A 70 MHz $\sim$ 270 MHz Electrical Tunable LC Bandpass Filter Based on Mixed Coupling and Cross-Coupling

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Abstract—This paper presents an electrical tunable bandpass filter based on tunable LC resonators loaded with semiconductor varactors. Magnetic dominated mixed coupling between the tunable resonators is utilized to compensate the bandwidth of the tunable filter. Cross coupling is created by using magnetic dominated mixed coupling between the resonators and source to load electrical coupling, and two transmission zeros are generated beside the passband. The tunable mechanism of the proposed filter is studied. The tunable filter is analyzed, designed, fabricated and measured. The measurement shows that the filter can be tuned from 70 MHz to 270 MHz with a fractional bandwidth from 27% to 21%.

1. INTRODUCTION

RF tunable filters, which can greatly increase the flexibility of filtering, are essential components for the radio frequency signal processing in wide-band/multi-band wireless communications [1–3]. Recently, microstrip tunable filters based on semiconductor varactor diodes have been widely developed because of their fast tuning speed and low cost [4–10]. However, these tunable filters based on microstrip technology are always applied to the frequency above 300 MHz. LC filter is another physical type filter with a simple structure and design theory. LC filters have a wide range applications in wireless communications from several MHz to GHz [11–17]. Recently, some tunable LC filters have been proposed. A two-stage Q-enhanced 4th-order LC band-pass filter is presented in [15] based on 55 nm CMOS technology, and its center frequency can be tuned from 2.35 GHz to 2.48 GHz [16] introduces a novel compact LC resonator-based bandpass filter with tunable transmission zeros. A tunable LC bandstop filter based on MEMS technology with wide tuning range is designed in [17]. The process of tunable LC filter can be by CMOS [12, 14, 15], PCB [13, 16], MEMS [17], etc. Therefore, it can be easily integrated with other components. On the other hand, cross coupling topologies have been widely used in microwave filter design to increase the selectivity of the filters [18, 19]. However, the tunable LC filter with cross coupling has been rarely reported.

In this paper, a tunable LC bandpass filter with cross-coupling and bandwidth compensation is proposed. The LC type cross-coupling is generated based on the source-load electric coupling and the magnetic dominated mixed coupling between the tunable LC resonators. Two transmission zeros are generated beside the passband to improve the frequency selectivity and out-band rejection. The LC type magnetic dominated mixed coupling is utilized to compensate the bandwidth of the tunable bandpass filter. The filter is analyzed, designed, fabricated and measured. The measurement shows that the filter can be tuned from 70 MHz to 270 MHz with a fractional bandwidth from 27% to 21%. In Section 2, the basic design theory is presented. Sections 3 and 4 discuss the simulation and measurement, respectively. The conclusion is given in Section 5.
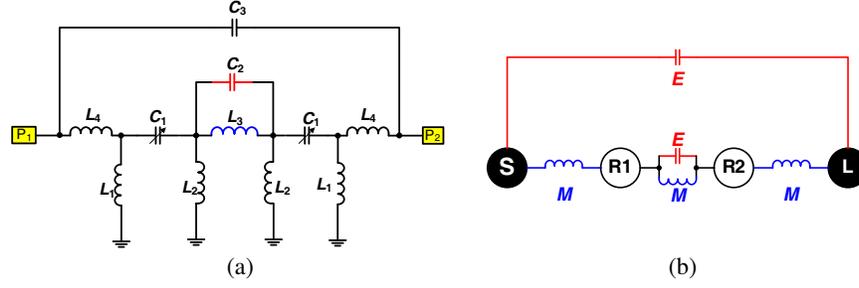
Figure 1. (a) Circuit model of the proposed tunable LC bandpass filter, (b) coupling topology.

2. FILTER DESIGN

Figure 1(a) shows the circuit model of the proposed bandpass filter based on LC resonators. In this model, L_1, C_1 and L_2 constitute two resonators on both sides, and the resonator can be tuned by changing the capacitance of C_1. The two resonators are coupled by C_2 and L_3. C_2 is very small; therefore, a magnetic dominated mixed coupling between the resonators can be achieved. The fractional bandwidth of the filter is proportional to the coupling coefficient $|k|$, and the mixed coupling coefficient can be written as [1]:

