The Compact UHF CT-Type BandPass Filter with a Mixed Coupling

Jeongho Ju · Sungtek Kahng

Abstract

This paper presents the design of the compact UHF Bandpass filter of CT-type with a novel mixed coupling topology. In further detail, this new coupling structure mathematically equals the transmission zero that enables particularly the better rejection performance in the stopband without increasing the filter's order. To envision this design concept, we fabricated a microstrip filter whose measurement proves to agree with the predicted performance and the suggested coupling structure reduces the overall size, and will be possibly used in the miniaturized UHF apparatus such as the RFID system.

Key words : Bandpass Filter, UHF Application, Cascaded Triplet(CT), Cross-Coupling, Mixed Coupling, Size Reduction.

Ⅰ. Introduction

The rapid growth of the wireless and mobile communication technologies has driven designers to cope with downsizing the components and minimizing their cost together with maximizing functionality. Especially, as a core part to the communication system, the bandpass filter(BPF) is guaranteed to achieve the lowest insertion and return loss and best selectivity possible. However, a good selectivity usually requires a higher ordered filter, which ends up with the longer guiding structure and the resultant increase of insertion loss, if the normal Chebyshev type of BPF is employed.

Alternatively, the cross-coupling mechanism with the Elliptic integral filter has been studied and suggested to achieve the sharp skirt(namely high selectivity) with relatively lower insertion loss[1]~[5]. Nevertheless, when it comes to the use of the microstrip line loop filter, the conventional structure has shown the shortcomings such as limited freedom in alignment of resonators due to either *E*- or *M*-coupling, and restrictions in feasible gap for a desired stronger coupling.

In this paper, we propose a microstrip BPF for a UHF communication system, with a mixed coupling equivalent to cross-coupling in the Cascaded Triplet(henceforth, 'CT') topology for assuring an excellent selectivity performance. Specifically, the mixed coupling is obtained by inserting a resonator to link the adjacent loops more tightly, which will generate a transmission zero without having to increase the order of the filter. This idea will be envisaged by the fabricated version of design that will show the reduction in the overall size and the performance of selectivity and insertion- and return loss which agree well with the requirements on the design. This will convince the UHF systems' engineers of possibly using the proposed technique.

Ⅱ. Design

Fig. 1 is a typical microstrip line two-loop filter[1],[2]. Each of the two loops plays a resonator.

The structure in Fig. 1 is simple to use: The gap between the loops and individual loops' open-gaps, line length and loop alignment. Changing the open gap of a resonator near its counterpart will introduce an electric coupling as a dominant linkage between the two loops. On the contrary, regarding the vertical parts of the loops where they are closest, the projection of one resonator's side to the other determines a magnetic coupling.

There are demands on designing the BPF included in the UHF application and the requirements of its electrical performances can be summarized as the table below.

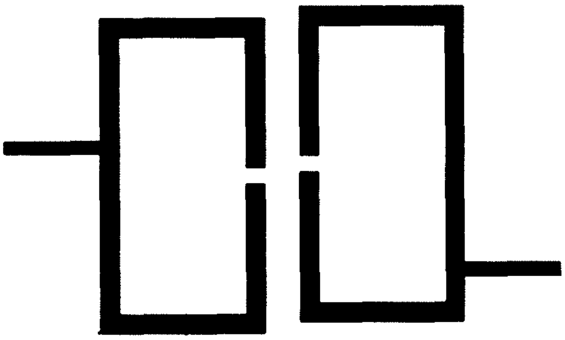

Fig. 1. Ordinary microstrip two-loop filter.

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Table 1. Requirements on the UHF filter to be designed.

Item	Spec.
Center frequency	920 MHz
Bandwidth	80 MHz
Insertion loss	<2 dB
Ripple level	<0.5 dB
Return loss	<−15dB
Attenuation(Asymmetric)	<−5 dB @ 850 MHz and <−15 dB @ 980 MHz

Seeing the insertion loss with the bandwidth as given in Table 1, the design is no big a deal to adopt the typical two loop filter. What is much more concerned is the specification on the asymmetric attenuation performance which considers the possible out-band spurious interference from the other RFID application in the nearby GHz regime. The attenuation level less than −15 dB at the 20 MHz-offset from the upper edge of the bandwidth is harsh and challenging with the geometry in Fig. 1 from the conventional design point of view. To meet this requirement and take into account the asymmetry in S_{21}, it is inevitable to choose a higher order and a transmission-zero other than Fig. 1.

