Low-Distortion Continuous-Time Integrated Filters for Low Frequency Applications

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Abstract A technique for designing high-linearity continuous-time integrated filters for low frequency applications using transconductance network is presented. With this technique, an area-efficient transconductance network is used to replace the large resistors in low-frequency integrator-based continuous-time filters. The key challenge with this technique is in determining and realizing the appropriate relationship between the transconductance circuit and the nonideal frequency-dependent properties of the opamp. Simulation results for a third-order Bessel low-pass filter designed in a 0.5u CMOS process with a cutoff frequency of 5 KHz provide a THD in excess of 70 dB for a 2V p-p output with the 5 V power supply.

I Background

As the media between the "digital world" and the "analog world", continuous-time filters are required in many mixed signal systems. Of particular interest are integrated continuous-time filters that can be used for antialiasing and reconstruction. Gm-C filter is the most popular technology to realize these continuous-time filters. However, there are several limitations along with Gm-C filters and some of them are very hard to be overcome. The first limitation is the process and temperature dependency. It is important to precisely control the Gm value when implement Gm-C continuous-time filters and that Gm values are strongly depend on the process and temperature. Although there are some techniques to compensate this dependency, the compensating circuit will increase the complicity of the overall design and thus increase the expense. Another limitation is its poor linearity. Because Gm-C filters use nonlinear elements, this limitation is inherent in this structure and no known solutions for high linear applications. Compared with the Gm-C approach, the RC filter has much higher linearity. However, when RC filter requirements specify low-frequency poles, the area for integrating the correspondingly large RC time constants will be unacceptably large. That is the why there it is seldom used in audio frequencies. The magnitude of the problem can be appreciated by considering, as an example, a lossy first-order integrator as depicted in Fig. 1.

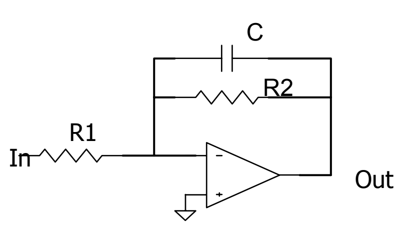


Figure 1 First Order Integrator

If zero DC-loss and a low cutoff frequency are required, the large RC time product necessitates either a large resistor or a large capacitor.

As we know $f_0 = \frac{\omega_0}{2\pi} = \frac{1}{2\pi RC}$,

If choose f_0=1 KHz,

$$RC = \frac{1}{2\pi f_0} = 1.59 \times 10^{-4},$$

Assume that the capacitance density and resistance density are $0.9 fF/\mu m^2, 30\Omega/\square$, respectively. If we set $A_R = A_c$, and choose the minimum resistor sheet squire as 1.2μm×1.2μm, (for AMI05 process), then we can approximate the total area needed to realize this RC product as $18.42\times10^4 \mu m^2$. When considering the design of a third order filter, the area for RC product will be approximately $1.5\,mm^2$. That is too large to be integrated on chip. However, on the other hand, if this area can be managed, RC filter can be viable for low frequency applications.

Observation

We observed that in figure 1, R1 and R2 act as transconductors because the input signal is voltage and output signal is current. So the question is that if we can find some other transconductors with the same transconductance value, small area and high linearity. The following is several kinds of transconductance network:

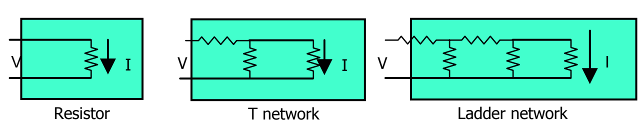


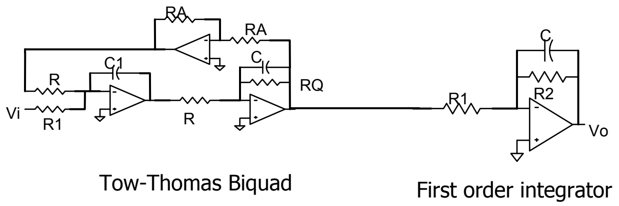
Figure 2 Three kinds of transconductance network

If the proper values are chosen, these three transconductance network will have the same transconductance value, high linearity and progressively small area.

II The design of A third-order Bessel Low-pass Filter

This design is presented as an example to show how this new technique, transconductance network, is used to build integrated continuous-time filters. The analysis of the relationship between the transconductance network and a non-ideal opamp is the high light in this design.

