A Passive-Mixer-First Acoustic-Filtering Chipset Using Mixed-Domain Recombination

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Abstract—A mixer-first acoustic-filtering high-intermediate-frequency (IF) superheterodyne radio frequency (RF) front end is presented, which utilizes a mixed-domain recombination architecture. By having a set of commutated switches or essentially a passive mixer before fixed-frequency acoustic filters, mixer-first acoustic filtering enables a widely tunable RF for the front end while preserving the acoustic filter’s high-order filtering response and high linearity. Compared to the prior work that uses IF-only recombination, the proposed IF-and-baseband mixed-domain recombination supports a wider instantaneous bandwidth (BW) and higher RF while reducing the number of IF passive components that are lossy and bulky. A proof-of-concept chipset is demonstrated; it consists of an RF frontend N-path commutated-LC passive mixer and an IF in-phase and quadrature (I–Q)-mismatch-compensating complex receiver in 65-nm CMOS as well as two 2.6-GHz Qorvo QPQ1285 bulk-acoustic wave (BAW) filters. In measurement, the chipset operates across 3.5–6.5 GHz RF with a 160-MHz instantaneous BW, 10-dB noise figure (NF) at 3.5-GHz RF, and an out-of-band IIP3 of +27 dBm at 1 × BW offset.

Index Terms—Bulk-acoustic wave (BAW), filter, impedance aliasing, interference, linearity, mixer, receiver.

I. INTRODUCTION

Radiofrequency (RF) front-end designs are becoming increasingly challenging as a modern mobile device has to support many frequency bands with numerous front-end switches and acoustic filters [1]. The cost and size of next-generation RF front ends are further stressed by the trend toward multi-in–multi-out (MIMO), broadband, and dynamic spectrum access.

Many silicon-based integrated circuit designs for monolithic reconfigurable RF front ends have been reported as possible alternatives to numerous fixed-frequency acoustic filters. High-order N-path filters in CMOS provide acoustic-filter-like selectivity at close-in offset frequencies but have limited tuning range and rarely operate above 2 GHz due to lossy coupling networks, parasitic effects, and the need of multichip square-wave RF clocks [2], [3]. Q-enhanced LC-resonator-based RF bandpass filters have been demonstrated with wide frequency tuning ranges, but they suffer from elevated noise levels and degraded linearity as active components are utilized for achieving high-Q on chip [4]. Mixer-first direct-conversion or low-intermediate-frequency (IF) receivers are also widely tunable and have excellent out-of-band (OOB) linearity in the presence of blockers at far-out frequency offsets; however, they have limited suppression and linearity for close-in interference [5]–[9]. Finally, filtering-by-aliasing receivers have been reported with superior close-in interference suppression but are limited to sub-1-GHz RF operation even when implemented in advanced CMOS nodes [10].

By combining a passive or parametric mixer and high-order filters in a mixer-first high-IF superheterodyne architecture, several recent works have demonstrated a new direction for reconfigurable RF front ends [11]–[14]. The idea is straightforward (see Fig. 1). The mixer converts the frequency of an incoming signal into a fixed high IF. Then, an IF high-order filter with a sharp filtering response suppresses interference before connecting to active components that are prone to high-power interference. By varying the mixer local oscillator (LO) frequency, these mixer-first superheterodyne reconfigurable front ends [11], [12] provide widely tunable RF operations with a much smaller filter count compared to an exhaustive filter bank design (e.g., [1]).

Using parametric varactors modulation, low-noise and input-matched superheterodyne mixer-first RF front ends have been demonstrated in [11] and [13]. However, in a parametric mixer, both the LO and signal voltages are across the same two terminals of each varactor, resulting in spurious intermediate-signal signals that are removed using bulky isolators in [11]. Also, parametric varactors modulation is inherently nonlinear, deteriorating the mixer linearity performance, especially for low-cost CMOS implementations [15].

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Unlike the nonlinear varactor-based parametric converters, linear periodically time-varying (LPTV) N-path passive mixers using transistor switches are highly linear [16]. A gigahertz-high-IF passive-mixer-first RF front end using a surface-acoustic-wave (SAW) IF filter—essentially a passive-mixer-first acoustic-filtering front end—has been demonstrated with matched RF input, low noise, and high linearity in [12]. The key challenge associated with a passive-mixer-first acoustic-filtering superheterodyne front end is the impedance aliasing that arises from 1) a high-IF superheterodyne architecture and 2) nonnegligible OOB impedance from IF acoustic filters [14]. An LC-based impedance shaping network has been introduced in [12] to suppress the impedance aliasing, concurrently achieving input matching, low noise, and high linearity. However, the work in [12] and [14] uses many off-chip IF passive components and only has an RF bandwidth (BW) of 65 MHz.

In [17], we presented a mixer-first acoustic-filtering front end with a new IF-and-baseband mixed-domain recombination architecture. Compared to the prior work [12], [14] that uses IF-only recombination, the proposed mixed-domain recombination supports a wider instantaneous BW and higher RF while reducing the number of IF passive components that are lossy and bulky.

A conceptual comparison to a conventional filter-bank-based front end is shown in Fig. 2; a more detailed treatment on our proposed architecture is given in Section III. In a conventional multiband RF front end, a bank of acoustic filters with different center frequencies connects to an antenna via a static RF single-pole-multithrow switch and matching networks (MNs) [18], [19]. Each acoustic filter is then connected to a dedicated radio frequency integrated circuit (RFIC) low-noise amplifier (LNA). A multiplexer switch selects one of the LNAs for further signal processing [20]. This approach requires numerous different filters and lacks flexibility to incorporate future frequency bands after field deployment.

Our design effectively makes the RF switch periodically rotate among a bank of identical filters, making the input frequency programmable and jointly defined by the switch rotational or commutation frequency \( f_C \) and the filter center frequency \( f_{IF} \).

This article is an expanded version from [17]. An overview of bulk-acoustic wave (BAW) filter basics and a new simplified analysis of N-path commutated-LC filters are described in Section II. Section III introduces the mixer-first acoustic-filtering RF front end using mixed-domain recombination with additional analyses and discussions on image rejection, frequency planning, IF choice, and noise performance. More implementation and simulation details are reported in Section IV. Finally, updated and additional measurement results with more detailed discussions are presented in Section V.

II. LPTV MIXER-FIRST ACOUSTIC-FILTERING FRONT END

We start with a brief review of the key characteristics and challenges associated with acoustic filters. Understanding of these plays a pivotal role in engineering an RF front-end architecture with the desired system-level performance. Then, we introduce LPTV passive-mixer-first acoustic-filtering front ends and its architecture evolution using a new energy-conservation-based analysis for switched-bandpass-filter circuits.

A. RF Acoustic Filters

Acoustic wave propagation in common RF materials has orders of magnitude smaller wavelength and lower loss compared to those in their electromagnetic counterparts [21]. Due to these fundamental advantages, acoustic filters have low loss, high selectivity, and compact form factors, resulting in their pervasiveness in modern commodity mobile devices.

A high-frequency acoustic filter often consists of several BAW resonators, while surface-acoustic wave (SAW) resonators are typically deployed for low-frequency (e.g., below 2 GHz) applications [18], [22]. Given that the fractional BWs of acoustic filters are fundamentally limited by the efficiency of their electromechanical transduction [21], [22], we focus on BAW filters in this article as their high-frequency operation results in wide instantaneous BWs.

A simplified BAW resonator structure is shown in Fig. 3(a), consisting of a layer of piezoelectric material sandwiched between two metal electrodes [18]. The resonance frequency is determined by the thickness of the piezoelectric material layer and the thickness and mass of the electrodes. The electrical behavior of a BAW resonator can be approximately captured by an equivalent circuit using the Butterworth–Van-Dyke (BVD) model [18], [19], [23] as shown in Fig. 3(b) where \( C_0 \) is the static capacitance and \( L_D \) and \( C_D \) are dynamic (also called motional) inductor and capacitor, respectively.

By connecting several BAW resonators in a ladder topology, an acoustic filter can be constructed, as shown in Fig. 3. Although on the same wafer and hence have the same thickness of the piezoelectric material layer, the shunt resonators have lower resonance frequencies compared to the series...
ones by introducing a mass loading layer [22]. Following the principles summarized in [22], we build an acoustic filter made of BVD-based resonators centering at 2.6 GHz with a BW of 200 MHz [see Fig. 3(c)]. The simulated filter $S$-parameters are shown in Fig. 3(d) with nearly zero insertion loss, excellent close-in and OOB rejection, and good input–output return loss.

Despite their pervasiveness, existing acoustic filter technologies face a couple of key challenges for future high-frequency and broadband RF applications. First, acoustic filters generally cannot be tuned across a wide frequency range and have somewhat fixed and predefined operation frequencies that are defined by the thickness and mass of the building materials. As more frequency bands are set to become available, e.g., the advent of sub-6-GHz 5G and Wi-Fi 6, a whopping 100 filters are expected in a next-generation mobile device [19]; this imposes significant challenges on RF front ends in terms of cost, size, and design complexity. Second, acoustic filters often provide superior performance only up to 3 GHz. Scaling acoustic filters beyond 3 GHz faces many fundamental challenges as the thicknesses of acoustic structures become too small at high frequencies to be manufactured in a low-loss and low-cost fashion [21], [24].

