Design and Realization of the Dual-mode Channel Filter and Group-Delay-and-Amplitude Equalizer for the Ka-band Satellite Transponder Subsystem

Sungtek Kahng · Man Seok Uhm · Seong Pal Lee

Abstract

In this paper, the design of a channel filter and its group-delay-and-amplitude equalizer is carried out for the Ka-band satellite transponder subsystem. The 8th order dual-mode filter is employed for high selectivity around the band-edges with an elliptic-integral function response and has an in-line configuration. The 2-pole, reflection-type, group-delay equalizer is designed and manufactured to reduce the group-delay and amplitude variation, which can be large for such a high order filter. It is noted that in both the filter and equalizer, adopting the dual-mode coupling mechanism leads to less mass and volume. Through measurement, the performance of the realized group-delay-equalized filter is shown to meet the equipment requirements and to be appropriate for the satellite input multiplexer.

Key words : Dual-mode Channel Filter, Group-delay Equalizer, Input Multiplexer, Satellite Communication, Passive Components.

Ⅰ. Introduction

No satellite system appears rich in frequency resource and a variety of design techniques have been developed for the components on a narrow-band use basis like channel filters in the input multiplexer.

The received signal enters the input multiplexer and is divided through channel filters of a higher order, each of which has a lower insertion loss in its passband and larger rejection in the other channels. If needed, a multiple-mode coupling mechanism is adopted to decrease the mass and volume. Unfortunately, the group-delay of its output varies abruptly in the passband and shows thermal degradation. To circumvent this matter, the group-delay variation should be lowered by adding a group-delay equalizer. Besides, it can be used to compensate for the amplitude variation.

In this paper, an in-line type 8th order dual-mode channel filter and a 2-pole reflection-type group-delay-and-amplitude equalizer are designed and implemented for a Ka-band satellite transponder. Both consist of circular waveguide cavities of a TE_{113} resonance mode, circular irises and tuning- and coupling screws. A space-qualified circulator is used as a path between the filter and the equalizer, and is not designed. To our knowledge, our work is the first to suggest the equalization of the amplitude as well as group-delay for a Ka-band application. The in-line filter is separated from the equalizer, which leads to significantly improved efficiency in tuning, mechanical handling and thermal stability. The measurement results show that the performance of the delay-and-amplitude equalized channel filter satisfies the design target for the Ka-band input multiplexer and validates the design method.

Ⅱ. Design

2-1 Configuration & Specifications

The channel of the Ka-band input multiplexer is composed of the channel filter and group-delay equalizer.

As seen in Fig. 1, the channel filter is followed by

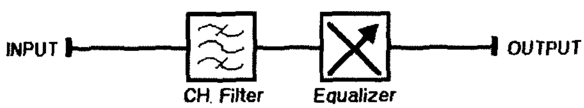


Fig. 1. Channel configuration.

Manuscript received August 21, 2003 ; revised October 15, 2003.
Communication Satellite Development Center, Electronics and Telecommunicatons Research Institute, Gajeong-dong, Yusong-gu, Daejeon, Korea.

the equalizer which lowers the group-delay variation. The signal from the filter enters and leaves the reflection-type of equalizer by way of a circulator which is space-qualified for thermal and mechanical performance. From the input port through the output port, waveguides of type WR42 are used. Based on the specified requirements of the whole satellite transponder, the levels of the principal electrical parameters for a component can be determined. Table 1 shows the specifications for the channel of the Ka-band input multiplexer.

Center frequency f_0 is 21 GHz and bandwidth $\triangle f$ is 100 MHz. The return loss will be at least greater than 13 dB, because the circulator chain with a higher return loss will be adopted for channel-branching, if necessary.

2-2 Channel Filter

As stated earlier, the 8th order dual-mode elliptic integral function type of coupling is chosen to best fit the amplitude of the specified frequency response. Similar to [1]~[3], its filter transfer function can be expressed as

$$S_{21} = u_0 \frac{s^4 + B_2 s^2 + B_0}{s^8 + A_7 s^7 + A_6 s^6 + A_5 s^5 + A_4 s^4 + A_3 s^3 + A_2 s^2 + A_1 s^1 + A_0} \quad (1)$$

where $s=\sigma+j\tau$ and $j=\sqrt{-1}$

Generally, manipulating equation (1) produces the network parameters of a canonical or folded coupling configuration, and the similarity transforms convert the network to what configuration is desired[1]~[3]. Fig. 2(a) and (b) show the signal flow and channel filter structure due to the in-line configuration desired in this work.

