High Frequency CMOS Gm-C Bandpass Filter Design

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ABSTRACT
This paper describes design of a high-frequency high-Q second-order Gm-C bandpass filter based on Nauta’s OTAs used for the RF sampling continuous-time bandpass ΔΣ AD modulator. By using 0.25μm CMOS process, a Gm-C bandpass filter with center frequency of 1GHz, bandwidth of 33MHz and Q factor of 30 may be feasible. We also discuss its nonlinearity, noise and power issues.

Keywords: CMOS, OTA, Gm-C Filter, Bandpass Delta-Sigma Modulator, RF Sampling

1. Introduction

A high-frequency, low-power continuous-time bandpass ΔΣ AD has been investigated for portable communication system applications such as WLAN, cellular phone [1, 2, 3]. A G_m-C bandpass filter (BPF) can be used as a high-frequency, low-power bandpass filter inside such a ΔΣ AD modulator.

For our design target of a RF direct sampling bandpass ΔΣ modulator [1, 2], its center frequency f_c should be as high as several hundreds MHz or higher, since f_c=2πg_m/C and C must be large enough for noise performance, the CMOS Operational Transconductance Amplifier (OTA) circuit has to have a large value of g_m with good high frequency characteristics. A suitable option of the OTA circuit with no internal nodes was presented by Nauta [5]. In this work, we will clarify high-frequency characteristics, Q, linearity, noise and power issues of a second-order bandpass filter as well as optimum Nauta’s OTA circuit design and tuning circuit considering parasitics in 0.25μm CMOS process. We have confirmed its center frequency of 1GHz and Q of 30 by SPICE simulation with BSIM3v3 parameters.

2. Analysis of Nauta OTA circuit

Fig.1 shows a Nauta’s OTA circuit [5, 6], where four inverters at the output provide both common mode stability and high differential mode output resistance. Since high-order poles are not introduced by the internal node parasitic capacitances, there an (almost) ideal integrator can be implemented; it is suitable for high frequency bandpass filter. By adjusting the power supply voltage Vdd for Inv4, Inv5 in Fig.1, the output resistance can be made large and very high DC gain can be realized. For the OTA design, nonlinearity and noise issues are also important and we will discuss them in the followings.

2.1 Nonlinearity Analysis of Nauta OTA

A dominant source of Nauta OTA nonlinearity is mobility reduction due to vertical field and parasitic capacitance in high frequency region. Electron and hole mobilities in MOSFETs can be expressed as

\[
\mu_n : = \frac{\mu_{n0}}{1 + \theta_n(V_{gm} - V_{thn})}
\]

\[
\mu_p : = \frac{\mu_{p0}}{1 + \theta_p(V_{gp} - |V_{thp}|)}
\]

The output currents from core inverters in Nauta
OTA (Fig. 2) are given by

\[ I_{od} = I_{o1} - I_{o2} = (I_{n1} - I_{p1}) - (I_{n2} - I_{p2}) = (I_{n1} - I_{n2}) + (I_{p2} - I_{p1}). \]

Using Taylor expansion, they are approximated by

\[ I_{n1} - I_{n2} \approx \frac{\beta_{0n} V_{on}(1 + \frac{1}{2}\theta_{n} V_{on})}{(1 + \theta_{n} V_{on})^2} V_{id} \]

\[ I_{p2} - I_{p1} \approx \frac{\beta_{0p} V_{op}(1 + \frac{1}{2}\theta_{p} V_{op})}{(1 + \theta_{p} V_{op})^2} V_{id} \]

Here

\[ V_{on} = V_{cm} - V_{thn}, \quad V_{op} = V_{DD} - V_{cm} - |V_{thp}| \]

\[ \beta_{0n} = \mu_{n0} C_{ox} \frac{W_n}{L_n}, \quad \beta_{0p} = \mu_{p0} C_{ox} \frac{W_p}{L_p}. \]

Now we have a differential output current as \( I_{od} \approx c_1 V_{id} + c_3 V_{id}^3 \), and we assume that \( \beta_{0n} \approx \beta_{0p} = \beta_0 \), \( V_{on} = V_{op} = V_o \), \( c_1 \approx 2\beta_0 V_o \), \( c_3 \approx -\frac{\theta_n}{8}(\theta_n + \theta_p) \), \( g_m(V_{id}) = \frac{\delta m}{V_{id}} = c_1 + 3c_3 V_{id}^2 \). Then we have the third-order distortion of Nauta OTA as

\[ \frac{c_3}{c_1} \approx -\frac{\theta_n + \theta_p}{16V_o} = -\frac{\theta_n + \theta_p}{8(V_{DD} - |V_{thp}| - V_{thn})}. \]  

2.2 Noise Analysis of Nauta OTA

In high frequency region, the dominant noise source is thermal noise from MOSFETs, and Fig. 3 is a noise model of Nauta OTA. The output equivalent current noise for one inverter can be expressed as

\[ \overline{i^2_{inv}} = 4\gamma_n kT g_m d f + 4\gamma_p kT g_p d f. \]

Here we assume that \( g_m n = g_m p \), \( \gamma_n = \gamma_p = \gamma_{inv} \) (noise factor), \( g_m n + g_m p = g_m_{inv} \). Then we have the total output equivalent current noise as follows:

\[ \overline{i^2_{od}} = 4\gamma_{inv} kT d f \sum_{i=1}^{6} g_m i \]

Considering the above-mentioned nonlinearity and noise performance, we have designed an OTA with 48dB DC gain (Fig. 4 shows frequency characteristics of an integrator using this OTA).

