A UHF 4TH-ORDER BAND-PASS FILTER BASED ON CONTOUR-MODE PZT-ON-SILICON RESONATORS

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ABSTRACT

A UHF 4th-order band-pass filter (BPF) based on the subtraction of two 2nd-order contour-mode resonators with slightly different resonance frequencies is presented. The resonators consists of a 1 µm pulsed-laser deposited (PLD) lead zirconate titanate (PZT) thin-film on top of a 3 µm silicon (PZT-on-Si). The resonators are actuated in-phase and their outputs are subtracted. Utilizing this technique, the outputs of the resonators are added up constructively while the feed-through signals are eliminated. The BPF presented a bandwidth of approximately 28.6 MHz and more than 30 dB stopband rejection at around 700 MHz.

INTRODUCTION

RF-MEMS filters are providing new opportunities for the next generation of wireless communication systems enabling low power consumption and high level of integration. FBAR [1] is most successful among available RF-MEMS filters and is already commercialized. The current demand requires the integrability of filters at different resonance frequencies compacted in the same die fabricated in a single process. Due to the lateral dependency of Lamb wave resonators as well as their high quality factor performance, they are the promising solution for future single-chip multi-frequency BPFs. Available Lamb wave BPF are based on mechanical and/or electrical coupling techniques [2, 3, 4] and they are still in the research stage.

In this paper, we propose a new technique of differentially readout of two in-phase actuated resonators, which is not based on the traditional methods such as mechanical and/or electrical couplings. The presented method has been presented recently operating at low frequencies [5], however this method requires to be developed further at UHF and SHF bands. In this paper, the technique is developed further at UHF band.

This method is effective for improving the stopband rejection by canceling the feed-through signal, which is more crucial in high-dielectric materials such as PZT especially at UHF and SHF [6].

THEORY AND MODELLING

Fig. 1 shows a simplified equivalent circuit of the filter concept. The transfer function of each resonator can be expressed as:

$$H_i(s) = \frac{V_{\text{out},\,i}}{V_{\text{in}}} = \frac{R_{\text{s}}}{L_{\text{m},\,i}} \times \frac{s}{s^2 + \omega_{\text{-3dB},\,i}\cdot s + \omega^2_{\text{res},\,i}},\ i = 1,2 \quad (1)$$

where $\omega_{\text{-3dB},\,i}$ and $\omega_{\text{res},\,i}$ are respectively the bandwidth $(2R_{\text{s}} + R_{\text{m},\,i})\,/L_{\text{m}.i}$ and the resonance frequency $1/\sqrt{L_{\text{m},\,i}C_{\text{m},\,i}}$ of the ith path.

Assuming $L_{\text{m},\,1} = L_{\text{m},\,2} = L_{\text{m}}$ and $R_{\text{m},\,1} = R_{\text{m},\,2} = R_{\text{m}}$, the total transfer function of the filter after subtraction can be expressed as

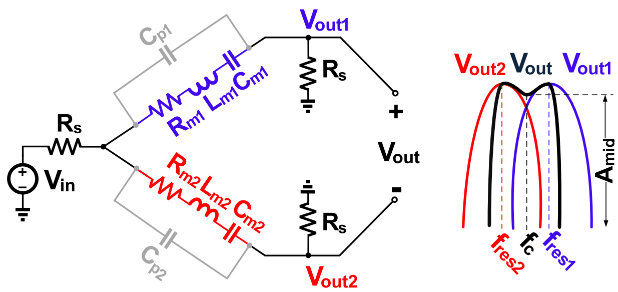

Figure 1: Equivalent circuit of 4th-order filter with differential readout of two in-phase actuated resonators.

$$H(s) = \frac{V_{\text{out1}} - V_{\text{out2}}}{V_{\text{in}}} = \frac{R_{\text{s}}}{L_{\text{m}}} \times \frac{2\omega_{\text{c}}\cdot\Delta\omega_{\text{res}}\cdot s}{\left(s^2 + \omega_{\text{-3dB}}\cdot s + \omega^2_{\text{res},\,1}\right)\left(s^2 + \omega_{\text{-3dB}}\cdot s + \omega^2_{\text{res},\,2}\right)}, \quad (2)$$

where $\omega_{\text{c}} = (\omega_{\text{res},\,1} + \omega_{\text{res},\,2})\,/2$ and $\Delta\omega_{\text{res}} = \omega_{\text{res},\,1} - \omega_{\text{res},\,2}$.

