Tunable Blocker-Tolerant On-chip Radio Frequency Front-end Filter with Dual Adaptive Transmission Zeros for Software Defined Radio Applications

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Abstract—This paper presents a tunable active bandpass filter with adjustable transmission zeros (TZ) close to the passband for software defined radio applications. The RF front-end frequency selectivity is enhanced by the creation of TZs which also improve the out-of-band (OOB) input-referred third-order intercept point (IIP3). The filter is based on two-path signal cancellation and consists of a tunable bandpass filter in parallel with two tunable bandstop filters. This combination ensures the correct amplitude and phase relationships across a wide tuning range to create adjustable TZs without sacrificing the gain of the passband. This paper presents in detail the design considerations and guidelines, as well as analysis of the filter performance in the presence of non-idealities such as parasitics and imperfect clock signal shape. The proposed filter is implemented with high-Q N-path filter blocks in a 65-nm CMOS process. The passband of the filter is tunable from 0.1 GHz to 1.4 GHz with a 3-dB bandwidth of 9.8–10.2 MHz, a gain of 21.5–24 dB, a noise figure of 3–4.2 dB, and a total power consumption of 50–73 mW. TZs are created on both sides of the passband with a minimal offset of 25 MHz and are tunable across a 20 MHz range with up to 60 dB rejection. The measured blocker 1-dB compression point is 8 dBm and the out-of-band IIP3 is 23 dBm. The reported filter provides a promising on-chip filtering solution for multi-standard, multi-frequency software-defined radio applications with improved interference mitigation capabilities.

Index Terms—software radios, active filters, tunable filters, CMOS, Bandpass, N-path, RFIC, transmission zero, switched-capacitor, carrier aggregation.

I. INTRODUCTION

In the face of an increasing demand for high data rate connectivity and the scarcity of the available frequency spectrum, the last two decades have witnessed the development of many innovations in wireless communication technologies. One example is software defined radios (SDRs) which can adapt their operating parameters, such as frequency and modulation scheme, by field programmability [2]–[4]. A common limitation of SDRs is a lack of high quality programmable/tunable front-end filter, which makes the SDR receiver susceptible to strong interferers from both fixed frequency and other SDR users [5].

This issue is exacerbated as the wireless communication industry moves to advanced long-time evolution (LTE-A) standards. The plethora of allocated frequency bands opens up demand for SDR based front-end transceivers, and at the same time, highlights the need for low-loss high-selectivity tunable filters. In particular, carrier aggregation (CA) which allows an aggregated higher data rate by simultaneously combining multiple transmit/receiver carriers presents new challenges in frontend filtering. Both intra-band aggregation (RF channels belong to the same band) and inter-band aggregation (different bands) are supported in 3GPP standard [6]. The channels can be adjacent to one another (contiguous) or not (non-contiguous) for intra-band aggregation. Suppressing near-by (in frequency) high power transmit signals (blockers) from either its own or near-by (in location) transmitter is necessary to avoid desensitizing the receiver.

To handle large out-of-band (OOB) blockers, various on-chip techniques, including current-mode architectures [7]–[12] and mixer-first architectures [13], [14], have been reported recently. These techniques translate the baseband low-pass impedance profile to the radio-frequency (RF) carrier frequency and creates an equivalent high-Q bandpass filter (BPF) centered at the local oscillator (LO) frequency. RF BPFs have also been reported based on the similar N-path filtering technique [15]–[18] which can emulate high-Q LC tank with a tunable center frequency and constant bandwidth. Frequency tunable high-Q BPF can be synthesized at the carrier frequency using discrete-time signal processing [19]–[21].

Even though the aforementioned techniques show promise in handling large OOB blockers, the performance may still be insufficient in carrier aggregation scenarios, especially when large blockers are located very close to, or even between, the weak desired RF channels. To this end, BPFs with transmission zeros (TZs) located at the blocker frequency may become a necessary solution [22].

This paper presents novel design techniques to create highly selective on-chip N-path tunable filters with adjustable transmission zeros very close to the passband edges. The central idea is to use multiple signal paths with appropriate gain and phase profiles (with respect to frequency) to create interference cancellation at select frequencies. The major advantage of this strategy is the ability to create close-in TZs, e.g. 25 MHz from a 1-GHz center frequency as demonstrated in this work, with a favorable trade-off in area, power consumption, and out-of-band rejection, than simply increasing the BPF filter order.
It is noted that there has been significant research devoted to developing wideband linear low noise amplifiers (LNA) to ensure co-existence of the weak received signal with strong blockers [22], [24], [25]. Due to smaller supply voltages used in modern CMOS processes and hence the voltage swing limitation at the LNA output, achieving high linearity is still challenging. In addition, due to the power level variations in RF carriers, a shared LNA with a fixed power gain may not present an optimum solution in carrier aggregation scenarios. The proposed filter design in this paper will be a compliment to high linearity LNA designs by providing significant blocker rejection up front. In fact, it has been shown in [26] that incorporating a bandpass filter in front of the LNA improves the linearity significantly. Several tunable duplexers have been demonstrated recently [27], [28] along the same line.

Built upon our previous work [1], this extension paper presents in detail the strategy, considerations, and guidelines for the design of the proposed tunable filter. Besides presenting detailed linearity and noise analysis, we also investigate the impact of non-idealities, such as parasitics and finite switching time, on the filter performance. Finally, the implementation details and measured results of the proposed filter are presented.

