Piezoelectric-Transducer-Controlled Tunable Microwave Circuits

Tae-Yeoul Yun, *Student Member, IEEE*, and Kai Chang, *Fellow, IEEE*

***Abstract*—This paper introduces a new method to tune microwave circuits of phase shifters, filters, resonators, and oscillators, controlled by a piezoelectric transducer (PET) with computational and experimental results. An optimized PET-controlled phase shifter is demonstrated to operate up to 40 GHz with a maximum total loss of 4 dB and phase shift of 480°. PET-controlled tunable bandpass filter, ring resonator, and one-dimensional photonic-bandgap resonator show a very wide tuning bandwidth of 17.5%–28.5% near 10 GHz with little performance degradation. A new PET-controlled or voltage-controlled dielectric-resonator oscillator (DRO) is demonstrated with a tuning bandwidth of 3.7% at the center frequency of 11.78 GHz. The tuning bandwidth is slightly less than that of a mechanical tuning using a micrometer-head-controlled tunable DRO with a tuning bandwidth of 4.7%. The new tuning method should have many applications in monolithic and hybrid microwave integrated circuits.**

***Index Terms*—DRO, PBG, perturbation, phase shifter, piezoelectric, resonator, tunable filter, VCO.**

I. Introduction

TUNABLE microwave circuits can be realized in many techniques, but whatever the method of tuning may be, they must conserve as much as possible their original S-parameter magnitudes or output characteristics over a tuning range [1]. Most tunable microwave circuits described in the literature fall into four basic types [2], i.e., mechanically [3]–[5], magnetically [6], [7], ferroelectrically [8]–[10], electrically [11]–[15], and optically [16], [17] tunable circuits. In addition, recently reported are microelectromechanical systems (MEMS) [18], [19], mechanical tuning with electrically controlled methods using electrostatic forces. A new method introduced in this paper is also a kind of electromechanical tuning method using a piezoelectric transducer (PET) or actuator [20]–[22]. The PET is a piezoelectric ceramic, deflected by an applied voltage [23]. The new PET tuning method will be demonstrated for the largest tuning range to date in a bandpass filter, resonator, and dielectric-resonator oscillator (DRO) based on a low-loss PET-controlled phase-shifter concept.

A wide-band and low-loss phase shifter is an important component in a phased array for beam steering and beam forming, timing recovery circuits, phase equalizers for data channels, etc. Published results [10], [14], [16] were narrow-band, lossy, or provided small phase shift. A new phase shifter was recently presented using dielectric perturbation on the microstrip line, electrically controlled by the PET [20]. The PET-controlled dielectric layer perturbs the electromagnetic fields of microstrip line. The dielectric perturber attached to the PET can be easily moved in z-axis direction, as shown in Fig. 1(a). In this paper, an optimized PET phase shifter will be presented with the largest phase shift with a low loss to date. In addition, from the PET-controlled phase-shifter idea, a new tunable bandpass filter and ring resonator are presented on the microstrip line with simulation and measurement data. The dielectric perturber produces a shielding effect on the microstrip line for reducing the radiation loss [24].

Recently, one-dimensional (1-D) photonic-bandgap (PBG) resonators have been reported using varactors for electronic tuning [25]. This resonator is based on a Fabry–Perot resonator consisting of a center resonant line with two sides of PBG reflectors. Even though a flip-chip varactor having very small parasitic effects was used, the quality (Q) factor was still degraded. In this paper a new electronically tunable 1-D PBG resonator is proposed using the PET-controlled perturber over the resonant line. The dielectric or metal perturber attached under the PET can perturb the electromagnetic fields at the center of the resonant line. This changes the line capacitance, which is exactly the same effect as with a varactor. Thus, the distance of the air gap between the perturber and resonant line determines the operating frequency. This new method is successfully demonstrated with wide-band tuning of the resonant frequency with little degradation of the Q factor.

The DRO is widely used for stable microwave sources because of its good temperature stability, small size, compactness, low price, etc. In addition, the electronically tunable DRO or voltage-controlled dielectric-resonator oscillator (VCDRO) has many applications in communication and radar systems and frequency-hopping spread-spectrum circuits. Several methods have been used to tune the resonant frequency of dielectric resonators (DRs) [2] and DROs, i.e., mechanical [5], ferrite, varactor [12], [13], p-i-n diode, and optical [17]. The mechanical method using a dielectric disc perturbation was able to maintain good performance over a wide tuning range [2]. The oscillation frequency of a DRO can be tuned by perturbing the DR's electromagnetic fields. This tuning can be achieved by varying the air gap between the DR and the dielectric (or metallic) disc above the DR using the PET. The PET-controlled VCDRO is demonstrated to obtain the largest tuning range up to date.

Manuscript received February 7, 2001; revised June 6, 2001. This work was supported in part by the National Science Foundation and by the U.S. Air Force.

T.-Y. Yun was with the Department of Electrical Engineering, Texas A&M University, College Station, TX 77843-3128 USA. He is now with TriQuint Semiconductor Inc., Dallas, TX 75083 USA.

K. Chang is with the Department of Electrical Engineering, Texas A&M University, College Station, TX 77843-3128 USA (e-mail: chang@ee.tamu.edu).

Publisher Item Identifier S 0018-9480(02)04059-0.

