A High-Performance Continuously Tunable MEMS Bandpass Filter at 1 GHz

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Abstract—This paper reports a continuously tunable lumped bandpass filter implemented in a third-order coupled resonator configuration. The filter is fabricated on a Borosilicate glass substrate using a surface micromachining technology that offers high-Q tunable passive components. Continuous electrostatic tuning is achieved using three tunable capacitor banks, each consisting of one continuously tunable capacitor and three switched capacitors with pull-in voltage of less than 40 V. The center frequency of the filter is tuned from 1 GHz down to 600 MHz while maintaining a 3-dB bandwidth of 13%-14% and insertion loss of less than 4 dB. The maximum group delay is less than 10 ns across the entire tuning range. The temperature stability of the center frequency from -50° C to 50° C is better than 2%. The measured tuning speed of the filter is better than 80 μ s, and the IIP₃ is better than 20 dBm, which are in good agreement with simulations. The filter occupies a small size of less than 1.5 cm \times 1.1 cm. The implemented filter shows the highest performance amongst the fully integrated microelectromechanical systems filters operating at sub-gigahertz range.

Index Terms—Micromachining, passive filters, RF microelectromechanical systems (MEMS), tunable bandpass filters, tunable capacitors, UHF filters.

I. INTRODUCTION

T HERE IS an increasing demand for high-performance RF front-end modules for advanced ground mobile radios. With the introduction of joint tactical radios as the next-generation system in the U.S. military, ground mobile radios have to support various waveforms, including VHF and UHF bands, which will require reconfigurable RF front-ends [1]. The key challenge in developing reconfigurable RF front-end modules is to reduce the size and weight while supporting multiple communication standards [2]. As the key component of the RF front-end, the band-select filter needs to satisfy the above-mentioned requirements, namely, multiple frequency band coverage and good RF performance, all in a small form factor. This calls for a single fully integrated frequency-tunable bandpass filter.

In the VHF or UHF range, lumped LC filters offer the smallest size compared to other alternative implementations

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such as cavity filters [3], [4] and distributed filters [5]–[7]. Using conventional CMOS technology, the size of the filter could be significantly reduced. However, the quality factor (Q) of CMOS-based passives is low, making it hard to achieve a sufficiently low insertion loss for the bandpass filter unless Q enhancement techniques using active components are utilized [8]. The power-handling capability of CMOS varactors is also limited, further constraining their application in RF systems [9]. Microelectromechanical systems (MEMS) technology can lead to low insertion-loss tunable filters with high RF power-handling capability, meeting all the requirements of ground mobile radios.

There are a few reports on tunable bandpass filters having all integrated components centered at frequencies below 1 GHz [10], [11]. The reported filters are designed in the second-order coupled resonator configuration. Due to the low order of the filter, the shape factor and out-of-band rejection of these filters are limited. In addition, in these filter implementations, a largevalue fixed capacitor is placed in parallel with a smaller-value MEMS capacitor to obtain the required capacitance value, reducing the tuning range of the filters to less than 25%. In this paper, a continuously tunable MEMS bandpass filter using a third-order coupled resonator configuration is proposed. Using continuous tuning, the frequency of the filter can be tuned to select any desired frequency in the tunable frequency range or altered to account for fabrication inaccuracies. Continuous tuning is achieved using MEMS tunable capacitors that exhibit high Q's (exceeding 100), fast tuning speed (less than 80 μ s), wide capacitance tuning range (5:1), and good temperature stability [12], [13].

DC and RF characteristics of the tunable capacitor plays an important role in defining the characteristics of the filter, such as the tuning range, tuning speed, power handling, and power consumption. Among different actuation mechanisms, electrostatic tuning is most commonly used because of its low power consumption [14]. A problem with electrostatic tuning is the pull-in effect, which limits the travel range of the moving element. The tuning range of electrostatic capacitors can be improved by using a capacitor gap smaller than the actuation gap. Such capacitors, called dual-gap capacitors, can exhibit high tuning ratios exceeding 5:1 and are employed in this work to tune the frequency [12], [13].