What we would like to do is to maintain the overall size of the filter as that of the two-loop filter and improve the attenuation performance with the increased order of the filter which leads to the eventual saving in size. This motivates us to come up with a cross-coupling mechanism. The technical details are as follows for the design. The coupling coefficient matrix $\overline{\overline{M}}$ as the unkowns can be solved via the generalized impedance and its scattering parameters as follows.

$$\overline{\overline{Z}}\cdot\overline{I}=\overline{e}, \tag{1}$$

where $\overline{e}^T=(1,\ 0\ ,0,\ \cdots,\ 0\ ,0)$ is the voltage excitation vector and $(\overline{I})^T=(i_1,\ i_2,\ i_3,\ \cdots,\ i_{n-1},\ i_n)$ is the current vector including all the resonators. Also, the generalized impedance $\overline{\overline{Z}}$

$$\overline{\overline{Z}}=j(\tau\overline{\overline{U}}+\overline{\overline{M}}) \tag{2}$$

has self- and mutual coupling values of the network.

$$\tau=\frac{f_0}{\triangle f}\left(\frac{f}{f_0}-\frac{f_0}{f}\right). \tag{3}$$

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

$$S_{21}=-2\sqrt{R_{in}R_{out}i_n}=\mathit{filtershape} \tag{4}$$

Table 2. Coupling coefficient matrix $[M]$ in use.

Mij	1	2	3
1	−0.276	1.045	−0.658
2	1.707	0.486	1.075
3	−0.658	1.075	−0.276

and

$$S_{11}=1-2R_{in}i_I. \tag{5}$$

where R_{in} and R_{out} mean input and output port resistance. Here is the coupling coefficient matrix *[M]* with the order of 3 to meet the things from Table 1 by way of the transfer function.

This has been obtained to have the transfer function that fits the amplitude of the required performance. The elements in Table 2 will be converted to the physical dimensions by way of the following equations[1]~[5].

$$K=\Delta f\cdot Mij/f_0 \tag{6}$$

$$K=(f_e{}^2-f_m{}^2)/(f_e{}^2+f_m{}^2) \tag{7}$$

where f_e and f_m denote the resonance frequencies of electrical and magnetic couplings, respectively. And the center frequency f_0 is with the bandwidth Δf. The geometrical dimensions are found by equating (7) from S_{21} of the Full-Wave Simulation to that from the Table 2.

Now we propose a new mixed coupling structure of the BPF to meet the expectation above.

The M-shaped line is enclosed by the left loop and right loop and crosses the two loops through the individual open gaps. Consequently, this newly proposed structure plays the cross-coupling that enables the transmission zero to occur. This zero induces the out-of-phase difference between the direct signal from resonator

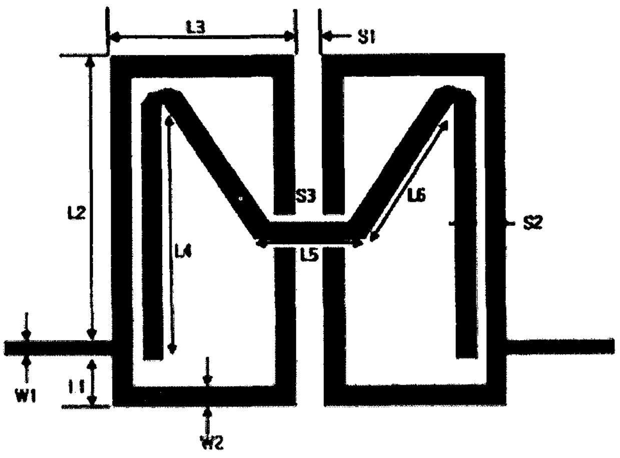

Fig. 2. Proposed mixed coupling structure with two loops.

1 to resonator 3 and that along the resonator 2. In other words, the destructive interference in obtained. If we name the left and right loops resonators 1 and 3, respectively, with the crossing line as resonator 2, without loss of generality, (S_2, S_3) and (S_1, L_5) will affect (M_{12}, M_{23}) and M_{13}, respectively. W_1 and L_1 are treated essential to the 50 Ω-impedance matching that is critically important to make sure good return loss performance. L_1, L_2 and L_3 are exploited to adjust the resonance frequency of resonator 1 or 3, while L_4, L_5 and L_6 decide f_0 of resonator 2. Finding the values of the design parameters in Fig. 2, the filter's functions are optimized to follow the requirements. When the design is done as desired, we will see the significant improvement in the electrical performance with the unchanged size from the simple two-loop filter in Fig. 1.

Ⅲ. Results of Design

Entering 10.8 and 1.27 mm as the substrate's relative dielectric constant and thickness, respectively, the design parameters have been found: W_1=1, W_2=1.4, L_1=3.3, L_2=18.6, L_3=11.9, L_4=16.8, L_5=6, L_6=10.7, S_1=1.8, S_2 =0.53, S_3=2.1 and all are expressed in mm. With these physical dimensions, a world acclaimed commercial 3D-EM simulator has been used to visualize the field and current's exact distributions to unveil the proposed filtering and coupling mechanisms and plot the results of the transmission(S_{21}) and reflection(S_{11}) performances.

The resultant transmission property shows the center frequency and the bandwidth are obtained as required and at the center frequency the insertion loss of around 1 dB that complies to the specification. This is explained even though the higher order is chosen for the present design, compared to Fig. 2, the size of the entire guiding structure remains unchanged. Most importantly, the transmission zero has been obtained at approximately 1 GHz, showing the asymmetry in need, attenuation levels of less than −15 dB at the upper offset frequency of 980 MHz and less than −5 dB at the lower offset frequency of 850 MHz(They read less than −22 dB at 980 MHz and less than −8 dB at 850 MHz). With regard to the reflection, the average level of S_{11} throughout the bandwidth is less than −15 dB. However, since the first one of the three ideal poles is embedded near the lower band edge, measures need to be taken such as moving the central pole at 910 MHz slightly upward(the distance between the last two ideal poles is relatively big). Convinced of the validity of our design methodology through the findings from the simulation, we have fabricated the designed filter.

