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A New Measurement Technique on Inherent-Ring-Resonance Frequency and Effective Loss-Tangent using Ring Filters

Hee-Ran Ahn · Kwyro Lee

Abstract

As an application of ring filters, a new and simple method to determine an inherent-ring-resonance frequency is introduced. The ring filter consists of a ring and two short stubs. They are connected at 90° and 270° points of the ring and the ring filter may be seen in such way that two filters are connected in parallel. Therefore, if the two powers of the two filters are out-of-phase at the output, the power excited at the input can not be delivered. That can be done by making difference in length of the two short stubs, and when a certain condition is satisfied, a frequency exists where all the excited power is reflected. That is the very inherent-ring-resonance frequency. In the lossless case, the return loss with the condition reaches 0 dB at the inherent-ring-resonance frequency but does not with conductor, dielectric losses and so on. Therefore, the effective loss tangent at a frequency of interest may be obtained correctly. To verify the method, two ring filters have been fabricated in microstrip lines and the measured results show good agreement with the predicted ones.

Key words : Ring Resonance Frequency, New Measurement Technique on Inherent-Ring-Resonance Frequency, Effective Loss-Tangent.

Ⅰ. Introduction

It is well known that ring resonators have low radiation loss, high Q factors and two orthogonal resonant modes. Because of these special properties, ring resonators have widely been used for the measurements [1], [2], band-pass filters and duplexers [3]. Microstrip open- and closed-ring resonators were intensively discussed [4], [5] and mixers, oscillators and tuning filters have been realized based on circuit theory concepts [6]. For the design of ring-based circuits, to know an exact resonance frequency is very important [7] and there have been many trials to determine it, using gap coupling [8], simple cavity model with magnetic side wall [9], [10], a planar waveguide model[11], [12], and so on[13]. However, even though those proposed by [1], [2], [8] have been considered as a simple method, the ring-resonance frequency determined can be shifted by gaps, which may be very harmful for narrow-band filter designs. Also, those by [9]~[13] require much time and somewhat complicate mathematical programs based on field theory analyses.

In this paper, a new and simple method to determine an inherent-ring-resonance frequency will be introduced as an application of the ring filters. The ring filter was for the first time suggested as a wideband transmission line [14], [15]. It consists of a ring and two short-stubs but it may be seen in other sense that two filters are connected in parallel. The two are named "up-" and "down-filter" in this paper. Therefore, the power excited at the input is divided just like three-port power dividers or ring hybrids [16]~[20], flows into the up- and down-filter and finally combined at the output. If the two powers are out-of-phase at the output, the excited power can not be delivered to the output. That can be done by making difference in length of the two stubs, and the phase of transmitted power (S_{21}) is abruptly changed at the vicinity of the inherent-ring-resonance frequency, where all the excited power is reflected under a certain condition. Therefore, if the frequency is determined, it will be the exact inherent-ring-resonance frequency. In lossless case, the return loss reaches 0 dB at the inherent-ring-resonance frequency but does not with conductor, dielectric loss and so on. Therefore, the return losses at the inherent-ring-resonance frequency are investigated depending on the differences between the two short stubs and the effective loss tangent at a frequency of interest

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Division of EE(Electrical Engineering), Dept. of EECS(Electrical Engineering Computer Science) and MICROS Research Center, KAIST, Korea.

may be obtained correctly. By the method explained above, the gap capacitances for the coupling in conventional ring-based circuits can also be extracted exactly.

II. Ring Filters for Inherent-Ring-Resonance Frequency

A ring filter is depicted in Fig. 1(a) and its up-filter in Fig. 1(b). It is terminated in Z_1 and Z_2, and consists of a ring and two short stubs. The two short-stubs are located at 90° and 270° points of the ring. The point where each short stub is connected may be considered as a hypothetical port whose termination impedance is Z_h. The Z_h is needed to design the ring filters and may arbitrarily be chosen when $Z_1=Z_2$, and $Z_h=(Z_1+Z_2)/2$ or $\sqrt{Z_1Z_2}$ in the case of $Z_1 \neq Z_2$. The length of four transmission-line sections forming a ring is equally ℓ and their characteristic impedances are Z_{ca}, Z_{cb}, Z_{cc} and Z_{cd}, shown in Fig. 1(a). Since lengths of the two short stubs are about 90°, the power excited at port ① is divided depending on the power division ratio of d_1 to d_2 indicated in Fig. 1(a), and the divided powers are combined at port ②. Thus, it can be understood in such way the two filters, "up-filter" and "down-filter" shown in Fig. 1(a), are connected in parallel and the termination impedances of the up-filter may be derived as explained in Fig. 1(b) [18], [19]. Z_s, ℓ_{us} and ℓ_{ds} are characteristic impedance, lengths of the two short stubs of up- and down-filters, respectively.