Because in Tow-Thomas biquad, each resistor has terminals connected to virtual ground, a first order filter and Tow-Thomas biquad are chosen to realize this third order Bessel filter. The filter structure is like below:



$$\frac{V_o}{V_i}=\frac{\frac{R}{R_1}(\frac{1}{R^2C^2})}{S^2+\frac{S}{R_QC}+\frac{1}{R^2C^2}}(1)\qquad \frac{V_o}{V_i}=\frac{-1/R_1C}{S+1/R_2C}(2)$$

Figure 3 the structure of the third-order Bessel low-pass filter

The overall transfer function is the production of the equation (1) and (2). The cutoff frequency is designed at 5 KHz with zero DC loss. For this conventional RC filter, our strategy is to replace all the large resistors with the corresponding transconctance network.

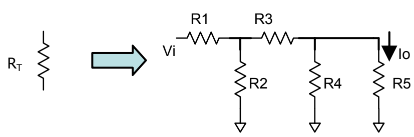


Figure 4 replacement of the transconductance network to huge R

$$R_T=R_1(R_3+R_5+\frac{R_3}{R_4}R_5)(\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3})$$

$$If\quad R_1>>R_2,R_3>>R_4,R_3>>R_2,R_5>>R_4$$

$$R_T\approx R_5\cdot\frac{R_3}{R_4}\cdot\frac{R_1}{R_2}$$

Although those huge resistors can replaced by transconductance networks, it dose not mean that any transconductance network that has the same transconductance value can work well with a non-ideal opamp. In practical design, it should be very careful to build the transconductance network with the consideration of the non-ideality of the opamp used in the filter. There is some non-ideality of opamp such as finite open loop gain, finite GB and limit linearity region will affect the performance of the filter. In order to understand that effect of non-ideal factors, the Y-Δ transformation is taken in this analysis.

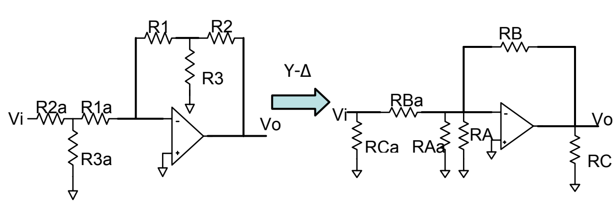


Figure 5 Y-Δ transformation of transconductance networks

$$R_A=\frac{R_1R_2+R_2R_3+R_1R_3}{R_2}\Rightarrow R_A\approx R_1\,(R_1>>R_3\;\;R_2>>R_3)$$

$$R_{Aa}=\frac{R_{1a}R_{2a}+R_{2a}R_{3a}+R_{1a}R_{3a}}{R_{2a}}\Rightarrow R_{Aa}\approx R_{1a}\;\;(R_{1a}>>R_{3a}\;\;R_{2a}>>R_{3a})$$

$$R_B=\frac{R_1R_2+R_2R_3+R_1R_3}{R_3}\qquad R_{Ba}=\frac{R_{1a}R_{2a}+R_{2a}R_{3a}+R_{1a}R_{3a}}{R_{3a}}$$

$$R_C=\frac{R_1R_2+R_2R_3+R_1R_3}{R_1}\qquad R_{Ca}=\frac{R_{1a}R_{2a}+R_{2a}R_{3a}+R_{1a}R_{3a}}{R_{1a}}$$

$$\frac{V_o}{V_i}=\frac{-G_{Ba}}{G_B+\frac{G_B+G_{Ba}+G_A+G_{Aa}}{A(s)}}$$

where A is the open loop gain of the OpAmp and G_B,G_{Ba},G_A,G_{Aa} are effective conductance of RB, RBa, RA and RAa, respectively. Form above equations, it can be shown that $R_B\,and\,R_{Ba}$ are much larger than $R_A\,and\,R_{Aa}$ and the total conductance $G_B+G_{Ba}+G_A+G_{Aa}$ will change the ideal transfer function and therefore will affect the performance of the filter as long as the open loop gain A cannot reach infinite value. Remember the equations form above analysis $R_T\approx R_5\cdot\frac{R_3}{R_4}\cdot\frac{R_1}{R_2}$, where RT is often above several Meg ohms when in the low frequency applications. The values of R2 and R1 should be chosen to be much bigger than the value of R3 to satisfy the assumption. In order to make these conductors negligible, however, R3 cannot be too small, or R1 will not be big enough to make R_A to some extend value and therefore G_B,G_{Ba},G_A,G_{Aa} will degrade the transfer function. After simulations, T-transconductrance network is used and the optimum ratio of the R1 over R2 is 10000/60.