In Fig. 2(a), sequential couplings are denoted as solid arrows, and dotted arrows indicate cross-couplings. The input or output port is coupled through a horizontal slot. The cross-shaped slots are used for intercavity couplings. Especially, with reference to [1]~[3], an 8th order two-port network is mathematically represented as follows:

$$\bar{\bar{Z}} \cdot \bar{I} = \bar{e} \quad (2)$$

where $\bar{e}^T = (1,0,0,\cdots,0,0)$ is the voltage excitation vector and $\bar{I}^T = (i_1, i_2, i_3, \cdots, i_7, i_8)$ is the current vector including all the resonators.

$$\bar{\bar{Z}} = j(\tau\bar{\bar{U}} + \bar{\bar{M}}) \quad (3)$$

has self- and mutual coupling values of the network.

$$\tau = \frac{f_0}{\Delta f}\left(\frac{f}{f_0} - \frac{f_0}{f}\right). \quad (4)$$

$\bar{\bar{U}}$ is the identity matrix. Using the above relation of the current and voltage, the major S-parameters can be represented as

$$S_{21} = -2\sqrt{R_{in} R_{out}}\, i_8 \quad (5)$$

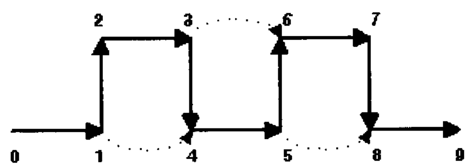
and

$$S_{11} = 1 - 2R_{in} i_1 \quad (6)$$

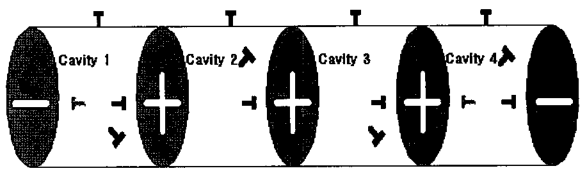
The normalized input(output) resistance $R_{in}(R_{out})$ is concerned with the input(output) port coupling. The

Table 1. Specifications for the channel.

Parameter	Frequency	Spec.
Amplitude variation	$f_0 \pm 30/38$ MHz	0.15/0.25(dBp − p)
	$f_0 \pm 45$ MHz	0.50(dBp − p)
	$f_0 \pm 50$ MHz	1.50(dBp − p)
Group-delay variation	$f_0 \pm$ 0 MHz	0.6(ns)
	$f_0 \pm 20$ MHz	1.8(ns)
	$f_0 \pm 30/38$ MHz	2.2/4.0 (ns)
	$f_0 \pm 45$ MHz	9.5(ns)
	$f_0 \pm 50$ MHz	22.5(ns)
Near-band rejection	$f_0 = 65/80/130$ MHz	− 19/ − 49/ − 55(dB)
Return loss	Within $\triangle f$	> 13(dB)
Insertion loss	Within $\triangle f$	< 4(dB)



(a) Signal flow



(b) Filter structure

Fig. 2. Signal flow and filter structure.

elements of $\overline{\overline{M}}$ or M_{pq} correspond to the intercavity couplings with M_{pp}=0. Also, M_{pq} can be related to inverter constant K_{pq} as in [4]

$$K_{pq} = M_{pq}\frac{3\pi \cdot \Delta f}{2f_0} \tag{7}$$

Once R_{in}, R_{out} and $\overline{\overline{M}}$ are obtained from equation (1) using the methods in [1]~[3] rotating $\overline{\overline{M}}$ is repeated until it reaches the wanted configuration coupling[4]~[5]. The final M_{pq} will be converted to the initial physical sizes of the slots.

As tips on the waveguides used here, the radius of the circular waveguide and its resonance mode should be determined as R_c and TE_{11s}, respectively with a WR42 (inner size = $a \times b$) in/output port. Also, the slot thickness and width are common to all the irises for convenience.

When we decide the initial sizes of the slots for the input(output) port and cavity 1(cavity 4) and the intercavity coupling, the scattering parameters, reactances, magnetic polarizabilities, etc can be used. In particular, the magnetic polarizabilities are used as goal values for searching the initial sizes. The approximate formula for the input(output) magnetic polarizability $P_{M,io}$[5] related to external Q is

$$P_{M,io} = \sqrt{\frac{abl_c^3\lambda_{gr}(3R_c^2)R_{in}}{4\pi\lambda_0^2 s^2}} \tag{8}$$

λ_0 is the free space wavelength at f_0, λ_{gr} is that of the WR42, and λ_{gc} is that of the circular waveguide. In mode TE_{11s}, cavity length l_c is preliminarily given as $s\frac{\lambda_{gc}}{2}$.

