A High-Speed TIA Based Programmable Broadband Complex Filter

By combining the functions of a poly phase filter and a variable gain amplifier, a 2nd order active complex filter is proposed in this paper for 5G and other broadband applications. Thanks to the TIA based structure, the complex filter has an operating speed up to 2.3GHz and a variable voltage gain of 0-20dB. Its center frequency is programmable from 500MHz to 1.85GHz by integrating an on-chip capacitor tank. An image rejection ratio of higher than 17dB is achieved when its gain is 14dB for all center frequency settings, consuming less than 40.6mW from a 1.4V supply. Fabricated in a 65nm CMOS process, the core circuit area is 0.25mm×0.2mm. To the best knowledge of the authors, this work is the first published GHz active complex filter, advancing the state-of-the-art.

key words: complex filter; poly phase filter; TIA; broadband filter; programmable center frequency
Classification: Integrated circuits (memory, logic, analog, RF, sensor)

1. Introduction

The 5th generation (5G) wireless communication for high throughput motivated the development of millimeter wave (mmWave) systems, such as 5G NR n257, n260. Such communication system requires both strong broad in-band IF signals and enough attenuation for adjacent channel interferers and image signals. For this purpose, complex filters can be utilized in the zero-intermediate frequency (zero-IF) and low-IF transceivers. They are also used to reduce the flicker noise and dc-offset. A complex filter can be realized by either a passive poly phase filter (PPF) or an active PPF. Passive PPFs suffer from large insertion loss and un-tunable center frequency. In broadband communications, high order passive PPFs are required, which increases the loss further. This increases the design difficulties in the whole communication chain. The most popular approaches of making active PPFs are using Op-amp RC structures [3]-[18]. But they are suffering from the very limited speed and poor capability of gain tuning.

A traditional low-IF transceiver has a structure shown in Fig.1 [1]. Mostly, a passive PPF and a variable gain amplifier (VGA) are both needed in the receiver for the image rejection and the voltage gain compensation. In a broadband transmitter, an amplifier is also required after the passive PPF to compensate the loss and relax the gain requirements of the following mixer and power amplifier (PA).

A programmable 2nd order active complex filter is proposed in this paper by combining the functions of a PPF and a VGA. This complex filter can be used to replace the two blocks (PPF and VGA) in a broadband receiver and relax the gain requirements of the transmitter. The active complex biquad is realized by using the high speed transimpedance amplifier (TIA) stage, paving a way to achieve high bandwidth and gain tuning capability at the same time. A capacitor tank is also implemented on-chip to tune the center frequency.

This paper is organized as follows. The TIA based active PPF biquad is presented in Section 2. Section 3 describes the design details of the TIA cell. Section 4 presents and evaluates the measurement results of the silicon prototype. It also provides a comparison between these work and other reported active complex filters. Finally, a conclusion is given in Section 5.
2. Active complex biquad

A complex filter is a shifted center frequency low-pass filter with a negative frequency stopband attenuation [1]. Fig.2 (a) shows the function diagram of a 1st order low-pass filter, whose transfer function is given by:

\[ H(s) = \frac{V_{\text{out}}}{V_{\text{in}}} = \frac{A_0}{s + \omega_0} \]  

(1)

After the frequency transformation, a 1st order complex filter can be realized. Its function diagram and transfer function are shown in Fig.2. (b) and in Equation (2), respectively.

\[ H(s)' = \frac{V_{\text{out},I}}{V_{\text{in},I}} = \frac{V_{\text{out},Q}}{V_{\text{in},Q}} = \frac{A_0}{s + \omega_0 \pm j\omega_c} \]  

(2)

Clearly, the center frequency of the low-pass filter is flatly shifted to a certain frequency \( \pm \omega_c \) from dc, i.e.

\[ j\omega \rightarrow j\omega \pm j\omega_c \]  

(3)

Then the image information at \( \mp \omega_c \) is rejected as shown in Fig.2. (c). The amplitude ratio between the frequency response at \( \pm \omega_c \) and \( \mp \omega_c \) is called the image rejection ratio (IRR).