Two-port capacitors are commonly used for matching or as the coupling elements in coupled resonator filters [15]. Using capacitive matching and coupling, it is hard to maintain a fixed bandwidth without tuning the value of the coupling capacitors. In this work, mutually coupled inductors and inductive

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 TABLE I

 TARGET SPECIFICATIONS OF THE PROPOSED FILTER

| Filter Characteristics | Target |
|--------------------------------------|---------------------------------|
| Center frequency range | 600 MHz ~ 1000 MHz |
| Insertion loss | $< 4 \text{ dB}$ at ω_c |
| 3dB-Bandwidth (BW3dB) | $12 \sim 16 \%$ of ω_c |
| Shape factor (BW30dB / BW3dB) | < 4 |
| Passband ripple | < 0.5 dB |
| Out-of-band rejection | > 40 dB |
| Group delay | < 10 ns |
| Impedance | 50 Ω |
| IIP ₃ | > 20 dBm |
| Tuning voltage | < 50 V |
| Tuning speed | < 100 µs |
| Temperature variation (ω_c) | < 5 % |
| Dimensions | $< 1.5 \times 1.5 \text{ cm}^2$ |

matching are utilized to provide a wider band matching and avoid complicated tuning control. Using broadband inductive matching and wide-range capacitive tuning in the resonator, a tunable filter is demonstrated with insertion loss of less than 4 dB and tuning range of 40%. To our best knowledge, this is the first single-chip implementation of a third-order tunable filter at sub-gigahertz range, which addresses diverse aspects of filter performance, such as insertion loss, tuning range, shape factor, linearity, tuning speed, and temperature stability.

This paper is organized as follows. First, the design and tuning configuration of the filter are discussed. Next, the design of each passive component and their 3-D electromagnetic simulation results are shown. The linearity analysis for the 1-dB compression point (P1 dB) and the third-order intermodulation intercept point (IIP₃) are also described. Finally, the measurement and characterization results of the fabricated filter are presented.

II. TUNABLE BANDPASS FILTER DESIGN

The target specifications of the pre-select tunable filter are listed in Table I. The filter is aimed to achieve frequency coverage from 600 MHz to 1 GHz with a 3-dB percentage bandwidth of 13%–14%. The insertion loss of the filter is targeted to be less than 4 dB to achieve a small noise figure for the entire radio. To obtain a shape factor (30-dB bandwidth ($BW_{30 dB}$)) to 3-dB bandwidth ($BW_{3 dB}$)) of less than 4, the order of the filter needs to be at least 3 [16]. A third-order Chebyshev filter with a 0.5-dB passband ripple is selected to achieve the desired shape factor. Using this configuration, the group delay is less than 10 ns, meeting all the specifications listed in Table I.

The filter design procedure is as follows. First, the low-pass prototype in Chebychev configuration is designed. The low-pass prototype values (g_1, g_2, g_3, g_4) of the third-order Chebyshev filter used here are 1.5963, 1.0967, 1.5963, and 1.0000, respectively. The low-pass filter (LPF) is then converted to a coupled resonator bandpass filter configuration using admittance inverters, as shown in Fig. 1(a). With this configuration, values of the lumped components are easily realizable using MEMS surface micromachining technologies [10]. The parameters of the



Fig. 1. Schematic views showing the design procedure of the third-order bandpass filter in this work. (a) Generalized bandpass filter using admittance inverter. (b) Conversion into inductive coupling. (c) Arrangement of inductance for mutual-inductive coupling. (d) Final schematic view of the filter. (e) Detail composition of the tunable bank included in each resonator.