The *ABCD* parameters of the up-filter in Fig. 1(b) are

$$A_{up} = \cosh^2\gamma\ell + \frac{Z_{ca}}{2Z_s}\sinh 2\gamma\ell \coth\gamma\ell_{us} + \frac{Z_{ca}}{Z_{cb}}\sinh^2\gamma\ell$$

$$B_{up} = \frac{Z_{ca}+Z_{cb}}{2}\sinh 2\gamma\ell + \frac{Z_{ca}Z_{cb}}{Z_s}\sinh^2\gamma\ell\coth\gamma\ell_{us},$$

$$C_{up} = \frac{\sinh 2\gamma\ell}{2Z_{ca}} + \frac{1}{Z_s}\cosh^2\gamma\ell\coth\gamma\ell_{us} + \frac{\sinh 2\gamma\ell}{2Z_{cb}},$$

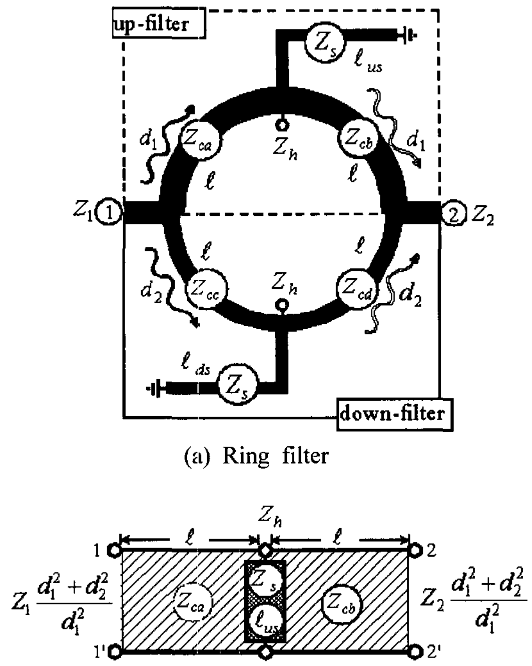
$$D_{up} = \frac{Z_{cb}\sinh^2\gamma\ell}{Z_{ca}} + \frac{Z_{cb}}{2Z_s}\coth\gamma\ell_{us}\sinh 2\gamma\ell + \cosh^2\gamma\ell, \quad (1)$$

where,

$\gamma = \alpha + j\beta$ (α and β : attenuation and phase constants),

$$Z_{ca} = \sqrt{Z_1 Z_h \frac{d_1^2+d_2^2}{d_1^2}},$$

$$Z_{cb} = \sqrt{Z_2 Z_h \frac{d_1^2+d_2^2}{d_1^2}}.$$



(a) Ring filter

(b) Up-filter

Fig. 1. Ring filter.

In the same way, those of the down-filter are

$$A_{do} = \cosh^2\gamma\ell + \frac{Z_{cc}}{2Z_s}\sinh 2\gamma\ell\coth\gamma\ell_{ds} + \frac{Z_{cc}}{Z_{cd}}\sinh^2\gamma\ell$$

$$B_{do} = \frac{Z_{cc}+Z_{cd}}{2}\sinh 2\gamma\ell + \frac{Z_{cc}Z_{cd}}{Z_s}\sinh^2\gamma\ell\coth\gamma\ell_{ds},$$

$$C_{do} = \frac{\sinh 2\gamma\ell}{2Z_{cc}} + \frac{1}{Z_s}\cosh^2\gamma\ell\coth\gamma\ell_{ds} + \frac{\sinh 2\gamma\ell}{2Z_{cd}},$$

$$D_{do} = \frac{Z_{cd}\sinh^2\gamma\ell}{Z_{cc}} + \frac{Z_{cd}}{2Z_s}\coth\gamma\ell_{ds}\sinh 2\gamma\ell + \cosh^2\gamma\ell. \quad (2)$$