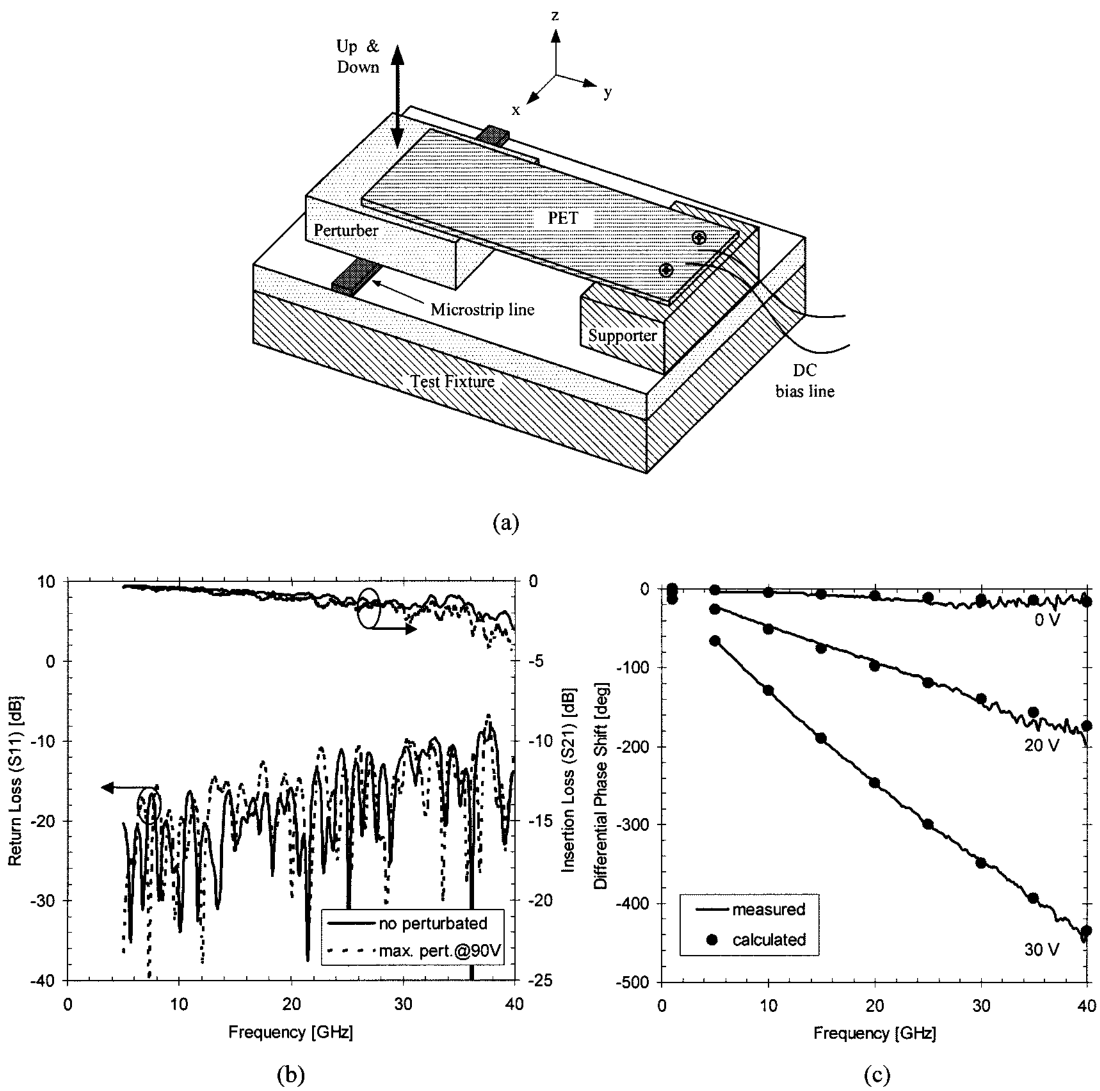


Fig. 1. PET-controlled phase shifter. (a) Configuration. (b) S-parameters. (c) Phase shifts versus frequencies with different PET voltages with a substrate of $\varepsilon_r = 10.8$, thickness $= 10$ mil, microstrip width $= 5$ mil, and perturber of $\varepsilon_r = 10.8$ and length $= 1.2$ in.

II. Formulation of Frequency-Dependent Phase Shift

A. Tunable Phase Shifter

The new phase shifter of Fig. 1(a) is consisted of microstrip line substrate (dielectric constant of ε_{r1}), air gap ($\varepsilon_{r2} = 1$), and perturber (ε_{r3}). To calculate the capacitance of perturbed microstrip-like transmission lines, the variational method was used [26]. The capacitance variations correspond to variations in effective dielectric constant, characteristic impedance (Z_c), and propagation constant. A dispersion formula is found from curve fitting [27], [28], which is used for optimizing the PET-controlled phase shifter to reduce the control voltage with smaller size and larger phase shift. The substrate optimized for microstrip-line is RT/Duroid 6010.8 with a dielectric constant of 10.8, thickness of 10 mil, and width of 5 mil. The dielectric perturber has a dielectric constant of 10.8, height of 50 mil, and length of 1.2 in. The microstrip linewidth of 5 mil is designed for a high Z_c of 64 Ω at 40 GHz to compensate for the decreased Z_c by dielectric perturbation. At maximum perturbation, i.e., when the dielectric perturber is placed on the microstrip line, Z_c is close to 50 Ω. This will maintain good impedance matching without any external matching circuit. The PET has a size of 2.75 in (length) $\times$ 1.25 in (width) $\times$ 0.085 in (thickness including a supporter) with a composition of lead–zirconate–titanate. This size makes a large capacitance of 290 nF and a relatively slow response time of 5 ms. Smaller size can be used to make a compact unit and improve the tuning time even though smaller phase shift or PET movement may be compromised to achieve this.

Thru-reflect-line (TRL) calibration was used to remove the coaxial connector-to-microstrip-line transition effect for S-parameters measurement of HP8510, but an imperfect calibration and/or surface wave generation caused a fluctuation in the insertion loss (S_{21}) near 37 GHz, as shown in Fig. 1(b). Except the fluctuation of S_{21}, the maximum perturbation added loss is less than 2 dB and, thus, total loss is less than 4 dB up to 40 GHz. The return loss (S_{11}) is less than -15 dB over all frequency range and about -10 dB near 40 GHz. Magnitude of S-parameters are not affected much by the dielectric perturbation.

Fig. 1(c) shows measured differential phase shifts with varying frequencies and PET voltages and how S_{21} of the microstrip line with PET-controlled perturber exhibits a phase

