Cross-Coupled Microstrip Combline Bandpass Filter Using Stepped-Impedance Resonators

Young-Ho Cho · Seung-Un Choi · Sang-Won Yun

Abstract

In this paper, a cross-coupled microstrip combline bandpass filter using stepped-impedance resonators(SIRs) is proposed. In order to improve the selectivity as well as the insertion loss, the SIR configuration is used. The cross coupling is also introduced to enhance the selectivity. The improvement of the insertion loss is demonstrated not only by deriving the quality factor of the SIR but through the measured performances. Both the proposed and the conventional combline bandpass filter with 5 % of fractional bandwidth at 2 GHz were fabricated and tested. Compared to the conventional combline bandpass filter, the proposed one exhibits the improved selectivity as well as the lower insertion loss characteristics.

Key words : Combline Filter, Stepped-Impedance Resonator, Cross Coupling, Microstrip.

Ⅰ. Introduction

The combline bandpass filter has been widely used in various types, such as cavity, ceramic and microstrip configurations. Furthermore, this structure can easily be applied to the tunable filter. As miniaturized designs, microstrip or stripline configurations with capacitive loading as shown in Fig. 1(a) are widely found, which can be based on LTCC process for the further reduction in size. However, the degradation of the insertion loss as well as the passband flatness is inevitable in the miniaturization process.

In order to overcome the aforementioned problems, several methods have been suggested. The filters using active resonator can improve the insertion loss by negative resistance, but the design depends on the complicated procedures and might introduce the potential instability[1]. The selectivity of filter also can be improved by the cross coupling method, however the insertion-loss is not improved under this scheme[2].

This letter presents capacitive loaded combline filter using the new cross coupling structure and the stepped-impedance resonators(SIRs). Many cross-coupled filters using SIRs were already introduced[3]~[7]. However, was not explained sufficiently why transmissions zeros are generated by the used cross coupled structures in the letters, nevertheless transmission zeros are not generated unconditionally by all cross coupled structures. Therefore, we will introduce the reason by using Darlington's method that the used cross coupled structure in Fig. 1 can generate transmission zeros[8]. As well as, these letters did not introduce that the quality factor of resonators can be improved by using SIRs. The insertion loss is also improved, because the SIR has the better quality factor than the conventional microstrip quarter-wavelength transmission line resonator. In this paper, it will be proved that the SIR has the better quality factor than the conventional microstrip quarter-wavelength transmission line resonator.

The new cross coupling structure is introduced in Section Ⅱ-A and quality factor of SIR is calculated in Section Ⅱ-B. The measured results of the proposed filter and the conventional combline filter are described in Section Ⅲ.

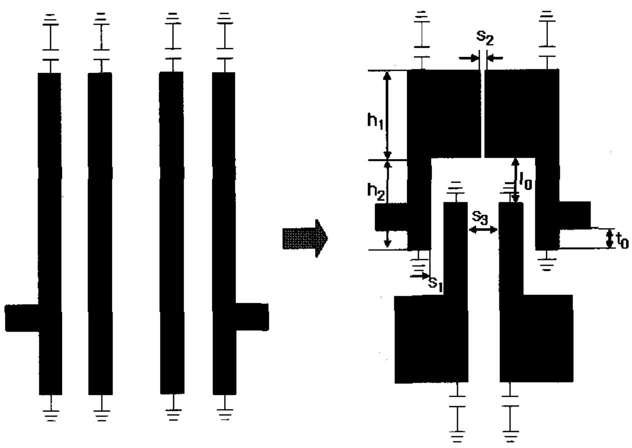

Fig. 1. (a) The conventional combline bandpass filter with lumped elements. (b) The proposed cross-coupled combline bandpass filter using capacitive loaded stepped-impedance resonator.

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Dept. of Electronic Engineering, Sogang University, Seoul, Korea.

Ⅱ. Analysis of Proposed Filter Configuration

2-1 New Cross Coupling Structure

It is difficult to generate the desired cross coupling in the conventional combline bandpass filter loaded by capacitors as shown in Fig. 1(a). To generate transmission zeros by using previous cross coupled methods, open ends of the 1st and the 4th resonators are coupled each other such as Fig 1(b)[9]. Also, it is difficult to apply SIRs in the structure of Fig. 2(b). However lumped capacitors of 2nd and 3rd resonators disturb cross coupling as shown in Fig. 2(b). So another cross-coupling method is needed. Another cross coupling method can be easily obtained by Darlington's procedure.