We assume two case studies for representation, where two resonators at resonance frequencies of f_{res1}=690 MHz and f_{res2}=710 MHz and $L_{\text{m,1}}$=$L_{\text{m,2}}$=6 µH are considered. For the first case, a bandwidth of $\omega_{\text{-3dB},\,i}$=8 MHz ($\omega_{\text{-3dB},\,i}<\Delta\omega_{\text{res}}$), and for the second case, a bandwidth of $\omega_{\text{-3dB},\,i}$=30 MHz ($\omega_{\text{-3dB},\,i}>\Delta\omega_{\text{res}}$) are considered. For the first case ($\omega_{\text{-3dB},\,i}<\Delta\omega_{\text{res}}$), the transmission of each resonator is presented in Fig. 2. By subtracting the output of two 2nd-order resonators, a 4th-order BPF centered at f_{res1} =700 MHz (Fig. 2) is obtained. As seen, the filter response shows the same insertion loss as each resonator with a ripple of 2.23 dB. For the second case ($\omega_{\text{-3dB},\,i}>\Delta\omega_{\text{res}}$), the transmission of each resonator as well as the subtracted filter response is presented in Fig. 3. In this case, the filter has zero ripple, however the insertion loss is lower than that of each resonator. Considering $\omega_{\text{-3dB},\,i}$=$(2R_{\text{s}} + R_{\text{m},\,i})\,/L_{\text{m}.i}$, in both cases, $L_{\text{m,1}}$ has been kept constant. Therefore by increasing the bandwidth of each resonator ($\omega_{\text{-3dB},\,i}$) in the second case, it is assumed that the motional impedance of the resonators is increased which will lead to increasing the insertion loss of the second case compared to the first case. For designing a BPF with zero ripple performance, there should be an optimum condition between the cases of $\omega_{\text{-3dB},\,i}>\Delta\omega_{\text{res}}$ and $\omega_{\text{-3dB},\,i}<\Delta\omega_{\text{res}}$. To investigate further on this issue, we will have a closer look at the operation of the filter at the center frequency of the filter, $\omega_{\text{c}} = (\omega_{\text{res},\,1} + \omega_{\text{res},\,2})\,/2$.

Assuming ω around the center frequency ($\omega = \omega_{\text{c}} + \delta\omega$) with a small frequency variation ($\delta\omega$), $\delta\omega << \omega_{\text{c}}$, allows (1) to be simplified as:

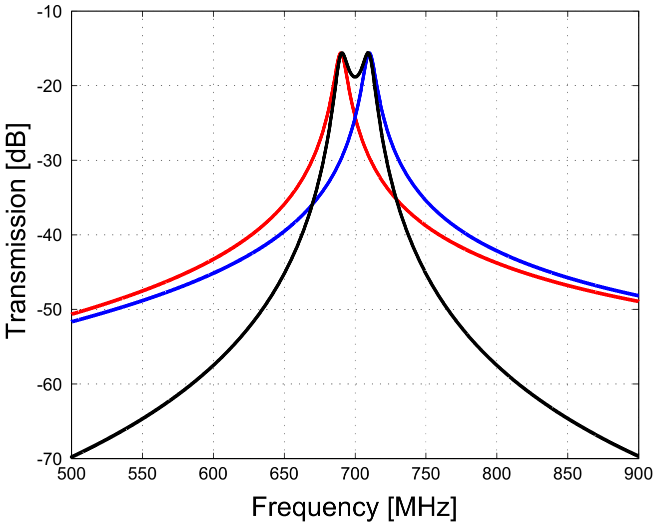

Figure 2: The simulation results of two 2nd-order resonators with $\omega_{\text{-3dB},\,i} = 8\,\mathrm{MHz}$ *and* $\Delta\omega_{res} = 20\,\mathrm{MHz}$ *(*$\omega_{\text{-3dB},\,i} < \Delta\omega_{res}$*).*