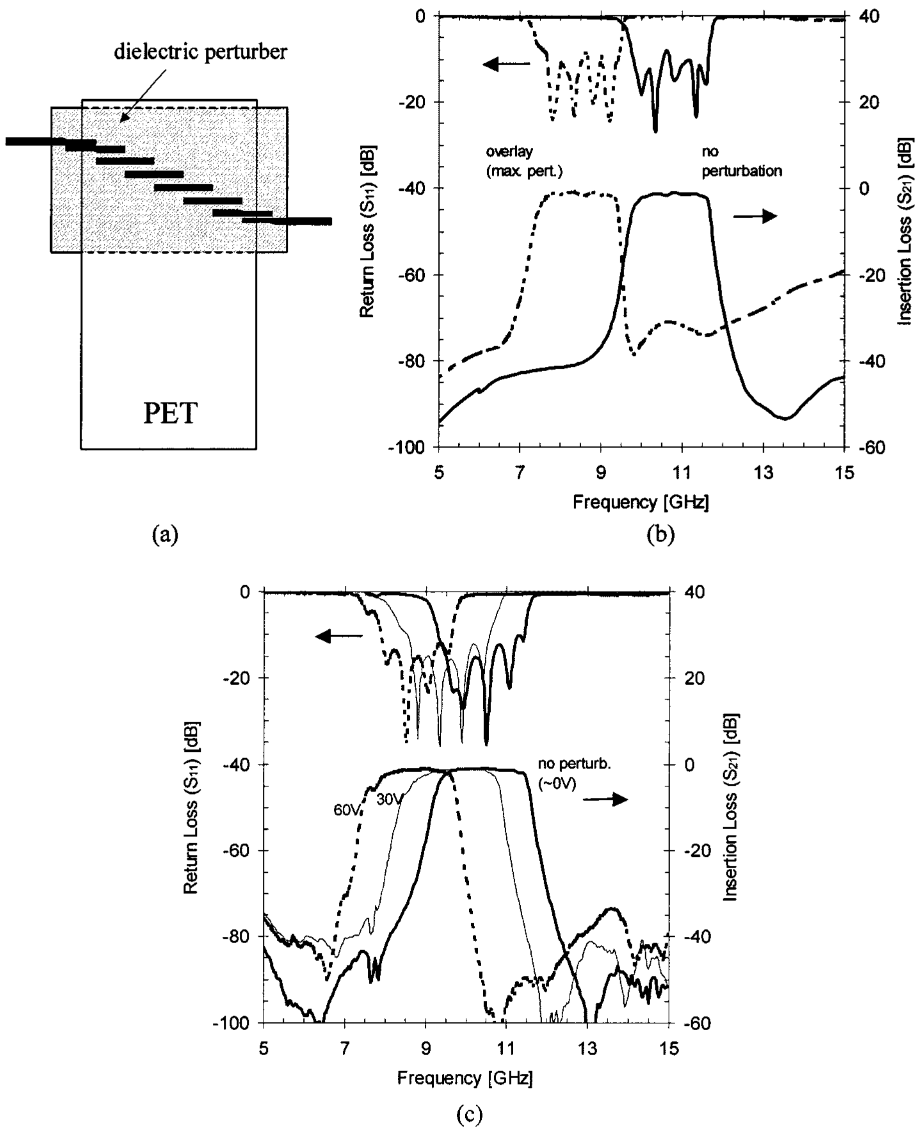


Fig. 2. Tunable bandpass filter. (a) Configuration—top view. (b) Simulated S-parameters. (c) Measured S-parameters with different PET voltages with a substrate of $\varepsilon_r = 10.8$, thickness $= 25$ mil, microstrip width $= 22$ mil, and perturber of $\varepsilon_r = 10.8$.

shift from $-450°$ to $-20°$ at 40 GHz with respect to the unperturbed condition. The amount of phase shift depends on PET deflection, which is controlled by varying the applied voltage from 0 to 30 V. There is no deflection (or no perturbation) at 0 V and full downward deflection (or maximum perturbation) at 90 V. This is called a top-down alignment, the reverse of the bottom-up alignment used in [20]–[22]. The calculated phase shift agrees very well with measured data at each PET control voltage. Between 30–90 V of the control voltage, the additional phase shifts becomes small. The phase shift at 90 V is almost saturated and shows $-480°$. Thus, the dc bias is required only up to 30 V. This dc voltage can be further decreased if the alignment is improved or a narrower microstrip line and thinner substrate are used.

The optimized PET phase shifter has three advantages compared with the previously reported results [20], [22]; the bias voltage range is reduced from 90 to 30 V, the size is reduced from 1.8 to 1.2 in, and the phase shift curves are more linear versus frequency, even though the S-parameters and phase-shifting performance are similar. In addition, phase-shift/insertion-loss ratio of 229°/dB at 25 GHz and 287°/dB at 35 GHz are achieved. The results are better than those reported in [15] and [19].

B. Tunable Bandpass Filter and Ring Resonator

A seven-section parallel-coupled microstrip bandpass filter was designed, as shown in Fig. 2(a). The dielectric perturber is large enough to cover the whole filter and attached under the PET. Details of design parameters for the filter are not described here, but can be found in [29]. The hand-calculated design is confirmed using IE3D, a moment-method full-wave electromagnetic simulator produced by Zeland Software Inc., Fremont, CA. In Fig. 2(b), multilayer simulation results of S-parameters are shown with and without perturbation, near the center frequency of 10 GHz. A wide tuning range of 2.315 GHz or 24% was achieved in the simulation without S-parameters degradation. The measured results are shown in Fig. 2(c) and agree very well with the simulation. The slight difference of