In the proposed filter configuration it can be easily realized as shown in Fig. 1(b). Fig. 1(a) can be transformed to Fig. 1(b) by applying the new cross coupling structure. In the following analysis, it is verified by means of Darlington's procedures that the proposed filter has the same transfer function as the conventional cross-coupled filter with the transmission zeros[8].

The matrix (1) is the general form of the cross-coupled filter with the transmission zeros[4]. The element M_{14} only has opposite sign compared the rest of the elements. Matrix (2) represents even-mode coupling matrix (1), and the Te is even-mode orthogonal-transformation matrix of (1).

According to [8], the poles of even-mode admittance are the eigenvalues of the even-mode coupling matrix, and the residues of even-mode admittance are the squares of the first row of even-mode orthogonal transformation matrix Te. If an arbitrary matrix has the same eigenvalues of even-mode coupling matrix and the squares of the first row of even-mode orthogonal transformation matrix, the transfer functions of these two matrices are identical to one another.

$$[M]=\begin{bmatrix} 0 & M_{12} & 0 & M_{14} \\ M_{12} & 0 & M_{23} & 0 \\ 0 & M_{23} & 0 & M_{12} \\ M_{14} & 0 & M_{12} & 0 \end{bmatrix} \tag{1}$$

$$[M]_e=\begin{bmatrix} M_{14} & M_{12} \\ M_{12} & M_{23} \end{bmatrix} = -[T_e]\begin{bmatrix} \lambda_{e1} & 0 \\ 0 & \lambda_{e2} \end{bmatrix}[T_e],$$

$$[T_e]=\begin{bmatrix} T_{e11} & T_{e12} \\ T_{e21} & T_{e22} \end{bmatrix} \tag{2}$$

The eigenvalues of (2) is

$$\lambda_{e1,2}=\frac{M_{23}-M_{14}\pm\sqrt{(M_{14}-M_{23})^2+4(M_{23}M_{14}+M_{12}^2)}}{2} \tag{3}$$

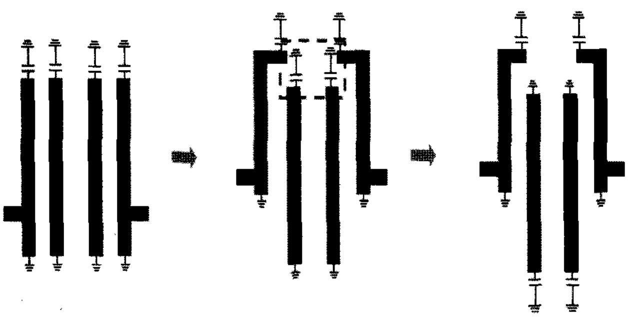
We can obtain (4) from (2).

$$M_{12}=-\lambda_{e1}T_{e11}T_{e21}-\lambda_{e2}T_{e12}T_{e22} \tag{4}$$

In (3), the sign of M_{12} does not affect the eigenvalues, because the M_{12} exists as a form of square. Moreover, the square of T_{e11} and T_{e12} in (2) can also be maintained even though the sign of M_{12} is changed in (4). Hence, a new coupling matrix which has an identical transfer function with the conventional cross coupling matrix can be derived such as follows:

$$[M]_{new}=\begin{bmatrix} 0 & M'_{12} & 0 & M_{14} \\ M'_{12} & 0 & M_{23} & 0 \\ 0 & M_{23} & 0 & M'_{12} \\ M_{14} & 0 & M'_{12} & 0 \end{bmatrix}, \quad M'_{12}=-M_{12} \tag{5}$$

The proposed structure is satisfied with matrix (5) as shown in simulated results of Fig. 2. The simulated coupling coefficients (M_{ij}) in Fig. 3 show that the phase responses of M_{12} and M_{14} are opposite to that of M_{23}.



