Wideband Hybrid Monolithic Lithium Niobate Acoustic Filter in the K-Band

Liuqing Gao®, Member, IEEE, Yansong Yang®, Member, IEEE, and Songbin Gong®, Senior Member, IEEE

Abstract—This article presents the design approach and the first demonstration of a wideband hybrid monolithic acoustic filter in the K-band, which exceeds the limitation of electromechanical coupling on the fractional bandwidth (FBW) of acoustic filters. The hybrid filter utilizes the codeesign of electromagnetic (EM) and acoustic to attain wide bandwidth while keeping the advantages of small sizes and high Q in the acoustic domain. The performance trade space and design flow of the hybrid filter are also presented in this article, which allows this technology to be applied for filters with different center frequencies and FBWs. The hybrid filter is simulated by hybridizing the EM and acoustic finite element analysis, which are carried out separately and combined at a system level. The fabricated filter built with resonators having an electromechanical coupling of 0.7% based on the seventh-order antisymmetric Lamb wave mode (A7) has a 3-dB FBW of 2.4% at 19 GHz and a compact footprint of 1.4 mm².

Index Terms—K-band, microelectromechanical systems (MEMSs), millimeter-wave devices, piezoelectric devices.

I. INTRODUCTION

As the sub-6-GHz spectrum becomes overcrowded with applications, the research community starts to explore beyond 6 GHz for new spectral venues to advance wireless capabilities. Several bands ranging from 12 to 27 GHz have been proposed [1], sharing the same challenge in scaling conventional front-end components well beyond their current operating frequencies. One indispensable front-end component that is particularly difficult to scale in frequency is the acoustic filters that have been commercially successful for 4G [2], [3]. Frequency scaling without compromising performance remains difficult due to various technical bottlenecks in material integration, device fabrication, and filter design for acoustic filters.

The scaling approaches so far can be classified into two categories. The first type resorts to the reduction in feature size for increasing the center frequency (e.g., electrode width for lateral mode or film thickness for thickness mode devices) [4], [5]. However, excessive thickness or size reduction often leads to poor film quality for the acoustic material or higher electrode resistance, consequently resulting in higher insertion loss (IL), degraded power handling, and more severe nonlinearity. The other type of approach adopts higher order of acoustic resonant modes (i.e., overmoding) while maintaining the sizes of the resonant cavity and other device features in the process of scaling [6], [7]. Unfortunately, these approaches face the loss of electromechanical coupling ($k_t^2$ $\sim$ 1/$f^2$ in scaling) and, hence, trade FBW off for a higher center frequency. In particular, scaling 4G acoustic filters based on surface acoustic wave (SAW), film bulk acoustic resonators (FBARs), or other compound modes [8] toward 20 GHz and beyond by overmoding might be more penalizing than rewarding. Their electromechanical coupling of, respectively, 10% and 6.5% would be reduced too much to recover from for anything other than extreme narrowband applications.

Recently, asymmetric Lamb wave microacoustic resonators based on LiNbO₃ have been shown with significantly larger $k_t^2$ of 30% [9], thus allowing better trades between $k_t^2$ and frequency scaling. Researchers have indeed demonstrated higher order asymmetric resonators up to 30 GHz with Qs of 400 and filter up to 10 GHz with IL of 3 dB. However, the fractional bandwidth (FBW) of the 10-GHz acoustic filter is only 0.7% due to the aforementioned design trade [10]. To overcome the $k_t^2$ loss in scaling and recover the FBW loss, this work codesigns acoustic and electromagnetic (EM) structures in tandem to enhance the FBW while still harnessing the small size and high Q in the acoustics domain. The principle behind our approach is to use the acoustic resonance in conjunction with an inductive element as the building blocks for constructing a ladder filter [11], [12]. The inductor can equivalently recover some of the $k_t^2$ loss, enhance the FBW, and introduce an additional antiresonance without harshly compromising IL and roll-off of the filter. Its effect has been similarly explored using coupled modes for lower frequency hybrid filters that combine acoustic resonator with inadequate $k_t^2$ and lumped elements [13]. However, the added value at lower frequencies for this approach comes at the expense of larger sizes as substantial inductances are typically needed. High-frequency adaptation of a hybrid filter has not been demonstrated (primarily due to the absence of 20-GHz acoustic resonators) despite it requires a much smaller inductance for the same purpose and does not add significant size.

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In this work, we develop a hybrid filter design that combines chip-scale reactive elements with A7 mode LiNbO3 resonators at 19 GHz. As a result of our effort in codesigning in the EM and acoustic domains, this work widens the FBW of acoustic filters. It demonstrates an acoustic filter with 2.4% FBW, only using resonators with 0.7% $k_t^2$ and occupying a small footprint of 1.4 mm$^2$.