B. Mixer-First Acoustic-Filtering RF Front-End

Mixer-first high-IF acoustic-filtering front ends address the aforementioned key challenges faced by acoustic filters.

As shown in Fig. 1, the input RF of a mixer-first acoustic-filtering front end is jointly defined by the LO frequency and the IF acoustic filter center frequency. By varying the LO frequency, the front-end RF can be made widely tunable with fixed-frequency acoustic filters. Moreover, a mixer-first acoustic-filtering front end allows relatively low-frequency acoustic filters being used at high-frequency bands as signals are frequency downconverted prior to entering the filters.

Mixer-first high-IF acoustic-filtering front ends are distinguishable from monolithic mixer-first direct-conversion receivers [5]–[9] by their IF choices [14], [17].

In a mixer-first direct-conversion receiver, a low or zero IF is used for a high integration level and low cost. For sub-6-GHz frequency bands, the upper frequencies of IF passbands are less than a couple of hundred megahertz. This relatively low upper frequency limit would make LC-based IF filters unacceptably bulky, and acoustic filters in this frequency range face very narrow BW (tens of kilohertz to a few megahertz, e.g., [25], [26]). Therefore, active RC-based baseband filters are often utilized in sub-6-GHz mixer-first direct-conversion receivers [5]–[9]. However, active filters limit the front-end frequency selectivity at RF and the linearity at close-in frequency offsets. Also, they can be power hungry with high-order filtering and broadband BWs [5].

In a mixer-first acoustic-filtering front-end, a high IF allows compact all-passive high-order filters with wide instantaneous bandwidth and low upper frequency limit would make acoustic filters beyond 3 GHz faces many fundamental challenges. As the thicknesses of acoustic structures become too small at high frequencies to be manufactured in a low-loss and low-cost fashion [21], [24].

C. Evolution and Analysis of Switched-BPF Circuits

Here, we present the evolution of mixer-first acoustic-filtering front ends using a new energy-conservation-based analysis for switched-bandpass-filter circuits.

A mixer-first acoustic-filtering front end can be simply constructed by having a double-balanced passive mixer in front of an IF filter [see Fig. 4(a)]. We use a second-order $RLC$ bandpass filter first and replace it with acoustic filters later.

Let us derive the conversion gain and the input impedance of a double-balanced RF mixer with a narrowband $RLC$ load where the mixer LO or clock frequency $f_C$ is significantly larger than the $RLC$ filter 3-dB BW. While double-balanced RF mixers have been studied extensively, most existing analyses have assumed either a purely resistive or an $RC$ load [28], [29].

Given a sinusoid source voltage $v_S = V_S \sin(\omega_S t + \phi)$, where $\omega_S$ is the sum of the mixer LO frequency $\omega_C$ and the $LC$ resonance frequency $\omega_O$, we find that the output voltage can be approximated as $v_O = V_O \sin(\omega_O t + \phi + \theta)$, i.e., a sinusoid with a constant amplitude $V_O$ and phase shift $\theta$. Intuitively, the output is a sinusoid as the high-$Q$ $RLC$ tank suppresses all the LO harmonics except at the resonance frequency $\omega_O = (LC)^{-\frac{1}{2}}$. In the steady state, the output amplitude can be approximated as a constant since the time constant of our high-$Q$ $RLC$ tank $2\pi RC$ is much larger than the LO period $1/f_C$.

To sustain a constant envelope sinusoid, the energy dissipated by the lossy $RLC$ tank and the source resistance has to be replenished by the voltage source over time. This results in

$$V_O^2 \frac{2MT_C}{R_L} = \int_0^{MT_C} \frac{\omega_O - \omega_S - \omega_{IN}}{R_S} V_{IN} \, dt$$

(1)
where \( M \) is the number of mixer LO periods, \( T_C = 1/f_C \), \( v_{IN} = v_{O}(t)sw(t) \), \( sw(t) \) is the square-wave LO waveform given in Fig. 4(a), and \( R'_S = R_S + 2R_{SW} \) is the sum of source and switches resistances.

Substituting \( v_{IN} = v_{O}(t)sw(t) = V_O \sin(\omega_{0}t + \phi + \theta) \cdot sw(t) \) and \( v_S = V_S \sin(\omega_{ST} + \phi) \) into (1) and letting \( M \to \infty \), we arrive at the double-balanced mixer conversion gain as

\[
CG_{DB} = \left| \frac{V_{O}(\omega_{0})}{V_s(\omega_{ST})} \right| = \frac{2\pi R_L}{\pi^2 R'_S + 8\alpha R_L} \tag{2}
\]

where \( \alpha = \sum_{h=1}^{H} 1/(2h - 1)^2 \) and \( H \) is the number of LO harmonics. It has been shown that given a high-\( Q \) \( RLC \) load with negligible load impedance at LO harmonics \( f_{O} \pm k f_C \) where \( k \) is a nonzero integer, the RF input impedance \( R_{in, DB} \) of the circuit in Fig. 4(a) is purely resistive [14], [30]. With that, \( \theta \) can be determined as \( \pi/2 \). Meanwhile, the RF input impedance can be expressed as

\[
R_{in, DB} = V_{in}(\omega_{ST}) = \frac{R'_S R_L}{\pi^2 R'_S/4 + (2\alpha - 1) R_L} + 2R_{SW}. \tag{3}
\]

A design example is used to verify (2) and (3) with \( R_S = 50 \Omega \), \( R_{SW} = 3 \Omega \), \( f_C = 2 \text{ GHz} \), \( f_{O} = 1/(2\pi \sqrt{LC}) = 2.6 \text{ GHz} \), and \( R_LLC \) tank \( Q \) of 30, and varying \( R_L \). The calculated and Cadence Spectre RF simulated conversion gains at 2.6-GHz IF and input impedance at 4.6-GHz RF are plotted in Fig. 4(a), showing a good match. Power conversion gain, defined as the ratio between the power delivered to \( R_L \) at \( f_o \) and the maximum available power from the source, can be readily calculated using (2). The simulated input impedance imaginary parts are much smaller compared to its real parts and hence are omitted in Fig. 4(a).

From (2) and (3), we know that a double-balanced passive mixer cannot achieve input matching and low power loss at the same time. As in Fig. 4(a), the input impedance increases with \( R_L \), and \( R_L \) of 800 \( \Omega \) results in an RF input impedance of 42 \( \Omega \). However, this results in a \(-10\text{-dB} \) power loss. Alternatively, a 50-\( \Omega \) input impedance can be obtained by increasing \( R_{SW} \), but this increases power loss as well as more power will be dissipated by \( R_{SW} \).

An \( N \)-path switched-\( LC \) circuit breaks the matching-loss tradeoff as the multiple paths result in reduced harmonic components at RF [14], similar to that in an \( N \)-path switched-\( RC \) circuit [32], [30], [31]. Here, conservation of energy is applied to the four-path switched-\( LC \) circuit in Fig. 4(b), resulting in

\[
\frac{V_O^2}{2R_L}MT_C = \sum_{m=0}^{M-1} \int_{mT_C+(k-1)T_c/4}^{mT_C+kT_C/4} \frac{v_S - v_{O,k}}{R'_S} \, dv \tag{4}
\]

where \( k \) corresponds to the \( k \)th path of the four-path switched-\( LC \) circuit in Fig. 4(b).

Substituting \( v_{O,k} = V_O \sin[\omega_{0}t + \phi + \theta + (k-1)\pi/2] \) and \( v_S = V_S \sin(\omega_{ST} + \phi) \) into (4) and letting \( M \to \infty \), we arrive at the conversion gain and the input impedance of the four-path switched-\( LC \) circuit as

\[
CG_{4path} = \left| \frac{V_{O}(\omega_{0})}{V_s(\omega_{ST})} \right| = \frac{2\sqrt{2}}{\pi} \frac{R_L}{R_L + 4R'_S} \tag{5}
\]

\[
R_{in,4path} = R_{SW} + R_{sh}(|\gamma R_L|) \tag{6}
\]

where \( R_{sh} = R'_S(N\gamma / (1-N\gamma)) \), \( \gamma = (\text{sinc}^2(\pi/N))/N \), and we have assumed that \( 2f_{IF} \) does not equal integral multiples of \( f_C \) and let \( \theta = \pi/4 \) for resistive input impedance. The generalized analysis in [14] yields the same results when loaded with high-\( Q \) \( RLC \) tanks.