3. Bandpass Gm-C Filter Design

Active and passive filters can be used for bandpass filter design. Active filters can be classified into active RC filters which use operational amplifiers, and Gm-C filters which use OTAs [3]. RC phase filters [7] and LCR filters are passive-type, and a passive LCR filter can be a model of a Gm-C filter where
L and R are equivalently realized by OTAs to save chip area. The *Gm-C* filter operates in open-loop which is suitable for high frequency operation, and *gm* value can be adjusted automatically by an embedded tuning circuit. Our design target here is a *Gm-C* bandpass filter with the center frequency \( f_c = 1 \text{GHz} \), \( Q = 30 \) and \( A_v(1GHz) = 20 \text{dB} \).

### 3.1 Second-Order Bandpass Filter Design

Fig. 5 shows a second-order *Gm-C* filter which models an LCR filter; \( g_{m1} \) is the input OTA while the inductor is equivalently made by two OTA cells (\( g_m \)) and a capacitor \( C_2 \), and a resistor is by an OTA \( (g_{m2}) \). Then its transfer function is given by

\[
H(s) = \frac{V_{out}}{V_{in}} = \frac{g_{m1} s C_2}{g_m + g_{m2} s C_2} + 1. \tag{3}
\]

Here \( \omega_0 = \frac{g_m}{\sqrt{C_1 C_2}} \), \( Q = \frac{g_m}{g_{m2}} \sqrt{\frac{C_2}{C_1}} \), \( A = \frac{g_{m1}}{g_{m2}} \).

### 3.2 High Frequency Operation

For high frequency performance of the filter, parasitic capacitance effects cannot be ignored; the parasitic capacitances of OTA circuits used in the filter may be virtually magnified due to Miller effect. Assume that parasitic capacitances are concentrated at the input and output nodes of OTA circuits (Fig 6).

Parasitic capacitances and capacitors \( C_1, C_2 \) of the bandpass filter determine its high frequency performance, and the whole capacitances \( C_{eff1}, C_{eff2} \) are expressed as

\[
C_{eff1} \approx C_1 + C_{g_{s, gm}} + C_{g_{s, gm2}} \\
C_{eff2} \approx C_2 + C_{g_{s, gm}}.
\]

The transfer function of the filter is given by

\[
H(s) = \frac{g_{m1} s C_{eff2}}{g_m s C_{eff1} + g_{m2} s C_{eff2} + 1}. \tag{4}
\]

Thus parasitic capacitances deviate \( \omega_0 \) and \( Q \) values as \( \omega_0 = \frac{g_m}{\sqrt{C_{eff1} C_{eff2}}} \), \( Q = \frac{g_m}{g_{m2}} \sqrt{\frac{C_{eff2}}{C_{eff1}}} \). From our calculation, \( C_{eff1} \approx 0.89pF \), \( C_{eff2} \approx 0.87pF \), and we found that to achieve our design target, \( g_m > 5.58mS \), \( g_{m1} > 1.86mS \), \( g_{m2} > 0.186mS \) are necessary. Fig 7 shows SPICE simulated AC characteristics of a modeled LCR filter and our designed *Gm-C* BPF filter.

### 3.3 Linearity Performance

Intermodulation distortion is the most important distortion factor for the bandpass filter, and the third-order intermodulation distortion \( (IM3) \) determines its linearity because the designed filter circuit is fully differential. When the input signal is given by

\[
V_{in}(t) = A cos(2\pi f_1 t) + A cos(2\pi f_2 t),
\]

we obtained \( IM3 \approx -41.2dB \) for the input frequencies \( f_1 = 1 \text{GHz}, f_2 = 1.01 \text{GHz} \) and the input ampli-
3.4 Noise Performance

The noise power of an RC circuit in whole band (0 < f < \infty) is known as kT/C, and we have found that of an LCR circuit is also kT/C in whole band (which is independent of R, L values). Also the noise power of a bandpass filter in whole band is given by $Q \cdot kT/C$ [5]. However we do not have to consider the noise power in whole band, but we consider just the noise in the signal band ($f_0 - BW/2 < f < f_0 + BW/2$) and we have derived its expression.