II. Codesign of Acoustics and EM Structures

A. Inductor Induced BW Enhancement for Acoustic Resonator

The mock-up of a Z-cut thin-film LiNbO3 resonator is shown in Fig. 1(a) and (b). The resonance of the antisymmetric Lamb wave mode is governed by the mode order $(m)$, film thickness $(t)$, the separation between adjacent interdigital electrodes $(l)$, and acoustic phase velocities in the vertical $(v_t)$ and longitudinal $(v_L)$ directions, which can be expressed by

$$f_{r,m} = \sqrt{\left(\frac{m v_t}{2l}\right)^2 + \left(\frac{v_L}{2l}\right)^2}.$$  \hspace{1cm} (1)

In this work, $t$ of 650 nm and $l$ of 4 $\mu$m are selected considering the power handling and fabrication capability. As the acoustic phase velocities $v_t = 3480$ m/s and $v_L = 6300$ m/s for Z-cut LiNbO3, the resonator to operate in the K-band, A7 mode is selected. The ratio of $k_t^2$ of the higher ordered antisymmetric mode to that of A1 is inversely proportional to the mode order’s square. A simple ladder acoustics filter would have an FBW that narrows at a rate inversely proportional to the increasing center frequency square. As a result, a ladder filter constructed with A7 will have an FBW of 0.35%, leading to few applications due to the extremely narrow bandwidth.

Connecting a shunt inductor to an acoustic resonator is a well-known technique to enlarge the spectral separation between the series and parallel resonances of the circuit, equivalent to the effect of increasing $k_t^2$ of the resonator but without the benefit of a larger impedance ratio $(Z_{ar}/Z_0)$. Such an effect can be modeled by representing the acoustic resonator with the modified Butterworth–Van Dyke (MBVD) model in a circuit comprising an acoustic resonator and shunt inductor, as shown in Fig. 2(a). $L_{shunt}$ denotes the shunt inductor, $C_0$ denotes the static capacitance, $R_0$ represents the loss in LiNbO3 film, and $R_m$, $C_m$, and $L_m$ electrically represent the motional branch and mechanical resonance. $C_f$ denotes the parasitic capacitance, which is an EM effect to be considered in the later layout design. The admittances of a resonator with and without a shunt resonator are plotted in Fig. 2(b). The antiresonance ($f_{ar+}$) of the circuit is shifted by $L_{shunt}$ to a higher frequency than the antiresonance of the acoustic resonator ($f_{ar}$), producing a larger BW that can be similarly achieved with a higher $k_t^2$ acoustic resonator. The inclusion of $L_{shunt}$ also induces an additional antiresonance at a lower frequency ($f_{ar-}$) that results from a second solution where $L_{shunt}$ and $L_m$ collaboratively tune out $C_m$ and $C_0$.

The resonance ($f_r$) and antiresonances ($f_{ar\pm}$) of the circuit can be theoretically calculated. Due to the low loss tangent of LiNbO3, the influence of substrate loss on $f_r$ and $f_{ar\pm}$ is negligible, and the input admittance ($Y_{11}$) looking into the port shown in Fig. 2(b) can be expressed by

$$Y_{11} = \frac{1}{j\omega L_m + L_{shunt}} \frac{1}{\frac{1}{\frac{1}{L_{shunt}} + \frac{1}{C_m}}}.$$  \hspace{1cm} (2)

$$\chi = \frac{L_m}{C_0} - \frac{1}{\omega^2 C_0 C_m}$$  \hspace{1cm} (3)

where $f_r$ is the frequency at which the denominator equates zero, and $f_{ar\pm}$ are the frequencies at which the numerator equates zero. The closed-form expressions of $f_r$ and $f_{ar\pm}$ are

$$f_r = \frac{1}{2\pi \sqrt{L_m C_m}}$$  \hspace{1cm} (4)

$$f_{ar\pm} = \frac{1}{2\pi} \sqrt{\frac{b \pm \sqrt{b^2 - 4L_m L_{shunt} C_0 C_m}}{2L_m L_{shunt} C_0 C_m}}$$  \hspace{1cm} (5)

$\Delta f$ is defined as the spectral separation between $f_{ar+}$ and $f_{ar}$. For a resonator at 19.15 GHz with $k_t^2$ of 0.7% and $C_0$ of 150 fF, its $\Delta f$ versus $L_{shunt}$ is plotted in Fig. 2(c), which is used to determine the $L_{shunt}$ value during the filter design for a target $\Delta f$.