The next step is to find the relationship between the finite GB and frequency response.

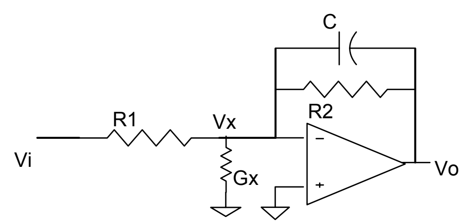


Figure 6 analysis of the effect due to the finite GB

$$\begin{cases}V_x(SC+G_1+G_2+G_x)=V_iG+V_o(SC+G_2)\\ V_o=\frac{GB}{-S}V_x\end{cases}$$

$$\therefore\frac{V_o}{V_i}=\frac{-G_1}{SC+G_2+\frac{S}{GB}(SC+G_1+G_2+G_x)}$$

from above equation, if Gx>>G1 and Gx>>G2,

$$\frac{V_o}{V_i}\approx\frac{-G_1}{S(C+\frac{G_x}{GB})+G_2}$$

It is clear that if GB is not big enough to satisfy the assumption that Gx/GB<<C, the frequency response will be deteriorated. Our goal is to minimize the error due to this finite GB. We need to make a compromise the considerations of the GB and capacitor values. After simulation, it shows that when Gx/GB less than 20% of C, this error can be neglected. The following table shows the capacitor values and corresponding GB value that satisfy that assumption.

C (pF)	GB needed for opm(M)
0.5	195.9
1	115.2
4	56.9
8	46.8
12	43.1
16	41.5

Table 1 the relationship between C and GB of opamp

Because this filter is designed to drive large load, 10 K resistor paralleled with 30 fF capacitor, we don't want to put big pressure on the GB design. Finally, the capacitor 8 pF and 46.8 MHz Gb are chosen. Out simulation will show the GB we design is much larger than this value.

All the specifications are presented; high DC gain, large linear region, high GB and large driving ability. We choose two stages, cascode large swing structure. For bias circuit, we choose non-sensitive-to-power supply circuit because of its tolerance to large deviation of the power supply and resistor. The schematic as below:

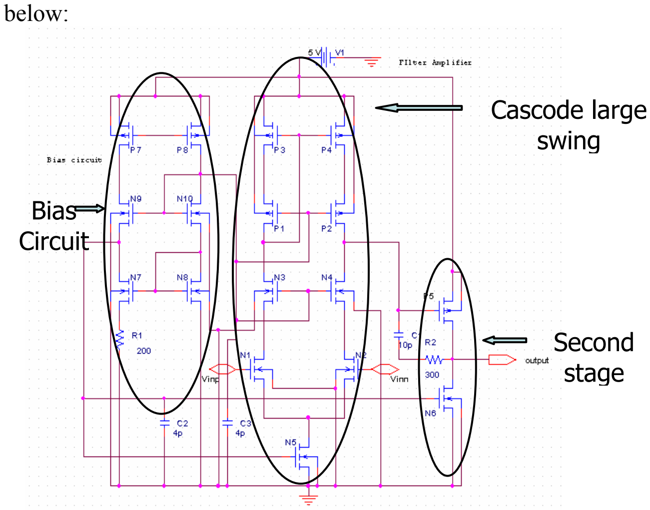


Figure 7 OpAmp schematic

The last step is to determine all the filter coefficients. The standard parameters are provided by the normalized Bessel transfer function. The frequency scale is used to de-normalize the normalized transfer function and our design cutoff frequency is 5 KHz. After the de-normalization, the equation set is shown as below:

$$\begin{cases} R_Q = \dfrac{1}{a_1\omega_0 C} \\ R = \dfrac{1}{\sqrt{a_0\omega_0^2 CC_1}} \\ R_2 = \dfrac{1}{b_0\omega_0 C} \\ R_1 = \sqrt{\dfrac{1}{a_0 b_0\omega_0^3 RC^2C_1}} \end{cases}$$

where a_0, a_1, b_0 are transfer parameters and

$a_0 = 2.09482, a_1 = 2.095588$

$b_0 = 1.32268, \omega_0 = 2\pi \times 5000$

C is chosen as 8 pF as above analysis. Because when opamp operates at the edge of the linear region, the linearity will be deteriorated seriously, we need to limit the gain of each stage. After the calculation of Matlab, all the values are set.

