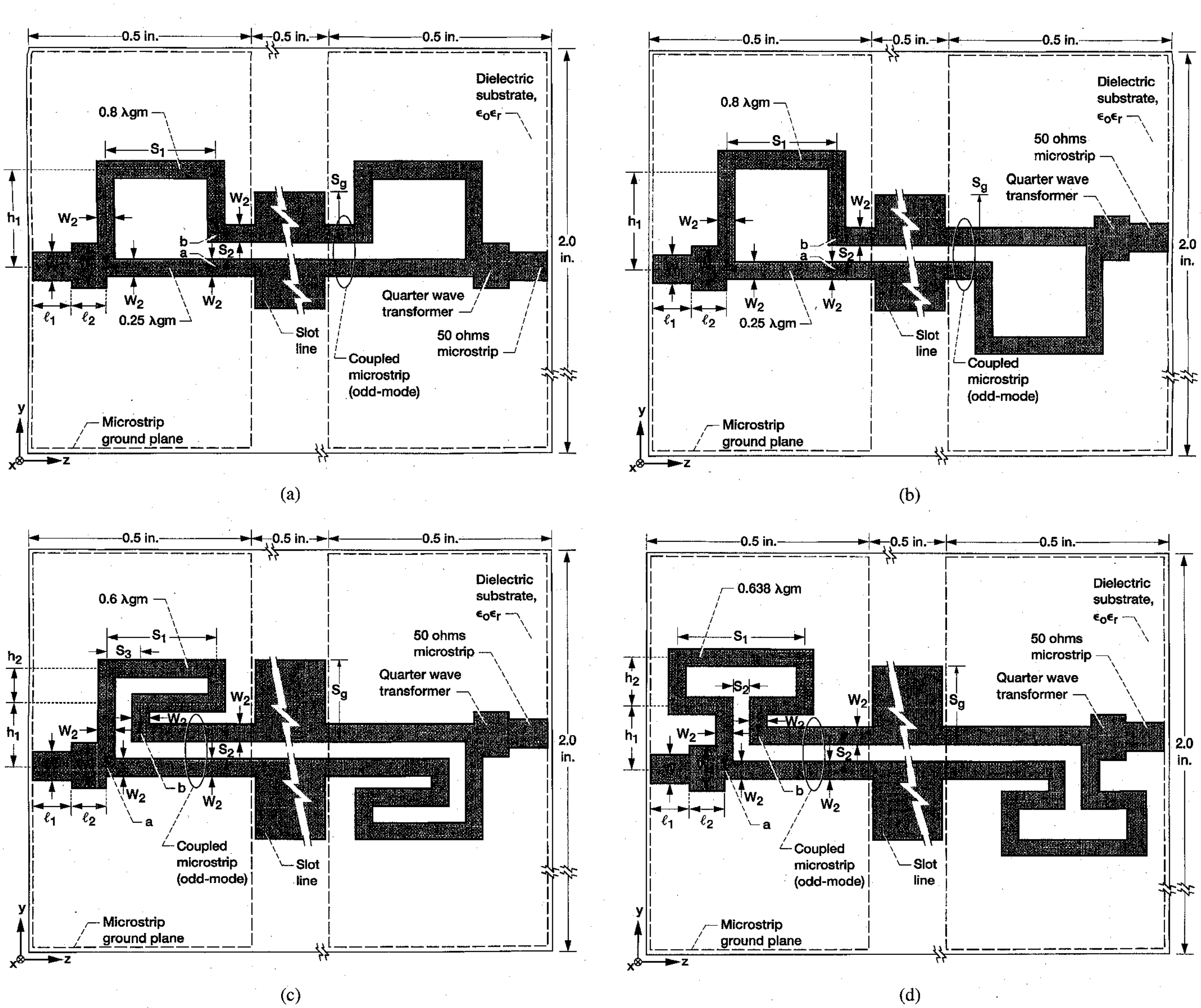


Fig. 1. Uniplanar back-to-back microstrip-to-slotline transitions (dimensions in inches): $W = 0.009, W_1 = 0.0129, W_2 = 0.004, \ell_2 = 0.1091, S_2 = 0.004, S_g = 0.3, \epsilon_r = 10.5$ and substrate thickness = 0.01. (a) unbalanced configuration: $S_1 = 0.1168, h_1 = 0.1285$ (b) balanced configuration: $S_1 = 0.1168, h_1 = 0.1285$ (c) balanced folded transition: $S_1 = 0.0855, S_3 = 0.014, h_1 = 0.0353, h_2 = 0.0273$ (d) balanced folded transition: $S_1 = 0.0826, h_1 = 0.0213, h_2 = 0.0527$.

may be adjusted by slightly changing the lengths of the microstrip lines. Hence, in this paper, the two microstrip paths are chosen to be of lengths 0.8 and 0.25 $\lambda_{g(\text{microstrip})}$ at the design frequency of 10 GHz.