admittance inverters are derived using the following equations [17]:

$$J_{01} = J_{34} = \frac{1}{2\pi f_c} \cdot \frac{1}{\sqrt{g_0 g_1 G_A L_{R1}}} \tag{1}$$

$$J_{12} = J_{23} = \frac{1}{2\pi f_c} \cdot \frac{BW_{3 dB}}{f_c} \cdot \frac{1}{\sqrt{g_1 g_2 L_{R1} L_{R2}}}$$
(2)

where f_c is the center frequency of the bandpass filter at initial state, G_A is the input impedance, and L_{R1} and L_{R2} are inductor values in each *LC* tank. To ease the characterization and tuning scheme, the initial value of all three capacitors (C_{R1}, C_{R2}, C_{R3}) are set to be the same. An initial value of 2.3 pF is chosen for the tunable capacitors, considering that the inductance value needs to be in the range of 1–15 nH to provide a high *Q* of more than 40. Using these values for the capacitors, the required inductance value for L_{R1}, L_{R2} , and L_{R3} is 11 nH and the unloaded Q_0 of each resonator would be about 40. As shown in Fig. 1(b), the admittance inverter is implemented using inductive coupling with $L_{12} = L_{23} = 1/(2\pi f_c J_{12})$. Values of L_{01} and L_{34} are not set at this step as they also depend on the matching condition.

The equivalent inductor of the second resonator $(-L_{12}//L_{R2}//-L_{23})$ in Fig. 1(b) is split into two inductors $[L_2$'s in Fig. 1(c)]. In Fig. 1(c), L_1 in the first and third resonators can be approximated as $L_{R1}//(-L_{12})$. To achieve more feasible inductance values, the inductive π -network of Fig. 1(c) is converted into mutually coupled inductors, as shown in Fig. 1(d). The matching inductance (L_M) is derived considering g_1 and g_4 and the loaded Q_n of the resonator. The mutual inductance (M) and resonator inductances (L_{C1}', L_{C2}') are derived from L_1 , L_2 , and L_K using well-known equations in [16].

To obtain the effective impedance of g_1 and g_4 at input and output nodes and impedance matching to 50 Ω , impedance transformation using two inductors (L_{C1}, L_M) in Fig. 1(d) is obtained using the following expressions:

$$L_{C1} = \frac{L_M L'_{C1} R_S}{R_S \left(L_M - L'_{C1} \right) + R_T L'_{C1}} \quad L_M = \frac{\sqrt{R_S R_T}}{2\pi f_c} \quad (3)$$

$$L_{C2} \approx L_{C2}^{'} \quad R_{S} = \frac{2\pi f_{c} \cdot L_{R1} Q_{0} \cdot L_{R1} Q_{n}}{L_{R1} Q_{0} - L_{R1} Q_{n}}$$
(4)

where R_T is the target input impedance (50 Ω) and R_S is the input impedance looking into the resonator.

To achieve frequency tuning, a tunable capacitor bank, which consists of one fixed capacitor (MIM capacitor), one continuously tunable capacitor, and three capacitive switches [see Fig. 1(e)], is employed in each resonator section. The tuning control mechanism is as follows. First the continuously tunable capacitor is tuned. When this capacitor reaches its maximum value, a switch will be turned ON and the value of the continuously tunable capacitor will be reset to set the frequency as required. To further tune the center frequency, again the continuously tunable capacitor will be tuned to finally reach its maximum value. At this state, another switch will be turned on. Therefore, to achieve continuous tuning, only one continuously tunable capacitor is required. Other capacitors are switched to ease the tuning control. The initial capacitance value of all tunable capacitors is set to 200 fF with tuning bias of less than 40 V, and tuning speed of less than 80 μ s including the stabilization time. The capacitance value and the corresponding filter frequency range at each tuned state are listed in Table II. The mechanical design of the tunable capacitors is reported elsewhere [12]. The values of the passive components in Fig. 1 are listed in Table III. In Section III, the simulation results and 3-D electromagnetic layout of the filter are presented.