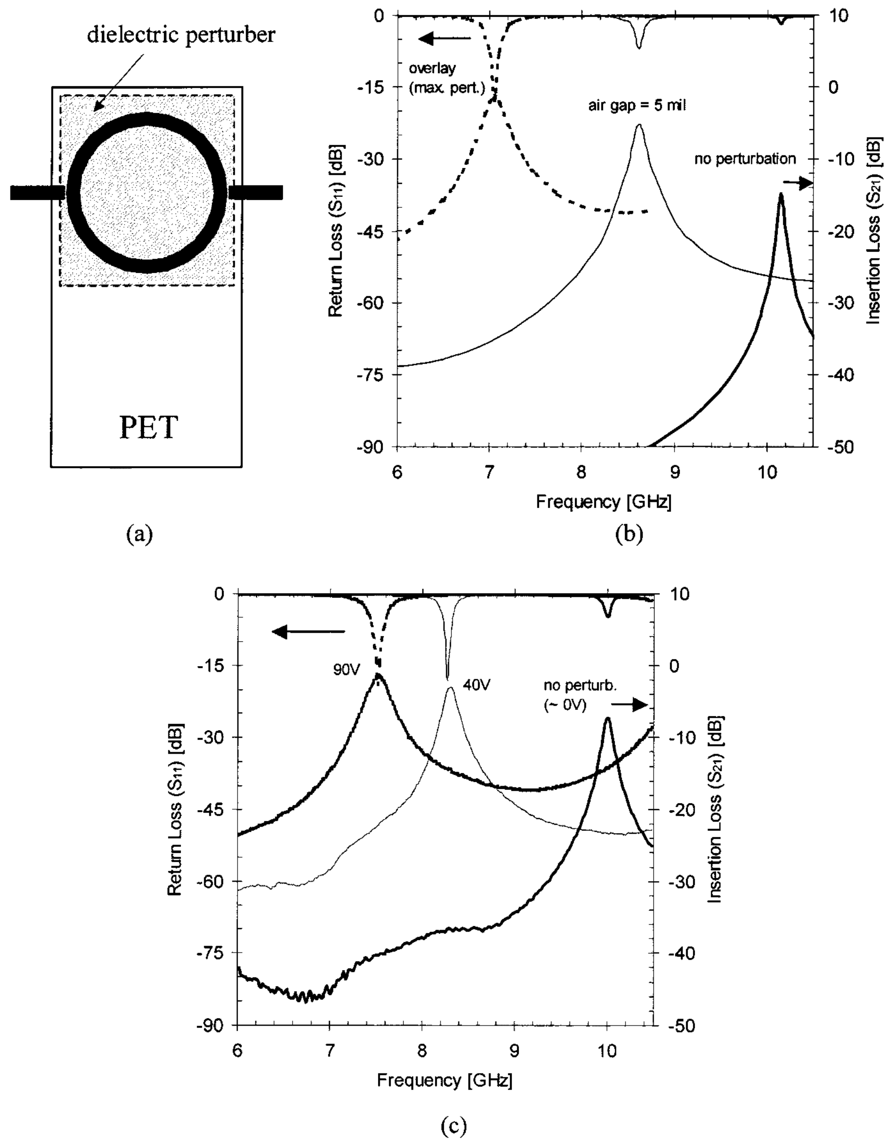


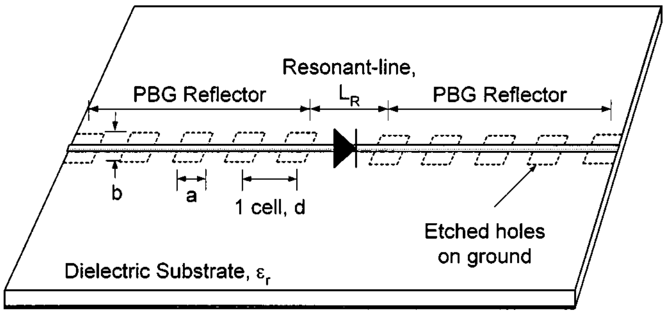
Fig. 3. Tunable ring resonator. (a) Configuration—top view. (b) Simulated S-parameters. (c) Measured S-parameters with different PET voltages with a substrate of $\varepsilon_r = 2.33$, thickness $= 20$ mil, microstrip width $= 50$ mil, ring coupling gap $= 5$ mil, and perturber of $\varepsilon_r = 10.8$.

filter responses between simulation and measurement may be caused by fabrication inaccuracy and long bent 50-Ω input and output microstrip lines to balance two ports on a test fixture that were not considered in the simulation. A measured tuning range is 1.675 GHz or 17.5%, which is slightly smaller than the simulated result due to nonperfect overlay or perturbation. The tuning range should be increased if the filter substrate is designed using a lower dielectric-constant material. This idea is demonstrated in the following PET tuned ring resonator.

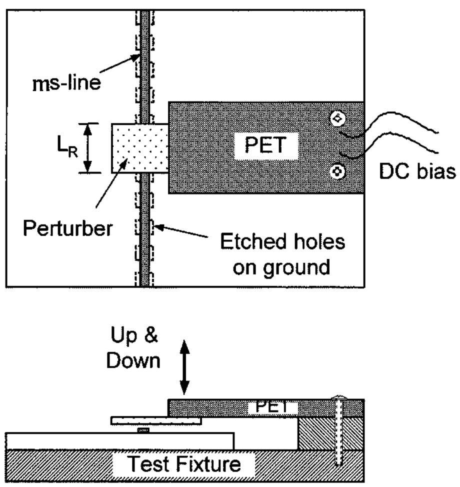
In Fig. 3, the PET tuning feasibility for the microstrip ring resonator is demonstrated with simulation and measurement. As mentioned above, a lower dielectric constant of $\varepsilon_r = 2.33$ was used for the microstrip substrate to increase the tuning range. The second resonant frequency of the ring resonator was designed near 10 GHz and tuned over 3.1 GHz or 36% with maximum perturbation in the simulation, as shown in Fig. 3(b). The measured results in Fig. 3(c) show a tuning range of 2.5 GHz or 28.5%; the range is greatly improved by using the lower permittivity material substrate. Although not shown, experiments indicated a tuning range of 1.35 GHz or 14.7% with a microstrip substrate of $\varepsilon_r = 10.8$. Note that the radiation and mismatch losses are gradually reduced by overlaying of dielectric perturber, as shown in Fig. 3(b) and (c), and as suggested in [24]. In addition, as a part of explanation, the dielectric overlay increases the capacitance between the ring and open-ended microstrip line, thus, increasing the coupling.

C. Tunable 1-D PBG Resonator

The 1-D PBG structure consists of a 50-Ω microstrip conductor line and etched square holes on the ground plane, as shown in Fig. 4(a). This has similar results with an alternated high–low characteristic impedance line or with air holes drilled through the substrate along the microstrip line [25]. When a varactor is mounted in a resonant line, the variable capacitance is used to change the resonant frequency (f_o). The substrate used is a RT/Duroid 6010.5 with dielectric constant of 10.5, length of 2 in, thickness of 25 mil, and linewidth of 22 mil. Slot size $(a = b)$ of 100 mil, spacing (d) of 200 mil, and resonant line



(a)



(b)

Fig. 4. Tunable 1-D PBG resonator configurations. (a) Varactor tuned. (b) PET tuned.

