# Decoupled Solutions for Design of Filters Having TEM Resonators with Attenuation Poles in Upper Stopband

Yoonmi Shin, Kyungdong Lee, and Bomson Lee, *Member, IEEE*

***Abstract*—The equivalent circuits together with decoupled equations are provided as a solution to the problem of designing the filters having TEM resonators with an attenuation pole in the upper stopband. The proposed method simplifies the design of inserting attenuation poles and gives us much flexibility. The presented design example verifies the exactness of the solutions and shows the usefulness of the approach.**

***Index Terms*—Attenuation poles, bandpass filters, TEM resonators.**

## I. Introduction

THE BANDPASS filters with attenuation poles using a capacitor connected in series with a short-circuited coaxial transmission line have been introduced in [1]–[3]. The authors in the paper [4] propose a modified Chebyshev bandpass filter design method by inserting attenuation poles into the upper or lower stopband using a lumped inductor or a capacitor in series with a resonator. In all of the mentioned papers, several coupled equations must be solved iteratively in order to obtain all circuit values necessary for the filter design. In a new approach proposed in [5], it is suggested that, in order to obtain the required filter element values with ease without going through the complicated procedures of iterations, the values of the adjacent inverters on both sides of the resonator with an attenuation pole be optimized using the linearity of the inverter element values of the conventional bandpass filter. This is just an approximated approach. In this letter, we propose a simple procedure of finding the exact values of the circuit elements necessary for the design of filters having TEM resonator sections with an attenuation pole in the upper stopband, using decoupled solutions. We also show that the structures (RUP3 and RUP7) which were mentioned to be inadequate for a filter structure in [4], are as good filter structures as other ones presented in the paper [4] and can be designed easily with the proposed method. To show feasibility of the proposed method and correctness of the decoupled equations, a design example is presented at the end of this letter.

Manuscript received June 29, 2001; revised December 13, 2001.
The authors are with the Department of Radio Engineering, Kyunghee University, Kyungki-do, South Korea (e-mail: rfshin@khu.ac.kr; kieslow@khu.ac.kr; bomson@khu.ac.kr).
Publisher Item Identifier S 1531-1309(02)02281-X.

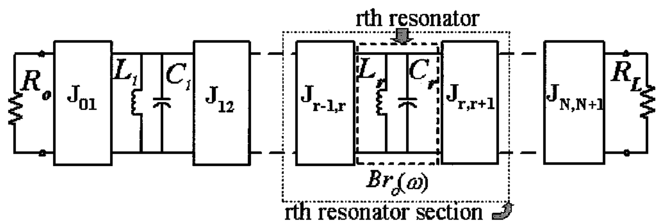


Fig. 1. Bandpass filter network using admittance inverters.

## II. Bandpass Filter Design Using TEM λ/4 Resonator With Attenuation Pole in Upper Stopband

The lowpass prototype filters can be transformed to the bandpass filters that use only parallel tuned resonators consisting of inductors ($L_r$s) and capacitors ($C_r$s) ($r = 1, 2, \ldots, N$), input resistance $R_O$, and load resistance $R_L$, using admittance inverters ($J$s) as shown in Fig. 1 [6], [7].

The angular resonant frequency of each resonator is identical and given by $\omega_o = 1/\sqrt{L_1 C_1} = \cdots = 1/\sqrt{L_r C_r} = \cdots = 1/\sqrt{L_N C_N}$. The susceptance $Br_o(\omega)$ can be expressed as $br_o(\omega/\omega_o - \omega_o/\omega)$, where $\omega$ is the angular frequency and $br_o$ is the susceptance slope parameter given by

$$br_o = \frac{\omega_o}{2} \frac{dBr_o(\omega)}{d\omega}\bigg|_{\omega=\omega_o} = \sqrt{\frac{C_r}{L_r}} = \omega_o C_r. \quad (1)$$