III. 3-D ELECTROMAGNETIC SIMULATION

Filters are designed and fabricated using a multiple-metal surface micromachining process technique [18]. This technology offers three metal layers (0.5 μ m Au/4 μ m Au/40 μ m Cu), one dielectric layer [aluminum oxide (Al₂O₃)] and two sacrificial layers (PMMA/Shipley 1813 photoresist). Using this process, each tunable or fixed passive component can be optimized for

TABLE II TUNING CONFIGURATION

| Tuning Controls | Capacitance (C_B) | Center Frequency (f_c) |
|--|---------------------|--------------------------|
| $V_{Bl} = 0.40 \text{ V}$ | 2.3-3.5 pF | 811-1000 MHz |
| $V_{B1} = 0.40 \text{ V}$ $V_{B2} = 40 \text{ V}$ | 3.4-4.6 pF | 707-824 MHz |
| $V_{B1} = 0.40 \text{ V}$ V_{B2} , $V_{B3} = 40 \text{ V}$ | 4.5-5.7 pF | 635-715 MHz |
| $V_{B1} = 0.40 \text{ V}$ $V_{B2}, V_{B3}, V_{B3} = 40 \text{ V}$ | 5.6-6.8 pF | 582-640 MHz |

TABLE III DESIGN VALUES OF THE MEMS FILTER

| Symbol | Description | Value | |
|-----------------------------------|---------------------------------|--------------------------------------|--|
| J ₀₁ , J ₃₄ | Admittance inverter parameters | 8.49e-6 | |
| J_{12}, J_{23} | Admittance inverter parameters | 1.75e-3 | |
| L_I | Inductively coupled inductor 1 | 12.5 nH | |
| L_2 | Inductively coupled inductor 2 | 29.0 nH | |
| L_K | Coupling inductor | 90.9 nH | |
| L_M | Matching inductor | 32.0 nH | |
| L_{CI} | Mutually coupled inductor 1 | 16.4 nH | |
| L_{C2} | Mutually coupled inductor 2 | 22.6 nH | |
| M | Mutual inductance | 3.2 nH | |
| R_T | Target Input / Output Impedance | 50 | |
| C_B | Tunable capacitor bank (TCB) | $2.3~\text{pF}\sim6.8~\text{pF}$ | |
| C_{B-F} | Fixed capacitor in TCB | 1.5 pF | |
| C_{B-C} | Continuous capacitor in TCB | $0.2 \text{ pF} \sim 1.4 \text{ pF}$ | |
| C_{B-S} | Switched capacitor in TCB | $0.2 \text{ pF} \sim 1.3 \text{ pF}$ | |



Fig. 2. Layout of the MEMS tunable filter.

the highest performance with a selection of metals, dielectrics, and sacrificial layers. The performance of individual passive components, as well as the tunable filter, is simulated using the ANSOFT HFSS 3-D electromagnetic simulation tool [19]. The material properties, such as conductivity, dielectric constant, and loss tangent, are characterized and the extracted values from measurements are used in the simulations.

The integrated filter layout is shown in Fig. 2. The detailed design of the matching inductor and coupled inductor is discussed in [18]. The lengths of the RF interconnecting lines are minimized and the ground connections are optimally placed to



Fig. 3. Simulation results of the MEMS filter. (a) S-parameters. (b) Group delay.

 TABLE IV

 Simulated Filter Specifications Using HFSS 3-D EM Tool

| Tuned Capacitors | Insertion Loss | 3dB Bandwidth | Shape Factor | Group Delay |
|---------------------|----------------------|---------------------|-----------------|----------------|
| No tuning | 2.3 dB @ 1005 MHz | 139 MHz (13.8 %) | 3.1 | < 9 ns |
| Tuning_1 | 3.2 dB @ 825 MHz | 120 MHz (14.5 %) | 3.5 | < 9 ns |
| Tuning_2 | 3.1 dB @ 734 MHz | 102 MHz (13.9 %) | 3.8 | < 8 ns |
| Tuning_3 | 3.4 dB @ 650 MHz | 90 MHz (13.8 %) | 4.1 | < 9 ns |
| Tuning_4 | 3.8 dB @ 580 MHz | 92 MHz (15.9 %) | 4.1 | < 10 ns |

reduce loss and parasitics. The bias lines of the three tunable capacitors in each resonator tank are connected together. One bias line also controls all corresponding capacitive switches. Therefore, only one analog bias line and three digital (0 V/40 V) bias lines are needed to tune the filter (instead of 12 control lines).