The intercavity coupling magnetic polarizability $P_{M,\text{int},pq}$ is expressed as

$$P_{M,\text{int},pq} = \frac{l_c^3(3R_c^2)}{\lambda_0^2}\frac{\Delta f}{f_0}M_{pq} \tag{9}$$

$P_{M,io}$ and $P_{M,int,pq}$ are equated to the counterpart of the full-wave analysis methods[4],[5] or the approximate formula for P_M' (infinitesimal thickness) and (considering the slot thickness) which include the initial size of a rounded slot[6]

$$P_M = \frac{P_M'}{1-(\frac{\lambda_l}{\lambda_0})^2}10^{-(\frac{2.73t_sA}{\lambda_l})\sqrt{1-(\frac{\lambda_l}{\lambda_0})^2}} \tag{10}$$

where

$$P_M' = \frac{L_s^3[0.187+0.052(\frac{W_s}{L_s})(1-\frac{W_s}{L_s})]}{\ln(1+2.12\frac{L_s}{W_s})} \tag{11}$$

L_s, W_s, t_s and λ_l denote the slot length, width, thickness and slot resonance wavelength, respectively. And constant A takes 3. So far, it has been talked how to determine the initial sizes.

Now the initial sizes or the coupling values of initial sizes need to be verified through measurement such as monitoring the network analyzer. A number of measurement approaches to this have been introduced, but here we use the scheme in [2], which has a short-ended 1-port configuration. As far as the input port coupling is concerned, after cavity 1 is tuned to f_0, the frequency span from $-90°$ to $+90°$ in the phase-axis is measured and denoted as $\Delta f_{\pm 90°}$. This will be compared to the computed value R_{in} by way of

$$\Delta f_{\pm 90°} = \text{Rin} \cdot \Delta f \tag{12}$$

Different from the case of input port coupling, measurement of the intercavity coupling requires cavities 1 and 2. Cavity 2 is short-ended. The two cavities are synchronously tuned to f_0, and the length of the frequency points for $-180°$ and $+180°$ of the phase is taken as $\Delta f_{\pm 180°, pq}$. This is concerned with M_{pq} as in

$$\Delta f_{\pm 180°,\ pq} = M_{pq} \cdot \Delta f \tag{13}$$

If the slot sizes are determined the cavity lengths can be computed. The electrical length of a cavity tends to increase by $\frac{\phi_{pq}}{2}$, when coupling occurs through a slot. Therefore, the real cavity should be shortened by the same amount. The well-known formula is used

$$\phi_{pq} = \frac{\lambda_{gc}}{2\pi}\tan^{-1}(\frac{2X_{pq}}{Z_0}) \tag{14}$$

where normalized reactance $X_{pq}/Z_0 = 4\pi P_{M,\text{int},pq}/(3R_c^2\lambda_{gc})$ and Z_0 is the impedance in the waveguide.

Eventually, tuning- and coupling screws are properly positioned considering the relations of the couplings. The tuning screws are used for removing the difference of resonance lengths for two orthogonal modes in one cavity. The two orthogonal modes are coupled through the coupling screw. Their depths are empirically determined, while the measurement is made.

2-3 Group-Delay Equalizer

In the reflection-type of the group-delay equalizer, the input signal is reflected from the shorted end. Fig. 3 shows one such cavity with 2 poles.

It has the same radius and resonance mode as the channel filter. Unknowns R_{eq}(normalized input resistance) and K_{12} or M_{12} will be decided, since only one cavity is needed. Because its frequency response is not represented by an elliptic-integral or Chebyscheff function, the following equation is alternatively used for design[2].

$$Z_{11} = j\frac{\upsilon^2 - M_{12}^2}{\upsilon}, \qquad (15)$$

where $\upsilon = \frac{1}{\Delta f}\left(\frac{f}{f_0} - \frac{f_0}{f}\right)$ and Z_{11} is the input impedance of the one-port circuit. S_{11} or reflection coefficient Γ becomes

$$\Gamma = \frac{Z_{11} - R_{eq}}{Z_{11} + R_{eq}}. \qquad (16)$$

Hence, the group-delay of the equalizer is expressed as

$$\zeta = -\frac{1}{2\pi}\frac{d}{df}(Phase(\Gamma)). \qquad (17)$$

The unknowns are found as soon as ζ compensates for the group-delay of the filter, optimized similar to [7] and then the initial-sizes are computed.