Fig. 2 shows the equivalent TEM resonant circuits with an attenuation pole in the upper stopband. The susceptance $Br_1(\omega)$ of the circuit in Fig. 2(a) is given by

$$Br_1(\omega) = \frac{-1}{Z_r \tan\left(\frac{\omega l_r}{v}\right) + \omega L_p} \quad (2)$$

where, $Z_r$ is the characteristic impedance given by $\pi/(4\omega_o C_r)$, $l_r$ is the length of the TEM resonator given by $\lambda/4$ ($\lambda$: wavelength), $v$ is the phase velocity, and $L_p$ is the inductance of an inductor given by $[-Z_r \tan(\omega_p l_r/v)]/\omega_p$ to realize an attenuation pole angular frequency at $\omega_p$. The susceptance slope parameter $br_1$ of $Br_1(\omega)$ turns out to be that of the resonator shown in Fig. 1 and is given by

$$br_1 = \frac{\omega_o}{2} \frac{dBr_1(\omega)}{d\omega}\bigg|_{\omega=\omega_o} = \frac{\pi}{4Z_r} = \omega_o C_r = br_o. \quad (3)$$

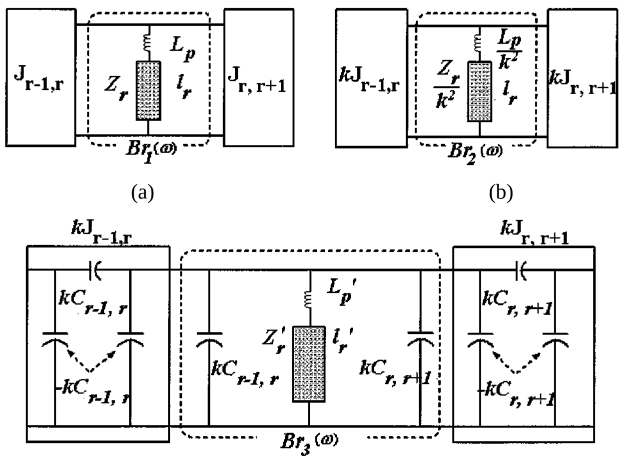

Fig. 2. Equivalent TEM resonant circuits with an attenuation pole in the upper stopband. (a) Basic circuit, (b) equivalent circuit, and (c) equivalent circuit to absorb the negative capacitances.

The circuit in Fig. 2(b) is equivalent to that in Fig. 2(a) for any constant $k > 0$, since the susceptance $Br_2(\omega)$ are be expressed as

$$Br_2(\omega) = \frac{-k^2}{Z_r \tan\left(\frac{\omega l_r}{v}\right) + \omega L_p} = k^2 Br_1(\omega) \qquad (4)$$

and $J$-inverter values on both sides of the resonator are multiplied by $k$ [6]. The introduction of $k$ in Fig. 2(b) and (c) renders more flexibility in choosing realizable circuit elements, maintaining the same circuit performances. The $J$-inverters are usually given by $\pi$ networks as shown in Fig. 2(c). In this case, we add positive capacitances in parallel with the resonator with an attenuation pole to absorb the negative capacitances of the adjacent $J$-inverters. The susceptance $Br_3(\omega)$ is now given by

$$Br_3(\omega) = \frac{-1}{Z_r' \tan\left(\frac{\omega l_r'}{v}\right) + \omega L_p'} + k\omega C$$
$$C = C_{r-1,\,r} + C_{r,\,r+1} \qquad (5)$$

where $C$ is the total capacitance of the added positive capacitors when $k = 1$.

The primed values $l_r'$, $Z_r'$ and $L_p'$ are the new ones changed from $l_r$, $Z_r/k^2$ and $L_p/k^2$, respectively, because of the added capacitances $kC_{r-1,\,r}$ and $kC_{r,\,r+1}$ and have to be determined such that $Br_3(\omega_o) \to 0$, $Br_3(\omega_p) \to \infty$, and $br_3 = br_2 = k^2 br_1 = k^2 br_0$. For a narrowband design, these requirements are sufficient enough for the practical equivalence of the circuits in Fig. 2(b) and (c). Thus, we have the following three coupled equations

$$k\omega_o C(Z_r' \tan x' + \omega_o L_p') - 1 = 0 \qquad (6)$$
$$Z_r' \tan(x'\omega_p/\omega_o) + \omega_p L_p = Z_r \tan y + \omega_p L_p = 0 \qquad (7)$$
$$br_3 = \frac{\omega_o}{2}\left[k^2\omega_o C^2 (Z_r' x' \sec^2 x' + \omega_o L_p') + kC\right]$$
$$= k^2 br_o = k^2 \omega_o C_r \qquad (8)$$

where $x' = \omega_o l_r'/v$, $y = \omega_p l_r/v$, and $C_r$ is the capacitance of the $r$th resonator without an attenuation pole shown in Fig. 1. After lengthy but straightforward algebraic manipulations, (6)–(8) are decoupled to give (9)–(11), shown at the bottom of the page. The value of $l_r'$ is the only unknown variable in (9), and other values have already been properly defined. It is easily found with a desired accuracy using a simple root-finding algorithm and substituted into (10) and (11) to obtain the values of $Z_r'$ and $L_p'$. We have many sets of solutions for $l_r'$, $Z_r'$, and