To make a broadband complex filter, clearly a broadband low-pass filter is needed according to the above analysis. As shown in Fig.3, 2 cascading TIA stages are used to make the wideband low-pass biquad. The 2 TIA stages have the same gain setting and intrinsic pole position. In this way, an extremely high bandwidth and a gain tuning capability are obtained.

In theory, the frequency transformation can be performed by quadrature signals, either by voltage signals or current signals. Most published works are implementing the frequency transformation by quadrature voltage signals [3]-[18]. In this paper, the transformation is done by current signals, as shown in Fig.4. A quadrature current is added to the input of the TIA gain stage. Together with the original current, it is converted to a voltage output by a Gm block and a bypass resistor \( R_f \). The complex output can be approximately expressed as:

\[ V_{\text{out}} \approx (I_I + jI_Q)R_f \]  

(4)

The detailed architecture of the proposed active complex filter is shown in Fig.5. It consists of 2 same low-pass biquads, realizing the I and Q path. By inserting extra Gm blocks to generate quadrature currents, these two low-pass biquads stages are converted to a complex filter, accomplishing the frequency shifting. Note that the \( g_{m1} \) value is the same as that of the main path.
It can be clearly seen from Fig. 5 that the current through resistor \( R_f \) is given by:

\[
I_{R,f} = g_{m1} \cdot V_{in,j} + jg_{m1} \cdot V_{out,q,i}
\]

Assuming \( g_{m3}R_f \gg 1 \), the output voltage of the first stage can be given by:

\[
\frac{V_{out}}{V} = \frac{I_{R,f}}{V} \cdot \frac{R_f}{1 + \frac{sc}{g_{m2}}} = (g_{m1} \cdot V_{in,j} + jg_{m1} \cdot V_{out,q,i}) \cdot \frac{R_f}{1 + \frac{sc}{g_{m2}}}
\]

(6)

Accordingly, the transfer function of the 1 stage complex filter is calculated as:

\[
\frac{V_{out,j}}{V_{in,j}} \approx \frac{g_{m1}R_f}{1 + \frac{sc}{g_{m2}}} \cdot \frac{g_{m3}R_f g_{m2}/C}{1 + \frac{sc}{g_{m2}}} = \frac{g_{m3}R_f g_{m2}/C}{1 + \frac{sc}{g_{m2}}} \cdot \frac{g_{m1}R_f}{1 + \frac{sc}{g_{m2}}}
\]

(7)

Equation (7) shows that the center frequency of the filter is shifted by \( g_{m1}R_f \cdot g_{m2}/C \). Clearly, the filter center frequency can be tuned by the load capacitance \( C \). Moreover, the transconductance \( g_{m1} \) and \( g_{m2} \) are also tunable in this work to vary the gain. Pretty high design freedom can be obtained with this architecture.

Since a 3-bit capacitor bank is implemented on chip, the center frequency can easily be controlled. If this variable gain complex filter is used in a transceiver system, the \( R_f \), \( g_{m1} \), and \( g_{m2} \) values can be first selected with the system requirements. Then use the capacitor bank to shift the center frequency to the required frequency. In this design, both \( g_{m1} \) and \( g_{m2} \) are set to be 2.6mS.

3. TIA cell design

The TIA cell in Fig. 3 is realized by a modified Cherry-Hooper amplifier, which is shown in Fig. 6 [2].

![Fig. 6 The structure of the TIA cell.](image)

In this structure, the shunt resistance \( R_f \) is implemented by a PMOS transistor. With its inherent parasitic capacitance \( C_f \), a positive zero is produced, improving the TIA’s bandwidth and stability performances. Note transistors \( M_5 \)–\( M_8 \) are used to keep the equivalent impedances large enough at the nodes A and B.