Ⅲ. Realization and Test Results

With regard to the specifications shown in Table 1, $\tau = \pm 1.4$ and $\tau = \pm 1.66$ turn out appropriate transmission zeros and equation (1) can be identified with $u_0 = -j0.04$, $B_2=4.72$, $B_0=5.40$, $A_0=0.24$, $A_1=1.11$, $A_2=2.91$, $A_3=4.82$, $A_4=0.24$, $A_5=5.80$, $A_6=4.72$, and $A_7=2.08$. Also, considering the symmetric filter structure, $R_{in}=R_{out}=1.1$, $M_{12}=M_{78}=0.905$, $M_{23}=M_{67}=0.737$, $M_{34}=M_{56}=0.523$ and $M_{45}=0.549$ as sequential couplings and $M_{14}=M_{58}=-0.196$ and $M_{36}=-0.00543$ as cross couplings. External Q is 170. When the coupling coefficients are used for the slot sizes, 0.7 mm is chosen for W_s and 0.4 mm for t_s. Slot lengths can be obtained via Equations (8) and (11). $P_{M,io}$ and $P_{M,int,pq}$ are 26.5 and 6.8 and slot lengths 5.9 mm and 4.7 mm.

These are converted to $\Delta f_{\pm 90°}$=123 MHz and, $\Delta f_{\pm 180°,pq}$=75 MHz, and verified in Fig.'s 5(a) and 5(b).

And then, cavity lengths are decided. Including mechanical factors, the filter is manufactured, going

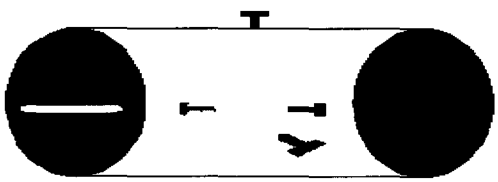


Fig. 3. Group-delay equalizer.

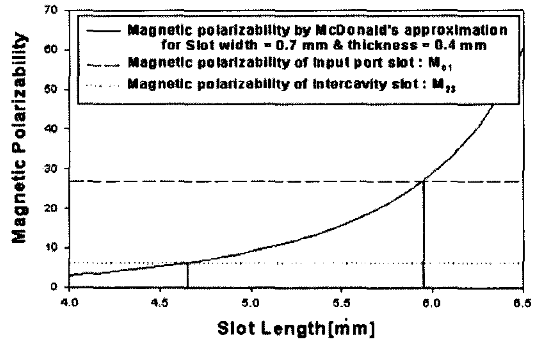
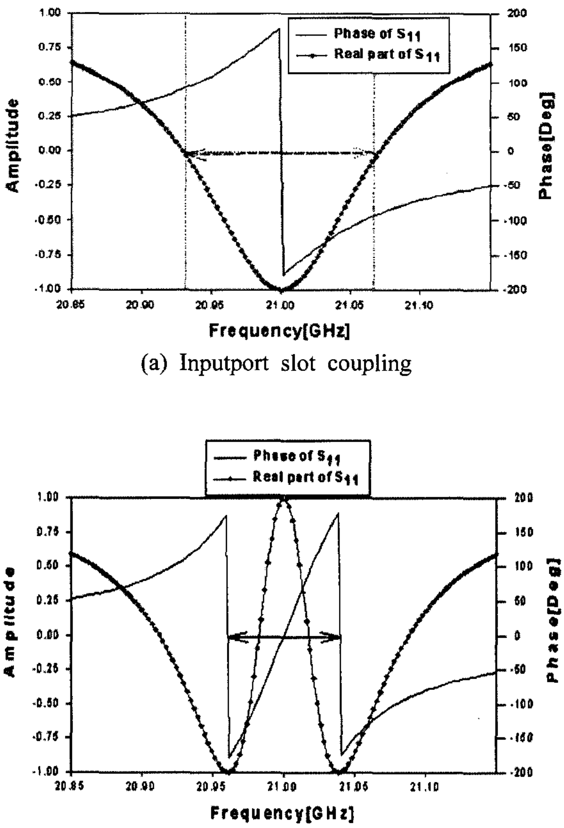


Fig. 4. Initial slot sizes VS magnetic polarizabilities.



(a) Inputport slot coupling

(b) Intercavity slot coupling

Fig. 5. Measurement of the slot coupling.

through wire-cutting, milling and silver-plating for irises, cavities and screws.