Fig. 1(c) and (d) show two balanced back-to-back folded microstrip-to-slotline transition topologies. In these transitions, the 0.25 $\lambda_{g(\text{microstrip})}$ line is eliminated and the delay path is made 0.6 and 0.638 $\lambda_{g(\text{microstrip})}$ for the transitions in Fig. 1(c) and (d), respectively, at the design frequency of 10 GHz to provide the 180° phase difference and to compensate for the right angle bends parasitics. The rest of the design is the same as above. These compact transitions offer as much as 45% reduction in size over the one shown in Fig. 1(a).

Impedance matching at the junction between the 50 Ω microstrip line and the two 70 Ω microstrip lines is achieved by a 41.8 Ω quarter wave impedance transformer. The transitions are fabricated on a 1.5 × 2.0 inch RT-Duriod6010.5 of 10 mils thickness and ϵ_r of 10.5. The measurements were done using an HP8510C Network Analyzer with the Wiltron Universal test fixture.

As an extension of the above design, a microstrip-to-CPS transition is realized by allowing the slotline ground plane width S_g to approach the strip width W_2. Furthermore, a uniplanar microstrip-to-CPW transition is realized by having the input microstrip line divide into two orthogonal paths as shown in Fig. 2. In Fig. 2, the electric field lines between (a) and (b) and (a) and (c) are 180° out of phase and consequently excites the coplanar waveguide mode. The longer path length is taken to be around 0.78 $\lambda_{g(\text{microstrip})}$ in order to compensate for the right angle bends parasitics.

III. FDTD Method

In this method, Maxwell's curl equations are expressed in discretized space and time domains and are then used to

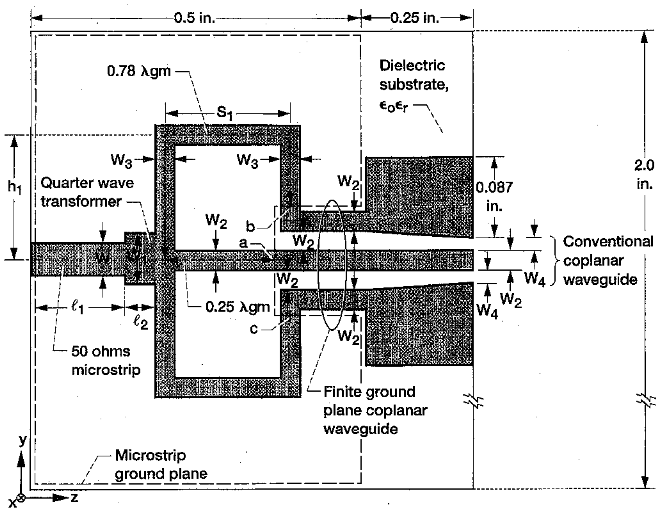


Fig. 2. Uniplanar microstrip-to-CPW transition (dimensions in inches): $W = 0.009, W_1 = 0.0184, W_2 = 0.01, W_3 = 0.004, \ell_2 = 0.107, S_1 = 0.1168, h_1 = 0.1297, \epsilon_r = 10.5$ and substrate thickness = 0.01.