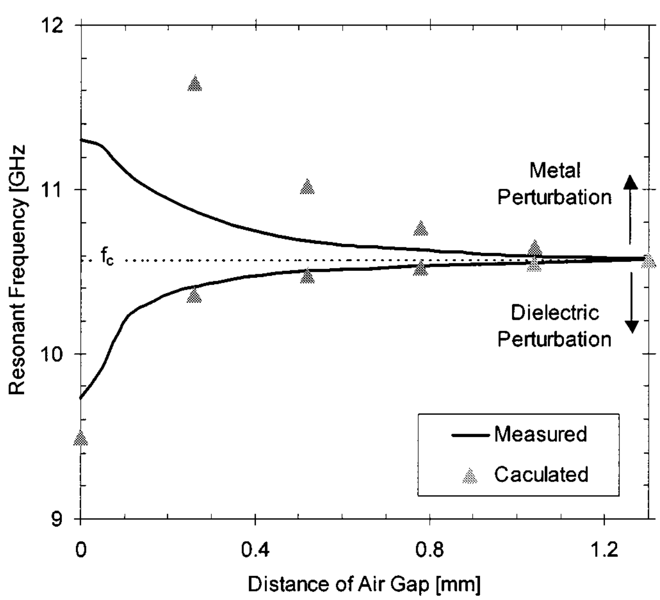


Fig. 5. Micrometer head controlled perturbation on 1-D PBG resonator, to compare measurement with calculation data.

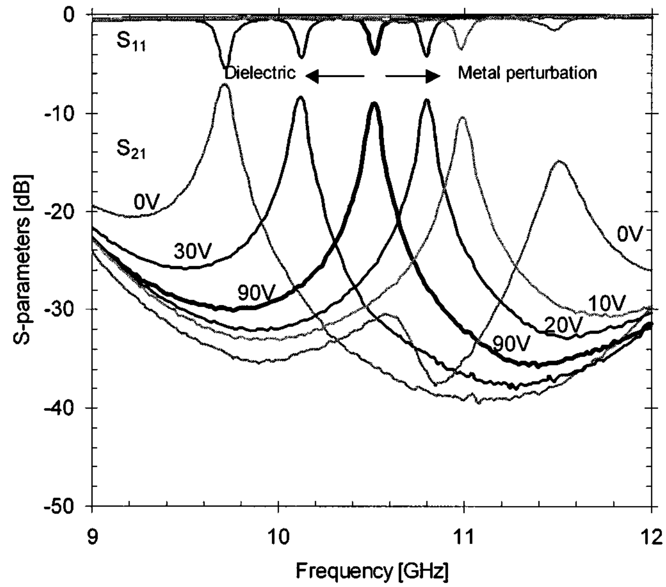
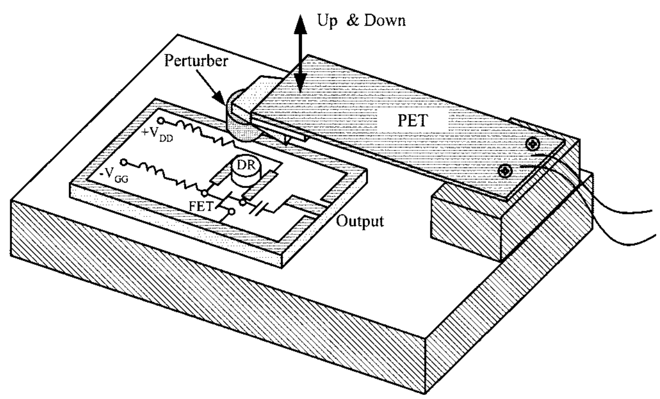
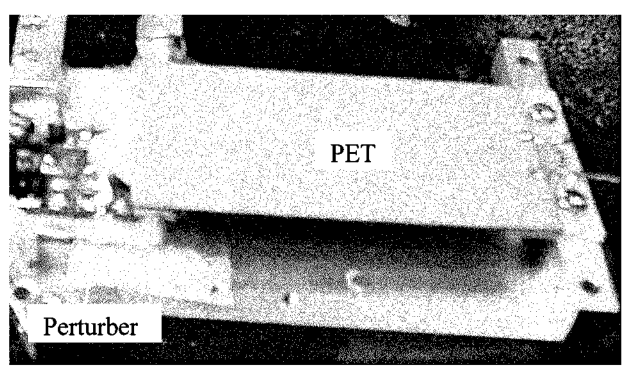


Fig. 6. PET-controlled tunable 1-D PBG resonator, measured S-parameters with different PET voltages with a substrate of $\varepsilon_r = 10.8$, thickness $= 25$ mil, microstrip width $= 22$ mil, and perturber of $\varepsilon_r = 10.8$ and thickness $= 50$ mil.



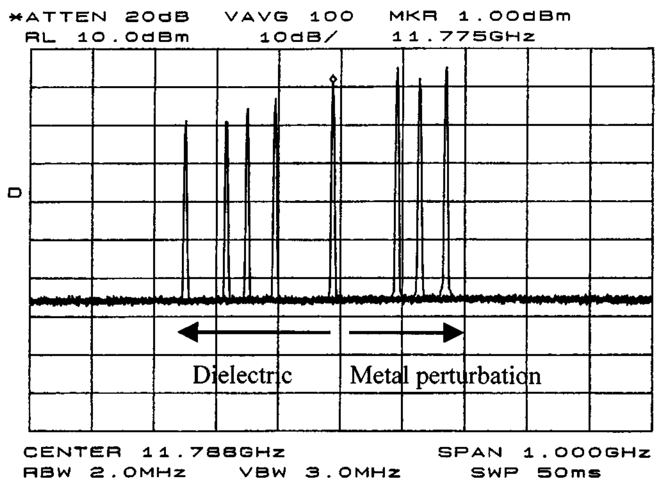
(a)



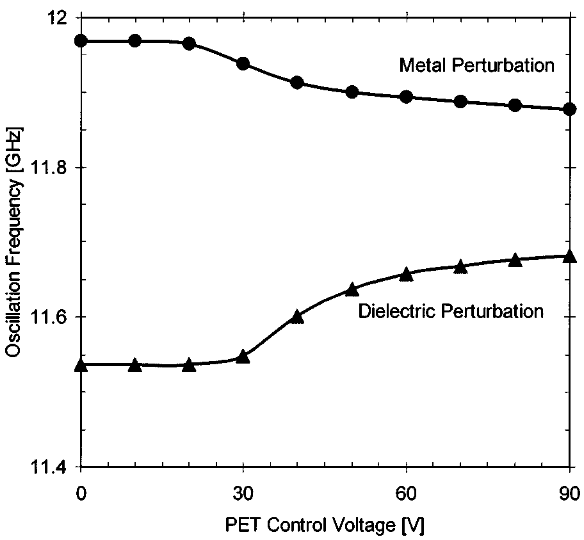
(b)

Fig. 7. PET-controlled VCDRO. (a) Configuration. (b) Photograph.

length $(L_R = 1.5d)$ of 300 mil are chosen. As a result of varactor-tuned PBG resonator, a wide-band tuning capability of approximately 20% near 10 GHz is measured with a maximum loaded-Q (Q_L) of only 77. The low intrinsic Q of the varactor near 10 GHz is the main reason for the low Q_L. With the same structure of a 1-D PBG resonator, the PET is used instead of the varactor to change the resonant frequency by perturbation of the electromagnetic fields on the resonant line, as shown in



(a)



(b)

Fig. 8. PET-controlled VCDRO results with the dielectric and metal perturbation. (a) Spectrum analyzer display of tuning the oscillation frequency. (b) Oscillation frequencies versus PET control voltages.