where

$$Z_{cc} = \sqrt{Z_1 Z_h \frac{d_1^2+d_2^2}{d_2^2}},$$

$$Z_{cd} = \sqrt{Z_2 Z_h \frac{d_1^2+d_2^2}{d_2^2}}.$$

2-1 Lossless Case

As mentioned above, if the two powers of up- and down-filter are out-of-phase at port ②, the excited power at port ① can not be delivered to port ②. If a certain condition is satisfied between the two short stubs, a frequency where all the excited power is reflected exists. That is the inherent-ring-resonance frequency. To derive the condition, lossless(α=0) is assumed and different values of ℓ, ℓ_{us} and ℓ_{ds} are set as written in (3),

$$\beta_o \ell = \pi/2,$$
$$\beta_o \ell_{us} = \pi/2 + \mu$$
$$\beta_o \ell_{ds} = \pi/2 + \nu, \tag{3}$$

where β_o is a propagation constant at a design center frequency.

After some calculations with (1)~(3), the admittance parameters of the ring filter in Fig. 1(a) are calculated as

$$Y_{11} = -j\frac{Z_s}{Z_{ca}^2}\cot\mu - j\frac{Z_s}{Z_{cc}^2}\cot\nu,$$
$$Y_{12} = Y_{21} = -\frac{jZ_s}{Z_{ca}Z_{cb}}\cot\mu - \frac{jZ_s}{Z_{cc}Z_{cd}}\cot\nu,$$
$$Y_{22} = -j\frac{Z_s}{Z_{cb}^2}\cot\mu - j\frac{Z_s}{Z_{cd}^2}\cot\nu, \tag{4}$$

which are the derived values at the design center frequency.

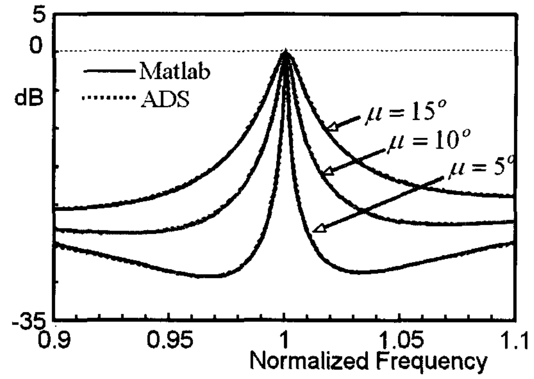
In the case of an equal power division $d_1 = d_2$, (4) is simplified as

$$Y = -jZ_s\frac{\sin(\mu+\nu)}{\sin\mu\sin\nu}\begin{bmatrix}\frac{1}{Z_{ca}^2} & \frac{1}{Z_{ca}Z_{cb}}\\ \frac{1}{Z_{ca}Z_{cb}} & \frac{1}{Z_{cb}^2}\end{bmatrix} \tag{5}$$

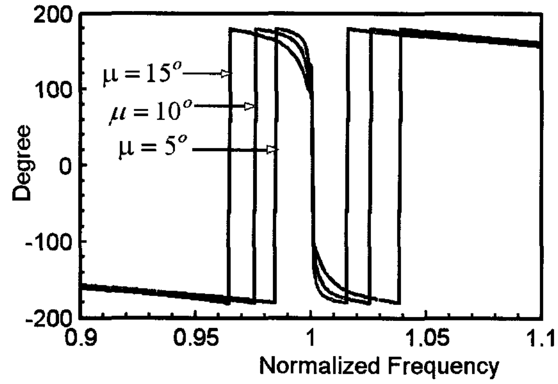
$\mu \neq 0$, $\nu \neq 0$ and $\mu + \upsilon = 0$ in (5) result in