II. TRANSMISSION ZERO CREATION MECHANISM

Creating TZs near the passband has been a popular strategy in passive high frequency filter designs to improve the rejection of close-in interferers (blockers) [29]. More recently, this idea has also been investigated for active filter designs.

Feed-forward techniques [30]–[34] or cascading of a bandpass filter (BPF) and a bandstop filter (BSF) with different center frequency [35]–[37] have been proposed to suppress transmit (TX) leakage and TX noise in the receive band. In [22], a TZ is created on one side of the passband by using a BPF+BSF cascade in the feedback path. A second TZ can also be obtained by adjusting the clock bias voltage. However, in these architectures, the bandwidth (BW) and TZ position are coupled with each other. The passband gain and noise figure degrade significantly if the TZs are placed close by. For example, in [22], the closest TZ is ∼250 MHz offset from the passband, which is insufficient for CA applications.

A second school of thought is to split the signal into multiple paths and achieve out-of-phase cancellation at select frequencies. For two signals at the same frequency, the criteria is that the two signals should have the same magnitude but differ in phase by an odd multiple of $\pi$. This simple idea forms the basis for TZ creation in cross-coupled passive resonator filters [38], and has also given rise to signal-interference type microwave filters [39]–[41].

Along this line, it has been proposed that two tunable bandpass filters with different amplitudes can be subtracted to create two TZs beside the passband [23]. However, the insertion loss degrades due to the subtraction of the two filter gains as illustrated in Fig. 1(a). Following our previous ideas [1], [42], we propose a novel and more effective TZ creation strategy by combining a tunable bandpass filter in parallel with a tunable bandstop filter to satisfy the correct magnitude and phase relationships for the creation of TZs close to the passband (Fig. 1(b)). The choice of using a bandstop filter in the auxiliary path is based on two main considerations. First, the use of two filters ensures that the correct phase relationship can be met over a wide frequency tuning range as both filters could be tuned synchronously. Second, due to the use of the bandstop filter in the parallel path, the passband gain is not reduced when the signals from the two paths are subtracted.

As a vehicle to illustrate the proposed concept, a simple

![Fig. 1. (a) Two BPF with different amplitude can be used in parallel to create dual transmission zeros on both side [23]. The insertion loss is increased due to the subtraction of two BPFs. (b)&(c) Magnitude and phase response respectively for the case shown in (a). (d) The proposed transmission zero creation concept. A BPF in the main path along with a BSF in the auxiliary path creates two TZs, one on each side of the passband, by judicious choice of the magnitude and phase responses. The insertion is not increased due to the use of the BSF in parallel path. (e)&(f) Magnitude and phase response respectively for the case shown in (d).](image)
by setting the numerator of one on each side of the passband as shown in Fig. 2(a). A parallel LC section (bandstop) is connected in parallel with the static passive filter circuit shown in Fig. 2(a) is first analyzed. The TZ renders the out-of-band rejection unusable as a practical filter. (b) A quasi-elliptic frequency response using acoustic wave lumped element resonators. However, as shown in Fig. 2(a), the out of band rejection is poor due to the cancellation of poles.

In the simulation, the two TZs can be created by putting a parallel LC section in parallel with the bandpass filter, as shown in Fig. 3(a). The impedance inverters can be implemented with active gyrators formed by back-to-back $g_m$ cells [17] as shown in Fig. 3(b). A network connected to one port of a gyrator is seen as its dual network in the second port. The active gyrator ensures that the impedance inverting property can be realized over a wide frequency tuning range.

The grounded LC resonators can be implemented with shunt configuration of N-path filters [16] and the floating bandstop resonator can be implemented with series configuration of N-path filters [44] as shown in Fig. 3(c). Assuming the same bandwidth for the RLC resonator and the N-path switched capacitor resonator, the corresponding $R_B, C_B$ and $L_B$ values for the N-path bandpass filter and $R_N, C_N$ and $L_N$ values for the N-path bandstop filter can be obtained using state space analysis [16], [36] as follows,

$$R_B = \frac{N [1 - \cos(2\pi D)] R_s + 2D^2 [R_s (1 - ND) + 2R_{sw}]}{2N(D)^2 - N [1 - \cos(2\pi D)]}$$

$$C_B = \frac{N D^2}{m \sin^2(\pi D) + D^2 (1 - ND)}$$

$$L_B = \frac{1}{(2\pi f_s)^2 C_B}$$

$$R_N = \frac{R_T N \sin^2(\pi D) + D^2 (1 - ND)}{N \left[(\pi D)^2 - \sin^2(\pi D)\right]}$$

$$C_N = \frac{N D^2}{m \left[N \sin^2(\pi D) + D^2 (1 - ND)\right]}$$

$$L_N = \frac{1}{(2\pi f_s)^2 C_N}$$

where $R_T = R_s + R_{sw,N} + R_L$, $D$ is the duty cycle of the clock and $m = 2$ for single-ended circuits and $m = 8$ for the differential circuits [44].

To increase the gain, reduce the noise figure (NF), and increase the dynamic range of the filter, the first gyrator is replaced with a single $g_m$ cell (amplifier) [17]. This effectively transforms the 6th-order (3-pole) bandpass filter prototype to a cascade of a 2nd-order (1-pole) filter with a 4th-order (2-pole) filter. The OOB roll-off of the BPF still follows a 3-pole characteristic due to the impedance isolation by $g_{m1}$.