$$H_i(j\omega) \cong \frac{A_0}{1 + \frac{j}{\omega_{\text{-3dB}}/2} \times (\delta\omega \pm \Delta\omega_{\text{res}}/2)}, \tag{3}$$

where $A_0 = R_{\text{s}} / \left(2R_{\text{s}} + R_{\text{m}}\right)$.

Using (3), the amplitude and the phase of each resonator, can be expressed as:

$$A_{\text{mid.}} = |H_i(j\omega)| = \frac{A_0}{\sqrt{1 + \frac{\Delta\omega_{\text{res}}^2}{\omega_{\text{-3dB}}^2}}}, \tag{4}$$

$$\varphi = \angle H_i(j\omega) = \pm\, tan^{-1} \frac{\Delta\omega_{\text{res}}}{\omega_{\text{-3dB}}}. \tag{5}$$

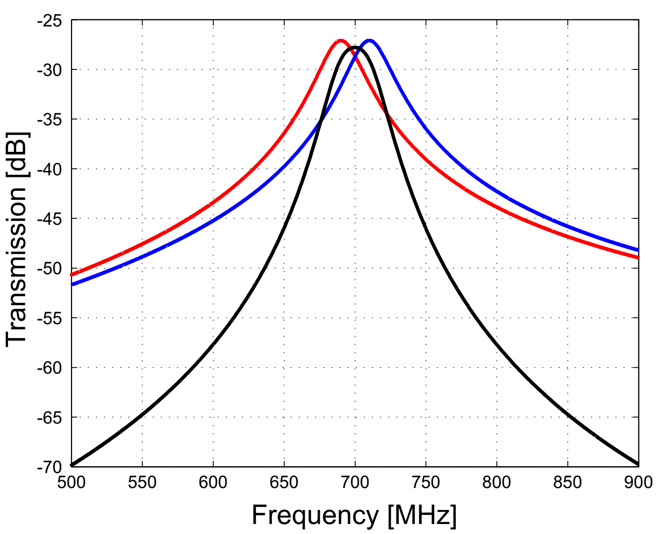

Figure 3: The simulation results of two 2nd-order resonators with $\omega_{\text{-3dB},\,i} = 30\,\mathrm{MHz}$ *and* $\Delta\omega_{res} = 20\,\mathrm{MHz}$ *(*$\omega_{\text{-3dB},\,i} > \Delta\omega_{res}$*).*

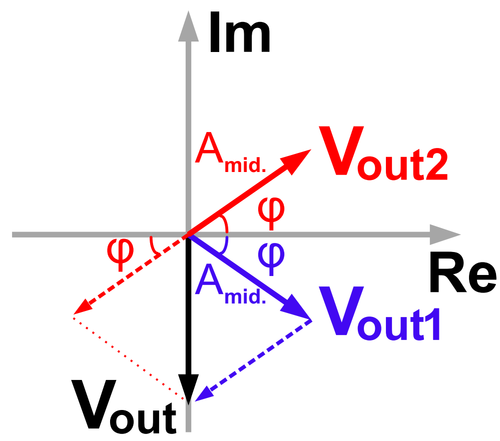

Figure 4: Complex plane representation of two in-phase actuated resonators (V_{out1} *and* V_{out2}*) and subtracted* V_{out} *at center frequency (*f_c*).*