The HFSS simulated insertion loss, return loss, and group delay of the integrated filter at each tuned state are shown in Fig. 3. In the simulations, nonideal conditions such as reduced capacitance tuning range and additional loss from the substrate are reflected from the characterization results of the tunable capacitor banks and the inductors [18]. The group delay in Fig. 3(b) is derived using the formula suggested in [20]. The simulated performance of the filter is summarized in Table IV.

 TABLE V

 Parameters of the Varactor Used in Nonlinear Simulations

| Symbol | Description | Value |
|---------------|---------------------------------------|--------|
| $C_{S\theta}$ | Initial capacitance in sensing node | 200 fF |
| $C_{A\theta}$ | Initial capacitance in actuation node | 220 fF |
| gso | Initial air gap in sensing node | 0.5 μm |
| g_{A0} | Initial air gap in actuation node | 2.0 µm |
| f_m | Mechanical resonant frequency | 22 kHz |
| k | Spring constant | 75 N/m |
| Q_M | Mechanical quality factor | 2 |



Fig. 4. Schematic view of the nonlinear electromechanical model for the: (a) varactor and (b) entire filter.

IV. LINEARITY ANALYSIS

The Agilent ADS simulation tool is used to analyze the nonlinear performance of the filters [21]. To estimate the P1 dB and IIP₃ values, the nonlinearity of the varactor and capacitive switches are taken into account using nonlinear electromechanical (EM) models [22]. The simulation parameters such as initial capacitance, air gap, and mechanical properties of the varactor are summarized in Table V. All values are carefully extracted from HFSS electromagnetic simulations and modal/displacement analysis in ANSYS [23].

Since the integrated varactor has separate electrodes for actuation and capacitance sensing, the total force can be approximated as the sum of actuation force from dc bias applied to the actuation electrode and the RF self-actuation force from the capacitance sensing electrode. Since the varactor has a dual-gap configuration, the equations in [22] are modified to take into



Fig. 5. IIP_3 value extracted from the nonlinear electromechanical model at frequency offset of 20 kHz: (a) without dc bias and (b) with 25 V of dc bias.



Fig. 6. IIP_3 value extracted from the nonlinear electromechanical model at different frequency offsets with and without dc bias. Tuning characteristics of the tunable capacitor is shown in the inset.

account both the sense and actuation gaps. The following equations are applied to the EM model of the varactor in Fig. 4(a) in order to calculate the *n*th-iterated total force applied to the top membrane and the sense capacitance, respectively,

$$F_{n} = \left(C_{S0} \cdot \frac{g_{S0}}{g_{S0} - \Delta g_{n-1}}\right) \cdot \left(\frac{V_{\rm RF \ n-1}^{2}}{2(g_{S0} - \Delta g_{n-1})}\right) + \left(C_{A0} \cdot \frac{g_{A0}}{g_{A0} - \Delta g_{n-1}}\right) \cdot \left(\frac{V_{\rm DC}^{2}}{2(g_{A0} - \Delta g_{n-1})}\right)$$
(5)

$$C_{Sn} = C_{S0} \cdot \frac{g_{S0}}{g_{S0} - \Delta g_n} = C_{S0} \cdot \frac{g_{S0}}{g_{S0} - (F_n/k)}$$
(6)

where Δg_{n-1} , $V_{\text{RF }n-1}$, and V_{DC} are the change of air gap and the equivalent RF bias from the (n-1)th iteration, and the dc tuning bias, respectively.

As shown in Fig. 4(a), a four-port symbolically defined device (SDD4P) is utilized to implement (5), where the value of Port 2 output, F_n , is derived from the other three port values, Δg_{n-1} , $V_{\text{RF}n-1}$, and V_{DC} . Likewise, an SDD2P on the right side is utilized to calculate the value of Port 2 output, C_{Sn} , from the Port 1 input value, Δg_n , using (6). The LPF polynomial reflects the mechanical response of the MEMS capacitor with parameters f_m , k, and Q_m . The schematic of the filter configuration taking into account the nonlinear models of the varactor and switched capacitors is shown in Fig. 4(b). The initial air gap at the sensing node is set as 1.5 μ m.