$$|k| = |k_M - k_E| \tag{1}$$

where k_M is the magnetic coupling coefficient, and k_E is the electric coupling coefficient. Since the filter is tuned by the capacitor C_1, the electric coupling decreases with the decrease of resonant frequency [6]. Therefore, the magnetic dominated mixed coupling is utilized to make the overall coefficient $|k|$ increase with the decrease of resonant frequency [1], which can compensate bandwidth and meet the requirement of controllable fractional bandwidth. The input ports feed to the resonator through L_4. Meanwhile, C_3 forms electric coupling between source and load. The coupling topology of the filter is shown in Figure 1(b). **R1** and **R2** represent the two resonators. A $\pi/2$ or $-\pi/2$ phase shift can be obtained through the serious capacitive (electric) or inductive (magnetic) coupling, respectively. The phase shift of the resonator below resonance is $\pi/2$, and the phase shift of the resonator above the resonance is $-\pi/2$. Therefore, the phase shift from source to load through the coupling of resonators above the resonance is $-\pi/2 - \pi/2 - \pi/2 - \pi/2 - \pi/2 = -\pi/2$, and the phase shift from source to load through the coupling of resonators below the resonance is $-\pi/2 + \pi/2 - \pi/2 + \pi/2 - \pi/2 = -\pi/2$. Since the phase shift from source to load through the capacitive coupling is $\pi/2$, two transmission zeros can be generated beside the passband [23].
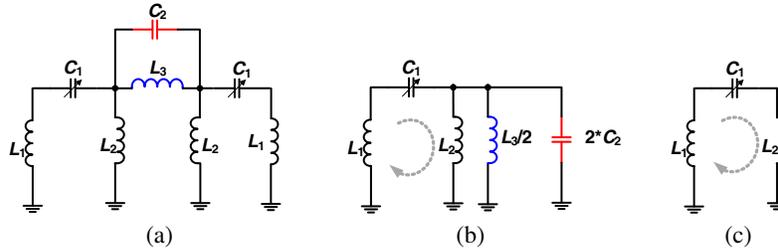

Figure 2. (a) Coupling model of the resonators, (b) odd-mode, (c) even-mode.

The coupling model of the resonators is shown in Figure 2(a), and the odd-mode and even-mode circuit model are shown in Figures 2(b) and (c), respectively. Based on the proposed circuit model, the odd- and even-mode impedances $Z_{r(e)}$ and $Z_{r(o)}$ can be expressed as:

$$Z_{r(e)} = j\omega L_1 + \frac{1}{j\omega C_1} + \frac{1}{j\omega 2C_2} \left\| j\omega \frac{1}{2} L_3 \right\| j\omega L_2 \tag{2}$$

$$Z_{r(o)} = j\omega L_1 + \frac{1}{j\omega C_1} + j\omega L_2 \tag{3}$$

When $Z_{r(e)} = 0$ and $Z_{r(o)} = 0$, the odd- and even-mode equivalent circuit work in resonant state, so the odd- and even-mode resonant frequencies can be obtained as

$$\omega_o = \sqrt{\frac{-b - \sqrt{b^2 - 4ac}}{2a}} \tag{4}$$

$$\omega_e = \sqrt{\frac{1}{C_1(L_1 + L_2)}} \tag{5}$$

where:

$$a = 2 \cdot L_1 L_2 L_3 C_1 C_2,$$
$$b = -[2C_2 L_2 L_3 + 2C_1 L_1 L_2 + C_1 L_1 L_3 + C_1 L_2 L_3],$$
$$c = 2L_2 + L_3.$$

The center frequency and mixed coupling coefficient can be written as:

$$\omega_c = \sqrt{\omega_e \cdot \omega_o} \tag{6}$$

$$|k| = \left| \frac{\omega_e^2 - \omega_o^2}{\omega_e^2 + \omega_o^2} \right| \tag{7}$$