The amplitude and the phase of the total output of the filter at ω_{c} are shown in Fig. 4. By increasing $\Delta\omega_{\text{res}}$, φ increases and the output vectors of each resonator are adding up more constructively. However, at the same time, the magnitude of each resonator decreases. Therefore, there must be an optimum $\Delta\omega_{\text{res}}$ to maximize the total output of the filter at ω_{c} ($A_{\text{mid.}}$) and hence reducing the ripple of the filter. It can be shown that the optimum $\Delta\omega_{\text{res}}$ equal the bandwidth of each resonator ($\Delta\omega_{\text{res}} = \omega_{\text{-3dB},\,i}$), which leads to an optimum value of $A_{\text{mid.}} = A_0$ and $\Delta\varphi = \pi/2$. The optimum condition has been shown in Fig. 5 assuming $\Delta\omega_{\text{res}} = \omega_{\text{-3dB},\,i} = 20\,\mathrm{MHz}$. If $\Delta\omega_{\text{res}} > \omega_{\text{-3dB},\,i}$, the filter response will contain a certain ripple which increases with increasing $\Delta\omega_{\text{res}}$. If $\Delta\omega_{\text{res}} < \omega_{\text{-3dB},\,i}$, the filter response will have zero ripple but the insertion loss of the filter will increase.

By assuming the same equal quality-factors ($Q_1 = Q_2$) for the resonators, they will have about the same bandwidths ($\omega_{\text{-3dB},\,1} \approx \omega_{\text{-3dB},\,2}$). Based on the formula presented in (4) and (5), this leads to $\varphi_1 \approx -\varphi_2$. However, if the fabrication variation of the resonators is leading to a considerable quality-factor differences, then the bandwidths will be different as well as the insertion losses of the resonators. This can lead to some distortion on the filter performance. To investigate further on this issue, a mismatching between the resonators has been applied to the ideal case presented in Fig. 5. By assuming an exaggerated value of 20% mismatch between the amplitude of the resonators ($V_{\text{out},\,2} = 0.8 \times V_{\text{out},\,1}$), there will be a certain distortion in the filter response as illustrated in Fig. 6. As seen, the mismatch between the resonators is leading to a distortion in the filter behavior and still shows an acceptable performance as a BPF. As seen, in Fig. 6, if there is an intersection between the responses in the region where the resonators are in phase, there will be a notch in the resulted filter response.

EXPERIMENTS

The proposed technique is realized using two contour-mode resonators, Fig. 7, with slightly different resonance frequencies approximately at 700 MHz. The configuration and the SEM image of the device are shown in Fig. 7(a) and (b). To cancel the feed-through signal of each resonator, the bottom-electrode patterning method [7] has been utilized for both resonators. Each resonator consists of 6 fingers with 3 fingers for the input and output ports. Each finger has

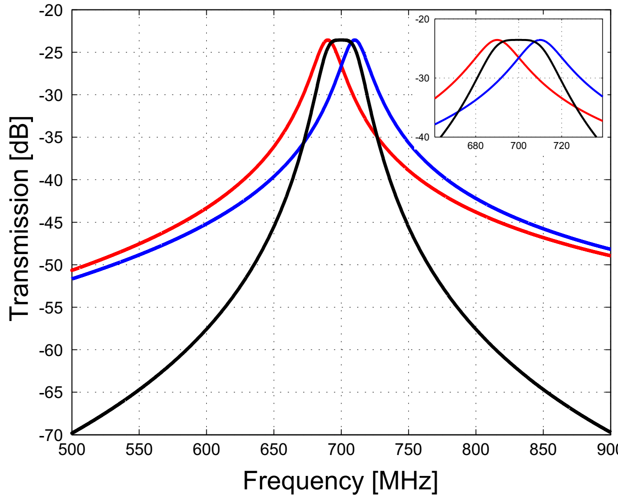

Figure 5: The simulation results of two 2nd-order resonators with $\omega_{\text{-3dB},\,i} = \Delta\omega_{res} = 20\,\text{MHz}$.