Fig. 4(b) shows a design example with \( R_S = 50 \Omega \), \( R_{SW} = 5 \Omega \), \( f_C = 2 \text{ GHz} \), \( f_{O} = 1/(2\pi \sqrt{LC}) = 2.6 \text{ GHz} \), \( R_LLC \) tank \( Q \) of 10, and a varying \( R_L \). The calculated and simulated conversion gains at 2.6-GHz IF and input impedance at 4.6-GHz RF are plotted in Fig. 4(b), showing a good match. Power conversion gain, the ratio between the power delivered to all load resistors at \( f_o \) and the maximum available power from the source, can be readily calculated using (5). The simulated input impedance imaginary parts are much smaller compared to its real parts and hence are omitted in Fig. 4(b).

As shown in Fig. 4(b), a matched input impedance of around 50 \( \Omega \) and a low power loss of 1.4 \( \text{dB} \) can be achieved simultaneously in a four-path switched-\( LC \) circuit with \( R_{SW} \) of 5 \( \Omega \) and \( R_L \) of 250 \( \Omega \).

Next, let us replace the \( LC \) filter in Fig. 4(b) with the 2.6-GHz BVD-based BAW filter given in Fig. 3(c). The switched-BAW-filters circuit is shown in Fig. 5(a) with \( R_S = 50 \Omega \), \( R_{SW} = 5 \Omega \), \( f_C = 2 \text{ GHz} \), \( f_{O} = 1/(2\pi \sqrt{LC}) = 2.6 \text{ GHz} \), and ideal 2:2:1 transformers to boost the in-band impedance from 50 to 250 \( \Omega \).

However, directly replacing the \( LC \) filter with the BAW filter results in poor RF input matching of around \(-5 \text{ dB} \) and
excess power loss of 7 dB, as shown in Fig. 5(a). We attributed the sharp performance degradation to impedance aliasing as detailed in [14]. Compared to a second-order RLC filter, a high-order BAW filter together with its interchip connections have significant impedance at OOB LO harmonic frequencies. Through the switching-based mixing operation, the load impedance $Z_L$ components at the LO harmonic frequencies are all translated to RF, becoming indistinguishable, or aliases of one another. Therefore, impedance aliasing degrades input matching and introduces excess loss; this is especially prominent with a wide RF range as a large number of RF impedances can be aliased.

We found in [12] and [14] that it is possible to mitigate impedance aliasing by suppressing the $Z_L$ at OOB LO harmonic frequencies via an impedance shaper, as shown in Fig. 5(b). The shunt LC circuit with a tank $Q$ of 10 provides low impedance path at OOB LO harmonic frequencies, restoring the RF input matching and reducing the power loss from 7 to 2.5 dB. The simulated input reflection and power loss are slightly worse compared to those predicted by (5) and (6) due to the finite impedance suppression provided by the LC impedance shaper.

### III. Mixer-First Acoustic Filtering Using Mixed-Domain Recombination

Despite favorable input matching and power loss, the mixer-first acoustic-filtering front end in Fig. 5(b) requires four IF paths and lacks analog-domain image rejection. Each IF path consists of an acoustic filter and an IF receiver. Having too many IF paths increases the system cost, size, and power consumption. Without analog-domain image rejection, an image-band blocker could stress the receiver baseband and analog-to-digital converter (ADC) dynamic range requirements, again, adding cost, size, and power consumption.

To reduce the IF paths count and suppress image-band interference, an IF recombination network was introduced in [12] (see Fig. 6). With the IF recombination, the filtering front end resembles a Hartley image rejection receiver, which needs a 90° phase shift and a signal summation before the IF filter [28], [29]. For high linearity, a lumped CLC phase shifter and a transformer balun were used, as shown in Fig. 6.

However, the mixer-first acoustic filtering with the IF-only recombination in Fig. 6 comes with two drawbacks.

First, the usage of many lossy IF passive components degrades the sensitivity of the front end. To reduce the loss of passives, a low IF is preferred as electromagnetic-induced losses typically increase with frequency. However, a low $f_{IF}$ results in narrow acoustic filter BW $f_{BW} = k_{FBW} \times f_{IF}$ as the fractional BW $k_{FBW}$ is fundamentally limited by the electromechanical coupling [22], [33]. Also, a low IF reduces image-band blocker filtering, as detailed in Section III-C.

Another drawback lies in the capability of coping with in-phase and quadrature ($I-Q$) imbalance, which needs to be kept small for high image rejection. Tuning IF passive components can compensate for the $I-Q$ imbalance, but it comes with significant loss penalty, especially with inductance tuning.

#### A. Architecture

In this work, we propose a new IF-and-baseband mixed-domain recombination architecture for mixer-first acoustic fil-
tering, as shown in Fig. 7. Following a four-path switched-\(LC\) passive mixer, there are two on-chip transformer baluns. Each I- or Q-path balun acts as the \(LC\) impedance shaper inductor and the first-stage IF recombination that merges four paths into 2, halving the IF filter and receiver count. After the IF BAW filters suppressing OOB interference, two IF quadrature direct-conversion receivers are adopted. The \(I-Q\) baseband outputs of the IF receivers are connected, acting as the second-stage recombination at baseband.

Quantitatively, the RF front end operates as follows.

Modeling an incoming desired signal as a sinusoid \(V_S \cos(\omega_d t)\) and the RF front-end mixer differential LOs as \(\cos(\omega_{LO1} t + \phi_{LO1})\) and \(\sin(\omega_{LO1} t + \phi_{LO1})\), the outputs at IF after the RF mixer and IF recombination can be expressed as

\[\begin{align*}
x_I &= g_1 V_S \cos(\omega_{LO1} t - \phi_{LO1}) \quad \text{and} \\
x_Q &= -g_1 V_S \sin(\omega_{LO1} t - \phi_{LO1}),
\end{align*}\]

where \(\omega_{RF} = \omega_d - \omega_{LO1}\) and \(g_1\) is the RF front-end conversion gain.

Since the IF receivers and the baseband recombination in Fig. 7 form a complex mixer [34], let us define \(x = x_I + j x_Q\), \(y = y_I + j y_Q\), and \(w = w_I + j w_Q\), where \(y_I\) and \(y_Q\) are the outputs after the IF BAW filters and \(w_I\) and \(w_Q\) are the IF mixer \(I-Q\) LOs. In this way, the final outputs after the baseband recombination can be written as \(z = z_I + j z_Q = g_2 y \cdot w\), where \(g_2\) is the IF receiver conversion gain.

\[
given x = g_1 V_S e^{-j(\omega_{LO1} t - \phi_{LO1})}, \quad we \ have \ y = g_1 g_F (\omega_{RF}) V_S e^{-j(\omega_{LO1} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))}, \quad \text{where} \ g_F(\omega_{RF}) \ \text{and} \ \phi_{RF}(\omega_{RF}) \ \text{are the magnitude and phase responses of the BAW filter at} \ \omega_{RF}, \ \text{respectively.}
\]

Finally, assuming the IF mixer LOs as \(w = w_I + j w_Q = e^{j(\omega_{LO1} t + \phi_{LO1})}\), the final complex output is

\[
z_S = g_2 y \cdot w = g_1 g_F (\omega_{RF}) g_2 V_S e^{j(\omega_{RF} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))}
\]

where \(\omega_{RF} = \omega_{LO2} - \omega_{RF} \approx 0\) and \(\phi_{LO1} = \phi_{LO1} + \phi_{RF} = \phi_{RF} / Delta 1\).

From (7), we see that the desired incoming signal at \(\omega_S\) is received at baseband, while a strong close-in blocker is suppressed significantly due to the high-frequency selectivity in BAW filter gain \(g_F\).

If an image-band interference \(V_{IM} \cos(\omega_{IM} t)\) enters the front end, where \(f_{IM} = f_{LO1} - (f_S - f_{RF}) = 2 f_{LO1} - f_{LO1}\), the complex output after the IF filters becomes \(y_{IM} = g_1 g_F V_{IM} e^{j(\omega_{LO1} t + \phi_{LO1} + \phi_{RF})}\). The output at the receiver baseband is

\[
z_{IM} = g_2 y_{IM} \cdot w = g_1 g_F (\omega_{RF}) g_2 V_{IM} e^{j(2 \omega_{RF} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))},
\]

This means that the image signal is frequency translated to a much higher frequency \((2 \omega_{RF} \gg \omega_{BB})\) and can be subsequently filtered along the receiver baseband chain.

When compared with the prior IF-only recombination design in Fig. 6, the new mixed-domain architecture pushes the 90\(^\circ\) phase shift and final stage recombination from the IF signal path to IF receiver LO path and baseband, respectively.

By having LO-path phase shift and baseband recombination after the IF receiver LNA, their associated noise penalty is minimized. Also, compared to the intrinsically narrowband signal-path CLC phase shifter, LO-path 90\(^\circ\) phase shifting is broadband and readily available in high-performance IQ receivers.

Compared to the IF-only recombination design, it is less lossy to incorporate a higher IF in the proposed architecture as the mixed-domain recombination eliminates the IF CLC phase shifter and final stage IF transformer balun. This high IF enables a wider acoustic filter BW as it is proportional to its center frequency. A high IF also eases the filtering of image-band blocker, as detailed in Section III-C [35].