$$Y = \begin{bmatrix}0 & 0\\ 0 & 0\end{bmatrix} \tag{6}$$

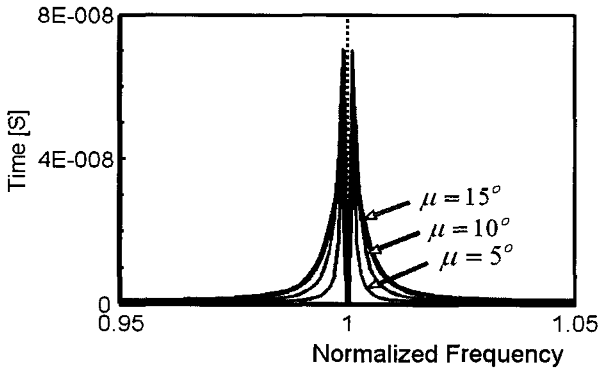
(6) implies that matching and power transfer can not occur with the condition $\mu+\upsilon=0$. The fact may be used to determine an inherent-ring-resonance frequency. Frequency responses of the return loss, phase responses of the insertion loss and group delay depending on $\mu$are plotted in Fig. 2 where those of return loss are in Fig. 2(a), those of phase response in Fig. 2(b) and those of group delay in Fig. 2(c). They are all satisfied with the condition of $\mu+\upsilon=0$. The return losses have been programmed based on (1)~(6), working on a mathematical program, Matlab 6.1 and compared with those by a commercial program, ADS 2002, in Fig. 2(a) where solid lines are those by Matlab 6.1 and dotted ones those by ADS 2002. Two types of results are almost identical and all the excited power is reflected at a frequency, which indicates the inherent-ring-resonance frequency. Fig. 2(b) shows sudden phase inversions with the condition of $\mu+\upsilon=0$ and $\mu \neq 0$. Due to the sudden phase inversions, high values of the group delay in Fig. 2(c) will occur, which is in consistency with (6). Therefore, the condition $\mu+\upsilon=0$ is a sufficient and necessary condition to determine the inherent-ring-resonance frequency.



(a) Return losses



(b) Phase responses



(c) Group delays

Fig. 2. Frequency responses.

2-2 Loss Case

If loss is taken into consideration for the microstrip lines, the complex propagation constant is

$$\gamma = \alpha + j\beta = j\omega\sqrt{\mu\varepsilon'(1 - j\tan\delta_e(f))}\,, \tag{7}$$

where $\omega=2\pi f$, $\tan\delta_e(f)$ is effective loss-tangent including dielectric, conductor losses and so on at a given frequency and f an operating frequency.

Since $\tan\delta_e(f) \ll 1$, (7) may be formulated as

$$\gamma\ell = \frac{\pi}{2}\frac{f}{f_0}\cdot\frac{\tan\delta_e(f)}{2} + j\left(\frac{\pi}{2}\frac{f}{f_0}\right),$$

$$\gamma\ell_{us} = \left(\frac{\pi}{2}+\mu\right)\frac{f}{f_0}\cdot\frac{\tan\delta_e(f)}{2} + j\left(\frac{\pi}{2}+\mu\right)\frac{f}{f_0},\,,$$

$$\gamma\ell_{ds} = \left(\frac{\pi}{2}+\nu\right)\frac{f}{f_0}\cdot\frac{\tan\delta_e(f)}{2} + j\left(\frac{\pi}{2}+\nu\right)\frac{f}{f_0}\,.\,. \tag{8}$$

Based on (1)~(8), frequency responses of ring filters with $\mu+\upsilon=0$ have been calculated. The calculated reflection coefficients are plotted in Fig. 3 where $d_1=d_2$, Z_{ca}, Z_{cb}, Z_{cc}, Z_{cd} and Z_s in Fig. 1(a) are equally 70.71 Ω, and X-axis is normalized to 3 GHz. The $\tan\delta_e(f)$ is considered as 0.0064 at 3 GHz for these calculations. According to the calculated results in Fig. 3, it may be seen, as the μ grows, the amount of reflected power at the inherent ring-resonance frequency is larger. Substituting (7) and (8) into (1) and (2) and after some calculations, the admittance parameters with loss are derived at $f=f_o$ as

$$Y_{11} = \frac{\frac{Z_{cb}}{Z_{ca}}\cosh^2\alpha\ell + K_u\frac{Z_{cb}}{2Z_m}\sinh 2\alpha\ell + \sinh^2\alpha\ell}{\frac{1}{2}(Z_{cb}+Z_{ca})\sinh 2\alpha\ell + K_u\frac{Z_{ca}Z_{cb}}{Z_s}\cosh^2\alpha\ell} + \frac{\frac{Z_{cd}}{Z_{cc}}\cosh^2\alpha\ell + K_d\frac{Z_{cd}}{2Z_s}\sinh 2\alpha\ell + \sinh^2\alpha\ell}{\frac{1}{2}(Z_{cc}+Z_{cd})\sinh 2\alpha\ell + K_d\frac{Z_{cc}Z_{cd}}{Z_s}\cosh^2\alpha\ell}$$