(a) (b) (c)

Fig. 2. (a) The conventional combline bandpass filter with lumped elements, (b) The conventional combline bandpass filter using the conventional cross-coupled structure, (c) The conventional combline bandpass filter using the proposed cross-coupled structure.

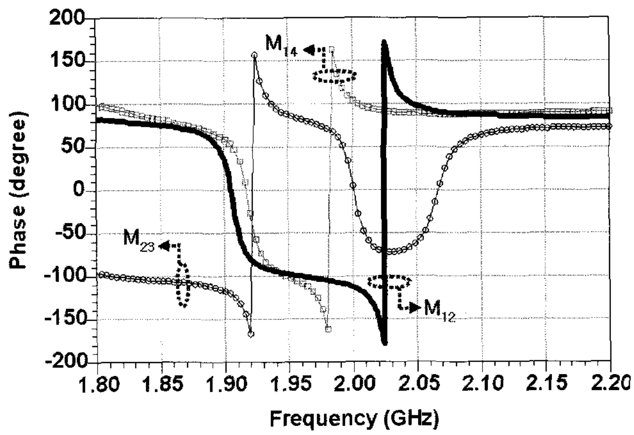


Fig. 3. Simulated phase responses of the coupling coefficients of the proposed combline filter in Fig. 1(b).

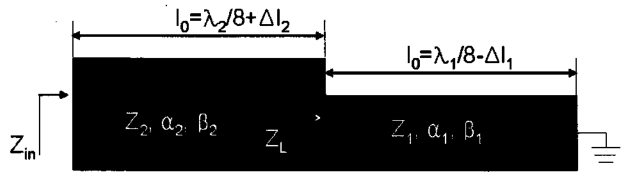


Fig. 4. Parameters of a stepped-impedance resonator.

Therefore, it can be known that transmission zeros can be generated by the cross coupled structure such as Fig. 1(b).

2-2 Quality Factor of Stepped-Impedance Resonator

Here is the procedure that proves the better quality factor can be obtained when general quarter-wave resonators transfer into SIRs in combline filter. Fig. 4 shows the typical SIR whose one end is open and the other is short circuited, and each section has the same length 'l_0' [10].

The impedance of short-end transmission line is expressed as

$$Z_L = Z_1 \frac{\tanh \alpha_1 l_0 + j \tan \beta_1 l_0}{1 + j \tanh \alpha_1 l_0 \tan \beta_1 l_0} \qquad (6)$$

If we let $w=w_0+\Delta w$ [6] and $\Delta l_1 \approx \Delta l_2 = \Delta l$, we obtain

$$\tan \beta_1 l_0 = \frac{1 - \tan\left(-\frac{w_0 \Delta l}{v_p} + \frac{\pi \Delta w}{4 w_0} \right)}{1 + \tan\left(-\frac{w_0 \Delta l}{v_p} + \frac{\pi \Delta w}{4 w_0} \right)} = \frac{1 - \tan \theta_1}{1 + \tan \theta_1} \qquad (7)$$

Also, $\tanh \alpha_1 l \approx \alpha_1 l$ is used for small loss[6]. Substitution of (7) into (6), and applying the condition $\theta_1 \ll 1$ leads to

$$Z_L \doteqdot Z_1 \frac{\alpha_1 l_0 + j \frac{1-\theta_1}{1+\theta_1}}{1 + j \alpha_1 l_0 \frac{1-\theta_1}{1+\theta_1}} \doteqdot \frac{2\alpha_1 l_0 Z_1}{(1-\theta_1)^2} + j \frac{Z_1}{(1-\theta_1)^2} \qquad (8)$$

Then, assuming $\theta_2 = w_0 \Delta l_2 / v_p + \pi \Delta w / 4 w_0 \ll 1$, one can write

$$Y_{in} = \frac{1}{Z_{in}} = \frac{8 Z_1 Z_2 \alpha_1 l_0 + 4 \alpha_1 l_0 (Z_1^2 + Z_2^2)}{(Z_1 + Z_2)\{(Z_1 + Z_2) - 4(Z_1 \theta_2 + Z_2 \theta_1)\}} + j \frac{\{Z_1^2 - Z_2^2 + 2\theta_2 (Z_2 + 2 Z_1 Z_2 - Z_1^2) + 4 Z_2^2 \theta_1\}}{(Z_1 + Z_2)\{(Z_1 + Z_2) - 4(Z_1 \theta_2 + Z_2 \theta_1)\}} \qquad (9)$$