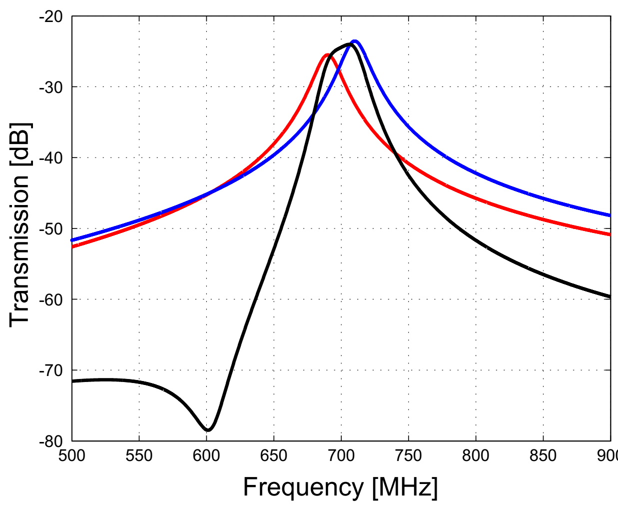

Figure 6: The result of 20% mismatch between the output of the resonators ($V_{out2} = 0.8 \times V_{out1}$*)*

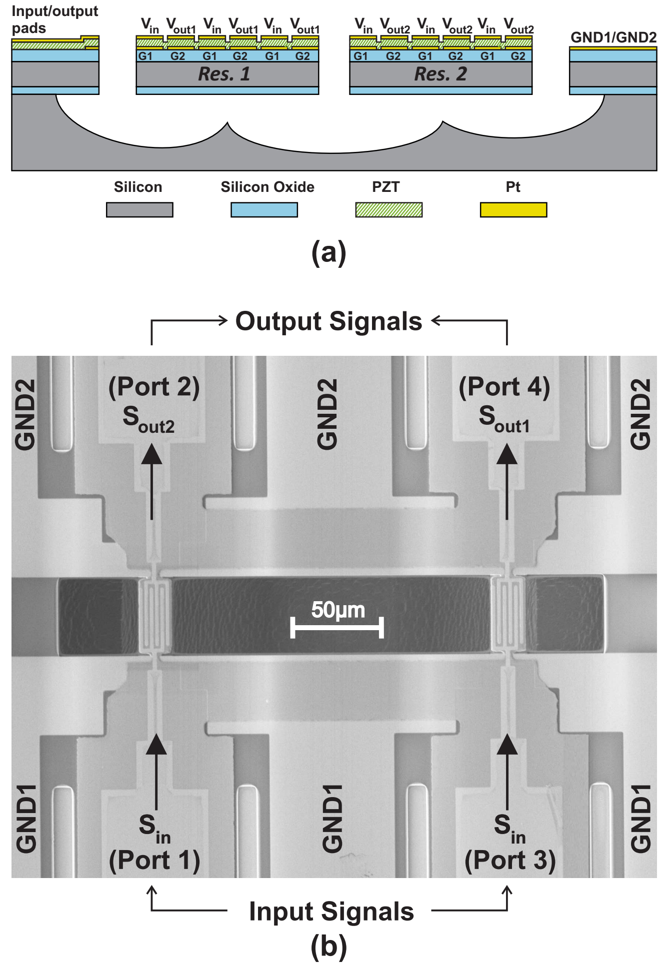

Figure 7: (a) The cross-section schematic of two bottom-electrode patterned resonators, (b) Scanning electron micrograph (SEM) of two 2nd-order PZT-on-silicon resonators with slightly different resonance frequencies around 700 MHz.

around 2.2 µm width with 1.8 µm spacing in between. The resonators consist of a composite stack of (bottom to top) SiO_2/Si/SiO_2/Ti-Pt/PZT/Pt. The fabrication process is similar to the earlier work presented in [7]. A 1 µm PLD-based PZT thin-film is utilized on top of an around 3 µm silicon layer using a highly resistive silicon-on-insulator (SOI) wafer with 0.5 µm buried oxide (BOX) layer.