A differential configuration is exploited for the filter design. The $g_m$ cells are implemented with self-biased inverters except $g_{m4}$ as shown in Fig. 4(b). A current starving self-biased inverter is used for $g_{m4}$ to control the $g_m$, which is necessary for better matching of the two paths to create TZs. Two negative resistors made of back-to-back inverters are added after the second and third BP resonators to increase the quality factor (Q) of the resonators. These negative resistors have a separate supply voltage ($V_{DD1}$) with a nominal value of...
1.2 V. However, large amount of negative admittance reduces the dynamic range (DR) as well as reduces the linearity and also increases the power dissipation. To make the gyrator stable, two diode connected inverters are added at the output of the filter to provide common mode feedback. To increase the gain and to reduce the noise, $g_{m2}$ is made bigger than $g_{m3}$. The various $g_m$ values, output impedance $r_o$ values, and the baseband capacitor $C_{BB}$ value are listed in Table I. The filter is capable of frequency tuning by changing the LO frequency. A snapshot of the simulated frequency response of the filter at 1 GHz (Fig. 3(f)) shows a high filter shape factor and two close-in (-40 MHz & 65 MHz) TZs with high rejection (>80 dB with respect to the passband gain).

The input impedance of the receiver, $Z_{in}$, is defined as the ratio of the fundamental components of voltage and current at the input port. Following the procedure in [11], $Z_{in}$ of the filter at the resonant frequency $f_s$ can be found as

$$Z_{in}(\omega) \simeq \left( R_{sw} + \frac{N}{\pi^2} \sin^2 \left( \frac{\pi}{N} \right) \right) \times \left[ Z_{BB}(\omega - \omega_{LO}) + Z_{BB}(\omega + \omega_{LO}) \right] / \left[ 1 + g_{m1} r_o \right],$$

where $N$ is the number of clock phases and $Z_{BB}$ is the impedance of baseband capacitor $C_{BB}$. The values of $R_{f1}$ (4.6 kΩ) and $g_{m1}$ (77.58 mS) are chosen properly for input impedance matching.

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<th>TABLE I</th>
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<td>Design Parameters</td>
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### IV. Linearity Analysis

In this section, we will analyze the benefits of creating TZs in terms of linearity. To determine the receiver input-referred third-order intercept point (IIP3), the level of in-band third-order intermodulation products (IM3) of any two blockers should be investigated. For two blockers with frequency of $f_{BLK1}$ and $f_{BLK2}$ (Fig. 5), the in-band IM3 product is

$$f_{RX} = 2f_{BLK1} - f_{BLK2},$$

FIG. 3. First iteration design of the proposed tunable filter with adjustable dual TZs. (a) Passive filter prototype; (b) Implementation of the impedance inverters by back-to-back $g_m$ cells; (c) Implementation of the bandstop resonator by series-connected N-path filter; (d) Implementation of the bandpass resonators by shunt-connected N-path filter; (e) Schematic diagram of the proposed filter; (f) An example simulated frequency response at 1 GHz showing a high filter shape factor and deep TZs on both sides of the passband.

FIG. 4. (a) The $g_{m1}$ cell; (b) The $g_{m4}$ cell; (c) The unit $g_m$ cell that is used in the filter with different scaling factors ($g_{m2}, g_{m3}$).
and the other term $2f_{BLK2}−f_{BLK1}$ falls outside of the receive (RX) band.

![Fig. 5. Comparison between proposed filter and an all pole filter with respect to linearity in presence of blockers.](image)

The output of a system can be modelled as $y(t) = α_1x(t) + α_2x(t)^2 + α_3x(t)^3$, where the input is $x(t)$. If two sinusoidal signals with amplitude $A_1$ and $A_2$ are applied to the system, the resulting IM3 and IIP3 are [45]

$$A_{IM3} = \frac{3}{4} \alpha_3 A_1^2 A_2 = \frac{\alpha_1}{A_{IIP3}^2} A_1^2 A_2,$$

and

$$A_{IIP3} = \sqrt{\frac{4}{3} \frac{\alpha_1}{\alpha_3}},$$

where $A_1$ and $A_2$ represents the amplitude of two blockers.

The IIP3 of the entire receiver unit can be found as follows

$$P_{IIP3-RX} (dBm) = P_{BLK} (dBm) + \frac{P_{BLK} (dBm) - [P_{1M3-out} (dBm) - A_{v-Rx} (dB)]}{2}, \quad (11)$$

where $P_{BLK}$ is the blocker power level, $A_{v-Rx}$ is the gain of the entire filter. The filter provides different rejections, $R_{BLK1}$ and $R_{BLK2}$ for the two blockers at $f_{BLK1}$ and $f_{BLK2}$ respectively. Fig. 5 shows the filtering behavior for the proposed filter and an all pole filter in the presence of two blockers. The IM3 level at the filter output is given by [26].

$$A_{IM3} = \frac{A_{v-Rx}}{R_{BLK1} R_{BLK2} A_{IIP3}^2} (12)$$

The goal is to make the $A_{IM3}$ smaller to increase the overall IIP3. The overall IIP3 improvement is

$$\uparrow P_{IIP3-RX} (dB) = R_{BLK1} (dB) + \frac{R_{BLK2} (dB)}{2}, \quad (13)$$

The improvement in linearity by adding the TZ can be seen by comparing the proposed filter with an all-pole filter with the same order and similar gain. As an example, if the proposed filter provides 15-dB extra attenuation at $f_{BLK1}$ and 8-dB worse attenuation compared to an all pole filter at $f_{BLK2}$, then according to (13), the IIP3 improvement is 11 dB. As a result, the proposed filter provides better linearity compared with an all pole filter with similar gain.