Fig. 9 shows the noise model of the bandpass filter, where $I_{nrx}$, $x = 1, 2, 3, 4$ are the output noise currents. The output noise density can be obtained as follows:

\[
\begin{align*}
V_{in}^2 & = \frac{1}{g_{m1}} \left[ I_{n1} + I_{n2} + I_{n3} + I_{n4} \left( \frac{g_m}{sC_2} \right)^2 \right].
\end{align*}
\]

For transfer function expressed by eq.(3), the input equivalent noise power spectral density can be written as

\[
\begin{align*}
V_{in}^2 & = \frac{1}{g_{m1}} \left[ I_{n1} + I_{n2} + I_{n3} + I_{n4} \left( \frac{g_m}{sC_2} \right)^2 \right].
\end{align*}
\]

In high frequency region, we consider only thermal noise and we assume that $I_{nrx}$, $x = 1, 2, 3, 4$ are independent of frequency. Then the input equivalent noise in the signal band ($f_0 - BW/2 < f < f_0 + BW/2$) can be written as

\[
\begin{align*}
\text{Noise}_{rms}^2 & = \int_{f_0 - BW/2}^{f_0 + BW/2} \text{Noise}_{den}^2 df.
\end{align*}
\]

Here, Noise$^2_{rms}$ is the input equivalent noise power spectral density. Note that output noise current can be expressed as $I_{nrx} = 4\gamma kTgmi$ ($x = 1, 2, 3, 4$, $gmi = g_m, g_{m1}, g_{m2}$), where $\gamma = \frac{2}{3}$ is a noise factor of $g_m$ cell which will be much higher for short channel device. Then the input equivalent noise of this filter is given by

\[
\begin{align*}
\text{Noise}_{rms}^2 & = 4\gamma kT \frac{BW}{g_{m1}} \left[ 1 + \frac{g_{m2}}{g_{m1}} + \frac{g_m}{g_{m1}} \right] \left( \frac{g_m}{sC_2} \right)^2.
\end{align*}
\]

From eq. (3) 

\[
\begin{align*}
\frac{g_{m2}}{g_{m1}} & = \frac{1}{A}, \\
\frac{g_m}{g_{m1}} & = \frac{Q}{A} \sqrt{\frac{C_2}{C_1}}, \\
\left( \frac{g_m}{C_2} \right)^2 & = (\omega_0 \sqrt{\frac{C_1}{C_2}})^2.
\end{align*}
\]

Input equivalent noise in the signal band can be written as

\[
\begin{align*}
\text{Noise}_{rms}^2 & = \frac{8}{3} \gamma kT \frac{BW}{g_{m1}} \left[ 1 + \frac{1}{A} + \frac{Q}{A} \sqrt{\frac{C_1}{C_2}} \right] \\
& + \frac{1}{A} Q \omega_0^2 \sqrt{\frac{C_1}{C_2}}.
\end{align*}
\]

Fig. 8. Transient response of our $Gm-C$ BPF.
Since $BW << f_0$ in this work, the equation can be approximated as

$$\text{Noise}_{rms}^2 \approx \frac{4}{3} \gamma kT \frac{BW}{g_{m1}} [1 + \frac{1}{A} + \frac{Q}{A} \frac{C_1 + C_2}{\sqrt{C_1 C_2}}]. \quad (6)$$

We see from eq.(6) that as $g_{m1}$ increases, noise decreases and gain (A) increases. Note that increase of $g_{m1}$ leads to increase of power; noise and power are trade-off. Also remark that if a high Q filter is required, it suffers from noise increase. In this work, we obtain the output noise density of $17.2 \frac{nV}{\sqrt{Hz}}$ from SPICE simulation, where we replace $C_1$, $C_2$ with $C_{\text{eff1}}$, $C_{\text{eff2}}$ respectively.

### 3.5 Tuning of Center Frequency and Q

On-chip $Gm-C$ bandpass filter requires tuning circuit for its center frequency and Q value to absorb process variation, supply voltage change and temperature change. In this work, we use PLL for center frequency tuning and Modified-LMS topology for Q tuning.

Fig.10 shows the whole tuning circuit for the center frequency and Q. Since tuning of the center frequency and that of Q may interact each other when both tunings operate simultaneously, only one tuning operates at one time (first center frequency tuning, then Q tuning are performed). Fig.11 and Fig.12 show the simulation result of a bandpass filter after tuning where 1% mismatch of MOSFETs is included.

![Fig. 10. Tuning circuit for center frequency and Q of our Gm-C BPF.](image)

### 3.6 Whole $Gm-C$ Bandpass Filter

Table 1 shows the SPICE simulation results of our whole $Gm-C$ bandpass filter.

![Fig. 11. Simulated gain characteristics of our Gm-C BPF after tuning.](image)

### 4. Conclusion

We have shown the feasibility of a very high frequency CMOS $Gm-C$ bandpass filter for an RF sampling continuous-time bandpass $\Delta\Sigma$ modulator. Usually we use an LCR filter to implement a very
high frequency bandpass filter on CMOS IC, but it is area consuming and hence costly. We have investigated to replace passive components with active ones in a very high frequency filter where nonlinearity, noise and power issues should be taken care of. We have shown by simulation with 0.25\(\mu\)m CMOS BSIM3v3 parameters that by use of Nauta OTS circuit a bandpass filter with center frequency of 1 GHz and Q of 30 would be feasible. As a next step, we will realize this filter as a real chip.

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References