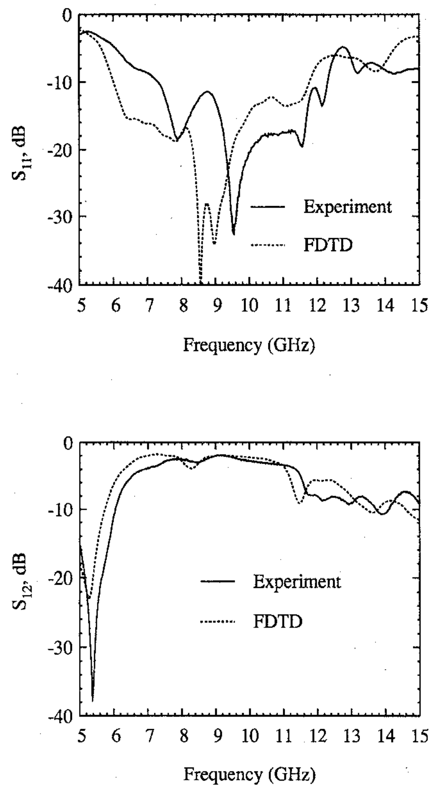


Fig. 3. S-parameters of the unbalanced back-to-back transition shown in Fig. 1(a).

simulate the propagation of an initial excitation in a "leapfrog" manner [12]–[14]. In order to characterize any planar discontinuity, propagation of a specific time-dependent function, usually a Gaussian pulse, through the structure is simulated using the FDTD technique. The space steps, $\Delta x, \Delta y$, and Δz, are carefully chosen such that integral numbers of them can approximate the various dimensions of the structure. As a rule of thumb and in order to reduce the truncation and grid dispersion errors, the maximum step size is chosen to be less than 1/20 of the smallest wavelength existing in the computational domain (i.e., at the highest frequency represented in the pulse). Then, the Courant stability criterion is used to select the time step to insure numerical stability. For the transitions analyzed here, the following parameters are used: $\Delta x = \Delta y = 50.8\,\mu\mathrm{m}, \Delta z = 101.6\,\mu\mathrm{m}$, and $\Delta t = 0.1128$ ps. The super-absorbing first-order Mur boundary condition is utilized to terminate the FDTD lattice at the front and back planes in order to simulate infinite lines. On the other hand, the first-order Mur boundary condition is used on the other walls to simulate an open structure.

It is worth mentioning that the two microstrip ground planes, shown in Fig. 1(a), were connected by a thin metal strip in the FDTD simulations to insure that they are at the same potential. On the other hand, it is found experimentally that the results are the same whether they are connected or not. This is due to the fact that both ports are connected to coaxial lines in the experiments which insure that both ground planes are at the same potential. However, in the FDTD simulation, only one port is excited while the other is matched to an absorber which makes it necessary to connect the two ground planes.

IV. Results and Discussion

Fig. 3 shows the S-parameters of the unbalanced back-to-back configuration shown in Fig. 1(a). The overall agreement between the experimental and theoretical results is satisfactory considering the differences between the theoretically analyzed structure and the experimentally characterized one. Specifically, the FDTD method considered a transition printed on an infinite substrate (due to the use of absorbing boundary conditions) while the experimentally measured transition was printed on a finite substrate. Furthermore, the FDTD method implemented a slightly different transition geometry due to limitations of the uniform grid in discretizing arbitrary dimensions without introducing an impossibly large number of nodes. The measurement data are calibrated to the connectors and include the connector insertion loss which amounts to 0.2 dB per connector. The measured insertion loss of the back-to-back configuration is better than 3 dB in the frequency range 7.5–10.5 GHz and translates into 1.3 dB per transition over a 44% bandwidth around a center frequency of 9 GHz.

Fig. 4 shows the S-parameters of the balanced back-to-back configuration shown in Fig. 1(b). The measured insertion loss is better than 3 dB, that is 1.3 dB per transition, from 7–11.5 GHz (49% bandwidth around a center frequency of 9.25 GHz). The minimum insertion loss achieved by a single transition is 0.6 dB at 10.5 GHz.

Fig. 5 shows the S-parameters of the balanced back-to-back configuration shown in Fig. 1(c). It can be seen that this transition has a narrower bandwidth than the previous ones. This is due to the existence of higher parasitics due to more right angle bends in the longer microstrip path. Specifically, the measured insertion loss is better than 3 dB from 8.625–10.35 GHz (18% bandwidth around a center frequency of 9.5GHz). Fig. 6 shows the S-parameters of the balanced back-to-back configuration shown in Fig. 1(d). It can be seen that this transition has a narrower bandwidth than the one shown in Fig. 1(c).