Fig. 4(b). The air-gap distance between the perturber and resonant line is controlled by varying the dc bias of the PET from 0 to 90 V.

Both metal and dielectric perturbations can be used to tune the 1-D PBG resonator. To investigate effects of metal and dielectric perturbers on the resonant frequency, a variational analysis [26] was developed and compared with measured results, as shown in Fig. 5. A micrometer head was used for measuring the accurate distance of the air gap. The resonant frequency depends on the effective length of resonant line, which can be easily calculated from the line capacitance and effective dielectric constant obtained from variational analysis. As can be seen from Fig. 5, metal perturbation increases the resonant frequency, whereas dielectric perturbation decreases the resonant frequency. This is because metal perturbation decreases effective dielectric constant and dielectric perturbation increases effective dielectric constant of the perturbed resonant. The calculated and measured results of dielectric perturbation agree very well. However, metal perturbation results are somewhat different, which

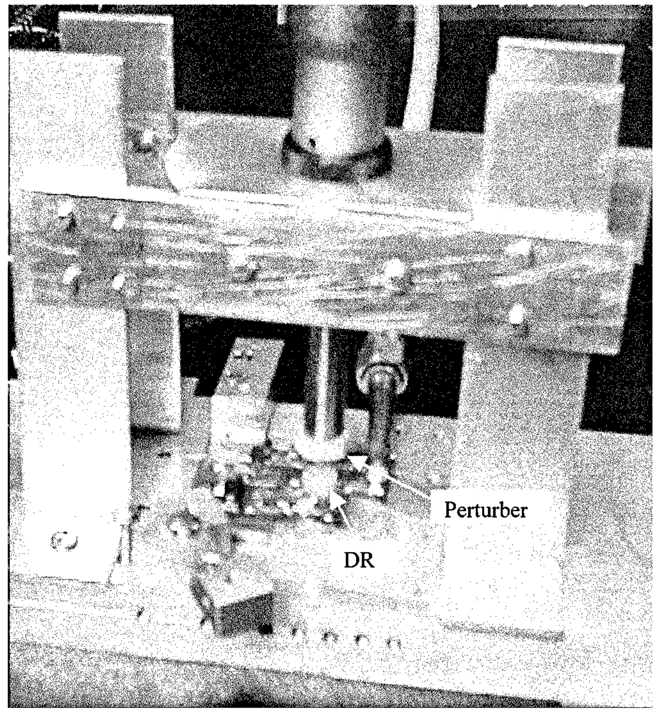


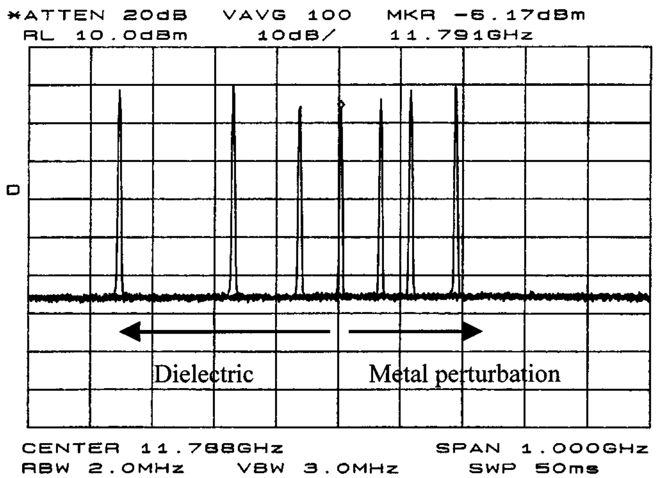
Fig. 9. Photograph of mechanical tuning using a micrometer-head controlled DRO.

is caused by the fact that a quasi-TEM of microstrip line is so distorted when the metal perturber is very close to the resonant line and the above variational analysis may be not accurately predicting this situation.

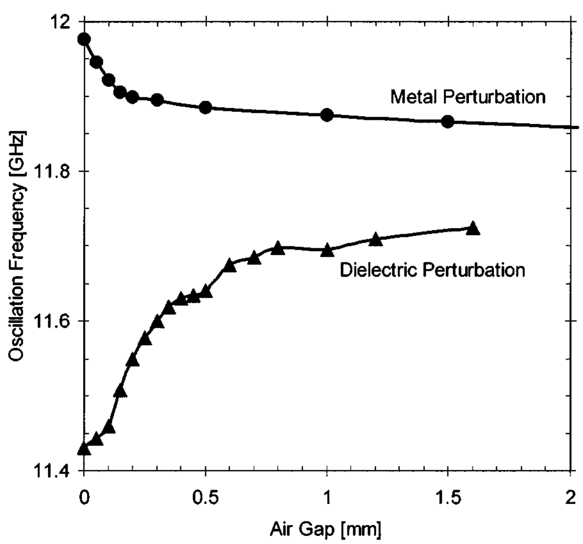
Insertion loss (S_{21}) and return loss (S_{11}) are measured for the PET-controlled metal and dielectric perturbation on the 1-D PBG resonator, as shown in Fig. 6. In order to avoid shorting the metal perturber and microstrip line at 0 V, an air gap of approximately 0.1 mm is added for the metal perturbation. Tuning capability is from -8.2 to $+9\%$ at 10.6 GHz. A maximum insertion loss increase is 1.6 dB and Q_L is larger than 100 for both perturbations. Measured Q_L is 148.5 at 10.6 GHz for the unperturbed condition under 90-V applied bias.