V. NON-IDEAL EFFECTS

A. Parasitic Capacitance

Due to the parasitic capacitance at different nodes of the filter, the expected dual-TZ behavior cannot be achieved as shown in the post-layout simulation (Fig. 6) of the proposed filter structure (Fig. 3(e)).

In this section we will analyze the parasitic capacitance effect on both the bandpass and bandstop filters. Even though the effect of parasitic capacitance on bandpass N-path filters has been analyzed in [17], the resulting expressions do not provide much design insight due to their complexity. In this paper, we will analyze the parasitic capacitance effect for both bandpass and bandstop resonators using a passive RLC resonator as an equivalent circuit. Fig. 7(a&b) show the RLC and N-path bandpass and bandstop resonators with parasitic capacitances ($C_p$, $C_s$, and $C_L$).

The transfer function for the grounded LC resonator with input parasitic capacitance $C_p$ is

$$H_{BP} (s) = \frac{s^2 L B C B R_B R_{SWB} + s L_B (R_B + R_{SWB}) + K}{T(s)}, \quad (14)$$

where $K = R_B R_{SWB}$ and

$$T(s) = s^2 L B C B C P R_S R_B R_{SWB} + s^2 [L_B C_B R_B (R_S + R_{SWB}) + L_B C_P R_S (R_B + R_{SWB})] + s [L_B (R_B + R_{SWB}) + C_P R_S R_B R_{SWB} + L_B R_s] + R_B (R_s + R_{SWB}). \quad (15)$$

Assuming the switch on resistance $R_{SWN}$ is small in Fig. 7(b), the transfer function for the floating LC resonator with source and load parasitic capacitance $C_s$ and $C_L$ can be found as

$$H_{BS} (s) = \frac{s^2 L_N C_N R_N R_L + s L_N R_L + R_N R_L}{T_1 (s)}, \quad (16)$$

and

$$T_1 (s) = \frac{s^2 L_N R_N R_S R_L [C_L (C_N + C_s) + C_s C_N]} + s^2 L_N [C_L R_L (R_s + R_N) + R_s R_N (C_s + C_N)] + C_N R_N R_L + C_s R_s R_L + s [R_s R_N R_L (C_s + C_L) + L_N (R_N + R_s + R_L)] + R_N (R_s + R_L). \quad (17)$$
Fig. 7. Equivalent parasitic capacitance at input and output nodes of the (a) bandpass and (b) bandstop filter.

![Diagram of equivalent parasitic capacitance](image)

Fig. 8. Symmetric phase responses for both the bandpass and bandstop filters create dual TZs on both sides of the passband (simulation). However, a shifted phase response of the bandstop filter creates a single TZ.

![Graph showing phase responses](image)

From (16), we can find the frequency shifting of the bandpass resonator due to $C_P$

$$\omega_{new,BP} = \omega_{LO} - \frac{\omega_{LO} \times C_P R_s}{2 C_B R_B} \times \frac{R_B + R_{SWB}}{R_s + R_{SWB}}. \quad (18)$$

Similarly, the frequency shifting of the bandstop (BS) resonator can be obtained as

$$\omega_{new,BS} = \omega_{LO} - \frac{\omega_{LO} \times (C_L R_N) + C_S R_s + R_s R_N C_S}{2 C_N R_N (R_s + R_L)}. \quad (19)$$

The parasitic capacitance for the bandstop filter include contributions from the top side of the switches, RF signal routing line, top and bottom plates of the metal-insulator-metal (MIM) capacitors, and the input capacitance of the main signal path. As a result, there is crosstalk among the 4 clock paths since $C_P$ is interacting with $C_1 - C_4$ during each clock phase. Fig. 9(a&c) show the simulated and theoretical behavior (14) for the bandpass resonator with parasitic capacitance. Here we assume $R_s = 50 \Omega$, $C_B = 121.8 \text{ pF}$, $C_{BB} = 4/\pi^2 \times C_B = 49.3 \text{ pF}$, $C_N = 3 \text{ pF}$, $R_{SW} = 10 \Omega$ and $R_L = 1 \text{k}\Omega$. Due to the impact of $C_P$, the center frequency decreases slightly as shown in Fig. 9(a) and the insertion loss increases due to crosstalk.

In contrast to the bandpass filter, the center frequency of the bandstop filter does not change much (Fig. 9). The baseband capacitor for the BS is small (1.13 pF) in our proposed filter implementation. However, both the magnitude and phase response show asymmetric behavior due to the parasitic capacitance as shown in Fig. 9(b&d).

Parasitic-induced phase shifting of the bandstop filter is shown in Fig. 9(f). Due to the phase asymmetry, it is difficult to generate two TZs as the phase requirement does not fulfill simultaneously on both sides of the passband (Fig. 8). Section VI-A will discuss techniques to overcome this problem.

### B. Clock Duty Cycle

In this section, we will consider the clock duty cycle effect. The finite rise and fall time are considered as part of the overall duty cycle reduction (Fig. 10).