Finally, having two IF receiving paths in the mixed-domain recombination architecture allows one to trade dc power with improved noise figures (NFs), as discussed in Section III-D.

The benefits of our proposed architecture come at the expense of requiring one more IF acoustic filter and receiver. However, since only two identical acoustic filters are needed, they can be fabricated together using the same process and hence have a significantly lower cost compared to having two acoustic filters at different frequencies [1]. In fact, it is essential to use two adjacent acoustic filters on the same die to reduce the \(I-Q\) mismatch in the proposed mixed-domain recombination architecture, as discussed shortly in Section III-B. The additional IF receiver does consume more power and chip area, but it reduces the front-end NF (see Section III-D) and modern inductorless receivers in nanoscale CMOS processes are compact and power efficient.

Our proposed architecture resembles a Weaver image rejection receiver but has two distinctions compared to prior works (e.g., [36]). First, eliminating the RF LNA and having a mixer-first design significantly enhances the front-end dynamic range. Second, the choice of a gigahertz IF allows us to use high-linearity passive acoustic filters to replace active filters.

### B. \(I-Q\) Mismatch and Image Rejection Compensation

One challenge in our architecture is the \(I-Q\) mismatch, which leads to degraded image rejection. While it has been shown that image rejection can be obtained in the digital domain [37], an image-band blocker could stress the dynamic range requirement of the analog front end and saturate the receiver.

Let us recalculate the image-band response in the presence of \(I-Q\) mismatches. Assume that the \(I-Q\) mismatch is dominated by the BAW filters, including the BAW-RFIC interconnects, and the \(I-Q\) BAW filters have magnitude and phase responses of \([1 \pm \Delta g(f)] g_F(f)\) and \(\phi_F(f) \pm \Delta \phi_F(f)\), respectively. The receiver normalized complex baseband output can be found as

\[
\frac{z_{IM,\text{mis}}}{g_1 g_F (\omega_{RF}) g_2 V_{IM}} \approx e^{j(2 \omega_{RF} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))} - \Delta g(\omega_{RF}) \cdot e^{j(\omega_{RF} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))}
\]

\[+j \Delta \phi_F(\omega_{RF}) \cdot e^{j(\omega_{RF} t + \phi_{LO1} + \phi_{RF}(\omega_{RF}))}
\]
in Fig. 8, the worst case gain and phase mismatch of 0.2 dB and 7° are observed, respectively. This mismatch results in 24-dB image rejection based on (9).

To suppress the spurious tones in (9) due to $I–Q$ mismatches, we introduce $I–Q$ mismatch compensation circuitry at baseband akin to that in [36]. As shown in Fig. 9, after each IF receiver mixer, there is a vector modulator, $VM_{ij}$, where $i, j = I, Q$. Each vector modulator consists of a main input and an auxiliary input. The signal for the main input is first directly fed to the vector modulator output with unity gain, corresponding to the same component from that (see Fig. 7) without $I–Q$ mismatch compensation. In addition, each vector modulator imparts adjustments to its main and auxiliary path magnitudes for $I–Q$ mismatch compensation, through weights $M_{ij}$ and $A_{ij}$, respectively.

Assuming that the IF receiver LO $I–Q$ mismatch is negligible after calibration, the vector modulator $I–Q$ mismatch compensation conditions at the $I$ path are

$$
\sqrt{M_{II}^2 + A_{II}^2} = \frac{1}{1 + \Delta g}, \quad \arctan \frac{M_{II}}{A_{II}} = -\Delta \phi_F
$$

$$
\sqrt{M_{QQ}^2 + A_{QQ}^2} = \frac{1}{1 - \Delta g}, \quad \arctan \frac{M_{QQ}}{A_{QQ}} = \Delta \phi_F. \tag{10}
$$

$I–Q$ mismatch at the $Q$ path can be similarly compensated using vector modulators $VM_{IQ}$ and $VM_{QI}$.

Satisfying (10) across a wide instantaneous BW is challenging due to the frequency selectivity difference between the vector modulators and the $I–Q$ mismatches $\Delta g$ and $\Delta \phi_F$. This challenge associated with wideband $I–Q$ mismatch compensation is known (e.g., see [35], [36]) and similar to the challenge of wideband self-interference cancellation using frequency-flat vector modulators [38], [39]. Using two adjacent acoustic filters from the same die should reduce $I–Q$ mismatches, increasing the image rejection BW. Also, replacing frequency-flat vector modulators with multitap analog filters should also increase the image rejection BW as demonstrated in broadband interference cancellation [40], [41].

Fortunately, in addition to widening the instantaneous BW, a high IF allows us to use a fixed-frequency high-pass filter to provide additional image filtering, as discussed in Section III-C.

C. Image Filtering, Frequency Planning, and IF Choice

The high-IF architecture of the proposed widely tunable filtering front end allows us to insert a fixed-frequency RF high-pass filter for additional image suppression. Given an IF and an image filtering requirement, we can determine the front-end RF range.

We start with a high IF that satisfies the instantaneous BW requirement. With an IF of 2.6 GHz, a 6.5% filter fractional BW results in an instantaneous BW of 170 MHz.

The proposed widely tunable RF front end with an input fixed-frequency high-pass filter is shown in Fig. 10. As shown in Fig. 10(b), when $f_{LO1} < f_F$, the signal and image bands are located at $f_S = f_{LO1} + f_F$ and $f_{IM} = f_F - f_{LO1}$, respectively. On the other hand, when $f_{LO1} \geq f_F$, the signal remains at $f_S = f_{LO1} + f_F$, but the image band changes to $f_{IM} = f_{LO1} - f_F$.

To ease the RF high-pass filter design requirements, the separation between the lowest signal frequency $f_{S,L}$ and the highest image frequency $f_{IM,H}$ should be maximized. If the lowest signal frequency is given as $f_{S,L} = f_{LO1,L} + f_F$, where $f_{LO1,L}$ is the lowest front-end mixer LO frequency, the highest image frequency is $f_{IM,H} = f_{LO1,L} - f_F = 2f_F - f_{S,L}$. When front-end mixer operates with an LO frequency of $f_{LO1,L}$ (the signal and image interference at $f_{S,L}$ and $f_{IM,H}$, respectively),
we can write the worst case image filtering ratio (IFR) as

$$\text{IFR}_{\text{min}} = \alpha \cdot 20 \log_{10} \frac{f_{S.L}}{2f_{\text{IF}} - f_{S.L}} \quad (11)$$

where $\alpha$ is the order of the RF high-pass filter. Given $\alpha = 5$ and an IFR$_{\text{min}}$ of 30 dB, we have $f_{S.L} = (4/3)f_{\text{IF}}$. Given $f_{\text{IF}} = 2.6$ GHz, $f_{S.L} = (4/3)f_{\text{IF}} = 3.47$ GHz and $f_{\text{IM,H}} = 2f_{\text{IF}} - f_{S.L} = 1.73$ GHz.

Since a fixed-frequency high-pass filter is used, the highest signal frequency $f_{S.H}$ is set by $f_{S.H} - 2f_{\text{IF}} = 2f_{\text{IF}} - f_{S.L}$ to retain the worst case IFR IFR$_{\text{min}}$ in (11). This gives us $f_{S.H} = (8/3)f_{\text{IF}}$. Given $f_{\text{IF}} = 2.6$ GHz, $f_{S.H} = (8/3)f_{\text{IF}} = 6.94$ GHz.

Like other high-IF superheterodyne receivers, LO feedthrough can potentially saturate the subsequent IF receivers [35]. Specifically, the RF LO can leak to the IF filters, and if the RF LO is inside the IF filter passband, it can reach the IF receiver causing possible saturation. The most significant LO feedthrough happens when $f_{\text{LO1}} = f_{\text{IF}}$, corresponding to a 5.2-GHz RF. Symmetrical mixer designs and an LO leakage cancellation circuitry (e.g., [42]) can be used to reduce the LO feedthrough.

The RF LO also can leak to the antenna port similar to mixer-first direct-conversion receivers. Unlike a direct-conversion mixer-first receiver, the LO leakage can be filtered by the input high-pass filter in our high-IF mixer-first superheterodyne front end. For the RF range of 3.5–6.1 GHz, the front-end mixer LO frequency is from 0.9 to 3.5 GHz. This $f_{\text{LO1}}$ range is outside of the high-pass filter passband, and the corresponding LO leakage will be suppressed.

Finally, let us summarize various tradeoffs associated with choosing an IF. A higher IF results in a higher RF and a larger RF range as $f_{S.L} = (4/3)f_{\text{IF}}$ and $f_{S.H} = (8/3)f_{\text{IF}}$. Also, a higher IF provides a wider instantaneous BW $f_{\text{BW}} = k_{\text{FBW}} \times f_{\text{IF}}$ as the acoustic filter fractional BW $k_{\text{FBW}}$ is limited by the electromechanical coupling [22], [33]. However, a higher IF may result in degraded acoustic filter performance as acoustic filters are generally more challenging to design and hence often have more insertion loss and/or less OOB rejection at higher frequencies, as discussed in Section II-A.