$$Y_{12} = Y_{21} = \frac{1}{\frac{1}{2}(Z_{cb}+Z_{ca})\sinh 2\alpha\ell + K_u\frac{Z_{ca}Z_{cb}}{Z_s}\cosh^2\alpha\ell} + \frac{1}{\frac{1}{2}(Z_{cc}+Z_{cd})\sinh 2\alpha\ell + K_d\frac{Z_{cc}Z_{cd}}{Z_s}\cosh^2\alpha\ell}$$

$$Y_{22} = \frac{\sinh^2\alpha\ell + K_u\frac{Z_{ca}}{2Z_s}\sinh 2\alpha\ell + \frac{Z_{ca}}{Z_{cb}}\cosh^2\alpha\ell}{\frac{1}{2}(Z_{cb}+Z_{ca})\sinh 2\alpha\ell + K_u\frac{Z_{ca}Z_{cb}}{Z_s}\cosh^2\alpha\ell}\,, + \frac{\sinh^2\alpha\ell + K_d\frac{Z_{cc}}{2Z_s}\sinh 2\alpha\ell - \frac{Z_{cc}}{Z_{cd}}\cosh^2\alpha\ell}{\frac{1}{2}(Z_{cc}+Z_{cd})\sinh 2\alpha\ell + K_d\frac{Z_{cc}Z_{cd}}{Z_s}\cosh^2\alpha\ell} \tag{9}$$

where $\alpha\ell = \frac{\pi}{4}\tan\delta_e(f_0)$,

$K_u = \left[\frac{e^{2(\alpha\ell_{us}+j\mu)}+1}{e^{22(\alpha\ell_{us}+j\mu)}-1}\right]$ and $K_d = \left[\frac{e^{2(\alpha\ell_{ds}-j\mu)}+1}{e^{2(\alpha\ell_{ds}-j\mu)}-1}\right]$ with $\alpha\ell_{us} = \left(\frac{\pi}{2}+\mu\right)\frac{\tan\delta_e(f_0)}{2}$ and $\alpha\ell_{ds} = \left(\frac{\pi}{2}-\mu\right)\frac{\tan\delta_e(f_0)}{2}$.

To calculate attenuations due to conductor loss α_c, dielectric loss α_d and so on for most microstrip lines, there are so many factors to be considered, for example, relative dielectric constant, dielectric loss, conductivity, characteristic impedance and width of a microstrip line and so on [21]. However, any of them can not be ensured to be correct, which causes errors. Therefore, if the return loss of the ring filter at the inherent ring-resonance frequency is known or measured, the $\tan\delta_e(f)$ may be calculated correctly at a given frequency using (9). To verify the method to determine the inherent ring-resonance frequency, two ring filters with $\mu+\upsilon=0$ have been fabricated on a substrate(ε_r=2.88, h=508 μm) in microstrip technology, where all the fabricated data are the same as those in Fig. 3. They have been designed at 3 GHz, the $\tan\delta_e(f)$ is 0.0064 and the loss due to feeding lines is ignored. The photo with μ=20° is shown in Fig. 4 and measured results are compared with those predicted in Fig. 5. Measured return loss with μ=20° is −2.26 dB and the simulated one −2.28 dB at 3 GHz. In the case of μ=15°, the measured one is −3.4 dB and the simulated one −3.71 dB. Due to conductor and dielectric losses, the measured and simulated values show not 0 dB but −2.26 dB or −3.4 dB at the inherent-ring- resonance frequency, 3 GHz.

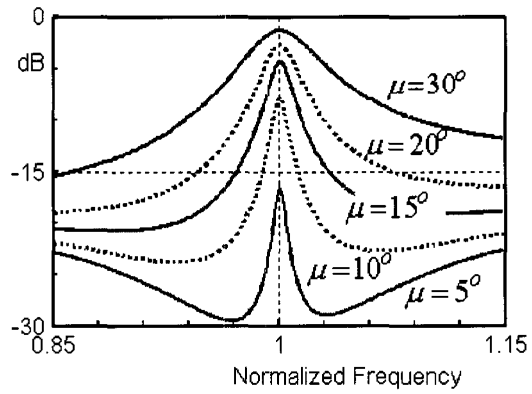


Fig. 3. Frequency responses of reflection coefficients considering dielectric and conductor losses.