To find the transfer function $H_{tol}(s)$ of the filter, the switches and capacitors are substituted with an equivalent...
baseband capacitor of $C_{BB}/D$. After that we find the transfer function $H_{L,PF}(s)$ of the baseband equivalent circuit. Then, we translate $H_{LPF}(s)$ in frequency to $\omega_{LO}$ and multiply the resulting transfer function by $\text{sinc}^2(\pi/D)$ [16], [17], where $D$ is the effective duty cycle of the clock. The effective duty cycle $D = PW + (t_r + t_f) / 2$ because the switches are turned on halfway in both the clock rise and fall times.

According to the principle mentioned earlier, Fig. 11 is used to find the transfer function of the corresponding baseband structure. Assuming $C = C_1/D = C_2/D = C_3/D$, $C_4' = C_4/D$ and $R_{L}' = R_L || r_o$, we find

$$H_{LPF}(s) = \frac{N(s)}{M(s)},$$

where

$$N(s) = s^2 CC_4' g_{m4} r_o r_{o4} R_{L}' + s C r_{o4} R_{L}' (g_{m1} g_{m2} r_{o1} + g_{m4}) + g_{m1} g_{m2} r_{o1} R_{L},$$

$$M(s) = (1 + s C R_o) [s^3 C^2 C_4' r_o r_{o4} R_{L}' + s C (C R_{L}' r_{o1} + C_4' k')] + s (C R_{L}' + C_4' g_{m2} g_{m3} r_{o1} r_{o4} R_{L}') + g_{m2} g_{m3} r_{o1} R_{L}' + 1],$$

and

$$k' = r_{o1} R_{L}' + r_{o1} r_{o4} + R_{L}' r_{o4}.$$  

Therefore

$$H_{tot}(s) = ND \times \text{sinc}^2\left(\frac{\pi}{D}\right) \times [H_{L,PF}(s - j\omega_{LO})]$$

$$+ (1 - ND) \times H_{DC},$$

where $H_{DC} \approx g_{m1}/g_{m3}$ is the transfer function without the baseband capacitors in Fig. 11.

The effect of the clock duty cycle on the proposed filter is shown in Fig. 12. If the duty cycle of the clock $D$ is smaller than $1/N$, all switches are periodically OFF for some time in each cycle and the output signal of the N-path filter tracks the input signal, resulting in higher input impedance for frequencies far away from the switching frequency and its odd harmonics, which then translates to less OOB rejection [35]. On the other hand, if $D > (1/N)$, two switches will be ON at the same time, which results in undesired charge sharing between capacitors and hence a degradation of the insertion loss. Careful layout is necessary to ensure $D$ close to $1/N$ and to increase the matching between different clock phases.

### VI. IMPROVED ARCHITECTURE

#### A. Dual-TZ Creation Strategy

As mentioned above in Section V-A, it is difficult to simultaneously control the TZs on both sides of the passband due to the parasitics. The TZ on one side of the passband can be made deeper but at the penalty of a shallower TZ on the other side (Fig. 6). To overcome this problem, a second BS section is added to give us more degrees of freedom in adjusting the frequency responses of the two signal paths. This addition can be justified in terms of circuit area because the bandstop resonators are purposefully designed to have a larger bandwidth, and therefore uses much smaller capacitors. To avoid mutual interaction between two BS sections, the second BS section is connected before the output buffer. Putting a buffer after second BSF make the performance slightly better.
however there is associated power dissipation with the buffers. Without the buffer, the phase response is still acceptable and two notches can still be generated.

It is well known that in a N-path filter, the input signal experiences down-conversion along with low-pass filtering passing through the switches to the capacitors and the same switches up-convert the filtered capacitor voltages to the output node [16]. As a result, it is possible to analyze the corresponding baseband version of the filter with a second high pass filter as shown in Fig. 14. The LPF is considered as a single pole for simplicity. The transfer function of the filter can be written as

\[ H(j\omega) = \frac{\alpha}{1 + \frac{j\omega}{P_1}} + \frac{\beta_1}{1 + \frac{j\omega}{P_2}} + \frac{\beta_2}{1 + \frac{j\omega}{P_3}} e^{j\pi}, \]  

(22)

where \( P_1, P_2 \) and \( P_3 \) are the poles of low pass filter (LPF) and high pass filters (HPF) respectively. To create TZ at a particular frequency, the real part and the imaginary part of the numerator of (22) should be zero at that particular frequency. However, this condition can be accomplished by multiple different gain (\( \alpha, \beta_1 \) and \( \beta_2 \)) and pole (\( P_1, P_2 \) and \( P_3 \)) locations. For example, one possible solution is to set the LPF gain as

\[ \alpha = \frac{-X \pm \sqrt{X^2 - 4P_1(P_2 + P_3)Y}}{2P_1(P_2 + P_3)}, \]  

(23)

where \( X = \beta_1(P_2 + 2P_1P_3 + P_1P_2) - \beta_2(P_2^2 + 2P_1P_2 + P_1P_3) \) and \( Y = \beta_2^2P_3(P_1 + P_3) - \beta_1\beta_2(P_1P_2 + 2P_2P_3 + P_1P_3) + \beta_2^2P_2(P_1 + P_2) \).