Based on the above discussion, numerical calculation and simulation are used to verify the tunable capability. The simulation is carried out by using *Keysight* ADS (*Advanced Design System*). Capacitor C_1 is used as the tunable capacitor, and the element parameters of the proposed filter shown in Figure 1(a) are: $C_2 = 2\,\text{pF}$, $C_3 = 0.2685\,\text{pF}$, $L_1 = 45\,\text{nH}$, $L_2 = 30\,\text{nH}$, $L_3 = 100\,\text{nH}$, $L_4 = 92\,\text{nH}$. Figures 3(a) and (b) show the center frequency and coupling coefficient of the filter when capacitor C_1 changes from 6 pF to 24 pF, respectively. It is shown that the central frequency of the filter can be tuned by the capacitor C_1. The coupling coefficient can be increased when the central frequency decreases, and the slope of the coupling coefficient k versus the central frequency can be controlled by the coupling capacitor C_2 and coupling inductor L_3. Therefore, the bandwidth of the filter can be compensated.
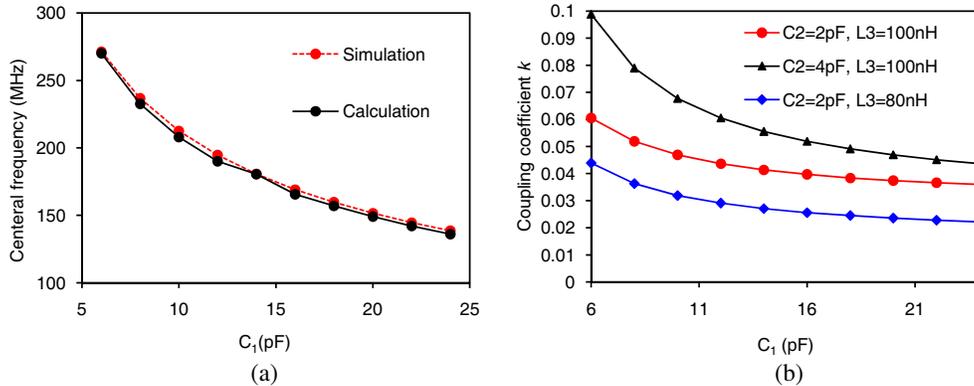

Figure 3. (a) Tunable center frequency, (b) coupling coefficient k.

3. FILTER SIMULATION

The filter is designed on a substrate of F4B-2 ($h = 0.8\,\text{mm}$, $\varepsilon_r = 2.65$, $\tan\theta = 0.001$), as shown in Figure 4. The lumped elements are connected by the 50 Ω microstrip lines, as shown in Figure 4. Skyworks SMV1212 varactor diode is chosen as a tunable capacitor. The cathode terminal of each varactor diode is biased through a 10-kΩ resistor. The capacitance for single varactor diode is 72.4 pF to

5.1pF at 0 V to 8 V reverse bias voltage. The inductors and capacitors are from *Coilcraft* and *American Technical Ceramics*. The passive microstrip structure is simulated by *SONNET*, and co-simulation is done by *Keysight* ADS (*Advanced Design System*). Figure 5(a) shows the simulated *S*-parameters of the tunable filter. It is shown that the passband of the filter can be tuned, and two transmission zeros are located at both sides of the passband to improve out-of-band rejection. The tuning range of center frequency is from 60 MHz to 300 MHz. Figure 5(b) shows the bandwidth versus tunable center
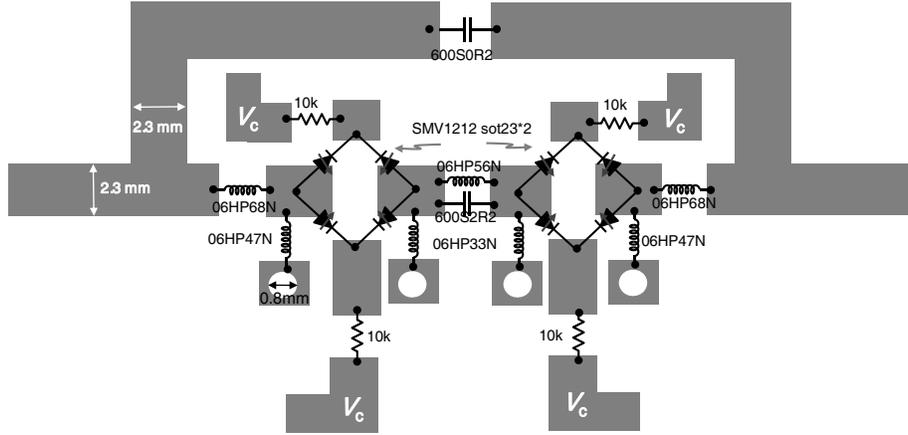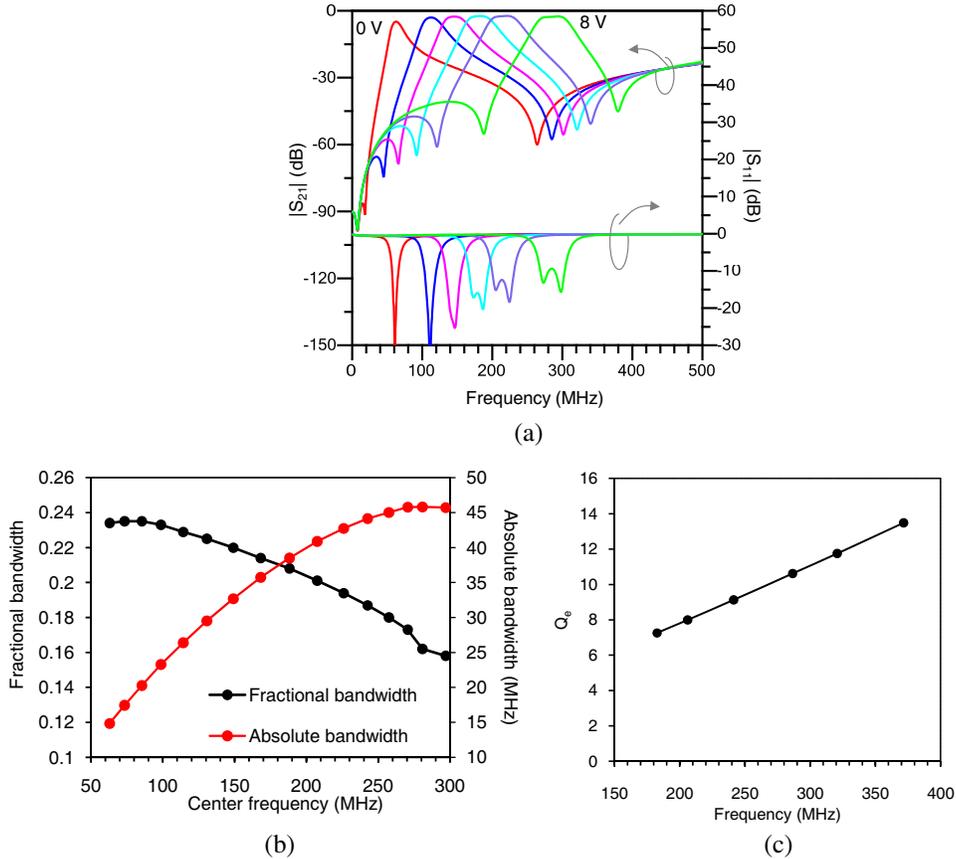