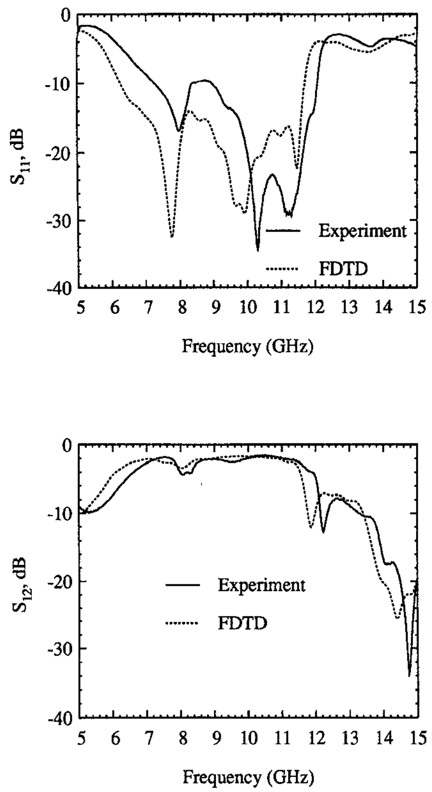


Fig. 4. S-parameters of the balanced back-to-back transition shown in Fig. 1(b).

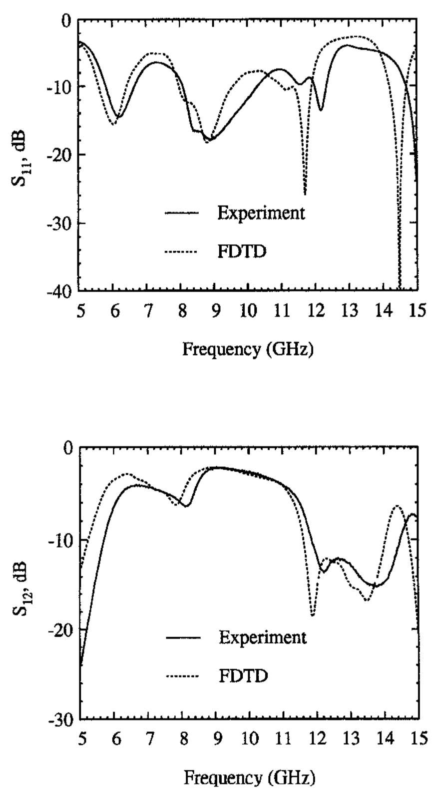


Fig. 5. S-parameters of the balanced back-to-back transition shown in Fig. 1(c).

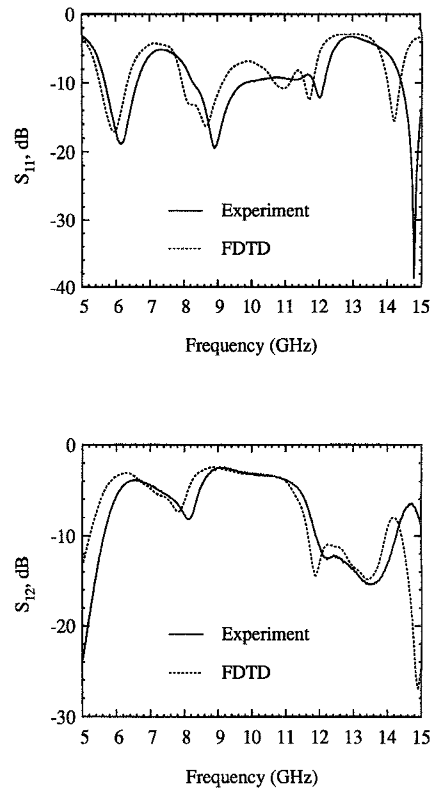


Fig. 6. S-parameters of the balanced back-to-back transition shown in Fig. 1(d).