To obtain a positive \( \alpha \) (which is necessary for real frequency TZ creation) two condition need to be fulfilled, either i) \( X < 0 \) and \( \sqrt{X^2 - 4P_1(P_2 + P_3)Y} > 0 \); or ii) \( X > 0 \) and \( Y < 0 \). However, there are many possible choices to fulfil the criteria. Assuming \( \beta_1 > \beta_2 \), the TZ frequency can be chosen

\[ \omega_{TZ} = \pm \sqrt{\alpha P_1P_2P_3}, \]  

(24)

As an example, if we choose \( \alpha = 20, \beta_1 = 10, \beta_2 = 8, P_1 = 10 \text{MHz}, P_2 = 20 \text{MHz} \) and \( P_3 = 15 \text{MHz} \), then according to (29), two TZ frequencies will be \( \pm 17 \text{MHz} \). After up-conversion to \( f_{LO} = 1 \text{GHz} \), the two TZ frequencies will be \( 1.017 \text{GHz} \) and \( 983 \text{MHz} \). The simulated filter responses with one and two bandstop paths are compared in Fig. 15.

We plot the magnitude and phase responses of BP+1BS filter with or without parasitics along with BP+2BS filter with parasitics in Fig. 16. Adding an extra BS section helps to maintain the symmetric phase at some particular frequencies and dual TZs can still be obtained. In the BP+2BS configuration, it is always possible to create sharp TZs on both sides of the passband, suitable for the desired application of suppressing strong blockers close to the desired channel. The \( g_m \) values used in the simulation are \( g_{m1} = 27.58 \text{mS}, g_{m2} = 29.36 \text{mS}, g_{mc} = 24.5 \text{mS} \) and \( C_p = 40 \text{pF}, C_{p1} = 20 \text{pF} \). The Q of the tank is specified as 35.

B. Improved Filter Design

The schematic of the improved tunable filter with dual TZs is shown in Fig. 18. By setting \( s = 0 \) and \( D = 1/N \) in (20), the center frequency gain of the filter can be found

\[ A_v = \frac{g_{m1} g_{m2} r_{oa} R_L}{1 + g_{m2} g_{m3} r_{oa} R_L} \text{sinc}^2 \left( \frac{\pi}{4} \right). \]  

(25)

To increase the gain of the filter, \( g_{m1} \) needs to be increased and \( g_{m3} \) needs to be reduced. However, increasing \( g_{m1} \) increases the power consumption and reduces IIP3. Reducing
$g_{m3}$ reduces the BW of the filter. A compromise between gain, NF, area, and power consumption needs to be reached when determining the $g_{m1}$ and $g_{m3}$ values.

Fig. 19 shows a simplified schematic of the filter to calculate the stopband rejection $A_{sb}$ (difference between passband and stopband voltage gain) of the filter. $A_{sb}$ can be found, assuming a perfect duty cycle, as

$$
\frac{1}{|A_{sb}|} = \sqrt{2} \frac{g_{m1}}{g_{m3}} \times \text{sinc}^2 \left( \frac{\pi}{4} \right) \times \frac{1 + g_{m2}g_{m3}R_{sw}^2}{g_{m1}g_{m2}R_{sw}^2 + (g_{m4} - g_{m5})R_{sw}} \times \left( 1 + \frac{R_{s}}{R_{sw}} \right).
$$

(26)

We would like to reduce $R_{sw}$, $g_{m2}$ and $g_{m3}$ to increase the stopband rejection. The BW shrinks if either $g_{m2}$ and $g_{m3}$ is reduced. From noise point of view, we would like to increase $g_{m2}$ and reduce $g_{m3}$.

Shown in Fig. 20, the proposed filter is compared with a 6th order all-pole bandpass filter. The filter has more close-in selectivity (~20 dB better at 40 MHz offset) compared with the all-pole filter due to the creation of close-in TZ. The ultimate out-of-band rejection is worse due to the cancellation of the BP poles. However, the rejection is still relatively high (40–50 dB) and additional low-Q bandpass filters could be easily integrated to provide further suppression.

Slight tuning of the clock bias voltage also helps to create deeper TZs because it can compensate for the non-ideal duty cycle due to parasitics, layout imperfections, and finite switching time (Fig. 10). However, this does not shift the TZ frequency significantly. The simulated frequency response of the filter with different clock bias voltage is shown in Fig. 17.

A common drawback of N-path filtering is the harmonic folding associated with their time-variant nature [16]. Due to the comb-filter like characteristic, an N-path filter folds back from input frequencies around $(K N \pm 1) f_s$ to the desired band around $f_s$, where $f_s$ is the switching frequency. The strongest term which result in down-conversion in a 4-path filter is the 3rd harmonic. As a result, increasing the number of paths will increase the distance between $f_s$ and the first folded component around $(N-1)f_s$. However, generating non-overlapping clocks with a high number of phases is difficult at RF as it generally requires very high frequencies (N times the switching frequency). Also, the dynamic power dissipation and parasitic capacitance also increases due to the use of high number of phases.