In conjunction with the fabrication, we conducted the measurement on the properties of the filter's transmission(S_{21}) and reflection(S_{11}).

In Fig. 5, we have compared the measured data(last two curves) with the simulated ones(first two curves). They both are in excellent agreement as showing almost the same transmission zero and attenuation level, though there has occurred a little discrepancy that the first ideal pole has been retrieved in the tuning process in the measurement. Therefore, the bandwidth is securely obtained in the measurement showing the three ideal poles. So we can judge the whole design and experimental work on the newly proposed structure has been successfully done, in compliance with the requirements.

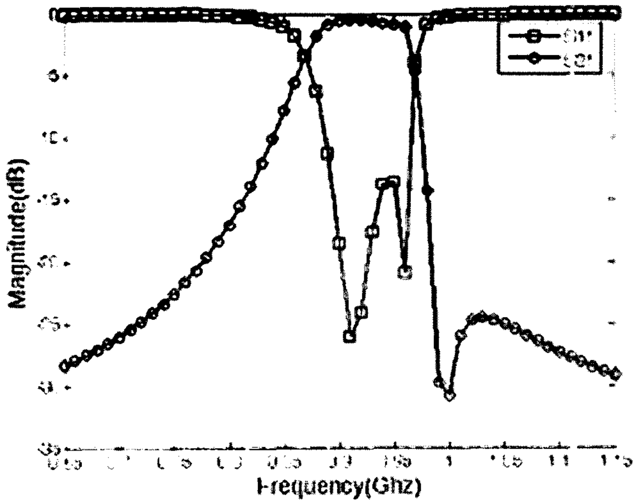

Fig. 3. Simulated S_{21} and S_{11} results of the proposed bandpass filter.

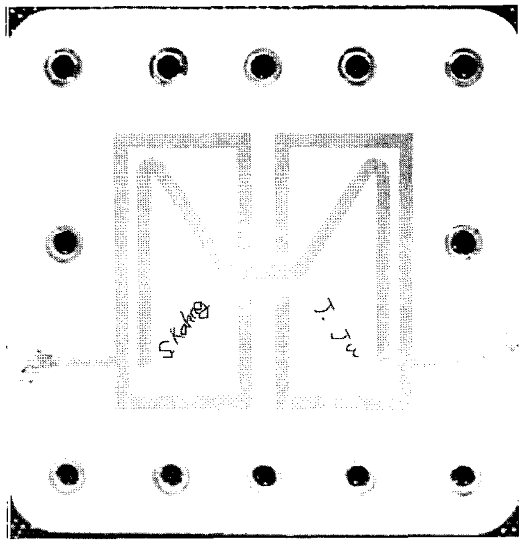

Fig. 4. Photo of the fabricated filter.

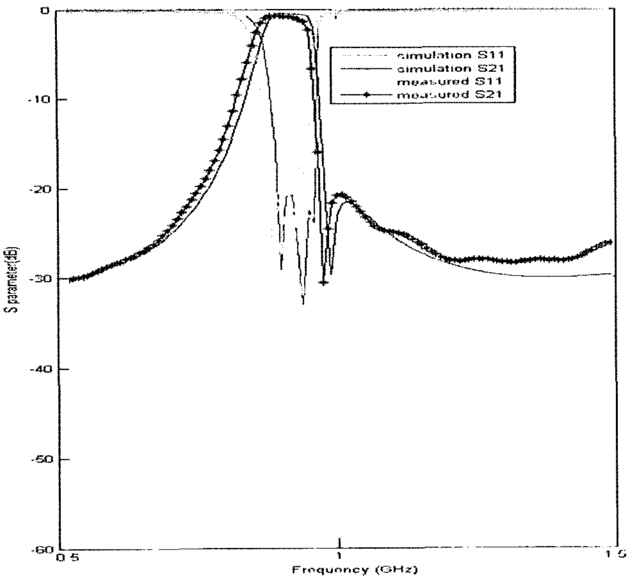

Fig. 5. Measurements with the simulated results.

IV. Conclusion

A new structure for the cross-coupling realized in a bandpass filter has been suggested to have a transmission zero. This enables the filter to show a high selectivity feature with meeting a challenging requirement on asymmetric attenuation levels, to reduce the spurious interference from the immediate neighbor to the UHF application like the RFID. Using the proposed structure, it can save the total length of the filter and enhance the selectivity, insertion and return loss characteristics.

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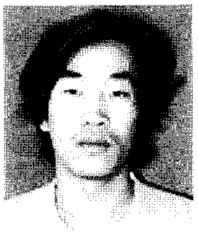

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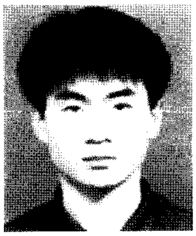

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