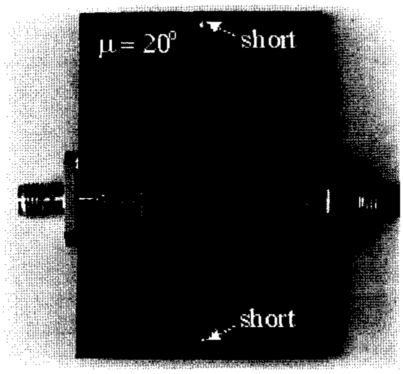


Fig. 4. A photo of the ring filter with $\mu+\nu=0°, \mu=20°$.

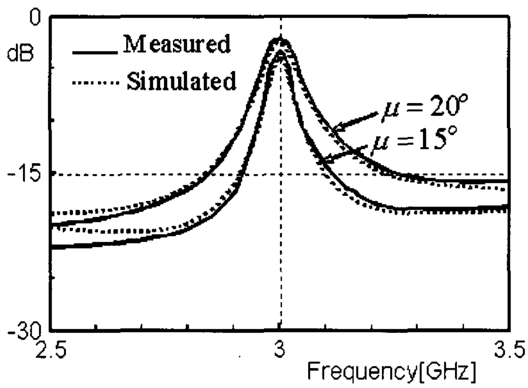


Fig. 5. Measured and simulated return losses.

III. Conclusions

A ring filter consists of a ring and two short stubs, and It may be seen in such way that two filters are connected in parallel. If the two output powers are out-of-phase, the excited power can not be delivered, which can be done by making difference in length of the two short stubs. If the frequency, where all the excited power is reflected, is determined, it will be the inherent-ring-resonance frequency. By knowing the exact inherent ring-resonance frequency, the capacitances due to gap coupling in conventional ring-based circuits may be extracted exactly. By knowing the return loss at the inherent-ring-resonance frequency, the effective loss tangent may be obtained as well.

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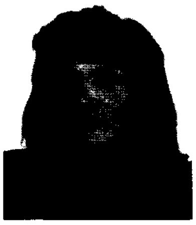
Hee-Ran Ahn



received the B.S., M.S., and Ph.D. degrees in electronic engineering from Sogang University, Seoul, Korea, in 1988, 1990 and 1994, respectively. From 1991 to 1995, she was a Part-Time Lecturer at Sogang University, Seoul, Korea and a Post-Doctoral Fellow from 1996 to 1997 at Duisburg-Essen University, Duisburg, Germany. From February 1997 to 2002, she was with the Department of Electrical Engineering at Duisburg-Essen University, Duisburg, Germany where she worked for the Habilitation dealing with asymmetric passive components. She is currently with Division of Electrical Engineering, Dept. of EECS(Electrical Engineering and Computer Science) at KAIST(Korea Advanced Institute Science and Technology), Daejeon, Korea as a Visiting professor. She is a Senior Member of IEEE since 1999 and in editorial board at IEEE on MTT(Microwave Theory and Technology). Her interests include high frequency and microwave circuit design, and biomedical application of microwave theory and techniques.

Kwyro Lee

received the B.S. degree in electronics engineering from Seoul National University, Seoul, Korea, in 1976, and the M.S. and Ph.D. degrees from the University of Minnesota, Minneapolis, in 1981 and 1983, respectively, where he did many pioneering works for characterization and modeling of AlGaAs/GaAs hetero-junction field-effect transistor.

He was an Engineering General Manager at Gold-Star Semiconductor, Inc. Korea, from 1983 to 1986, responsible for development of the first polysilicon CMOS products in Korea. He joined the Korea Advanced Institute of Science and Technology in 1987 as an Assistant Professor in the Department of Electrical Engineering, where he is now a Professor. He has written more than 150 publications in major international journals and conferences. He is the principal author of the book *Semiconductor Device Modeling for VLSI* (Englewood Cliffs, NJ: Prentice Hall, 1993) and one of the developers of AIM-Spice, the world's first SPICE run under windows. Since 1997, he has been the Director of MICROS(Micro Information and Communication Remote Object-oriented Systems) Research Center, an Engineering Center of Excellence supported by Korea Science and Engineering Foundation.

Dr. Lee is Senior Member of IEEE since 1991 and Life Member of IEEK. During 1990-1996, he served as the Conference Co-Chair of Int. Semiconductor Device Research Symposium, Charlottesville, VA. In 998-2000, he served as the KAIST Dean of Research Affairs as well as the Dean of Institute Development and Cooperation. During the same periods, he also served as the Chairman of IEEE Korea Electron Devices Chapter, and is currently serving as the elected member EDS AdCom.