There are two options to counteract this problem. A low-pass pre-filter can be used in front of the N-path filter to sufficiently suppress harmonic folding, which does not necessarily have to be tunable. In a differential configuration, second harmonic folding is cancelled which relaxes the pre-filter transition band. The alternate option is to use harmonic rejection mixers instead of simple switches that are commonly used in N-path filter. However, each harmonic rejection mixer requires roughly N switches and impedances. This means that the area required for switches and impedances scales with $N^2$ which may be problematic for large N [47].
where $A$ receiver is given by

$$ F \simeq 1 + \frac{\gamma g_{m1} R_s A_1^2}{g_{m1}} + \frac{\gamma g_{m1} R_m g_{m2} r_{o1}^2 A_1^2}{g_{m2}} + \frac{R_{sw}}{R_s} \times \left( \frac{g_{m3} g_{m1} A_1}{g_{m1} A_1} \right)^2 $$

$$ + \frac{\gamma g_{m3} g_{m4} r_{o2}^2}{g_{m3}} \times \left( \frac{1}{R_{BS1}} \right)^2 $$

$$ + \frac{g_{m,neg1} (g_{m1} A_1)^2 g_{m2} r_{o3}^2}{g_{m,neg2}} $$

$$ + \frac{g_{m,neg3} (g_{m1} A_1)^2 g_{m3} r_{o4}^2}{g_{m,neg4}} $$

(27)

where $\gamma$ is the noise parameter of the MOSFET and $R_{BS1}, R_{BS2}$ are the insertion loss of the BS filters at the center frequency.

There is an additional contribution to the noise factor due to noise folding. The noise folding mechanism is shown in Fig. 23. The noise sources located at $(1 + kN)f_{LO}$ fold back to $f_{LO}$ with a gain of $1/|1 + KN|$, where $k = \pm 1, \pm 2...$ and $k \neq 0$. For $N = 4$, the first folding start from $3^{rd}$ harmonic and the folding back gain is $1/3 = -9.54$ dB. For each of the noise source, the additional factor due to noise folding is

$$ \sum_{n=-\infty}^{\infty} |H_{n,RF}(j\omega) V_{n,N, s} [j(\omega - n\omega_s)]|^2, $$

(28)
Fig. 16. One BS section with a $g_m$ cell (a) without parasitic capacitance (b) with parasitic capacitance. (c) A second BS section is added to adjust the frequency response. (d) Magnitude and (e) phase response with and without parasitics for one and two BS sections. Adding the second BS path allows one to fine tune the magnitude and phase to create deeper notches.

Fig. 17. Simulated filter performance at various clock bias levels.

where $n$ indicates the harmonic number of \( f_s \), \( H_{n,RF}(j\omega) \) is the \( n^{th} \) harmonic transfer function associated with the frequency \( n f_s \), and \( N_s \) represents different noise sources. However, the output noise around the LO harmonics is much smaller compared with the LO first harmonic [13], [48]. The noise contribution from $g_{m4}$ and $g_{m5}$ is negligible at the center frequency due to the use of bandstop filters after those blocks. The noise transfer function of $R_{sw}$ for the bandpass resonators is a notch function and therefore the noise contribution of switches used in the bandpass resonators is relatively suppressed at the center frequency of the filter. On the other hand, the noise transfer function of $R_{sw}$ for the bandstop resonators is a bandpass structure. However, the

Fig. 19. A simplified schematic of the filter to calculate the stopband rejection. The baseband capacitors are shorted to ground for frequencies far from the passband of the filter and $\sqrt{2}$ is the voltage gain of the balun.

Fig. 20. Simulated comparison between all pole filter and the proposed filter.

Fig. 21. Monte Carlo simulation of the maximum TZ depth with PVT variations in 100 runs.

Fig. 23. Folding back of noises located at $|1 + 4k|f_{LO}$ to $f_{LO}$.

Fig. 24. Simulated group delay of the proposed filter.
VII. FILTER IMPLEMENTATION DETAILS

Large NMOS RF switches of \( W/L = 80 \mu m/60 \text{nm} \) (on-resistance of approximately \( 4.2 \Omega \)) are used to reduce the noise, non-linearity and mismatch between them. The switches are driven by 25\% duty cycle 4-phase non-overlapping clocks. Baseband capacitors are made differential in the bandpass section to save area. MIM capacitor with underlying metal are used for the baseband capacitors. The resistors are realized with N+ poly resistor without silicide. Each switch is biased at 900-mV DC voltage (VCLK) to provide full 1.2-V swing to maximize the linearity of the switches. The \( g_{m5} \) cell is shown in the inset of Fig. 18. The negative sign in Fig. 14 can be easily obtained by flipping the connection of the top bandstop filter in a differential configuration.

Monte Carlo simulation of TZ depth with PVT variations in 100 runs is shown in Fig. 21. The TZ depth varies from 66 dB to 73 dB at different process corners.

The simulated group delay of the proposed filter is shown in Fig. 24. The addition of the transmission zeros does not significantly affect the group delay variation for in-band signals. However, the group delay variation at the transmission zero frequencies is not very important as the filter provides high attenuation at those frequencies.

VIII. MEASUREMENT RESULTS

The filter is fabricated in 65-nm CMOS technology. The chip has an area of 2.4 mm\(^2\) including bonding pads (Fig. 25). The chip is wirebonded and tested on a printed circuit board. An off-chip balun (MACOM MABA-010247-2R1250) is used to transform a single-ended signal to a differential signal. The typical insertion loss of the balun is 1 dB.