Fig. 5 shows the harmonic simulation result at different input power levels. The frequency difference of the two input tones is



Fig. 7. P1 dB value extracted from the nonlinear electromechanical model: (a) without dc bias and (b) with 25 V of dc bias.





Fig. 8. (a) Scanning electron microscope (SEM) view. (b) Photographic view of a fabricated filter.

20 kHz. The extracted IIP₃ value is 30 dBm when no dc bias is applied to the varactors/switches. With the application of 25-V dc bias, the IIP₃ is reduced to 20 dBm. At this bias point, the capacitance value of the varactors is most sensitive to the applied RF power as the C-V curve has the sharpest slope at this point (see inset of Fig. 6). Therefore, 20 dBm is a pessimistic estimation of IIP₃ when a dc bias is applied. The extracted IIP₃ at different input frequency offsets is shown in Fig. 6. At both



Fig. 9. Measured filter response at each tuned stage. (a) Insertion loss. (b) Return loss.

initial and tuned states, the IIP₃ value is better for larger frequency offsets. This is due to the low-pass filtering behavior of the MEMS capacitors, which is taken into account in the nonlinear EM model of Fig. 4(a) by considering a cutoff frequency of 22 kHz (the mechanical resonant frequency (f_m) of the tunable capacitor). A similar trend is expected in the measured IIP₃.

The P1 dB is also simulated using the same EM models. Fig. 7(a) and (b) shows the simulated P1 dB when 0 and 25 V of actuation bias are applied, respectively. The extracted P1 dB is 22 dBm when no dc bias is applied to the varactors/switches, whereas it is reduced to 15 dBm when 25 V of dc bias is applied. The power-handling capability of these filters is thus limited to P1 dB and not to the IIP₃ value.

V. MEASUREMENT RESULT

Insertion loss and return loss are measured using Cascade Microtech ground–signal–ground (GSG) ACP probes and an N5214A Agilent PNA-X network analyzer. The dc bias is applied to each bias line using Microtech dc probes. The images of the fabricated device are shown in Fig. 8. The footprint of the entire filter is around 1.5 cm \times 1.0 cm, which is much smaller than other UHF filters using microstrip lines and SMT passive components [5]–[7].

 TABLE VI

 Measured Filter Specifications at Each Tuned Stage

| Tuned Capacitors | Insertion Loss | 3dB Bandwidth | Shape Factor | Group Delay |
|---------------------|----------------------|---------------------|-----------------|----------------|
| No tuning | 3.0 dB @ 1011 MHz | 135 MHz (13.4 %) | 3.2 | < 7 ns |
| Tuning_1 | 3.3 dB @ 833 MHz | 113 MHz (13.6 %) | 4.5 | < 8 ns |
| Tuning_2 | 3.4 dB @ 735 MHz | 99 MHz (13.5 %) | 4.6 | < 9 ns |
| Tuning_3 | 3.6 dB @ 650 MHz | 87 MHz (13.4%) | 4.7 | < 9 ns |
| Tuning_4 | 3.6 dB @ 602 MHz | 81 MHz (13.5 %) | 4.7 | < 8 ns |



Fig. 10. Measured filter response at different temperatures. (a) At initial state. (b) When a dc bias of 25 V is applied to C_{B-C} .

A. Insertion Loss and Return Loss

Fig. 9 shows the measured insertion loss and return loss at each tuned state when dc bias of 0 to 40 V is applied to the varactors and switched capacitors. The center frequency is tuned from an initial value of 1011–602 MHz by applying a maximum of 40 V to the capacitors. Across the entire tuning range, the insertion loss is less than 4 dB and the return loss is greater than 15 dB. The measurement results are summarized in Table VI. The 3-dB bandwidth shows good agreement with the electromagnetic simulation. However, the measured shape factor at most tuned states is above 4. This is caused by an unwanted



Fig. 11. Measured group delay at each tuned stage.



Fig. 12. Measured input power versus output power. (a) Without de bias. (b) With 25 V of de bias.