B. Codesign Procedure

To fully harness the benefit of a virtually increased $k_t^2$ of a resonator by a shunt inductor for constructing a filter with wider FBW, the offset between the resonances of series and parallel resonators is increased by $\Delta f$. This creates a challenge
in implementing the design, as adjusting the lateral dimensions of the interdigital electrodes only gives a limited amount of offset. Thus, we opt for trimming the film thickness, which dominantly affects the resonance, as shown in (1). This is the first demonstration of this monolithic technique to create an increased resonance offset.

Consider a filter constructed with only shunt inductors \(L_{p_{shunt}}\) connected to the parallel resonators, and as can be seen in Fig. 3(a), the stopband filter performance deteriorates. To improve the stopband filter performance, shunt inductors \(L_s_{shunt}\) are connected to series resonators to decrease the admittance through the series branch in the stopband, as shown in Fig. 3(b). As a result, most of the input power gets reflected in stopband due to impedance mismatching. The stopband performance is greatly improved, as shown in the filter IL and RL in Fig. 3(c).

### C. Finite Element Simulation of MEMS Resonators

At high frequencies, the self-inductance from the lead lines and electrodes of a microelectromechanical system (MEMS) resonator is no longer negligible. Along with the static capacitance, it introduces a self-resonance that masks acoustic resonant response. Consequently, the power transmitted through the mechanical coupling between the interdigital electrodes can be significantly reduced.

The self-resonance frequency of a MEMS resonator is dependent on the number of electrodes, electrode length, width, and separation. To increase the self-resonance, the high-frequency EM response from the electrode layout of the acoustic resonator is studied with a high-frequency structure simulator (HFSS). Since our target operating frequency is at 19 GHz, as shown in Fig. 4, an electrode layout with an electrical self-resonance at 24 GHz is targeted for mitigating its influence on filter performance. The resonator has 30 pairs of electrodes, with electrode width of 3 \(\mu\)m and separation of 4 \(\mu\)m, resulting in a \(C_0\) of 150 fF, a self-inductance \((L_s)\) of 0.28 nH, and a resistance \((R_s)\) of 4.8 \(\Omega\).

### D. Cosimulation of Hybrid MEMS Filters

In addition to the self-inductance, the parasitic effects and couplings in the layout can also be significant, which must be considered in the filter design. However, commercial modeling solutions do not support finite element analysis (FEA) that couples EM and acoustic simulations at drastically different scales. Therefore, the codesign of the filter layout is done by first performing EM simulation in HFSS and then adding motional branches derived from acoustic FEA to fully capture effects in layouts and acoustics in the advanced design system (ADS). \(C_0\) and \(R_0\) of a standalone resonator from EM simulations are consistent with those from COMSOL-based FEA, which validates this method.

The whole layout of the hybrid filter is first simulated in HFSS, and the model is shown in Fig. 5(a). The yellow rectangles represent lumped ports, with six ports assigned for adding motional branches to each resonator in ADS in the following step and four ports corresponding to the two GSG probes for measurement. The HFSS-simulated S-parameter matrix of the ten-port system is compiled into a touchstone file. The file is then imported into ADS, and the motional branches in the MBVD models of the resonators are added to their corresponding ports. The simulated IL and RL by this cosimulation method are plotted in Fig. 5(b) and (c) as Co-simulation #1. Co-simulation #1 matches the measurement results, proving that the cosimulation method provides a valid prediction of the filter performance.

Due to the small feature size (3 \(\mu\)m) of the interdigital electrodes, the mesh size needs to be very small to get accurate results. Significant computing resources are required, and it might still take a long time for one simulation. This makes the iterative layout design process tedious and time-consuming.
To this end, a much more time-efficient cosimulation approach is developed. Considering the drastic size difference between the interdigit electrodes and the rest of the layout, the mesh size can be increased when the layout is simulated without the electrodes. As a result, the simulation time can be reduced seven times. Assuming that the coupling between the interdigit electrodes and the bus lines is negligible, the filter layout is simulated without the electrodes, while the electrodes are simulated in a separate HFSS model. The result is then combined with the filter layout in ADS at a system level. The simulated results using this method are plotted in Fig. 5(b) and (c) as Co-simulation #2. Comparing the cosimulation result with the measurement, method #2 does not match measurement, as well as Cosimulation method #1.

However, it provides adequate accuracy for adjusting the layout for reducing the parasitic coupling caused by the bus lines.