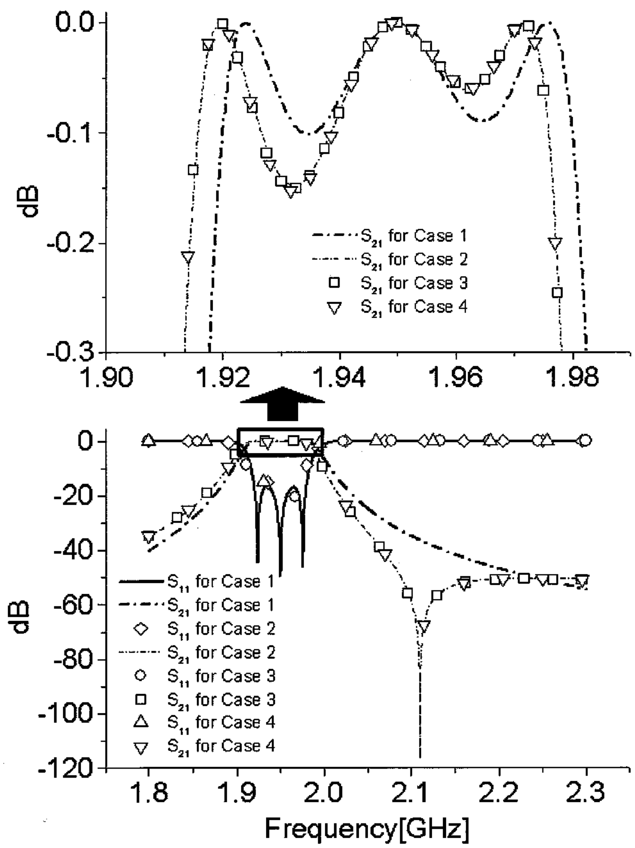


Fig. 3. $S_{11}$s and $S_{12}$s for case 1 to 4. Case 1: without attenuation pole; case 2: with attenuation pole ($k = 1$); case 3: with attenuation pole ($k = 2.5$); case 4: with attenuation pole ($k = 10$).

$$2\omega(kC - C)\tan(x'\omega/\omega) + \omega[Cx' \sec^2 x' + \tan x'(C - 2kC)] = 0 \qquad (9)$$

$$Z_r' = \frac{2(kC_r - C)}{k\omega_o C^2 (x' \sec^2 x' - \tan x')} \qquad (10)$$

$$L_p' = \frac{Cx' + (C - 2kC_r)\sin x' \cos x'}{k(\omega_o C)^2 (x' - \sin x' \cos x')} \qquad (11)$$

TABLE I
SPECIFICATIONS FOR DESIGN EXAMPLE

| Description | Specifications |
|---|---|
| Passband | 1920-1980 MHz |
| Passband ripple | 0.1 dB |
| Attenuation pole frequency | 2110 MHz |
| Number of resonators (N) | 3 |
| Relative permittivity of TEM transmission line ($\epsilon_r$) | 45 |

TABLE II
VALUES OF FILTER ELEMENT WITHOUT ATTENUATION POLE

| | |
|---|---|
| $J_{01}$, $J_{34}$ | 0.00719 ℧ |
| $J_{12}$, $J_{23}$ | 0.00245 ℧ |
| $Z_1$, $Z_3$ | 9.1212Ω |
| $Z_2$ | 9.0779Ω |
| $l_1$, $l_3$ | 5.4262mm |
| $l_2$ | 5.5681 mm |

$L'_p$ as $k$ changes as a parameter. Actually, for any $k$, the circuit in Fig. 2(c) is practically equivalent to that in Fig. 2(b), and thus to that in Fig. 2(a). The equivalent circuit in Fig. 2(c), together with the values of $l'_r$, $Z'_r$, and $L'_p$ obtained from the equations given by (9)–(11), needs only to be substituted into the $r$th resonator section in Fig. 1 for realization of an attenuation pole in the upper stopband. Furthermore, we can choose a convenient $k$ value without affecting the overall filter performance.