Figure 4. Layout of PCB.

(a)

(b)

(c)

Figure 5. (a) *S*-parameters of the filter, (b) bandwidth versus tunable center frequency, (c) Q_e versus the frequency.

frequency. It shows that the absolute bandwidth decreases from 46 MHz to 15 MHz, and the fractional bandwidth increases from 16% to 24% when center frequency decreases.

The external quality factor Q_e can be obtained by [21]:

$$Q_e = \frac{\omega_0 \cdot \tau_{s_{11(\omega_0)}}}{4} = \frac{2\pi f_0 \cdot \tau_{s_{11}(f_0)}}{4} \quad (8)$$

where $\tau_{s_{11}(f_0)}$ is the group delay of S_{11} at the resonant frequency f_0. To study the external quality factor of the filter, we extract half circuit of the filter, and the group delay $\tau_{s_{11}(f_0)}$ can be obtained via simulations through *ADS*. Figure 5(c) shows Q_e for the filter response. It is shown that Q_e increases while the frequency increases, which will be helpful for compensating the absolute bandwidth [22].

4. FABRICATION AND MEASUREMENT

The fabricated filter is shown in Figure 6, and the filter size is 44 mm × 30 mm. The frequency response of the tunable filter is measured by *Keysight* E5071C vector network analyzer. Figure 7 shows the measured and simulated S-parameters of the filter with the reverse bias voltage from 0 V to 8 V. Two transmission zeros are located at both sides of the passband. The rejection level is larger than 50 dB. Figure 8(a) shows the measured and simulated bandwidths versus center frequency. The simulation matches the measurement very well. It is shown that the absolute bandwidth is about 20 MHz to 55 MHz, the fractional bandwidth about 21% to 27%, and the fractional bandwidth compensated to a larger value at lower frequency. Figure 8(b) shows the measured insertion loss of the filter versus the center frequency. The center frequency tuning range is from 70 MHz to 270 MHz, and the insertion loss is less than 3 dB at all center frequencies. The comparison with related works is summarized in Table 1.
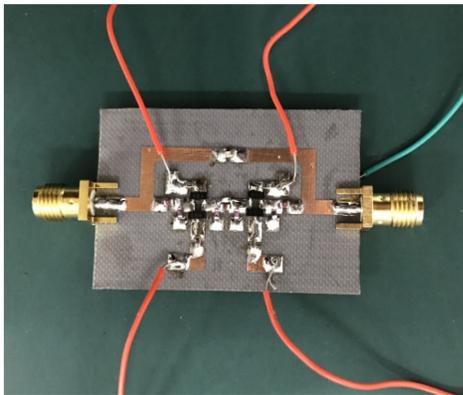