III. DESIGN TRADEOFF AND PROCEDURE

A. Performance Trade Space

A design tradeoff analysis is performed onto the design space of such a wideband hybrid monolithic acoustic filter, which guides efficiently optimizing the filter performance and sets a realistic performance objective. The S-parameters of the hybrid filter are calculated from the circuit model shown in the inset of Fig. 3(c). The circuit model involves $k_2^t$, $C_0$, and resonator $Q$ of the parallel and series resonators, as well as $L_p$ shunt, $L_s$ shunt, and the inductor $Q$ ($Q_{ind}$).

With the key circuit parameters listed in Table I, the passband IL and Out-of-Band Rejection (OoBR) of the hybrid filter with resonator $k_2^t$ varying from 0.005 to 0.02 and $Q_{ind}$ varying from 10 to 50 are plotted in Fig. 6. As shown in Fig. 6(a), the passband IL of the filter has a high dependence on $k_2^t$ and $Q_{ind}$ when both design parameters are small, whereas the dependence decreases at larger values of $k_2^t$ and $Q_{ind}$. Larger $k_2^t$ and $Q_{ind}$ result in smaller passband IL; however, with the resonator $Q$ of 500, the smallest passband IL is 2.4 dB, which results from the mechanical loss in resonators with resonator $Q$ of 500. Similarly, Fig. 6(b) indicates that the OoBR has a similar trend to the passband IL and also favors large $k_2^t$ and
$Q_{\text{ind}}$. However, due to $k_2^2$ being roughly proportional to $1/f^2$ in scaling, the value of $k_2^2$ is very small at high frequencies. Moreover, due to the skin effect, it is also challenging to get high $Q_{\text{ind}}$ at high frequencies.

### B. Design Procedure

As shown in Fig. 7, the design of a wideband hybrid monolithic acoustic filter starts with deciding the center frequency and designs of the comprising resonators. Next, the FBW of the filter is determined based on the needs of the application and performance tradeoff, which is used to calculate the LiNbO$_3$ thinning needed to get the desired offset between the resonances of series and parallel resonators. Then, the value of $L_{\text{shunt}}^p$ connecting to the parallel resonator is calculated from (7). The goal is to achieve the desired amount of $\Delta f$ such that the higher antiresonance ($f_{p\text{ar}}^+$) of the parallel resonator aligns with the resonance of the series resonator ($f_s^r$)

$$L_{\text{shunt}}^p = \frac{1}{(\omega_{p\text{ar}}^+)^2 C_0^P} - \frac{1}{(\omega_{p\text{ar}}^+)^2 L_m^P + \frac{1}{C_0^P}} \frac{1}{\frac{1}{C_m^P} - (\omega_{p\text{ar}}^+)^2 L_m^P + \frac{1}{C_0^P}}$$

(7)

where $\omega_{p\text{ar}}^+ = 2\pi f_{p\text{ar}}^+$, $L_m^P$ and $C_0^P$ are coming from the electrical representation of the motional branch and mechanical resonance in the MBVD circuit model, and $C_0^P$ is the static capacitance of the parallel resonator. Similarly, the value of $L_{\text{shunt}}^s$ connecting to the series resonator is calculated such that the lower antiresonances of the series and parallel resonators align [as shown in Fig. 3(b)] to improve the stopband filter performance

$$L_{\text{shunt}}^s = \frac{1}{(\omega_{s\text{ar}}^-)^2 C_0^S} - \frac{1}{(\omega_{s\text{ar}}^-)^2 L_m^S + \frac{1}{C_0^S}} \frac{1}{\frac{1}{C_m^S} + (\omega_{s\text{ar}}^-)^2 L_m^S + \frac{1}{C_0^S}}$$

(8)

where $\omega_{s\text{ar}}^- = 2\pi f_{s\text{ar}}^-$, $L_m^S$ and $C_0^S$ are coming from the MBVD circuit model, and $C_0^S$ is the static capacitance of the series resonator. Next, HFSS is used to translate the inductor values into physical dimensions of the single-turned on-chip inductors. Finally, a cosimulation is performed to the whole layout, and the layout is adjusted to reduce the parasitic effects.

A hybrid filter is designed following this procedure, and the design parameters are listed in Table II. The physical dimensions, $l_s$ and $w$, of $L_{\text{shunt}}^s$ are marked in Fig. 8(a).