In addition, From Fig. 5,

$$l_0 = \frac{2\pi}{8\beta_1} + \Delta l = \frac{2\pi}{8\beta_1} - \Delta l \qquad (10)$$

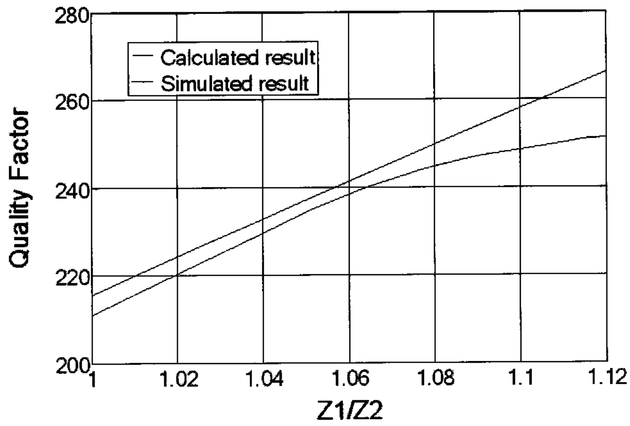


Fig. 5. Quality factor values calculated using (11) and simulated using ADS 2006(RO3003 20 mil, copper clad, f_0=3 GHz, Z_1=50 Ω).

Therefore, the length 'l_0' of SIR can be derived as

$$l_0 = \frac{\pi}{8} \left(\frac{1}{\beta_1} + \frac{1}{\beta_2} \right) \qquad (11)$$

as shown in Fig. 3.

Using (9) and (10), the quality factor of the SIR is derived as

$$Q = \frac{2 w_0}{\pi \Delta w} \frac{(Z_1 + Z_2)\left\{ Z_1 - Z_2 + (3 Z_2 - Z_1) \frac{\pi \Delta w}{2 w_0} \right\}}{\left(\frac{1}{\beta_1} + \frac{1}{\beta_2} \right) \{ 2 Z_1 Z_2 \alpha_1 + \alpha_2 (Z_1^2 + Z_2^2) \}} \qquad (12)$$

When the impedance condition is $Z_1 = Z_2$ in (12), the quality factor of resonator becomes $\beta / 2\alpha$, which is equal to that of typical short circuited $\lambda/4$ resonator[11]. Fig. 5 represents the graph of quality factor of SIR which is dependent directly upon the ratio of Z_1 and Z_2. These results show that the higher Z_1/Z_2 value gives the better quality factor. However, the error between the simulated and the calculated results is gradually increasing when the Z_1/Z_2 increases as in Fig. 5. It is mostly caused by parasitic effects at the discontinuity of the stepped-impedance line.

Ⅲ. Measurement Results

The proposed combline filter and the conventional combline filter in Fig. 1 were designed with 5 % of fractional bandwidth at 2 GHz. Initially, the characteristic impedance values of the SIR were selected as Z_1=50 Ω and Z_2=23 Ω considering coupling coefficient values, and capacitances at the ends of the resonators were chosen as 0.7 pF. The same values of the capacitances were also used for the 50 Ω conventional resonators in Fig. 1(a). Quality factors of the SIR with lumped capacitor and the conventional resonator with lumped capacitor were measured as 150 and 120, respectively. The coupling matrix of the proposed filter was given as

$$[M]=\begin{bmatrix} 0 & -0.040 & 0 & -0.010 \\ -0.040 & 0 & 0.037 & 0 \\ 0 & 0.037 & 0 & -0.040 \\ -0.010 & 0 & -0.040 & 0 \end{bmatrix} \quad (13)$$

The dimensions of designed filter are derived from (13)[9]. They are given by h_1=h_2=7 mm, s_1=0.8 mm, s_2=0.5 mm, s_3=1.5 mm, l_0=2 mm and t_0=1.5 mm. The length of resonators in the conventional combline bandpass filter of Fig. 1(a) is 9.2 mm. As shown in Fig. 6, the proposed bandpass filter has the enhanced insertion