D. Tunable DRO

As shown in Fig. 7(a), the DRO consists of a MESFET feedback-type oscillator with a DR. To perturb electromagnetic fields of the DR and thereby tune the oscillation frequency, a dielectric or metal perturber disc attached to the PET moves vertically above the DR puck as the dc-bias voltages is varied from 0 to 90 V. A commercial DR (Trans-Tech) of $\varepsilon_r = 30.09$ and a MESFET (Agilent, ATF-26836) are used. To obtain an optimum power level and a more stable oscillation, the location of the DR between two microstrip lines and bias conditions are adjusted. DC biases of $V_{GG} = -0.8$ V, $V_{DD} = 12$ V, and $I_{DS} = 30$ mA are applied. A drain resistor of 100 Ω is used. The dielectric perturber has a dielectric constant of 10.8, a diameter of 0.3 in, and a thickness of 0.1 in. The perturber's size affects the tuning bandwidth. So as not to change the normal circuit operation, the perturber diameter is chosen very similarly with the DR puck diameter. Fig. 7(b) shows a photograph of the PET-controlled VCDRO.



(a)



(b)

Fig. 10. Mechanically micrometer-head controlled DRO results with the dielectric and metal perturbation. (a) Spectrum analyzer display of tuning the oscillation frequency. (b) Measured oscillation frequencies versus PET control voltages.

Measured results are shown in Fig. 8. A bottom-up alignment method was used, which means the minimum perturbation and deflection at 0 V and the maximum perturbation and deflection at 90 V. The use of dielectric and metal perturbation results in the tuning of the oscillation frequency from 11.54 to 11.97 GHz or -2.0 to $+1.7\%$ about the center frequency of 11.78 GHz. There is a gap without tuning from 11.7 to 11.87 GHz in Fig. 8(b), which is caused by the fact that the DRO was not designed for PET tuning and it was difficult to align the PET perfectly on the DR. The metal perturbation may increase the stored magnetic energy in the DRO with respect to the electrical energy, which results in increasing the oscillation frequency. Increase in resonant frequency can be also explained with a metal cavity wall movement inward. The dielectric perturbation effect can oppositely be explained [2]. Higher dielectric-constant material than 10.8 for the perturber may produce a wider tuning range [5]. Ideally, the VCO output power level remains a constant, but practically, the PET and metal or dielectric perturber make the output power level fluctuated from -5 to $+2.9$ dBm. The output power level without perturbation is $+1.45$ dBm.

For comparison, a mechanical tuning using a micrometer head was set up, as shown in Fig. 9, and the results are given in Fig. 10. The micrometer-head controlled DRO produces more ideal perturbation results due to better alignment and the smaller effect on circuit by the micrometer-head movement. A tuning bandwidth of -3.1 to $+1.6\%$ with a power level variation of 4.5 dB has been achieved. The results are slightly better than the electronic tuning.

III. Conclusions

New PET-controlled tunable microwave circuits have been successfully demonstrated. Theoretical results agreed well with measured data. A new analog PET-controlled phase shifter using a perturbed microstrip line showed the largest phase shift with low loss over ultra-wide bandwidth up to 40 GHz. The new phase shifter should be useful for beam steering and beam forming of antenna arrays. In addition, a new PET tuned bandpass filter, ring resonator, 1-D PBG resonator, and DRO have been demonstrated with the largest tuning range to date. The proposed tuning method should have many applications in monolithic and hybrid microwave integrated circuits.

Acknowledgment

The authors would like to thank M.-Y. Li, Texas A&M University, College Station, C. Wang, Texas A&M University, and H. Tehrani, Texas A&M University, for technical assistance and helpful discussion, and R. S. Tahim, RST Scientific Research, Anaheim, CA, for providing a DRO. A U.S. patent entitled "electromagnetic perturbation using piezoelectric transducer" is pending.

References

[1] J. Uher and W. J. R. Hoefer, "Tunable microwave and millimeter-wave band-pass filters," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 643–653, Apr. 1991.

[2] B. S. Virdee, "Current techniques for tuning dielectric resonators," *Microwave J.*, pp. 130–138, Oct. 1998.

[3] G. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*. Norwood, MA: Artech House, 1980, ch. 17.

[4] M.-Y. Li and K. Chang, "New tunable phase shifters using perturbed dielectric image lines," *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1520–1523, Oct. 1998.

[5] T. Shen, K. A. Zaki, C. Wang, and J. Deriso, "Tunable dielectric resonators with dielectric tuning disks in cylindrical enclosures," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, 2000, pp. 1441–1444.

[6] W. S. Ishak and K. W. Chang, "Tunable microwave resonators using magnetostatic wave in YIG-films," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 1383–1393, Dec. 1986.

[7] J. Uher, J. Bornemann, and F. Arndt, "Magnetically tunable rectangular waveguide E-plane integrated circuit filters," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1014–1022, June 1988.

[8] D. Nicholson, "Ferrite tuned millimeter wave bandpass filters with high off-resonance isolation," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1988, pp. 867–870.

[9] F. A. Miranda, G. Subramanyam, F. W. V. Keuls, R. R. Romanofsky, J. D. Warner, and C. H. Mueller, "Design and development of ferroelectric tunable microwave components for Ku- and K-band satellite communication systems," *IEEE Trans. Microwave Theory Tech.*, vol. 48, pp. 1181–1189, July 2000.

[10] J. B. L. Rao, D. P. Patel, and V. Krichevsky, "Voltage-controlled ferroelectric lens phased arrays," *IEEE Trans. Antennas Propagat.*, vol. 47, pp. 458–468, Mar. 1999.

[11] K. Chang, *Microwave Ring Circuits and Antennas*. New York: Wiley, 1996, ch. 4.

[12] P. C. Kandpal and C. Ho, "A broadband VCO using dielectric resonators," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1988, pp. 609–612.

[13] J. Y. Lee and U. S. Hong, "Voltage controlled dielectric resonator oscillator using three-terminal MESFET varactor," *Electron. Lett.*, vol. 30, no. 16, pp. 1320–1321, Aug. 1994.

[14] S. Lucyszyn and I. D. Robertson, "Synthesis techniques for high performance octave bandwidth 180° analog phase shifters," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 731–740, Apr. 1992.