**D. Noise Analysis**

Here, we calculate the NF of the mixer-first acoustic-filtering front end with proposed mixed-domain recombination. We ignore all reactive components for our in-band noise analysis for simplicity.

Looking at Fig. 7, first, the total output noise at the I or Q path is $4K_{\text{fold}}^2g_1^2g_F^2V_{n,RS}^2$, where $g_1$, $g_F$, and $g_2$ are (conversion) gain of the front-end mixer, IF filter, and IF receiver, respectively, as used in Section III-A, and $K_{\text{fold}} \approx 1.13$ is a constant that factors in the noise folding effect of 25% duty-cycle four-path mixing [43]. The factor 4 is due to the fact that the summing noises at the receiver baseband are correlated as they both originate from the same source noise $V_{n,RS}^2$.

Second, the contribution of the IF receiver can be determined. The IF receiver consists of two paths that combine at the baseband output, and each path has its own LNA, mixer, and baseband transconductance cells, as shown in Fig. 7. Given the NF of a standalone single path to be $F_{\text{RX}}$, the total output noise at the I or Q output due to the entire IF receiver is $2(F_{\text{RX}} - 1)g_2^2V_{n,\text{IF}}^2$, where $g_2$ is the conversion gain of the standalone single-path receiver as in Section III-A and $V_{n,\text{IF}}^2$ is the source noise of $F_{\text{RX}}$. Since the thermal noise of an linear time-invariant (LTI) passive network can be calculated from its impedance [29], we have $V_{n,\text{IF}}^2 = 4kT\text{IF}$, where $R_{\text{IF}}$ is the impedance looking into the IF filter from the IF receiver input port. We have ignored noise folding for the IF receiver assuming that the IF filters have significantly suppressed the OOB noise. Unlike the source noise, the noises of the two IF receiver paths are uncorrelated, resulting in the factor of 2 in the total output noise.

Next, we will determine the noise contribution of the lossy RF front-end mixer and IF filters. The noise seen at the I or Q filter output is $V_{n,\text{IF}}^2 = K_{\text{fold}}g_1^2g_FV_{n,RS} + V_{n,\text{RMF}}^2$, where $V_{n,RS}$ is the RF source noise and $V_{n,\text{RMF}}^2$ is the thermal noise from the RF mixer and IF filter. Since $V_{n,RS}$ and $V_{n,\text{RMF}}^2$ are uncorrelated, we arrive at $V_{n,\text{RMF}}^2 = 4kT(R_{\text{IF}} - K_{\text{fold}}g_1^2g_F^2)$. Therefore, the total output noise due to the lossy RF front-end mixer and IF filters is $2g_2^2V_{n,\text{RMF}}^2$.

Finally, the total chipset noise factor $F_{\text{Total}}$ can be expressed as

$$F_{\text{Total}} = K_{\text{fold}}^2 + \frac{(F_{\text{RX}} - 1)g_2^2V_{n,\text{IF}}^2 + g_2^2V_{n,\text{RMF}}^2}{2g_2^2g_F^2g_1^2V_{n,RS}^2}$$

$$= \frac{1}{2} \left( K_{\text{fold}}^2 + \frac{F_{\text{RX}} R_{\text{IF}}}{g_1^2g_F^2 R_S} \right)$$

(12)

From (12), we see that the noise is halved, which is another benefit of having dual I-and-Q IF paths besides eliminating the lossy IF components in Fig. 6. This noise benefit is at the expense of increased dc power consumption using two IF receiving paths. In addition, interestingly, (12) tells us that reducing $R_{\text{IF}}$ with respect to $R_S$ can also reduce noise. The usage of I–Q mismatch compensation in Fig. 9 is expected to have a negligible impact on the NF as the IF receiver NF is dominated by its LNAs. In Section V-A, we compare simulated and measured NFs with those predicted by (12), showing good matches.

**E. Summary and Comparison to the IF-Only Recombination Architecture in [12] and [14]**

Comparing to the IF-only recombination in [14], the distribution of the multipath recombination to IF-and-baseband provides several key advantages, including a higher RF, wider instantaneous BWs, lower noise, and fewer IF LC components.

Having fewer IF LC components in the proposed mixed-domain recombination architecture (comparing Figs. 6 and 7) allows one to utilize a significantly higher IF, 2.6 GHz here compared to 1.6 GHz in [14], without suffering from the complexity and loss associated with the excessive LC components at IF.
As discussed in Section III-C, a higher IF $f_{IF}$ results in a higher RF and a larger RF range as $f_{S.L} = (4/3)f_{IF}$ and $f_{S.H} = (8/3)f_{IF}$, where $f_{S.L}$ and $f_{S.H}$ are the RF range lower and upper bound, respectively.

Also, a higher IF provides a wider instantaneous BW $f_{BW} = k_{FBW} \times f_{IF}$ as the acoustic filter fractional BW $k_{FBW}$ is fundamentally limited by the electromechanical coupling [22], [33].

Moreover, as detailed in Section III-B, unlike the narrowband IF-only image rejection, the proposed mixed-domain recombination architecture allows image rejection across a wider instantaneous BW due to the complex signal processing capability at analog baseband.

Finally, as quantified in Section III-D and (12), the proposed mixed-domain recombination architecture has a lower system NF. This low noise is achieved at the expense of having two IF receiving paths, instead of one in the IF-only recombination architecture, as discussed in Section III-A.

IV. CIRCUIT IMPLEMENTATIONS

We devised a proof-of-concept prototype of the proposed mixer-first acoustic-filtering front end with mixed-domain recombination using a 65-nm CMOS process and commodity BAW filters. The block diagram and schematic of the RF front-end mixer and IF receiver chipset two 2.6-GHz Qorvo QPQ1285 BAW filters are shown in Fig. 11.

A. RF Passive Mixer With Asymmetric IF Transformer Balun

Similar to many mixer-first receivers and N-path filters [2], [5], [8], a differential architecture is utilized for the RF mixer front end. A wideband 1:1 off-the-shelf transformer is served as a balun at the RF input to facilitate single-ended measurements. Also, the differential implementation reduces the source impedance seen by the mixer-first front-end, relaxing the impedance step-up transformation requirement [14].

Mixer switches are designed to have an ON-resistance of 8 Ω for a balance between front-end power conversion loss and LO path dc power consumption. Like in our prior work [14], the mixer switches are realized using deep-n-well transistors, allowing us to use bootstrapping resistors at their bulk nodes, as shown in Fig. 11(a). An on-chip divide-by-two circuitry is used to generate the 25% duty-cycle clocks that drive the mixer switches. Ac coupling capacitors are utilized at the RF input, which also acts as a first-order high-pass image filter.

While simultaneously acting as parts of the LC impedance shaper and the IF recombination, the on-chip transformer baluns could introduce significant loss in practice, desensitizing the receiver front end. We have derived the transformer loss or efficiency analytically using a simplified model as in Fig. 12. The power efficiency defined as the ratio between the input and output power shown in Fig. 12 can be calculated

$$\eta = \frac{P_{out}}{P_{in}} = \frac{V_{out}^{2}}{V_{in}^{2}} \times \eta_{k}$$

where $\eta_{k}$ is the transformer efficiency and $\eta_{k}$ is calculated by the following equation

$$\eta_{k} = \frac{1}{1 + \frac{\omega_{IF}L_{2}}{Q_{i}R_{IF}} + \frac{1}{k^{2}Q_{1}(\omega_{IF}L_{1})}}$$

where $Q_{i} = (\omega_{IF}L_{i})/R_{i}$, $i = 1$ or 2, and we have assumed that $C_{i}$ resonates with the inductive component at IF.

Based on (13), a low-loss or high-efficiency transformer requires high $Q$ and coupling factors. However, there exists a tradeoff between $Q$ and coupling factors in integrated transformers. A coplanar transformer features high $Q$ but has limited coupling. A stacked transformer has strong coupling but uses a lower thin metal layer, degrading the $Q$-factor.

Interestingly, based on (13), we find that the transformer efficiency is asymmetrical between the primary and secondary winding $Q$-factors, and the efficiency is mostly determined by the primary $Q_{i}$ (see Fig. 12).

Based on this insight, we adopt a 2:2 stacked transformer achieving a high coupling around 0.9. The top thick metal is
assigned to the primary resulting in $Q$ of 13.8 with 0.8-dB loss at 2.6-GHz IF, while the secondary has $Q$ of 5.6 and a loss of 0.3 dB. Both the primary and secondary windings of IF transformers have an inductance of 2 nH for a balance between area and power loss. The design considerations related to inductance choice are detailed in [14].

On each in-phase or quadrature path, a 5-bit switched capacitance bank is inserted between the mixer switches and IF transformer to tune the impedance shaper resonance frequency to 2.6 GHz. At each front-end balun output, an on-chip capacitor and a 2-mm bond wire form an L-shape MN further boosting the IF load impedance. Note that this MN is at IF and hence only needs to support a small fractional BW of <10% around 2.6 GHz.