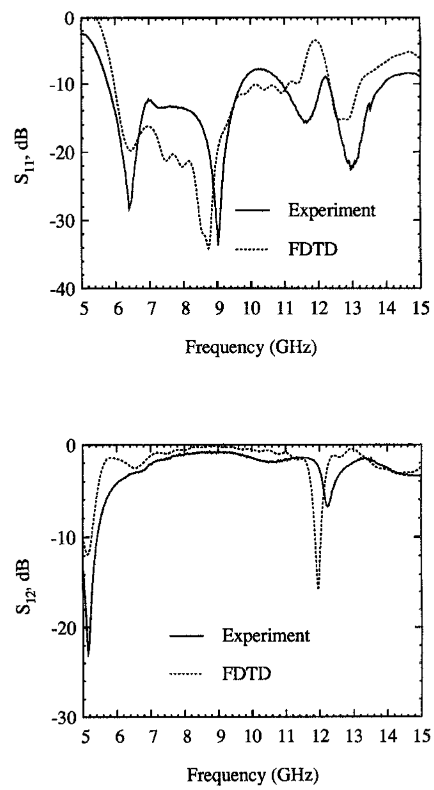
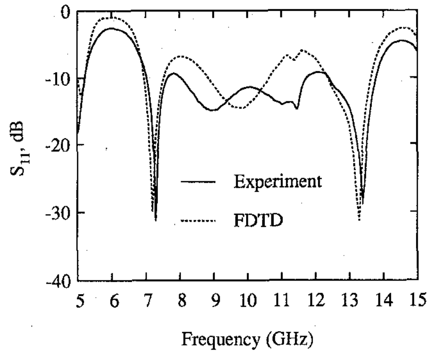


Fig. 7. S-parameters of a balanced back-to-back microstrip-to-CPS transition. The structure is the same as that shown in Fig. 1(b) except that $S_g = 0.03$ inches.

Fig. 7 shows the S-parameters of the balanced back-to-back microstrip-to-CPS transition with a coplanar strip width of 760 μm. The measured 3 dB insertion loss bandwidth is 59% around 9.3 GHz.



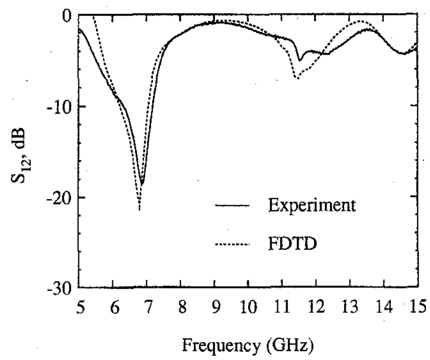


Fig. 8. S-parameters of a back-to-back microstrip-to-CPW transition (see Fig. 2 for a single transition).

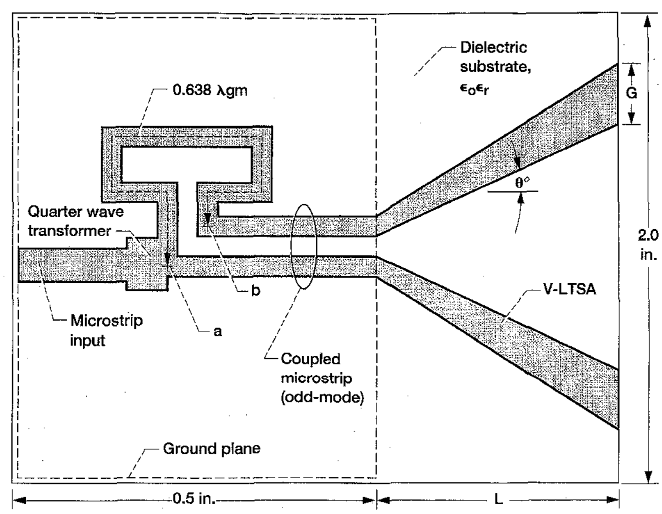


Fig. 9. A linearly tapered slotline antenna fed by a folded microstrip-to-slotline transition.

Fig. 8 shows the S-parameters of a back-to-back microstrip-to-CPW transition shown in Fig. 2. It is interesting to note that a dip as small as the one seen around 11.5 GHz has been captured by both FDTD and experiment. The measured 3 dB insertion loss bandwidth is approximately 40% around 9.2 GHz and the minimum insertion loss achieved by a single transition is 0.25 dB at 9.2 GHz.

It should be mentioned that while experimentally characterizing all of the transitions, a piece of microwave absorber was placed at the edges of the substrate to absorb the surface waves launched by the microstrip line discontinuities. Without the absorber, deep nulls in the insertion loss measurements were observed. Also, as mentioned earlier, the measured data include the losses from the coaxial connectors of the Wiltron Universal test fixture which were not practical to calibrate out.

From the above results, it can be seen that there is always a small frequency shift between the theoretically and experimentally derived results. This may be attributed to the following: a) The metallization thickness is not taken into consideration in the FDTD simulations; b) the length of the connectors used in the experiments which were not calibrated out; and c) the real microstrip lines widths may not be the same as those simulated due to fabrication errors.