The transfer function of this Cherry-Hooper amplifier is calculated as:

\[
A_v(s) \approx g_{m1}R_f \cdot \left( \frac{1}{g_{m3}R_f} \right) \cdot \frac{1 + \frac{scR_f}{g_{m1}}} {1 + \frac{scR_f}{g_{m1}} + \frac{scR_f}{g_{m2}}}
\]

(8)

where \( g_{m1} \) and \( g_{m3} \) are the transconductances of transistors \( M_1 \) and \( M_3 \); \( R_A \) and \( R_B \) are the equivalent resistances at nodes A and B; \( C_A \) and \( C_B \) are the parasitic capacitances at node A and node B, respectively. Equation (8) is only valid when \( R_A \gg R_f \), \( R_B \gg R_f \), \( C_B \gg C_f \), and \( C_B/g_{m3} \gg R_fC_f \). Since there is a capacitor bank added at node B, \( C_B \) is then considered much larger than \( C_A \) or \( C_f \). With this design, \( R_f \) is much smaller than \( R_A \) and \( R_B \). As a result, the capacitor bank can be used to tune the bandwidth of the TIA based low-pass biquad. As mentioned in Section 2, tuning the capacitor bank will also tune the center frequency of the complex filter. Note that when this complex filter is applied in a transceiver system, the bandwidth and the center frequency should be considered together.

From Equation (8), the DC gain of the TIA cell \( A_v0 \approx g_{m1}R_f \) can be obtained if \( g_{m3}R_f \gg 1 \) and \( g_{m3} \cdot R_A \cdot R_B \gg R_A + R_B + R_f \). With a gate tuning voltage, the \( R_f \) value of the PMOS transistor can be tuned linearly. As a consequence, the gain of the TIA cell can be tuned by two factors in this design: transconductance of transistor \( M_1 \) and the shunt resistance \( R_f \).

4. Implementation and Measurements results

Fig. 7 shows the photograph of the realized chip in 65nm CMOS technology. The area of the core circuit is 0.25mm×0.2mm.

![Fig. 7 Include all graphics in the manuscript file. Photograph of the 65nm implementation.](image)

The complex filter measurements were performed with Agilent N5247A network Analyzer and an oscilloscope 70604C. When the proposed complex filter is used in a
receiver, the I and Q outputs of the complex filter can be shortly connected. In this way, 4 input sinusoidal signals (0°, 90°, 180° and 270°) are applied to the filter and 2 output signals (0° and 180°) are generated. As a result, the image signal is rejected and the number of the ADCs can be reduced from 2 to 1. With 50mV peak-to-peak input signals, the output waveforms are shown in Fig.8, operating at (a) 1.85GHz and (b) -1.85GHz respectively. Clearly, an image rejection ration of about 17dB is obtained.

The output spectrum for a 1.85GHz sinusoidal signal is presented in Fig. 9, showing a more than 45dB SFDR performance of the complex filter.

Fig. 10 shows the measured gain response of the reported complex filter with a tuning center frequency from 500MHz to 1.85GHz. It achieves an approximately 14dB at the center frequency for all settings. The programmable capacitor bank, shown as in Fig. 11, has 7 settings from 50~350fF. Switches are added both on positive and negative nodes to maintain the symmetry characteristic of the circuit. Its layout has been carefully taken care to obtain more accurate value. Parasitic capacitance coming from the switches and metal paths has been considered as well. From the measurements, as shown in Fig 10, 7 different settings are obtained by using this capacitor bank.
The center frequency ($f_c$), the bandwidth (BW) and the center frequency IRR of each setting are listed in Table I. Almost all the ratios of $f_c$ and BW for each setting are around 1 except for $f_c > 1.4$GHz. This exception comes from the intrinsic bandwidth property of the TIA based low-pass biquad. The IRR can reach higher than 17dB at each center frequency for all settings.

Table I  $f_c$, BW and IRR for each setting

<table>
<thead>
<tr>
<th>load capacitor(FF)</th>
<th>$f_c$ (GHz)</th>
<th>BW(GHz)</th>
<th>IRR@$f_c$</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>1.85</td>
<td>1.02</td>
<td>17.6</td>
</tr>
<tr>
<td>100</td>
<td>1.56</td>
<td>1.22</td>
<td>25.5</td>
</tr>
<tr>
<td>150</td>
<td>1.32</td>
<td>1.37</td>
<td>24.4</td>
</tr>
<tr>
<td>200</td>
<td>1.09</td>
<td>1.21</td>
<td>18.8</td>
</tr>
<tr>
<td>250</td>
<td>0.7</td>
<td>0.7</td>
<td>18</td>
</tr>
<tr>
<td>300</td>
<td>0.61</td>
<td>0.61</td>
<td>22.5</td>
</tr>
<tr>
<td>350</td>
<td>0.53</td>
<td>0.52</td>
<td>17</td>
</tr>
</tbody>
</table>