In postlayout simulation, when loaded with 50-Ω termination, the RF mixer front end has a power conversion loss of 5 dB and an input matching of $13 \text{ dB}$ with a 3.5-GHz RF. Based on (5), the minimal achievable loss is 1.6 dB with $8-\text{Ω } R_{SW}$. We found additional 1.7-dB loss due to the RF switch parasitics, 1.4-dB loss from the on-chip transformer balun, and another 0.3-dB loss due to the on-chip first-order high-pass filter capacitor as well as finite LO rising and falling times.

### B. IF Complex Receiver With $I-Q$ Mismatch Compensation

As to the IF receiver, each I or Q path consists of a resistive feedback LNA as shown in Fig. 11(b) followed by a four-phase passive mixer. One clock generation circuit is shared among two receiving paths. In the presence of RF input bond wire and parasitic capacitance of on-chip pad, electrostatic discharge (ESD) diodes, and package lead, each IF receiver alone has an input matching of $-13 \text{ dB}$ with a double-sideband (DSB) NF of 2.8 dB at 2.6-GHz RF in simulation.

As discussed in Section III-B, we compensate the front-end $I-Q$ mismatch in the IF receiver baseband. As shown in Fig. 11(c), we implement a 7-bit vector modulator at each IF receiver I or Q baseband output. Switched-capacitor banks $C_{BB}$ at each vector modulator input form a first-order $RC$ filter together with the LNA and IF mixer output resistance. This $RC$ filter provides additional OOB interference suppression at large frequency offsets. The outputs of the vector modulators are added in the current domain for high linearity in the presence of image-band blockers. The vector modulator unit cell transconductance amplifier is similar to that reported in [44].

As illustrated by (10), the baseband compensation essentially uses the vector modulators to create a complex adjustment weight that mimics the $I-Q$ mismatch. This resembles a vector-modulator-based self-interference cancellation, and it has been shown that our 7-bit vector modulators together can compensate for a wide range of $I-Q$ amplitude and phase mismatches [39].

For testing purposes, a Mini-Circuits T4-6T-KK81+ transformer balun is used at each front-end I and Q output. The 1:4 balun translates the 50-Ω testing equipment impedance to a 200-Ω differential impedance that acts as the vector modulator load and performs current-to-voltage conversion.

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**V. MEASUREMENT RESULTS**

As shown in Fig. 13, the RF front-end mixer and IF receiver 65-nm CMOS chips are assembled using QFN packages and mounted on an FR-4 printed circuit board (PCB) with the two Qorvo QPQ1285 2.6-GHz BAW filters. A 1:1 balun (Johanson Technology Inc., 4400BL15A0050E) is used to facilitate single-ended measurements and its loss has been deembedded. The PCB trace between the BAW filter output and the IF receiver input has a measured loss of 0.4 dB. This trace is made long to support an IF chip orientation that allows a convenient access to the IF chip LO ports, as annotated in Fig. 13. However, in practice, an integrated LO is expected at the IF receiver side and the trace can be greatly shortened. Based on these, the long IF trace loss is deembedded in our measurements.

The RF front-end chip has a dc power consumption of 28–48 mW with 3.5-to-6.5-GHz RF, and the IF receiver chip draws 62-mW power in the nominal setting. Both the RF front end and the IF receiver chips use 1.2-V supplies.

**A. Fixed LO Small-Signal Measurements**

Using an RF LO and IF LO of 1.04 and 2.6 GHz, respectively, the filtering front-end chipset was first measured operating at 3.64 GHz near the lowest RF. This corresponds to the smallest frequency separation between the signal and its image with the worst case image filtering, as discussed in Section III-C.

As illustrated by (10), the baseband compensation essentially uses the vector modulators to create a complex adjustment weight that mimics the $I-Q$ mismatch. This resembles a vector-modulator-based self-interference cancellation, and it has been shown that our 7-bit vector modulators together can compensate for a wide range of $I-Q$ amplitude and phase mismatches [39].

For testing purposes, a Mini-Circuits T4-6T-KK81+ transformer balun is used at each front-end I and Q output. The 1:4 balun translates the 50-Ω testing equipment impedance to a 200-Ω differential impedance that acts as the vector modulator load and performs current-to-voltage conversion.
filtering primarily comes from the IF LNA and mixer output resistance and the vector modulator input switched-capacitor bank $C_{BB}$, as shown in Fig. 11. This results in an instantaneous BW of 160-to-170 MHz controlled by the IF receiver baseband switched-capacitor bank $C_{BB}$ setting.

The chipset has measured conversion gain of 20 dB, NF of 10 dB, and input $S_{11}$ of $<-10$ dB. Reasonable matches between simulated and measured gains and $S_{11}$ results are observed. Regarding NF, given simulated 5-dB RF passive mixer loss $g_1$ and 2.8-dB standalone IF receiver NF $F_{RX}$ mentioned in Section IV and BAW filter loss $g_2$ of 3 dB, the calculated chipset NF is 8.2 dB based on (12). This is lower than the simulated NF of 9.5 dB due to the nonideal impedance matching at various RF and IF interfaces. For example, given $R_{IF}$ of 68 $\Omega$ which corresponds to a $-16$-dB $S_{11}$, the calculated NF using (12) increases from 8.2 to 9.5 dB.

Without $I-Q$ mismatch compensation, in other words, the receiver baseband vector modulator weights $M_{ij}$ and $A_{ij}$ are all set to be zero (see Fig. 9), and the measured IRR is from 18 to 21 dB with an average of 19 dB across a 160-MHz signal BW.

Next, we used IF receiver baseband vector modulators to compensate $I-Q$ mismatch for improved IRR. The measured conversion gain and NF with $I-Q$ mismatch compensation are plotted in Fig. 15. As expected, the measured signal band conversion gain and NF are almost identical to those in Fig. 14. However, the average IRR is improved from 19 to 42 dB after $I-Q$ mismatch compensation. Because of the frequency selectivity difference between the vector modulators and the $I-Q$ mismatches [see Section III-B and (10)], the image rejection experiences a larger variation of 33-to-45 dB, compared to 18-to-21 dB.

Finally, we measured the chipset with a fixed-frequency RF input high-pass filter (Mini-Circuits, VHF-3100+) and the results are shown in Fig. 16. This high-pass filter further improves the average IRR to 81 dB across the 160-MHz BW while degrading the conversion gain and NF by 0.8 dB. As discussed in Section III-C and based on the measured high-pass filter response, this fixed-frequency high-pass filter would provide an additional 32-to-45-dB image rejection across the entire 3.5-to-6.5-GHz RF operation range.

**B. Small-Signal Measurements Across LO Frequencies**

The measured chipset conversion gain, NF, and input reflection coefficient ($S_{11}$) across seven different RF LO frequencies (IF receiver LO is fixed at 2.6 GHz) are plotted in Fig. 17. For beyond 6-GHz RF, the conversion gain is measured with a noise source, and hence, only in-band gain is reported. Across the entire RF range of 3.5-to-6.5 GHz, the chipset has 18-to-20-dB conversion gains, 10-to-12-dB NFs, and $<-6$-dB $S_{11}$. The simulated $S_{11}$ results are plotted in Fig. 17 showing a good match with those in measurements. $S_{11}$ is degraded beyond 5-GHz RF due to parasitics associated with the QFN package, bond wire, as well as on-chip pads and ESD diodes.

**C. Large-Signal Measurements**

The measured mixer-first acoustic-filtering chipset in-band linearity results are shown in Fig. 18 with RF at 3.5 and 6 GHz. The measured output-referred in-band 1-dB compression point
Fig. 17. Measured chipset RF input reflection coefficient, NF, and conversion gain across seven different RF LO frequencies (IF receiver LO is fixed at 2.6 GHz). The simulated input S11 is plotted in dashed lines.

Fig. 18. In-band linearity measurement results: (a) P1dB and (b) IP3 with 3.5-GHz RF, and (c) P1dB and (d) IP3 with 6-GHz RF.

(P1dB) is −4 dBm, while the in-band output-referred IP3 is +10 dBm. The in-band linearity is dominated by the IF receiver baseband vector modulators.

The measurement setups for OOB linearity measurements and their results versus normalized offset frequency (Δf/fBW) are shown in Fig. 19. For blocker-induced 1-dB compression point (B-1dB) measurement, a weak in-band sinusoid is applied at fS = 3.5 GHz (or 5.1 GHz) with RF mixer switching frequency fLO1 = 0.9 GHz (or 2.5 GHz). Another strong OOB sinusoidal blocker is fed to the front-end input at fS + Δf, causing compression of the weak in-band signal. Regarding the measurement of OOB IIP3, a two-tone blocker signal is used. The two-tone frequencies f1 and f2 are chosen such that the third-order intermodulation product falls inside the front-end passband, that is, 2f1 − f2 = fS, assuming f2 > f1. The frequency offset Δf is calculated as Δf = f1 − fS, which is also Δf = f2 − f1 [2], [3].