The resonators were characterized using GSGSG probes and a 4-port Agilent N5244A network analyzer. A 4-port SOLT calibration has been performed using Agilent Electronic Calibration Module up to the probes. All the measurements have been done by applying 0 dBm input power. The frequency response of each fabricated resonator (S_{21} and S_{43}) and the differential readout of the filter (S_{dc21}) at DC-bias voltage of 0 V are shown in Fig. 9(a). The phase change of each resonators is presented in Fig. 9(b). The phase of the resonators at f_c is around $\varphi_i = \pm 120^\circ$. The BPF is characterized using 50 Ω termination and shows the bandwidth of approximately 28.6 MHz with 30 dB stopband rejection. The transmission of the filter (S_{dc21}) at different DC-bias voltages, 0-4 V, with 1 V increments are presented in Fig. 9.

As presented in Fig. 8(a), by using the differential readout technique, the feed-through signals are canceled out and the stopband rejection of the filter stays quite constant compared to each single resonator response. Utilizing the presented method, there is more 20 dB improvement in the stopband rejection. As seen in Fig. 8(b), the phases ($\varphi_i = \pm 120^\circ$) are not in the optimum condition ($\Delta\varphi = \pi/2$), therefore it leads to a ripple of 3.52 dB at 0 V DC-bias voltage. By applying DC-bias voltage, the bandwidth of each resonator is increasing which causes the reduction of the ripple to 1.55 dB at 4 V.

By applying the DC-bias voltages, the transverse piezoelectric coefficient (e_{31}) of PZT layer increases and subsequently causes the reduction of motional impedance and insertion loss of the resonators. As seen in Fig. 9, the insertion loss of the filter is reduced from -23.75 dB up to -14.18 dB with DC-bias voltage of 0 V till 4 V. In-band IIP3 measurement [3] of the filter is presented in Fig. 10 using the DC-bias of 4 volt. The IIP3 at different DC-bias voltages is presented as a subfigure of Fig. 10.