Fig. 13. Measured power spectrum when 25 V of dc bias is applied to the continuously tunable capacitor. (a) Output power spectrum with frequency offset of 20 kHz and input power of -4 dBm. (b) Extracted IIP₃ with frequency offset of 20 kHz. (c) Output power spectrum with frequency offset of 500 kHz and input power of -3 dBm. (d) Extracted IIP₃ with frequency offset of 500 kHz.

resonance located at the lower side of the passbands. This resonance is presumably due to the coupling between the inductors



Fig. 14. (a) Setup used for tuning-speed measurements. The measured tuning speed when: (b) 40 V and (c) 25 V is applied to continuously tunable capacitor, respectively.

and the ground plane on the backside of the wafer and can be reduced by increasing the thickness of the substrate (i.e., 500- μ m thick).

Temperature stability of the filter is tested using a Microtech KV-230 cryogenic station and GGB RF probes. Short-open-load-thru (SOLT) calibration is done at each temperature. Fig. 10(a) shows the filter response from -50 °C to 50 °C when no dc bias is applied. The center frequency of the filter is shifted from 1035 to 1016 MHz, showing a variation of less than 2%. The frequency response of the filter when a dc bias is applied to one of the capacitors is shown in Fig. 10(b). Upon temperature change, the center frequency is shifted by 1.5% from 955 to 941 MHz, which is considered small for a MEMS device [24], [25]. The temperature stability of the filter is better than the temperature shift of the varactor itself. The capacitance variation of individual varactors is less than 7% over the same temperature range [12]. Since the varactor is placed in parallel with a more temperature stable fixed capacitor, the

| | Brown '00 [5] | Manuel '05 [6] | Rais-Zadeh '09 [10] | Xiu '10 [7] | This work |
|---------------------------|---|--|---|--|--|
| $\mathbf{f}_{\mathbf{c}}$ | 700-1330 MHz | 470-862 MHz | 786-836 MHz | 680-1000 MHz | 602-1011 MHz |
| Filter configuration | 5 th order discretely tunable combline filter | 3 rd order continuously tunable combline filter | 2 nd order continuously tunable lumped filter | 2 nd order continuously tunable combline filter | 3 rd order continuously tunable lumped filter |
| Insertion loss | 2.0-6.0 dB | 3.0-6.0 dB | 4.0-6.0 dB | 1.1-1.5 dB | 3.0-3.6 dB |
| BW_{3dB} | 8-22 % of f _c | 1-3 % of f _c | 5-6 % of f _c | 8-12 % of f _c | 13-14 % of f _c |
| BW300B/BW30B | 2.0-3.0 | 3.0-5.0 | > 7.0 | 5.0-6.0 | 3.2-4.7 |
| Tuning Speed | N/A | < 1 µs | < 1 ms | < 1 µs | 40-80 µs |
| IIP ₃ | 18-24 dBm | N/A | N/A | 13 dBm | 20-30 dBm |
| Technology | PCB + MEMS switch | PCB + CMOS varactor | MEMS (single chip) | PCB + CMOS varactor | MEMS (single chip) |
| Size | $< 31.0 \times 40.0 \text{ mm}^2$ | $< 50.0 \times 65.0 \text{ mm}^2$ | $< 5.0 \times 6.0 \text{ mm}^2$ | $< 30.0 \times 30.0 \text{ mm}^2$ | $< 11.0 \times 15.0 \text{ mm}^2$ |

 TABLE VII

 Comparison Between Tunable Front-End Filters in the UHF Range

temperature stability of the filter is improved at the initial state, as well as the tuned state of the varactors.

B. Group Delay

The group delay of the filter at each tuned state is extracted from the measured S-parameters. As shown in Fig. 11, the group delay is less than 10 ns, meeting the design requirement.

C. Linearity

The P1 dB measurements are carried out using an N5214A Agilent PNA-X network analyzer in the high-power narrowband detection mode, which supports up to 20 dBm of input source power. Fig. 12(a) shows the measured input power versus output power with zero dc bias. As expected from the simulations shown in Fig. 7(a), the filter shows no significant degradation up to 20 dBm of input power. The measured P1 dB with 25 V of dc bias is around 13 dBm [see Fig. 12(b)], which is similar to the simulated value in Fig. 7(b).