The frequency responses of the filter before equalization are presented in Fig. 6. In Fig. 6(a) and (b) show S_{21} and S_{11}.

The bandwidth and rejection turn out to be good. The insertion loss at f_0 is 0.97 dB. The return loss meets the specification. Apart from the larger scales, the variation of the amplitude and group-delay is investigated on a smaller observation scale in Fig. 7.

It is clear that the amplitude and group-delay vary more than their specified levels from CF(=f_0) through both the edges. This is why a group-delay equalizer is needed, which enables the amplitude- as well as group-delay variation to decrease.

The unknowns for the group-delay equalizer are determined as R_{eq}=1.005 and K_{12}=0.0146 where the 2-variable Secant concept is applied for optimization. These values are used for manufacturing the iris and the cavity of the equalizer and adjusting the tuning-screws. When manufactured, the equalizer has W_s and t_s the same as those of the filter. The simulated and measured results of the group-delay before and after equalization are in Fig. 8.

Seeing Fig. 8, the amplitude and group-delay have been effectively flattened at the frequency points near CF(=f_0). Especially, Fig. 8(b) presents the measurements have margins from the mask. Next, S_{21} and S_{11}

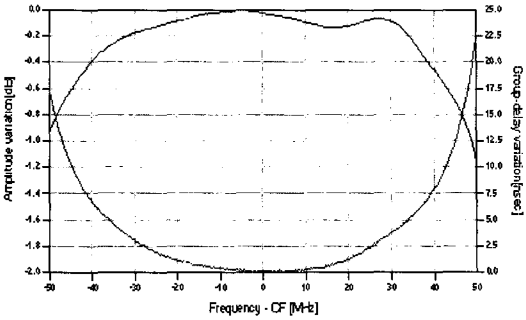
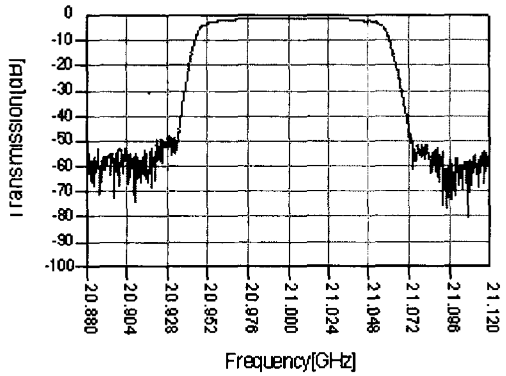
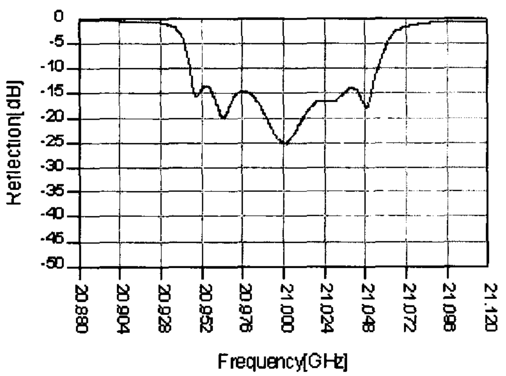


Fig. 7. Amplitude and group-delay variation of the filter before equalization.

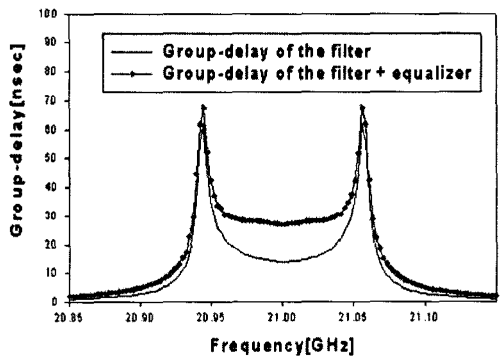


(a) Transmission of the filter before equalization

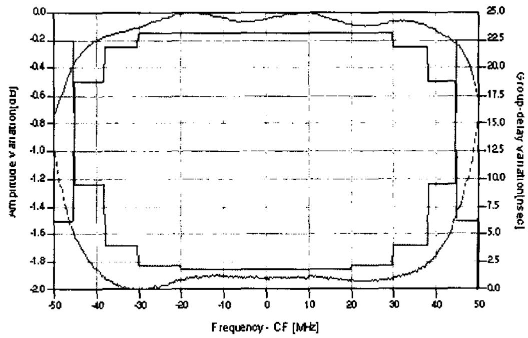


(b) Reflection of the filter before equalization

Fig. 6. Transmission and reflection of the filter before equalization.