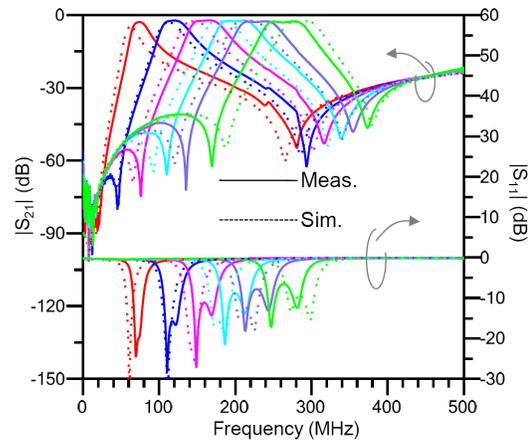**Figure 6.** Fabricated tunable bandpass filter.

Figure 7. Measured S-parameters.

Table 1. Comparison with related works.

Reference	Order	Tuning device	Tunable range	Bandwidth	Insertion loss	Size
[1]	$N = 2$	Varactor diode	2.30–2.72 GHz	8%–8.5%	3.2–1.5 dB	24 mm × 19 mm
[3]	$N = 3$	DTC	0.41–0.82 GHz	54 ± 12 MHz (9 ± 1%)*	6.5–4.9 dB	60 mm × 68.5 mm
[20]	$N = 3$	Varactor diode	1.75–2.25 GHz	9510 MHz+	7.2–3.2 dB	10.4 mm × 14.8 mm
This work	$N = 2$	Varactor diode	0.07–0.27 GHz	20–55 MHz (27%–21%)*	2.9–2.2 dB	44 mm × 30 mm

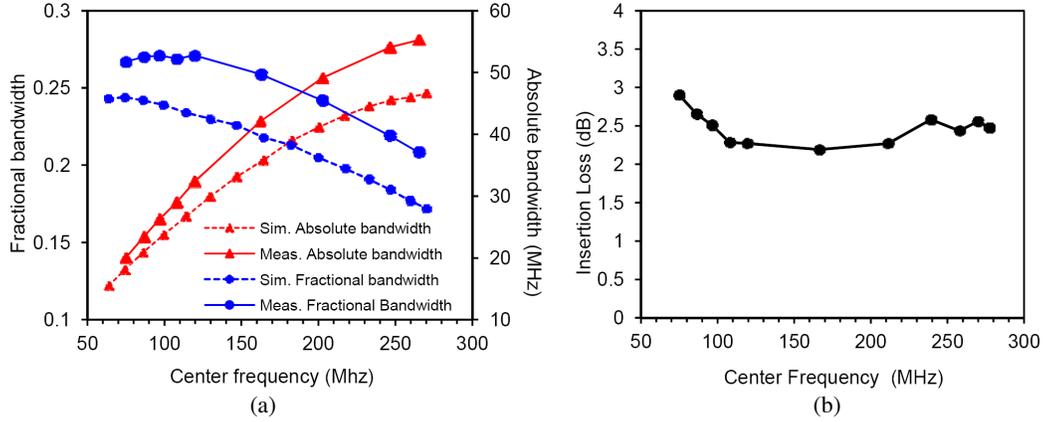* − 3 dB bandwidth, + − 1 dB bandwidth

Figure 8. (a) Measured and simulated bandwidth versus the center frequency, (b) measured insertion loss versus the center frequency.

5. CONCLUSION

This paper has presented a tunable LC filter with magnetic dominated mixed coupling and electrical coupling between source and load, and two transmission zeros are generated based on cross-coupling. The magnetic dominated mixed coupling can be used to compensate the fractional bandwidth of the filter at lower band. The measured results show that the filter has achieved a tuning range from 70 MHz to 270 MHz, and the fractional bandwidth is from 27% to 21%. The proposed filter can be used in the reconfigurable RF front-end of frequency agile communication systems and frequency synthesizers.

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