### IV. Measurement Results and Discussion

The devices were fabricated with a standard thin-film LiNbO$_3$ MEMS resonator fabrication process on a 650-nm Z-cut LiNbO$_3$ thin film on a high resistivity Si wafer. The desired offset between the resonances of series and parallel resonators is achieved by regionally thinning the LiNbO$_3$ film where the series resonators are situated [14]. SPR 220 is patterned as a mask for the thinning process. Chlorine-based inductively coupled plasma reactive ion etching (ICP-RIE) is used at a slow etching rate (5 nm/min) to precisely control the thinning process. Such a technique shifts the resonance to a higher frequency. The film is thinned down from 650 to 635 nm, creating a resonance offset of 400 MHz between the series and shunt resonators. To reduce electric loss, the

### Table II

| Design Parameter | Dimension |
|------------------|-----------|
| Electrode length | 35 μm     |
| Electrode width  | 3 μm      |
| Electrode separation | 4 μm   |
| Number of electrodes | 60   |
| Side length of $L_{\text{shunt}}^p$ | 337 μm |
| Width of $L_{\text{shunt}}^p$ | 20 μm |
| Side length of $L_{\text{shunt}}^s$ | 282 μm |
| Width of $L_{\text{shunt}}^s$ | 20 μm |

Fig. 7. Hybrid filter design procedure.

Fig. 8. (a) SEM image of an inductor in the first fabrication round (without release and partially electroplated). (b) Optical microscope image of an inductor in the second fabrication round (with release and fully electroplated). Measured (c) inductances and (d) $Q$s of inductors in the first and second fabrication rounds.
inductors, lead lines, and probing pads (i.e., regions other than the acoustic resonator cavities) are thickened with an additional step of copper electroplating, as shown in Fig. 9(a).

The electroplating process uses a 50-nm-thick copper layer as a seed layer and SPR 220 as the mold. The quality of the electroplated copper is highly dependent on the plating rate, which is controlled by the dc bias voltage and current to be 10 mV and 9 mA to be 500 nm/min. After electroplating, the seed layer is removed by dipping the sample into copper etchant.

In addition, the inductors, as shown in Fig. 9(d), are also suspended to reduce the substrate loss caused by the Eddy current in the Si substrate. For performance diagnosis purposes, comprising structures, such as a standalone resonator with or without a shunt inductor, were also fabricated along with the filter. The measurement results of these comprising structures and the MEMS filter are reported and analyzed in this section.

A. Standalone Inductors

From the trade space analysis, it is apparent that the inductor $Q$ influences the filter performance. Several well-known methods, including curved corners [15], Si substrate removal [16], and electroplating [17], have been employed to improve on-chip inductor $Q$.

In the first round of fabrication, the inductors and bus lines are partially thickened by electroplated copper. The electroplating avoids the anchors and suspended peripherals of the resonators so that electroplating-induced stress would not fracture the anchors and lower the yield. As a result of the partial electroplating, there is one region [as highlighted in Fig. 8(a)] in the inductor connected to a shunt resonator and two regions [as highlighted in Fig. 9(c)] in the inductor connected to a series resonator having only 150-nm thickness of Al. In the fabrication, the Al layer in this region is partially etched in the step of removing the seed layer for electroplating, which further increases the resistive loss. Thus, eliminating the gaps caused by partial electroplating could increase the inductor $Q$.

In the second fabrication round, the electrode length is shorter. The release radius is smaller, and the bus line between two adjacent resonators does not suspend after release, which allows all the inductors and bus lines to be electroplated. The optical microscope image of a fully electroplated inductor with Si substrate removal is shown in Fig. 8(b), and its measured inductance and $Q$ are compared to an inductor without release and with one gap in Fig. 8(c) and (d). By releasing the inductor, its self-resonance would occur at a higher frequency from reduced eddy current in the Si substrate. As shown in Fig. 8(d), removing the gap has largely improved the inductor $Q$ in the lower frequency range, and the peak $Q$ value increases from 11 to 33.

The inductance is close to the design value (0.53 nH); however, the resistive loss is 3.75 times larger than simulated at 19 GHz. This is caused by the electroplated copper having a significantly lower conductivity due to the nonoptimal plating recipe, as well as the increased surface roughness of the electroplated copper in the seed layer removal process. These issues can be fixed upon further optimizations in future fabrications.

B. Comprising Structures

The SEM image and measured response of the fabricated standalone resonator are shown in Fig. 9(a) and (b). The measured response is fitted by the MBVD model with additional branches taking the self-inductance ($L_s$) and parasitic capacitance ($C_f$) into account. The parameters extracted from the circuit fitting that quantifies the resonator A7 mode response are resonance at 19.6 GHz, $Q$ of 228, $k_f^2$ of 0.7%, $C_0$ of 152 fF, $L_s$ of 0.31 nH, and $C_f$ of 19 fF.