R=2.43 M ohm, R1=2.7 M ohm, R2=3 M ohm, RQ=1.9 M ohm
For the T-network design,

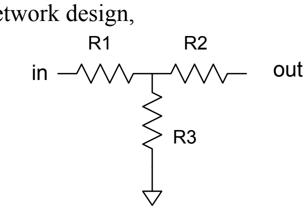


Figure 8 designed T-network

Set R1=10 K ohm and R3= 60 ohm, so

$$R_2 = \frac{R_{desire}}{\frac{R_1}{R_3}},,, R_1 >> R_3, R_2 >> R_3$$

III. The Simulation Results

The Hspice simulation shows that the opamp and filter works well and meets all our specifications. These simulations are done with the deviation of the resistors and capacitors.

Dc gain	GB	Output swing	PM
98 dB	105 M	>3 V p-p	80 degree

Table 2 the performance of the opamp in the filter

	DC loss	Cutoff Freq	Output swing	THD
Design	0	5 KHz	>2v p-p	>70 dB
Simulation	-210m dB	4.89 KHz	>2v p-p	73.3 dB

Table 3 the performance of the whole filter

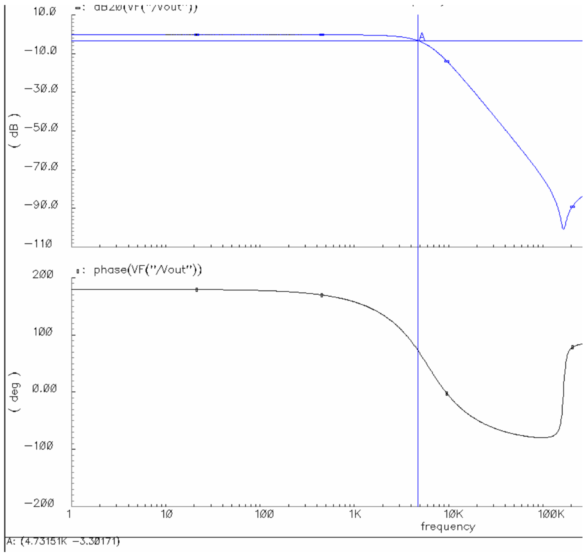


Figure 9 the frequency response

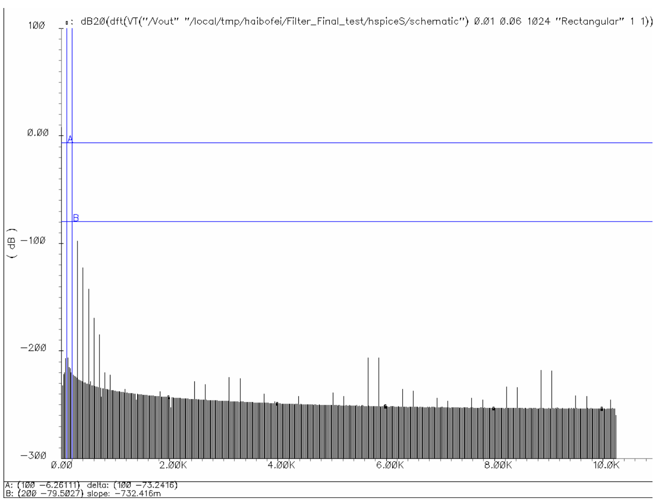


Figure 10 the spectral performance

IV Conclusion

A strategy for realizing integrated high-linearity low-frequency filters has been introduced. With this approach, a transconductance network replaces the large resistors which are needed for low frequency applications. Duo to this new technique, the total area used for realizing large effective RC time constants is dramatically reduced while maintaining high linearity.

A prototype third-order Bessel filter designed for a band-edge of 4 KHz was designed and simulated. The total area for implementing this filter is $685\mu m \times 587\mu m$. It is much smaller than the conventional RC filter as we approximated as 1.5 mm^2. Simulation results showed a THD of 73.3 dB with an output of 2V p-p can be obtained. This high linearity is obtained with the single ended structure and the third harmonic distortion is below 98 dB. So when using this technique in fully differential structure, the THD can be easily reach over 95 dB. Compared with most Gm-C filters with second harmonic at around 30 dB and total THD much worse than 60 dB, this technique proves a real big improvement. Another attractive advantage of this technique is that this approach can be used for other processes. It can be used for the process without switched capacitor, such as BiCMOS process. It can also be used for some other inexpensive process and that shows good economic potential.

V Reference:

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