## III. DESIGN EXAMPLE

To show the feasibility of the proposed method, we design a Chebyshev bandpass filter of which specifications are given in Table I. We plotted return losses ($S_{11}$s) and insertion losses ($S_{21}$s) for four cases in Fig. 3 based on simulations using Ansoft Serenade V8.5. For all simulations, $R_O$ and $R_L$ are 50 Ω. Case 1 is the one without an attenuation pole. The element values for Case 1 are summarized in Table II. Cases 2–4 are according to Fig. 2(c) with ($k = 1$, $Z'_2 = 5.1547$ Ω, $l'_2 = 5.6072$ mm, $L'_p = 4.1866$ nH), ($k = 2.5$, $Z'_2 = 1.1396$ Ω, $l'_2 = 5.6728$ mm, $L'_p = 0.7631$ nH) and ($k = 10$, $Z'_2 = 0.0846$ Ω, $l'_2 = 5.7147$ mm, $L'_p = 0.0512$ nH), respectively. These values $Z'_2$, $l'_2$, and $L'_p$ corresponding to a specific value of $k$ have been arbitrarily chosen from Fig. 4, which were drawn based on (9)–(11). Fig. 3 shows that $S_{11}$s and $S_{21}$s for the cases from two to four are shown to be exactly the same, having a resonant frequency at 1950 MHz and an attenuation pole frequency at 2110 MHz. Off the resonant frequency of 1950 MHz, they deviate somewhat from those for case 1 due to the effect of the inserted attenuation pole in the upper stopband. The designed filter structure is the combination of the structures (RUP3 and RUP7) which were mentioned to be inadequate for a filter structure in [4]. We have shown that they are as good filter structures as other ones presented in the paper [4] and can be designed easily with the proposed method.

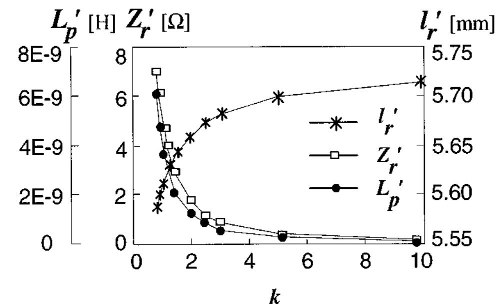


Fig. 4. Values of $Z'_r$, $l'_r$, and $L'_p$ as a function of $k$ (in the case of $r = 2$).

## IV. CONCLUSION

Convenient equivalent circuits together with decoupled equations have been provided as a solution to the problem of designing filters having TEM resonators with an attenuation pole in the upper stopband. The presented design example verified the correctness of the solutions and showed the usefulness of the approach. A similar approach can be applied to the case of inserting an attenuation pole in the lower stopband.

## REFERENCES


[1] H. Matsumoto, T. Yorita, Y. Ishikawa, and T. Nishikawa, "Miniaturized duplexer using rectangular coaxial dielectric resonators for cellular portable telephone," *IEICE Trans.*, vol. E74, no. 5, pp. 1214–1220, May 1991.

[2] T. Nishikawa, "RF front-end circuit components miniaturized using dielectric resonators for cellular portable telephones," *IEICE Trans.*, vol. E74, no. 6, pp. 1556–1562, June 1991.

[3] H. Matsumoto and T. Nishikawa, "Design of miniaturized dielectric duplexer with attenuation poles" (in Japanese), *IEICE Trans.*, vol. J76-C-1, no. 5, pp. 164–172, May 1993.

[4] J. S. Lim and D. C. Park, "A modified Chebyshev bandpass filter with attenuation poles in the stopband," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 898–904, June 1997.

[5] Y. J. Ko, J. H. Kim, and B. K. Kim, "A novel approach for the design of a bandpass filter with attenuation poles using a linear relationship," *IEICE Trans. Electron.*, vol. E82-C, no. 7, pp. 1110–1115, July 1999.

[6] R. E. Collin, *Foundations for Microwave Engineering*, 2nd ed. New York: McGraw-Hill, 1992.

[7] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks and Coupling Structures*. New York: McGraw-Hill, 1964.