The measured B-1dB and OOB IIP3 at 1 × BW offset are +5 and +27 dBm, respectively, with fS = 3.5 GHz. When fS = 5.1 GHz, similar results are seen, as shown in Fig. 19. This high linearity at this close-in frequency offset is achieved through the sharp filtering from the IF BAW filters. As the OOB blocker frequency offset increases, the B-1dB and IIP3 saturate to about +9 and +30 dBm, respectively, and they are limited by the RF mixer switches.

We also measured front-end NF with an OOB sinusoidal blocker. As shown in Fig. 20, the RF and IF mixer LO frequencies are 1.7 and 2.6 GHz, respectively, corresponding to an RF frequency fS = 4.3 GHz. The blocker is at 4.48 GHz, which is 1 × BW away. These RF and blocker frequencies are chosen based on the availability of a blocker noise filter that suppresses the signal generator receiver band noise. In this way, the NF degradation solely comes from front-end gain compression and reciprocal mixing due to the mixer LO phase noise [5], [9]. Among the RF and IF mixers, the reciprocal mixing is dominated by the RF mixer as the blocker is eliminated by the IF BAW filters. The signal generator that provides the RF mixer LO signal has a measured effective phase noise of around −160 dBc/Hz at the 180-MHz offset, dominating the LO path phase noise.
The measured blocker NFs are shown in Fig. 20. With a weak blocker, the NF is 10.5 dB, which is consistent with that from the small-signal NF measurement. When the blocker power $P_{Blk}$ is from $-5$ to $+5$ dBm, the NF increases almost linearly. This is as expected as the blocker NF due to reciprocal mixing can be calculated as $NF_{Blk} = P_{Blk} - 160 \text{dBc/Hz} + 174 \text{dBm}$, which is 9-to-19 dB with $-5$-to-$+5$-dBm $P_{Blk}$ [45]. When $P_{Blk} > 5$ dBm which is our measured front-end 1-dB gain compression point, the gain compression starts to contribute noticeably to the NF degradation.

### D. Comparison and Discussion

A measurement summary and comparison with recent state-of-the-art works is given in Table I. This work achieves $+27$-dBm OOB IIP3 and $>30$-dB rejection at $1 \times \text{BW}$ offset. This is a 6-to-11-dB improvement in OOB IIP3 at $1 \times \text{BW}$ offset compared to state-of-the-art monolithic mixer-first direct-conversion or low-IF receivers in [5]–[9]. While the $N$-path filter in [2] has similar linearity performance, this work operates at five-times higher frequencies with two-times wider tuning range and $>15$-dB higher OOB rejection. Clock-path bootstrapping akin to that used in [2] could further enable the front-end OOB linearity performance.

It should be noted that the works in [2], [11], and [14] do not include an IF receiver. To facilitate NF comparison between the filter-only and the filtering receiver front-end works, we have assumed a 3-dB-NF IF receiver for each filter-only work.

When compared to the mixer-first acoustic-filtering front end using the IF-only recombination in [14], this work achieves 2.6-times wider BW, operates at 1.4-times higher RF, and eliminates off-chip IF balun and inductor.

The 1.5-to-1.9-dB higher NF is mostly due to the loss difference between the acoustic filters used. In [14], a 1.6-GHz-RF SAW filter is utilized that has a BW of 65 MHz and a loss of 1.1 dB. Acoustic filters operating a higher RF with a wider BW are generally more difficult to design and hence often come with higher insertion loss [21]. Here, the 2.6-GHz 190-MHz BW BAW filter has a loss of 3 dB, which alone leads to a 1.9-dB worse NF.

Also, mixer switch parasitics introduce more losses and degrade NF at a higher RF or IF in a mixer-first receiver [46]. The mixer switches in this work introduce 3.3-dB losses in simulation, which is 0.9 dB more compared to that in [14] at a lower frequency.

Moreover, additional loss and hence more NF degradation will be introduced by the off-chip IF balun and inductor as well as their package parasitics in [14] when scaling them from 1.6-GHz IF to 2.6 GHz.

While the OOB linearity and interference rejection at $1 \times \text{BW}$ offset in this work compare favorably with most others in Table I, they are worse compared to those in [14]. However, it should be noted that the OOB linearity and interference rejection are set by the IF acoustic filters. Based on our measurements in Fig. 8, the high-frequency broadband BAW filter used here has around 10-to-20-dB less OOB and transition-band rejection compared to the narrower band SAW filter in [14]. Therefore, scaling the design in [14] to a higher RF and IF will lead to similar performance degradation.

Linearity-and-noise performances, calculated as OOB IIP3 at $1 \times \text{BW}$ offset divided by NF, of the aforementioned state-of-the-art works are plotted in Fig. 21 versus RF and instantaneous BWs. This work achieves favorable linearity-and-noise performance while operating from 3.5- to 6.5-GHz RF and supporting a $>160$-MHz instantaneous BW. The mixer-first acoustic-filtering front end with IF-only recombination in [14] has 2.4-dB higher IIP3 and 1.5-to-1.9-dB lower NF. However, the lower NF and higher IIP3 in [14] are primarily due to the usage of a lower frequency SAW filter.
filter. While the SAW filter has 1.9-dB lower loss and better OOB rejection, it results in a sub-4.5-GHz RF and a nearly three-times narrower instantaneous BW. As we discussed in Sections III-E and V-D, if scaling the work in [14] to 3.5-to-6.5-GHz RF with a high-frequency broadband acoustic filter, it will have a similar IIP3 and worse NF. The proposed mixed-domain architecture has a superior noise performance in part due to the usage of two IF receiving paths (but more IF dc power and chip area) and in part due to the elimination of the excessive IF LC components, as discussed in Section III.

Finally, let us briefly compare our work to a conventional filter-bank-based multiband front end, as shown in Fig. 2(a). To seamlessly cover the 3.5-6.5-GHz RF with an instantaneous BW of 160 MHz, 18 different filters are needed in a conventional filter-bank-based design. This large number of filters for one receiver comes with prohibitively large size and high cost for a small-form-factor mobile device, especially when MIMO operation is incorporated. Furthermore, it is known that a single-pole-N-throw switch with a large N suffers from high loss and/or small BWs. Given a switch ON-resistance that corresponds to an acceptable switch loss, the switch parasitic capacitance at the antenna side increases linearly with the number of paths N. With N = 18, the parasitic capacitance is prohibitively large, requiring an antenna MN that has high loss and/or narrow BWs.

VI. Conclusion

By fusing commutated-LC passive mixer and acoustic filters in a mixer-first acoustic-filtering front end with a new mixed-domain recombination architecture, we have demonstrated a reconfigurable wireless receiver front end with a widely tunable RF from 3.5 to 6.5 GHz for future wireless applications. The front end is capable of achieving very high linearity, especially at close-in frequency offsets, with a wide 160-MHz instantaneous BW and zero off-chip IF components except for acoustic filters.

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REFERENCES

[1] A. Hagelauer, G. Fattinger, C. C. W. Ruppel, M. Ueda, K. Hashimoto, and A. Tag, “Microwave acoustic wave devices: Recent advances on architectures, modeling, materials, and packaging,” IEEE Trans. Microw. Theory Techn., vol. 66, no. 10, pp. 4548–4562, Oct. 2018.

[2] P. Song and H. Hashemi, “RF filter synthesis based on passively coupled N-path resonators,” IEEE J. Solid-State Circuits, vol. 54, no. 9, pp. 2475–2486, Sep. 2019.

[3] M. Darvishi, R. van der Zee, E. A. Klumperink, and B. Nauta, “Wideband 4th order switch-GM-C band-pass filter based on N-path filters,” IEEE J. Solid-State Circuits, vol. 47, no. 12, pp. 3105–3119, Dec. 2012.

[4] F. Amin, S. Raman, and K.-J. Koh, “Integrated synthetic fourth-order Q-enhanced bandpass filter with high dynamic range, tunable frequency, and fractional bandwidth control,” IEEE J. Solid-State Circuits, vol. 54, no. 3, pp. 768–784, Mar. 2019.

[5] S. Krishnamurthy and A. M. Niknejad, “Design and analysis of enhanced mixer-first receivers achieving 40-dB/decade RF selectivity,” IEEE J. Solid-State Circuits, vol. 55, no. 5, pp. 1165–1176, May 2020.

[6] G. Pini, D. Manstretta, and R. Castello, “Analysis and design of a 260-MHz RF bandwidth ×22-dBm OOB-IIP3 mixer-first receiver with third-order current-mode filtering TIA,” IEEE J. Solid-State Circuits, vol. 55, no. 7, pp. 1819–1829, Jul. 2020.