A. Frequency Response

Fig. 26(a&b) show the measured \( S_{21} \) and \( S_{11} \) over the entire tuning range of 0.1–1.4 GHz. The filter shows a gain of 21.5–24 dB, and 3-dB bandwidth of 9.8–10.2 MHz. The parasitic capacitance at each node of the filter modifies the equivalent resistance of that node. As a result, the Q-factor of the filter decreases for higher center frequencies and reduces the gain. The filter draws 27.3 mA, each output buffer draws 4.7 mA and the LO chain draws 3.9 mA to 22.7 mA from 1.2 V in the whole tuning range. The LO feedthrough to the input port of the filter is \(-58 \text{dBm}\) at a center frequency of 1 GHz.

Fig. 26(c) shows the s-parameters of the filter at 1 GHz. At 1 GHz, the filter has a 23-dB gain with a passband ripple of 0.6 dB and a 3-dB bandwidth of 9.8–10.2 MHz. Transmission zeros are created on both sides of the passband with a minimal offset of 25 MHz and are tunable across 20 MHz range with up to 60 dB rejection. To the best of the authors knowledge, this represents the closest close-in TZ demonstrated in on-chip tunable filters.

The locations of the TZs can be tuned by changing the bank of capacitors \( (C_T) \) in the top and bottom BS sections as shown...
in Fig. 18. Example measurement results at a center frequency of 1 GHz are shown in Fig. 26(d). The measurement result shows approximately 20 MHz of TZ frequency tuning on both sides of the passband. Tuning of $g_{m4}$ also helps to shift the TZ frequency because it changes the bandstop gain. As the TZs move closer to the passband, the TZ rejection reduces because of a deterioration in the phase matching criteria.

B. Noise Figure

Over the tuning range of 0.1–1.4 GHz, the filter has a measured noise figure (NF) of 3–4.2 dB. The theoretical, simulated and measured NF of the filter over the tuning range is shown in Fig. 27(a). For the theoretical noise calculation, $R_{BS1} = R_{BS2} \approx 30$. The NF is higher at high frequencies as the gain of the filter decreases. The off-chip balun loss is de-embedded in the NF measurement.

Measured NF with various blocker power level is shown in Fig. 27(b) with $f_{LO} = 1$ GHz. In this test, the blocker is placed at the TZ frequency ($\sim 40$ MHz offset). The NF increases with increasing blocker power due to reciprocal mixing of the blocker and the LO phase noise. The proposed filter can tolerate an 8-dBm blocker with NF < 10 dB.

Simulations are carried out to demonstrate the effect of reciprocal mixing. When the blocker mixes with LO phase noise, it deposits additional noise in the receive channel proportional to the blocker amplitude. The simulated phase-noise of on-chip 4-phase LO signals ($S_1, S_2, S_3$ and $S_4$) with frequency of 1 GHz from a noisy external signal generator (Keysight E8663D) with frequency of $f_{io}$ is shown in Fig. 28(a). The SSB phase noise at various frequency offset (e.g. $-143$ dBc/Hz at 100 KHz offset) is obtained from the equipment data sheet and specified in Spectre to run the simulation. The simulated NF with noisy and noiseless external clock generator for different blocker power levels are shown in Fig. 28(b). The measured NF is very close to the simulation as shown in Fig. 28(b) and Fig. 27(b). The measured NF at various blocker frequencies with respect to the center frequency (1 GHz) is shown in Fig. 27(c). The blocker power was set to 0 dBm for this measurement. The measured NF shows a dip at TZ frequency ($\sim 40$ MHz) due to significant suppression of blocker at that frequency.

C. Linearity

To evaluate the linearity of the filter, two-tone IIP3 and compression measurements were carried out. Results are shown in Fig. 29(a&b). With the filter tuned to 1 GHz, the two tones are placed at 1.002 GHz and 1.003 GHz for the in-band IIP3 measurement, and 1.045 GHz and 1.092 GHz for the out-of-band IIP3 measurement. The out-of-band IIP3 and blocker 1-dB compression point (B1dB) are shown in Fig. 29(a). The B1dB is measured by placing an interfering signal at the TZ frequency and observing the compression behavior of the in-band signal while increasing the power of the interferer. B1dB is around 8 dBm over the tuning range. Out-of-band and in-band IIP3 measured at 1 GHz center frequency are shown in Fig. 29(b).

D. Comparison with other works

The proposed filter is compared with the state-of-the-art receivers and filters in Table II. The proposed filter shows advantages in creating two adjustable deep transmission zeros on both sides of the passband with low power consumption. A critical advantage is that TZs can be created extremely close (25 MHz) to the passband.

IX. Conclusion

A tunable bandpass filter with two adjustable transmission zeros has been presented in this paper. The TZs are created by the frequency selective cancellation of the transmitted signal from two parallel paths. Due to the tunability of the TZs, the proposed filter can handle strong dynamic blockers in close proximity to the passband and as such show promises as front-end filters for software-defined radio applications.

REFERENCES


TABLE II

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²From Maximum Gain to Notch depth
³Excluding clock buffer power consumption
⁴Active area

Fig. 28. (a) The simulated phase-noise of on-chip 4-phase LO signals ($f_{LO} = 1$ GHz) from a noisy external signal generator (at 4$f_{LO}$ (b) NF with noisy and noiseless LO signals. The phase noise of the external signal generator is specified in Spectre.

Fig. 29. (a) Measured out-of-band IIP3 and blocker 1-dB compression point over the whole tuning range. (b) Measured out-of-band IIP3 and in-band IIP3 at 1GHz.


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