The IIP₃ measurements are carried out in the two-tone source power mode. As the IIP₃ of the PNA-X receiver itself is around 30 dBm, IIP₃ values higher than 30 dBm cannot be accurately measured using this system. Therefore, the IIP₃ of the filter without dc bias could not be measured; it was only verified that the IIP₃ value is above 30 dBm at both 20 and 500 kHz of frequency offset.

Fig. 13(a)–(d) shows the linearity measurements when 25 V of dc bias is applied to the varactor. When dc bias is applied, IIP₃ degrades as the smaller capacitance gap becomes more sensitive to the RF signal power. As shown, the IIP₃ value is at the lowest at 20 kHz of frequency offset, i.e., the mechanical resonance frequency of the varactor membrane. The extracted IIP₃ at 20-kHz offset is about 20 dBm [see Fig. 13(b)], which is close to the simulated value of 22 dBm shown in Fig. 5(b). The IIP₃ with an applied voltage of 25 V at 500-kHz frequency offset is also above 30 dBm [see Fig. 13(d)].

D. Tuning Speed

The tuning speed of the filter is measured using the setup shown in Fig. 14(a). 10 dBm of a single-tone RF signal at the corresponding center frequency for the dc tuning bias is applied using the network analyzer. The RF signal at the output port is converted into dc voltage using a KRYTAR 201A power detector. The RF signal before applying the bias is zero; after application of bias, the filter tunes to the frequency of the input RF signal and a nonzero power is detected using the power detector. The tuning bias and power detector outputs are monitored with an Agilent MSO7104A oscilloscope to extract the tuning speed.

Fig. 14(b) shows transition of detected power level when a pull-in bias of 40 V is applied to the tunable capacitors. The measured transition time with this bias condition is better than 50 μ s, which is the maximum tuning speed of the filter. As shown in Fig. 14(c), with 25 V of dc bias, the transition time is around 80 μ s. At this bias, the membrane does not completely touch down and the stabilization time is longer.

E. Comparison

There has been extensive work on tunable front-end filters in the UHF band, implemented with several different configurations and integration methods. For successful adoption in the RF front-end system, a filter should satisfy wide frequency band coverage, low insertion loss, and high power-handling capability, all in a small size and low cost. Filter implementations using integration of passives with varactor diodes or employing MEMS capacitors on a printed circuit board (PCB) can satisfy only a few of these requirements [5]–[7] (Table VII). Integration of separately packaged passives can not only result in additional insertion loss, but also derive increased fabrication cost and size. The form factor of a reported filter fabricated using a singlechip MEMS technology [10] is much smaller than that implemented using the PCB technology. However, its tuning performance was limited due to the low order of the filter, and limited tuning range of the tunable capacitors. In this work, a significantly better performance is achieved using 12 wide-tuning range MEMS capacitors and a higher order filter in a Chebyshev configuration. Compared to the reported work, the filter presented in this paper is the highest performance single-chip filter in the sub-gigahertz frequency band.

VI. CONCLUSION

Design and measurements of a continuously tunable MEMS bandpass was reported in this paper. Insertion loss of the filter at all tuned states (from 600 MHz to 1 GHz) is less than 4 dB, while the 3-dB bandwidth is maintained within 13%–14 %. The shape factor of the filter is above 4 (less than 5) and can be improved by optimizing the layout of the inductors and reducing the substrate coupling. The measured shift in center frequency of the filter is less than 1.5% across 100 °C of temperature change and the tuning speed is better than 80 μ s. The worst case IIP₃ is around 20 dBm. However, considering the lower value of P1 dB, the practical range of power is limited to about 13 dBm.

Improvements in the design of tunable capacitors are required to achieve better power-handling capability. Future work will focus on such design optimizations, as well as characterization of other filter specifications such as phase noise and sensitivity to vibration. The presented filter technology could be extended to other applications in the UHF range such as TV tuners, which requires smaller channel selection bandwidth.

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