Fig. 9 shows a possible application of one of the proposed transitions in which it is used as a feed to a linearly tapered slotline antenna (LTSA). Such an antenna is the subject of study in [15].

V. Conclusion

New uniplanar microstrip-to-slotline transitions, microstrip-to-CPW transition and microstrip-to-CPS transition have been proposed and characterized both experimentally and theoretically. Such transitions are compact and easily adaptable to be used in MIC/MMIC and antenna applications. The response of the transitions is evaluated numerically using FDTD method and experimentally from 5–15 GHz and both results are in good agreement. For a single microstrip-to-slotline transition, an insertion loss better than 1.3 dB is achieved over a 49% bandwidth with a minimum insertion loss of 0.6 dB. On the other hand, a single microstrip-to-CPW transition achieved a 1.3 dB insertion loss bandwidth of 40% with a minimum insertion loss of 0.25 dB.

References

[1] K. Gupta, R. Garg, and I. Bahl, *Microstrip Lines and Slotlines.* Norwood, MA: Artech House, 1979.

[2] P. Soltysiak and J. Chramiec, "Design of broadband transitions from microstrip to slotline," *Electron. Lett.*, vol. 30, no. 4, pp. 328–329, Feb. 1994.

[3] Y. Antar, A. Bhattcharyya, and A. Ittipiboon, "Microstripline-slotline transition analysis using the spectral domain technique," *IEEE Trans. Microwave Theory and Tech.*, Mar. 1992, pp. 515–523.

[4] N. El-Minyawi, "A new microstrip-slotline transition," *Microwave J.*, vol. 26, no. 10, pp. 140–141, Oct. 1983.

[5] ——, "Single sided slotline microstrip transition," Technical Memorandum, *IEE Proc., Pt. H*, vol. 134, no. 1, pp. 101–102, Feb. 1987.

[6] J. Yook, N. Dib, L. Katehi, R. Simons, and S. Taub, "Theoretical and experimental study of microstrip-to-slot line uniplanar transition," in *1994 IEEE AP-S Int. Symp. Dig.*, pp. 1206–1209.

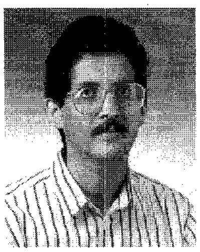
[7] R. Simons, S. Taub, R. Lee, and P. Young, "Microwave characterization of slot line and coplanar strip line on high-resistivity silicon for a slot antenna feed network," *Microwave and Optical Technology Lett.*, vol. 7, no. 11, pp. 489–494, Aug. 1994.

[8] T. Ho and S. Hart, "A novel uniplanar microstrip to slot-line transition," *Microwave and Optical Technology Lett.*, vol. 7, no. 13, pp. 619–620, Sept. 1994.

[9] B. Golja, H. Sequeira, G. Mendenilla, and N. Byer, "A coplanar-to-microstrip transition for W-band circuit fabrication with 100-μm-thick GaAs wafers," *IEEE Trans. Microwave and Guided Wave Lett.*, Feb. 1993, pp. 29–31.

[10] J. Yook, N. Dib, and L. Katehi, "Characterization of high frequency interconnects using finite difference time domain and finite element methods," *IEEE Trans. Microwave Theory and Tech.*, Sept. 1994, pp. 1727–1736.

[11] J. Izadian and S. Izadian, *Microwave Transition Design.* Norwood, MA: Artech House, 1988, chap. 5.
[12] K. Kunz and R. Luebbers, *The Finite Difference Time Domain Method for Electromagnetics.* Boca Raton: FL: CRC Press, 1993.
[13] X. Zhang and K. Mei, "Time-domain finite difference approach to the calculation of the frequency-dependent characteristics of microstrip discontinuities," *IEEE Trans. Microwave Theory and Tech.*, Dec. 1988, pp. 1775–1787.
[14] D. Sheen, S. Ali, M. Abouzahra, and J. Kong, "Application of the three-dimensional finite-difference time-domain method to the analysis of planar microstrip circuits," *IEEE Trans. Microwave Theory and Tech.*, July 1990, pp. 849–857.
[15] R. Simons, N. Dib, and L. Katehi, "Integrated transition for linearly tapered slot antenna," *IEEE Trans. Antenna Propagat.*, Special Issue on Packaging Technologies for Phased Array Applications, pp. 998–1002, Sept. 1995.