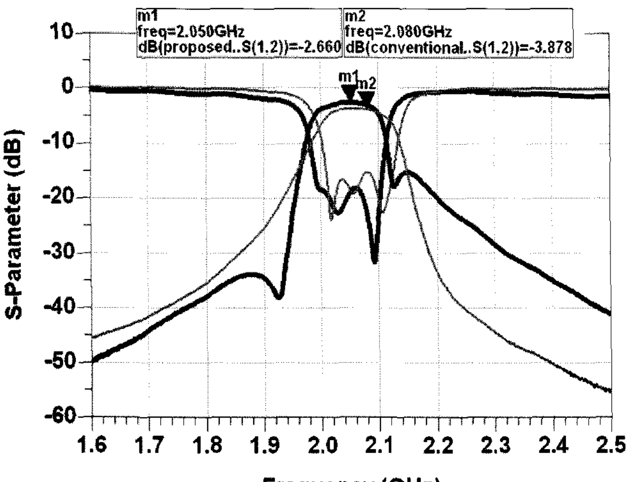


Fig. 6. Measured results of the conventional combline Bandpass filter(Gray-line) in Fig. 1(a) and the proposed filter(Black-line) in Fig. 1(b).

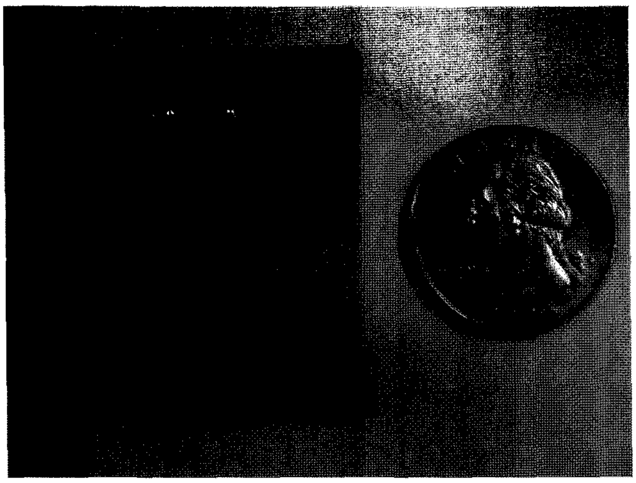


Fig. 7. Photograph of the cross-coupled microstrip combline bandpass filter using stepped-impedance resonators.

loss and selectivity characteristics in comparison with those of the conventional combline bandpass filter. The insertion loss is improved about 1 dB.

Ⅳ. Conclusion

The cross-coupled combline bandpass filter using the capacitive loaded SIR has been proposed. In order to apply the cross coupling in the combline filter configuration, a new coupling structure was introduced so that the cross coupling created the transmission zeros, which results in the improvement of the selectivity. It is also verified that the SIR has an improved quality factor in comparison with the conventional quarter-wave resonator. It is believed the proposed filter can be applied to the miniaturized designs.

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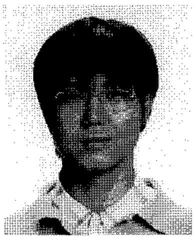
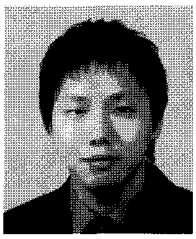
Young-Ho Cho



received the B.S. degree in electronic engineering from Sogang University, Seoul, Korea in 2005. He is currently working toward Ph.D. degree in electronic engineering at Sogang University. His research interests include synthesizers, filters and RF systems.

Seung-Un Choi



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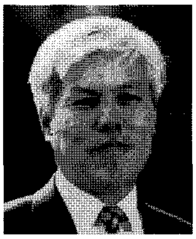
Sang-Won Yun



received the B.S. and M.S. degrees in electronic engineering from Seoul National University, Seoul, Korea in 1977 and 1979, respectively, and the Ph.D. degrees from University of Texas at Austin in 1984. Since 1984, he has been a professor in the department of electronic engineering, Sogang University, Seoul, Korea. From January 1988 to December 1988, he was a visiting professor at the University of Texas Austin. His research interests include microwave and millimeter-wave devices and systems.