Initially, the EM simulations are carried out in momentum for the ease of the system-level combination with the acoustic simulations. After the first fabrication round, the measurement results are compared with the simulation results. It is observed that the measured $L_s$ is three times larger than the momentum-simulated value and roughly the same as the HFSS-simulated value. As a result, the self-resonance of the resonator occurs at 24 GHz, much lower than the design value 40 GHz based on momentum simulation results. The SEM and optical microscope images of the fabricated resonator with a shunt inductor are shown in Fig. 9(c) and (d). Fig. 9(e) shows the measured response fitted with the circuit model to extract the inductance and $Q$ of $L_{\text{shunt}}$, which are 0.61 nH and 8, respectively.

C. MEMS Filter With Shunt Inductors

The SEM image and the measured and simulated performances of the fabricated filter are shown in Fig. 10(a)–(c). Simulation #1 represents the desired response with the key circuit parameters listed in Table I, with a resonator $Q$ of 500 and an inductor $Q$ of 30. Simulation #2 uses the self-inductance, parasitic capacitance, and resistive loss extracted from the comprising structures that are larger than the values predicted by EM simulations. Simulation #2 is consistent with the measurement results, whereas there is a 5-dB difference in the passband IL and a 6-dB difference in the stopband rejection.

Fig. 9. (a) SEM image and (b) measurement and circuit model of the fabricated resonator. (c) SEM and (d) optical microscope images. (e) Measured and circuit modeled responses of the fabricated resonator with a shunt inductor.
between the measurement and simulation #1. The discrepancy in the stopband is mainly caused by the parasitic capacitances between the lead lines. As shown in Fig. 10(a), the lead lines are long and cause considerable parasitic coupling. As mentioned in Section IV-A, the resistive loss in the fabricated inductor is 3.75 times larger than simulation, which contributes to a 3.7-dB drop in the passband IL. The second cause of the increase in passband IL is the resonator \( Q \) of 228 being much lower than the 500 in simulation (achieved in a previous fabrication run). This contributes to the other 1.3-dB difference in passband IL. The filter performance can be much improved in future designs by reducing \( L_s, C_f, \) and \( R_L \) and enhance the fabrication yield in producing resonators of \( Q \) around 500 and higher.

Based on the insights gained from the two fabrication rounds so far, several aspects can be optimized for improving the filter performance in future designs. The first part is to increase the inductor \( Q \) at 19 GHz from 10 to 30, which will reduce the passband IL by 3.7 dB. As shown in Fig. 6, IL and OoBR have a high dependence on \( k_f^2 \) and inductor \( Q \) (0.7% and 10 in the current fabricated device). An increase in inductor \( Q \) can be achieved by some recently developed technologies, such as the trap-rich layer method [18]. In addition, instead of LiNbO\(_3\) on the Si substrate, LiNbO\(_3\) on the glass substrate could also reduce the inductors’ self-resonance by reducing the parasitic capacitance of the inductors, which will shift the peak of inductor \( Q \) to a higher frequency (currently smaller than 10 GHz). Finally, the inductor \( Q \) can be further increased by increasing the thickness of the electroplated copper. The conductivity of the electroplated copper is measured to be 1/3 of the bulk conductivity, which corresponds to a skin depth of 0.47 \( \mu \)m at 19 GHz. For the miniaturized on-chip inductor in this article, the thickness of the metal trace is comparable to its width. As a result, increasing the metal thickness can effectively increase the cross-sectional perimeter, thus reducing the resistance of the inductor. Thicker metal results in higher capacitance between the inductor wires and reduces the self-resonant frequency, which reduces inductor \( Q \).

To validate these methods, an inductor with the same physical dimensions as the one in Fig. 8 is simulated in HFSS. The inductor is built on a trap-rich wafer and simulated with various Cu thickness: 3, 4, 6, and 8 \( \mu \)m. The simulated inductor \( Q \)s plotted in Fig. 11 indicate the feasibility of using the trap-rich layer method and thicker metal to improve inductor \( Q \). In the previous fabrication rounds, the thickness of the electroplated copper is limited by the thickness of the photoresist mold, SPR 220 with 4.8-\( \mu \)m thickness. If the electroplated copper is too thick, the spin-coated photoresist in the following fabrication step will be uneven due to the considerable thickness variation on the surface. It then causes issues in photolithography and the development of the photoresist. These fabrication challenges can be overcome in the future by developing new lithography recipes for thicker photoresist.