[7] C. Wu, Y. Wang, B. Nikolić, and C. Hull, “An interference-resilient wideband mixer-first receiver with lo leakage suppression and I/Q correlated orthogonal calibration,” IEEE Trans. Microw. Theory Techn., vol. 64, no. 4, pp. 1080–1091, Apr. 2016.

[8] E. C. Szoka and A. Molnar, “Circuit techniques for enhanced channel selectivity in passive mixer-first receivers,” in Proc. IEEE Radio Freq. Integ. Circuits Symp. (RFIC), Jun. 2018, pp. 292–295.

[9] Y.-C. Lien, E. A. M. Klumperink, B. Tenbroek, J. Strange, and B. Nauta, “Enhanced-selectivity high-linearity low-noise mixer-first receiver with complex pole pair due to capacitive positive feedback,” IEEE J. Solid-State Circuits, vol. 53, no. 5, pp. 1348–1360, May 2018.

[10] S. Ba, S. Hameed, and S. Pamarti, “Periodically time-varying noise cancellation for filtering-by-aliasing receiver front ends,” IEEE J. Solid-State Circuits, vol. 56, no. 3, pp. 928–939, Mar. 2021.

[11] Q. Wu, X. Zou, S. Qin, and Y. E. Wang, “Frequency translational RF receiver with time varying transmission lines (TVTL),” in IEEE MTT-S Int. Microw. Symp. Dig., Jun. 2017, pp. 1767–1769.

[12] H. Seo and J. Zhou, “A 2.5-to-4.5-GHz switched-LC mixer-first acoustic-filtering RF front-end achieving <6 dB NF, +30dBm IIP3 at 1×bandwidth offset,” in Proc. IEEE Radio Freq. Integ. Circuits Symp. (RFIC), Aug. 2020, pp. 283–286.

[13] M. Hedayati, L. K. Yeung, M. Panahi, X. Zou, and Y. E. Wang, “Parametric downconverter for mixer-first receiver front ends,” IEEE Trans. Microw. Theory Techn., vol. 69, no. 5, pp. 2712–2721, May 2021.

[14] H. Seo and J. Zhou, “A passive-mixer-first acoustic-filtering superheterodyne RF front-end,” IEEE J. Solid-State Circuits, vol. 56, no. 5, pp. 1438–1453, May 2021.

[15] K. Badiyari, N. Nallam, and S. Chatterjee, “An N-path band-pass filter with parametric gain-boosting,” IEEE Trans. Circuits Syst. I, Reg. Papers, vol. 66, no. 10, pp. 3700–3712, Oct. 2019.

[16] M. Soer, E. Klumperink, Z. Ru, F. E. van Vliet, and B. Nauta, “A 0.2-to-2.0 GHz 65 nm CMOS receiver without LNA achieving >11 dBm IIP3 and <6.5 dB NF,” in IEEE ISSCC Dig. Tech. Papers, Feb. 2009, pp. 222–223.

[17] H. Seo, M. Sha, and J. Zhou, “A 3.5-to-6.2-GHz mixer-first acoustic-filtering chipset with mixed-domain asymmetric IF and complex BB recombination achieving 170 MHz BW and +27 dBm IIP3 at 1× BW offset,” in Proc. IEEE Radio Freq. Integ. Circuits Symp. (RFIC), Jun. 2021, pp. 83–86.

[18] P. Warder and A. Link, “Golden age for filter design: Innovative and proven approaches for acoustic filter, duplexer, and multiplexer design,” IEEE Microw. Mag., vol. 16, no. 7, pp. 60–72, Aug. 2015.

[19] M. Zolfagharloo Koohi and A. Mortazawi, “Reconfigurable radios employing ferroelectrics: Recent progress on reconfigurable RF acoustic devices based on thin-film ferroelectric barium strontium titanate,” IEEE Microw. Mag., vol. 21, no. 5, pp. 120–135, May 2020.

[20] J. Lee et al., “A low-power and low-cost 14 nm FinFET RFIC supporting legacy cellular and 5G FR1,” in IEEE ISSCC Dig. Tech. Papers, Feb. 2021, pp. 90–92.

[21] S. Geng, R. Lu, Y. Yang, L. Gao, and A. E. Hassanien, “Microwave acoustic devices: Recent advances and outlook,” IEEE J. Microw., vol. 1, no. 2, pp. 601–609, Apr. 2021.
[22] F. Z. Bi and B. P. Barber, “Bulk acoustic wave RF technology,” *IEEE Microw. Mag.*, vol. 9, no. 5, pp. 65–80, Oct. 2008.

[23] The Piezoelectric Vibrator: Definitions and Methods of Measurements, Standard IRE 14.S1, IRE Standards on Piezoelectric Crystals, 1957.

[24] R. Aigner, G. Fattinger, M. Schaefer, K. Karnati, R. Rothemund, and F. Dumont, “BAW filters for 5G bands,” in *IEDM Tech. Dig.*, Dec. 2018, pp. 14–16.

[25] RF Front-End (RFFE) Filters: 313.85 MHz SAW Filter B3931B3768SZX10, Qualcomm, San Diego, CA, USA, 2012.

[26] 165 MHz SAW Filter SF2170D, RFMI, Carrollton, TX, USA, 2015.

[27] K.-Y. Hashimoto et al., “Moving tunable filters forward: A ‘heterointegration’ research project for tunable filters combining MEMS and RF SAW/BAW technologies,” *IEEE Microw. Mag.*, vol. 16, no. 7, pp. 89–97, Aug. 2015.

[28] T. H. Lee, *The Design of CMOS Radio-Frequency Integrated Circuits*, 2nd ed. Cambridge, U.K.: Cambridge Univ. Press, 2003.

[29] B. Razavi, *RF Microelectronics*. Upper Saddle River, NJ, USA: Prentice-Hall, 2011.

[30] C. Andrews and A. C. Molnar, “Implications of passive mixer transparency for impedance matching and noise figure in passive mixer-first receivers,” *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 57, no. 12, pp. 3092–3103, Dec. 2010.

[31] A. Mirzaei, H. Darabi, and D. Murphy, “Architectural evolution of integrated M-phase high-Q bandpass filters,” *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 59, no. 1, pp. 52–65, Aug. 2012.

[32] M. C. M. Soer, E. A. M. Klumperink, P. de Boer, F. E. van Vliet, and B. Nauta, “Unified Frequency Domain Analysis of Switched-Series-RC Passive Mixers and Samplers,” *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 57, no. 10, pp. 2618–2631, Oct. 2010.

[33] S. Gong and G. Piazza, “Design and analysis of lithium-niobate-based high electromechanical coupling RF-MEMS resonators for wideband filtering,” *IEEE Trans. Microw. Theory Techn.*, vol. 61, no. 1, pp. 403–414, Jan. 2013.

[34] T.-H. Wu and C. Meng, “5.2/5.7-GHz 48-dB image rejection GaInP/GaAs HBT weaver down-converter using LO frequency quadrupler,” *IEEE J. Solid-State Circuits*, vol. 41, no. 11, pp. 2468–2480, Nov. 2006.

[35] L. Gao, Q. Ma, and G. M. Rebeiz, “A 20–44-GHz image-rejection receiver with >75-dB image-rejection ratio in 22-nm CMOS FD-SOI for 5G applications,” *IEEE Trans. Microw. Theory Techn.*, vol. 68, no. 7, pp. 2823–2832, Jul. 2020.

[36] L. Sundstrom et al., “A receiver for LTE Rel-11 and beyond supporting non-contiguous carrier aggregation,” in *IEEE Int. Solid-State Circuits Conf. Dig. Tech. Papers*, Feb. 2013, pp. 336–337.

[37] S. C. Hwu and B. Razavi, “An RF receiver for intra-band carrier aggregation,” *IEEE J. Solid-State Circuits*, vol. 50, no. 4, pp. 946–961, Apr. 2015.

[38] J. Zhou, T.-H. Chuang, T. Dinc, and H. Krishnaswamy, “Integrated wideband self-interference cancellation in the RF domain for FDD and full-duplex wireless,” *IEEE J. Solid-State Circuits*, vol. 50, no. 12, pp. 3015–3031, Dec. 2015.

[39] J. Zhou, F. Dumont, “BAW filters for 5G bands,” in *IEEE ISSCC Dig. Tech. Papers*, Feb. 2017, pp. 100–102.

[40] Y. Lien, E. Klumperink, B. Tenbroek, J. Strange, and B. Nauta, “A high-linearity CMOS receiver achieving +44 dBm IIP3 and +13 dBm B1 dB for SAW-less LTE radio,” in *IEEE ISSCC Dig. Tech. Papers*, Feb. 2017, pp. 412–413.

[41] D. Murphy et al., “A blocker-tolerant, noise-cancelling receiver suitable for wideband wireless applications,” *IEEE J. Solid-State Circuits*, vol. 47, no. 12, pp. 2943–2963, Dec. 2012.

[42] D. Yang, C. Andrews, and A. Molnar, “Optimized design of N-phase passive mixer-first receivers in wideband operation,” *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 62, no. 11, pp. 2759–2770, Nov. 2015.

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