Nihad I. Dib received the B.Sc. and M.Sc. degrees in electrical engineering from Kuwait University in 1985 and 1987, respectively, and the Ph.D. degree in electrical engineering from the University of Michigan, Ann Arbor, in 1992.

From 1993 to 1995, he was an Assistant Research Scientist at the Radiation Laboratory at the University of Michigan, Ann Arbor. In 1995, he joined the Electrical Engineering Department at Jordan University of Science and Technology as an Assistant Professor. His research interests include the numerical analysis and modeling of planar discontinuities and dielectric lines.

Rainee N. Simons (S'76–M'80–SM'89) received the B.S. degree in electronics and communications engineering from the Mysore University, India, in 1972, and the M.Tech. degree in electronics and communications engineering from the Indian Institute of Technology, Kharagpur, in 1974. In 1983 he received the Ph.D. degree in electrical engineering from the Indian Institute of Technology (IIT), New Delhi.

He started his career in 1979 as a Senior Scientific Officer at IIT, New Delhi. At IIT he worked on fin-line components for mm-wave applications and also on toroidal latching ferrite phase shifters for phased arrays. From 1985 to present, he has been with the Space Electronics/Communications Division of NASA Lewis Research Center, Cleveland, Ohio, as a National Research Council Research Associate (1985 to 1987), Case Western Reserve University Research Associate (1987 to 1990), Senior Engineer with Sverdrup Technology, Inc. (1990 to 1993), and Senior Engineer with NYMA, Inc., (1994 to present). At NASA he worked on optical control of MESFET and HEMT devices, high temperature superconductivity, modeling of coplanar waveguide (CPW) discontinuities, CPW feed systems for printed antennas, linearly tapered slot antenna arrays for communications, packaging of MMICs and multichip modules with integrated radiators for phased arrays. He is the author of a book entitled "Optical Control of Microwave Devices," published by Artech House, Inc. He has also authored a chapter entitled "High-Temperature Superconducting Coplanar Waveguide Microwave Circuits and Antennas" which has appeared in "Advances in High-T_c Superconductors" published by Trans Tech Publications, Switzerland.

Dr. Simons has received the distinguished alumni award from his alma mater and several NASA Tech Brief and Group Achievement awards. He served as a member of the Technical Program Committee for the IEEE International Antennas and Propagation Symposium in 1993. He organized a Special Session on "Advances in MMIC Packaging for Phased Array Antennas" during the 1993 Antennas and Propagation International Symposium held at Ann Arbor, MI. He is a member of the Editoral Board of the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES.

Linda P. B. Katehi (S'81–M'84–SM'89–F'95) received the B.S.E.E. degree from the National Technical University of Athens, Greece, in 1977 and the M.S.E.E. and Ph.D. degrees from the University of California, Los Angeles, in 1981 and 1984, respectively.

In September 1984 she joined the faculty of the EECS Department of the University of Michigan, Ann Arbor. Since then she has been interested in the development and characterization (theoretical and experimental) of microwave, millimeter printed circuits, the computer-aided design of VLSI interconnects, the development and characterization of micromachined circuits for millimeter-wave and submillimeter-wave applications and the development of low-loss lines for Terahertz-frequency applications. She has also been studying theoretically various types of uniplanar radiating structures for hybrid-monolithic and monolithic oscillator and mixer designs.

Dr. Katehi was awarded the IEEE AP-S W. P. King Award in 1984, the IEEE AP-S S. A. Schelkunoff Award in 1985, the NSF Presidential Young Investigator Award and an URSI Young Scientist Fellowship in 1987, the Humboldt Research Award and The University of Michigan Faculty Recognition Award in 1994. She is a Fellow of IEEE, and an Associate Editor of the IEEE Antennas and Propagation Society, and Radio Science. She is a member of IEEE AP-S, MTT-S Sigma Xi, URSI Commission D and an elected member of the IEEE Antennas and Propagation Society Administrative Committee. She has graduated seven Ph.D. students and is presently supervising 15 Ph.D. graduate students. She is the author and co-author of more than 220 papers published in referred journals and symposia proceedings.