The second target is to increase the resonator \( Q \) from 228 to 500 or higher. It will reduce the passband IL by 1.3 dB. As shown in Fig. 9(b), the EM self-resonance of the resonator occurs at 24 GHz, which lowers the loaded \( Q \). As the HFSS simulation of the resonator matches well with the measurement, the resonator design can be further optimized with HFSS for higher self-resonance, consequently improving the loaded \( Q \). The mechanical \( Q \) of the resonator is dependent on the film quality of the LiNbO\(_3\) thin film, which can be enhanced by microfabrication advances.

The third venue is to optimize the width and orientation of the lead lines to reduce the parasitic coupling, which will improve the OoBR. Reducing the width of the bus lines and lead lines presents a tradeoff between resistive loss and the significant parasitic coupling at high frequencies.

For real-life applications, the temperature effects on the hybrid filter need to be considered. Although the temperature...
coefficient of frequency (TCF) of the devices presented in this article was not measured, there was a previously measured A1 mode resonator on 400-nm-thick LiNbO3 showing a TCF of $-63.85\,\text{ppm/K}$ [9]. With a 100 K variation in temperature, the operating frequency will vary by 0.64% about the center frequency. The temperature-induced variation can be greatly reduced by employing temperature compensation technologies, such as coating the resonator with a compensation oxide having a positive TCF.

### D. Hybrid MEMS Filter Employing A5

The fabricated K-band monolithic hybrid filter based on A7 demonstrated in this work has suboptimal filter performance due to the low inductor $Q$ and high parasitic coupling at 19 GHz. As indicated in Fig. 8(d), the inductor $Q$ increases when the operating frequency decreases from 19 to 14 GHz. As a result of the increased inductor $Q$ and reduced parasitic coupling at the lower frequency, the hybrid MEMS filter employing A5 on a LiNbO3 thin film of 650-nm thickness is expected to have better filter performance. Such a filter is designed and implemented using the methodology presented in this article, with details in [24].

As shown in Fig. 12, the filter employing A5 has a 3-dB FBW of 2.93% at 14.7 GHz, with A5 mode $k^2_t$ of 1.1%. It has a passband IL of 7.5 dB and a stopband rejection of 19.5 dB, as a result of the improved inductor $Q$ of 18 and reduced parasitic coupling at the lower frequency. Similar to the filter based on A7, the filter employing A5 also suffers from the suboptimal inductor and resonator $Q$s, and the passband IL can be reduced with improvements in the fabrication.

### E. Discussion on Incorporating Capacitors

In this work, a lumped-element inductor is used to compensate for the loss of $k^2_t$ in overmoded MEMS resonators. The filter is mainly based on acoustic components, whereas the hybridization between acoustics and EM domains can be expanded by adopting a larger portion of EM filters to attain a high OoBR. By adding two series resonators with shunt inductors to a third-order Elliptic filter, the roll-off and shape factor improve due to the four additionally transmission zeros. The circuit schematic is shown in Fig. 13(a).

With the circuit parameters listed in Table III, the simulated filter performances are shown in Fig. 13(b) and (c), which shows a passband IL of 10 dB under the assumption of an inductor $Q$ of 50, a capacitor $Q$ of 500, and a resonator $Q$ of 500. The main obstacle preventing the real implementation of this design is the lack of high-$Q$ on-chip inductors and capacitors. Due to the larger portion of EM components in this design, the passband IL is more susceptible to the electrical loss in the nonideal $LC$ components.

### F. Comparison to SoA

There have been several EM waveguide filters [19]–[21] and acoustic filters [10], [22], [23] in the comparable frequency range of this work. It is the first demonstration of a hybrid monolithic acoustic filter in the K-band and gives the largest FBW to the $k^2_t$ ratio for acoustic demonstrations.

As seen from the SoA EM filters listed in Table IV, the filter presented in this work is hundreds of times smaller in volume. This work can potentially fill the gap of lacking miniaturized filters in the K- and Ku-bands for handheld applications.

On the other hand, as seen from the SoA acoustic filters listed in Table V, the filter FBWs are limited by the $k^2_t$'s
of the resonators. This work has demonstrated a method to exceed the $k_2^2$ limitation on FBW. As the resonant frequency and $k_2^2$ of an FBAR are dependent on the thicknesses of the piezoelectric film and electrodes, it requires excessive reductions in thickness to scale to high frequencies. For instance, the FBAR filter operating at 24 GHz requires an AlN film thickness of 120 nm and an electrode thickness of 20 nm [24], both highly susceptible to the fabrication variations nonuniformity. However, the high resonance in this work is attained by overmoding, which does not require a reduction in physical dimensions.