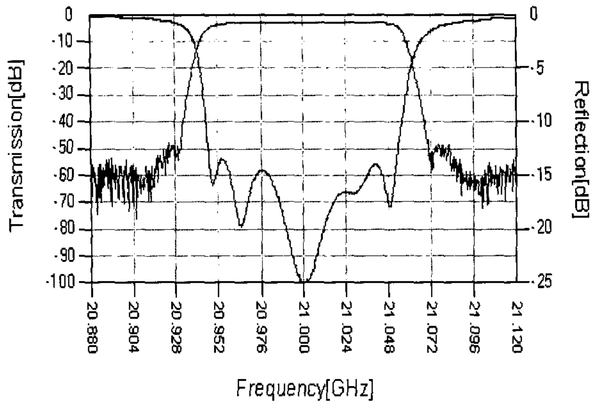


(a) Simulated effect of the group-delay equalization



(b) Measured amplitude and group-delay variation after equalization

Fig. 8. Amplitude and group-delay variation of the filter after equalization.



(a) S_{21} and S_{11} after equalization

(b) Photograph

Fig. 9. S_{21} and S_{11} after equalization and photograph.

after equalization are given in Fig. 9 with the photograph of the realized group-delay equalized filter.

The insertion loss of 3.06 dB at f_0 is less than 4 dB. Besides, as to the near-band rejection, S_{21} is less than -60 dB and complies to the design target.

Ⅳ. Conclusion

In this paper, an 8th-order narrow-band channel filter and its 2-pole group-delay equalizer are designed and manufactured for the Ka-band input multiplexer of the Satellite Transponder. Our measurements reveal that the channel filter has good rejection with the desired bandwidth, but has poor performance of amplitude and group-delay variation. However, the performance is much improved by the group-delay equalizer, and the entire channel results are in compliance with the specific requirements. Thus, our group-delay equalized channel filter is suitable for the satellite input multiplexer.

References

[1] A. E. Williams, "A Four-Cavity Elliptic Waveguide Filter", *IEEE Trans. on Microwave Theory and Technique*, vol. 18, no. 12, pp. 1109-1114, Dec. 1970.

[2] A. E. Atia, A. E. Williams, "Measurements of InterCavity Couplings", *IEEE Trans. on Microwave Theory and Technique*, vol. 23, no. 6, pp. 519-522, June 1975.

[3] R. J. Cameron, J. D. Rhodes, "Asymmetric Realizations for Dual-Mode Bandpass Filters", *IEEE Trans. on Microwave Theory and Technique*, vol. 29, no. 4, pp. 51-58, Jan. 1981.

[4] Ji-Fuh Liang, "Dual-Mode Dielectric or Air- Filled Rectangular Waveguide Filters", *IEEE Trans. on Microwave Theory and Technique*, vol. 42, no. 7, pp. 1330-1336, July 1994.

[5] J. Bornemann et al., *Wavguide Components for Antenna Feed Systems. Theory and CAD*, Norwood, MA: Artech House, 1993.

[6] McDonald, "Simple Approximation for the Longitudinal Magnetic Polarizabilities of Some Small Apertures", *IEEE Trans. on Microwave Theory and Technique*, vol. 36, no. 7, pp. 1141-1144, July 1988.

[7] C. M. Kudsia, "Synthesis of optimum reflection type microwave equalizers", *RCA Review*, vol. 31, pp. 571-595, Sept. 1970.

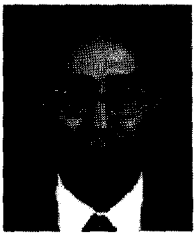
Sungtek Kahng

was with Hanyang University, Seoul, Korea and received the Ph. D. degree of Electronics and Communication Engineering in 2000, with the field of specialty in radio science and engineering. Since 2000, he has been with the Electronics and Telecommunications Research Institute, Daejon, Korea, as a senior staff of research and development, working on the development of microwave equipment and systems for satellites and numerical analysis in electromagnetic characterization of RF devices and circuits.

Man Seok Uhm

received his M.S. degree in electronic engineering from Chungang university, Korea, in 1987 and 1989, respectively. He has joined the Electronics and Telecommunications Research Institute, Daejon, Korea since 1982, as a senior staff of research and development.

Seong Pal Lee



received his Ph. D. degree in electrical and computer engineering from Polytechnic Institute of New York in 1990. Since 1980, he has been with the Electronics and Telecommunications Research Institute, Daejon, Korea, as a principal staff of research and development, and project manager of the communication satellite development center.