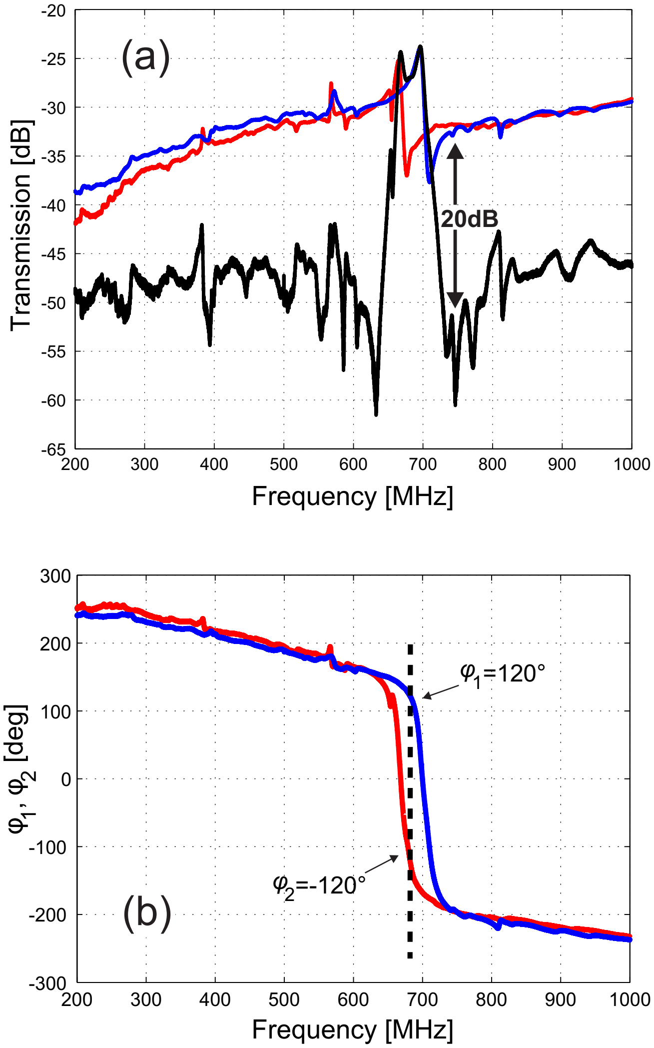

Figure 8: (a) Measured transmission gain of individual resonators and the filter (Sdc21) with 50 Ω termination at DC-bias of 0 V, (b) The measured phase change of the resonators.

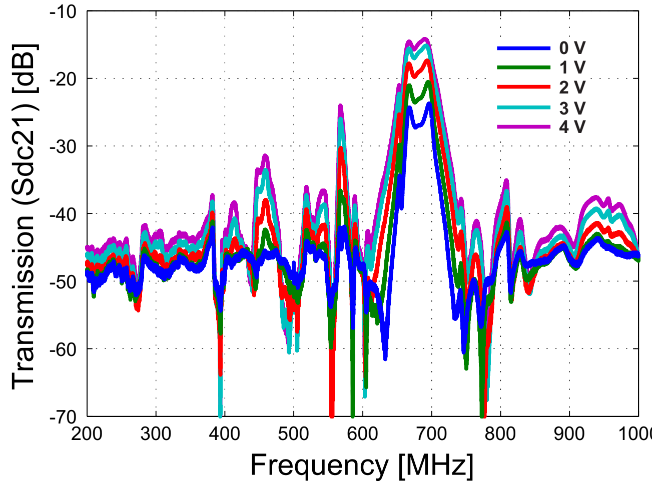

Figure 9: Measured transmission gain of the filter at different DC-bias voltages from 0-4 V with 1 V increments.

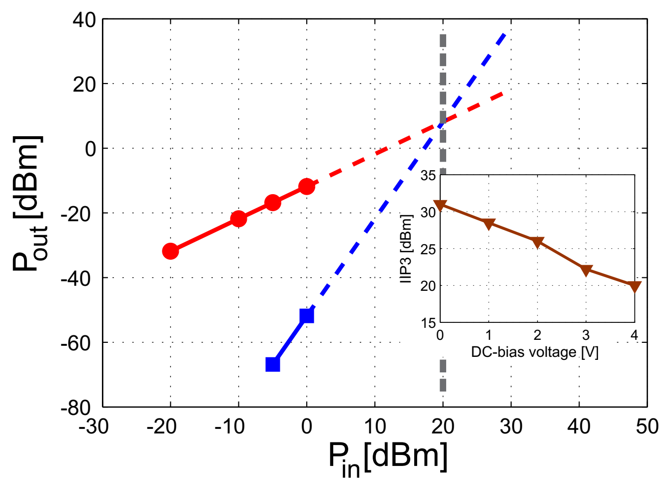

Figure 10: Measured IIP3 for the 700 MHz 4th-order filter at DC-bias of 4 V.

CONCLUSIONS

A new 4th-order BPF technique was demonstrated using the subtraction of two in-phase actuated 2nd-order contour-mode Lamb-wave resonators with slightly different resonance frequencies around 700 MHz. Utilizing this technique, the output of the resonators were added up constructively as the phase of the resonators was φ_1=-φ_2, while the feed-through signals were eliminated. The BPF was presented with a bandwidth of approximately 28.6 MHz and 30 dB stopband rejection. This technique is a powerful approach for RF-MEMS filter design as well as resolving design issues associated with feed-through at high frequencies for materials with high-dielectric constant such as PZT.

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