In a previously demonstrated MEMS filter, the FBW was limited by the frequency offset between the resonances of the series and parallel resonators. The LiNbO$_3$ MEMS filter [10] utilizing the third-order antisymmetric Lamb wave mode did not fully use the FBW allowed by the $k_2^2$. A greater frequency offset has been attained in this work by regionally thinning the LiNbO$_3$ film at the regions the series resonators situated using a finely controlled recipe.

V. CONCLUSION

In this work, a wideband hybrid monolithic acoustic filter in the K-band is codeigned in the EM and acoustic domains. The cosimulation carries out FEA in two domains separately and then combines the results at a system level. Using the cosimulation technique, this work has demonstrated an FBW exceeding the $k_2^2$ limitation on FBW typically seen in acoustic filters while maintaining a small footprint of 1.4 mm$^2$. The filter performance in this article is far from the limit and can be improved by resolving the high resistive loss currently plaguing the electroplated copper.

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Liuqing Gao (Member, IEEE) received the B.S. and M.S. degrees in electrical engineering from the University of Illinois at Urbana–Champaign (UIUC), Urbana, IL, USA, in 2016 and 2020, respectively, where she is currently pursuing the Ph.D. degree. Her research interests include design and microfabrication techniques of microelectromechanical systems (MEMS) resonators, filters, switches, and wireless communication systems. Ms. Gao won the Best Student Paper Award at the 2020 IEEE International Ultrasonics Symposium and the Third Place in the Best Paper Competition at the 2020 IEEE International Microwave Symposium. She was a recipient of the 2015 Omron Electrical Engineering Scholarship, the 2016 E. C. Jordan Awards, the 2016 Illinois Engineering Achievement Scholarship, the 2016 Highest Honors at Graduation, the 2017 ECE Distinguished Research Fellowship, the 2018 James M. Henderson Fellowship, the 2019 Dr. Ok Kyun Kim Fellowship, and the 2020 John Bardeen Graduate Research Award from the Department of Electrical and Computer Engineering, UIUC.

Yansong Yang (Member, IEEE) received the B.S. degree in electrical and electronic engineering from the Huazhong University of Science and Technology, Wuhan, China, in 2014, and the M.S. and Ph.D. degrees in electrical engineering from the University of Illinois at Urbana–Champaign (UIUC), Urbana, IL, USA, in 2017 and 2019, respectively. He is currently a Postdoctoral Researcher with UIUC. His research interests include design and microfabrication techniques of RF microelectromechanical system (MEMS) resonators, filters, switches, and photonic integrated circuits. Dr. Yang won the Second Place in the Best Paper Competition at the 2018 IEEE International Microwave Symposium and the Best Paper Award at the 2019 IEEE International Ultrasonics Symposium. He was a finalist for the Best Paper Award at the 2018 IEEE International Frequency Control Symposium. He was a recipient of the 2019 P. D. Coleman Graduate Research Award from the Department of Electrical and Computer Engineering, UIUC.

Songbin Gong (Senior Member, IEEE) received the Ph.D. degree in electrical engineering from the University of Virginia, Charlottesville, VA, USA, in 2010. He is currently an Associate Professor and the Intel Alumni Fellow with the Department of Electrical and Computer Engineering and the Micro and Nanotechnology Laboratory, University of Illinois at Urbana–Champaign, Urbana, IL, USA. His research primarily focuses on the design and implementation of radio frequency microsystems, components, and subsystems for reconfigurable RF front ends. In addition, his research explores hybrid microsystems based on the integration of microelectromechanical system (MEMS) devices with photonics or circuits for signal processing and sensing. Dr. Gong is also a Technical Committee Member of the MTT-21 RF-MEMS of the IEEE Microwave Theory and Techniques Society and the International Frequency Control Symposium. He was a recipient of the 2014 Defense Advanced Research Projects Agency Young Faculty Award, the 2017 NASA Early Career Faculty Award, the 2019 UIUC College of Engineer Dean’s Award for Excellence in Research, and the 2019 Ultrasonics Early Career Investigator Award. Along with his students and postdocs, he received the Best Paper Awards from the 2017 and 2019 IEEE International Frequency Control Symposium and the 2018, 2019, and 2020 International Ultrasonics Symposium and won the Second Place and the Third Place in the Best Paper Competition at the 2018 and 2020 IEEE International Microwave Symposium. He also serves as the Chair of the MTT TC6 and an Associate Editor for T-UFFC, IEEE JOURNAL OF MICROELECTROMECHANICAL SYSTEMS, and IEEE JOURNAL OF MICROWAVES.