Table II presents the comparison of the proposed complex filter to other published state-of-the-art results. The FoM used in Table II is shown in Equation (9). It takes both the BW and $f_c$ into account to evaluate the performances. Compared with other complex filters, clearly this work has the highest center frequency with the highest bandwidth and the smallest area. Accordingly, it is very suitable for broadband applications with a high area-efficiency, such as 5G systems.

$$\text{FoM} = \frac{\text{power}}{(N. \text{ poles}) \times (\text{BW}^2 + f_c^2) \times \text{SFDR}}$$  \hspace{1cm} (9)

Table II. Comparison with the state-of-the-art results

<table>
<thead>
<tr>
<th>Reference</th>
<th>[13]</th>
<th>[17]</th>
<th>[5]</th>
<th>[19]</th>
<th>[20]</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Process</td>
<td>CMOS (nm)</td>
<td>180</td>
<td>180</td>
<td>130</td>
<td>90</td>
<td>180</td>
</tr>
<tr>
<td>Topology</td>
<td>Active RC</td>
<td>Active RC</td>
<td>Active RC</td>
<td>Gm-C</td>
<td>Gm-C</td>
<td>TIA</td>
</tr>
<tr>
<td>Order</td>
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<td>3</td>
<td>4</td>
<td>4</td>
<td>7</td>
<td>2</td>
</tr>
<tr>
<td>SFDR(dB)</td>
<td>65.8</td>
<td>52.7</td>
<td>40</td>
<td>-</td>
<td>62</td>
<td>45</td>
</tr>
<tr>
<td>$f_c$ (MHz)</td>
<td>3</td>
<td>2</td>
<td>1.65</td>
<td>0-6</td>
<td>3</td>
<td>500-1850</td>
</tr>
<tr>
<td>BW (MHz)</td>
<td>1</td>
<td>1</td>
<td>0.9</td>
<td>0.25-2.5</td>
<td>2.5-3.5</td>
<td>500-1370</td>
</tr>
<tr>
<td>Gain</td>
<td>0.01</td>
<td>0.81</td>
<td>8.1</td>
<td>-</td>
<td>0</td>
<td>0.2-20</td>
</tr>
<tr>
<td>IRR(dB)</td>
<td>56</td>
<td>34</td>
<td>18.33</td>
<td>55</td>
<td>53</td>
<td>17</td>
</tr>
<tr>
<td>area(mm²)</td>
<td>0.4</td>
<td>0.51</td>
<td>0.24</td>
<td>0.7</td>
<td>0.37</td>
<td>0.05</td>
</tr>
<tr>
<td>VDD(V)</td>
<td>1.8</td>
<td>0.4</td>
<td>0.7</td>
<td>1.2</td>
<td>2.3</td>
<td>1.4</td>
</tr>
<tr>
<td>Power (mW)</td>
<td>1</td>
<td>0.0656</td>
<td>0.35</td>
<td>2.5-3.6</td>
<td>7.36</td>
<td>33.6-40.6</td>
</tr>
<tr>
<td>FoM(pJ)</td>
<td>0.041</td>
<td>0.023</td>
<td>0.46</td>
<td>-</td>
<td>0.18</td>
<td>0.049</td>
</tr>
</tbody>
</table>

5. Conclusion

A novel TIA based 2nd order broadband active complex filter up to 2.3GHz has been presented. The center frequency of the complex filter is programmable from 500MHz to 1.85GHz. The complex biquad has a voltage gain of 0-20dB. An IRR of higher than 17dB is achieved at each center frequency when the gain is 14dB, consuming less than 40.6mW power. The complex filter is implemented in a standard 65nm CMOS technology and has a core area of 0.05 mm².

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References