[15] A. S. Nagra and R. A. York, "Distributed analog phase shifters with low insertion loss," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1705–1711, Sept. 1999.

[16] S.-S. Lee, A. H. Udupa, H. Erlig, H. Zhang, Y. Chang, D. H. Chang, D. Bhttacharya, B. Tsap, W. H. Steier, L. R. Dalton, and H. R. Fetterman, "Demonstration of a photonically controlled RF phase shifter," *IEEE Microwave Guided Wave Lett.*, vol. 9, pp. 357–359, Sept. 1999.

[17] G. Molin and J. Renpei, "A study on the optical control of dielectric resonator stabilized FET oscillator," in *Asia–Pacific Microwave Conf.*, 1997, pp. 565–568.

[18] H.-T. Kim, J.-H. Park, Y.-K. Kim, and Y. Kwon, "Millimeter-wave micromachined tunable filters," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Anaheim, CA, 1999, pp. 1235–1238.

[19] A. Borgioli, Y. Liu, A. S. Nagra, and R. A. York, "Low–low distributed MEMS phase shifter," *IEEE Microwave Guided Wave Lett.*, vol. 10, pp. 7–9, Jan. 2000.

[20] T.-Y. Yun and K. Chang, "A low loss time-delay phase shifter controlled by piezoelectric transducer to perturb microstrip line," *IEEE Microwave Guided Wave Lett.*, vol. 10, pp. 96–98, Mar. 2000.

[21] ——, "An electronically tunable photonic bandgap resonator controlled by piezoelectric transducer," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, June 2000, pp. 1445–1447.

[22] ——, "A phased-array antenna using a multi-line phase shifter controlled by a piezoelectric transducer," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, June 2000, pp. 831–833.

[23] R. C. Buchanan, Ed., *Ceramic Materials for Electronics*. New York: Marcel Dekker, 1986, ch. 3.

[24] K. Chang and K. Klein, "Dielectrically shielded microstrip (DSM) lines," *Electron. Lett.*, vol. 23, no. 10, pp. 535–537, May 1987.

[25] T.-Y. Yun and K. Chang, "One-dimensional photonic bandgap resonators and varactor turned resonators," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Anaheim, CA, 1999, pp. 1629–1632.

[26] B. Bhat and S. K. Koul, "Unified approach to solve a class of strip and microstrip-like transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 679–686, May 1982.

[27] A. K. Verma and R. Kumar, "New empirical unified dispersion model for shielded-, suspended-, and composite-substrate microstrip line for microwave and millimeter-wave applications," *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1187–1192, Aug. 1998.

[28] M. Kirschning and R. H. Jansen, "Accurate model for effective dielectric constant of microstrip with validity up to millimeter-wave frequencies," *Electron. Lett.*, vol. 18, no. 6, pp. 272–273, Mar. 1982.

[29] T. C. Edwards, *Foundations for Microstrip Circuit Design*. New York: Wiley, 1981, ch. 8.

Tae-Yeoul Yun (S'99) received the B.S.E.E. degree from the Kyung-pook National University, Kyung-pook, Korea, in 1987, the M.S.E.E. degree from the Korea Advanced Institute of Science and Technology (KAIST), Seoul, Korea, in 1989, and the Ph.D. degree in electrical engineering from Texas A&M University, College Station, in 2001.

From 1989 to 1996, he was with the Optical Telecommunication System Group, Electronics and Telecommunications Research Institute (ETRI), Taejon, Korea, where he developed 2.5- and 10-Gb/s systems. Since 2001, he has been a MMIC Designer with Triquint Semiconductor Inc., Dallas, TX. He has authored or co-authored over 40 technical papers. His research interests are passive and active microwave circuits, phased-array antenna systems, and high-speed optical telecommunication devices and systems.

Kai Chang (S'75–M'76–SM'85–F'91) received the B.S.E.E. degree from the National Taiwan University, Taipei, Taiwan, R.O.C., in 1970, the M.S. degree from the State University of New York at Stony Brook, in 1972, and the Ph.D. degree from The University of Michigan at Ann Arbor, in 1976.

From 1972 to 1976, he was with the Microwave Solid-State Circuits Group, Cooley Electronics Laboratory, The University of Michigan at Ann Arbor, where he was a Research Assistant. From 1976 to 1978, he was with Shared Applications Inc., Ann Arbor, MI, where he was involved with computer simulation of microwave circuits and microwave tubes. From 1978 to 1981, he was with the Electron Dynamics Division, Hughes Aircraft Company, Torrance, CA, where he was involved in the research and development of millimeter-wave solid-state devices and circuits, power combiners, oscillators, and transmitters. From 1981 to 1985, he was with TRW Electronics and Defense, Redondo Beach, CA, where he was a Section Head involved with the development of state-of-the-art millimeter-wave integrated circuits and subsystems, including mixers, voltage-controlled oscillators (VCOs), transmitters, amplifiers, modulators, upconverters, switches, multipliers, receivers, and transceivers. In August 1985, he joined the Electrical Engineering Department, Texas A&M University, College Station, as an Associate Professor, and became a Professor in 1988. In January 1990, he became an E-Systems Endowed Professor of Electrical Engineering. He has authored and co-authored several books, including *Microwave Solid-State Circuits and Applications* (New York: Wiley, 1994), *Microwave Ring Circuits and Antennas* (New York: Wiley, 1996), *Integrated Active Antennas and Spatial Power Combining* (New York: Wiley, 1996), and *RF and Microwave Wireless Systems* (New York: Wiley, 2000). He has served as the Editor of the four-volume *Handbook of Microwave and Optical Components* (New York: Wiley, 1989 and 1990). He is the Editor of *Microwave and Optical Technology Letters* and the Wiley Book Series on "Microwave and Optical Engineering." He has also authored or co-authored over 350 technical papers and several book chapters in the areas of microwave and millimeter-wave devices, circuits, and antennas. His current interests are in microwave and millimeter-wave devices and circuits, microwave integrated circuits, integrated antennas, wide-band and active antennas, phased arrays, microwave power transmission, and microwave optical interactions.

Dr. Chang was the recipient of the 1984 Special Achievement Award presented by TRW, the 1988 Halliburton Professor Award, the 1989 Distinguished Teaching Award, the 1992 Distinguished Research Award, and the 1996 Texas Engineering Experiment Station (TEES) Fellow Award presented by Texas A&M University.