EE 508

Lecture 39

Transconductor Design and Applications
Simple single-ended OTA

\[ I_0 = (\beta_5 - \beta_6) V_{\text{in}}^2 \]
\[ + V_{\text{in}} \left( 2\beta_5 \left[ V_{\text{TN}} - V_{\text{TP}} - V_G \right] + 2\beta_6 \left[ V_{\text{TN}} - V_{\text{TP}} + V_{G1} \right] \right) \]
\[ + \beta_5 \left[ V_{\text{TP}} - V_{\text{TN}} + V_G \right]^2 - \beta_6 \left[ V_{\text{TP}} - V_{\text{TN}} + V_{G1} \right]^2 \]

If size devices so that \( \beta_5 = \beta_6 \) and \( V_G = V_{G1} \), this simplifies to

\[ I_0 = V_{\text{in}} \left( 4\beta_5 \left[ V_{\text{TN}} - V_{\text{TP}} - V_G \right] \right) \]

Note this behaves as a linear transconductor!

\[ g_m = 4\beta_5 \left[ V_{\text{TN}} - V_{\text{TP}} - V_G \right] \]

- Since both \( M_2 \) and \( M_3 \) are driven, this is a power-efficient method for generating a given \( g_m \)
- Behavior will degrade with bulk-dependent threshold voltages of n-channel devices
- Would like to generate \( V_G \) and \( V_{G1} \) independent of \( V_{\text{DD}} \)
**V\textsubscript{DD} Independent Bias Generators**

Two widely-used V\textsubscript{DD} independent bias generators (start-up ckts not shown)

\[ I_{D2} = I_{D1} = MI_{D3} \]

\[ I_{D4} = I_{D3} = \frac{\mu C_{OX} W_4}{2L_4} (V_{01} - V_{Tn})^2 \]

\[ I_{D1} = \frac{\mu C_{OX} W_1}{2L_1} (V_{01} - V_X - V_{Tn})^2 \]

\[ V_X = I_{D1} R_1 \]

Define:

M is the M\textsubscript{3}:M\textsubscript{2} mirror gain

\[ \beta_k = \frac{\mu C_{OX} W_k}{2L_k} \]

\[ V_X = \frac{1}{R_1} \left( \sqrt{\frac{1}{M\beta_4}} - \sqrt{\frac{1}{\beta_1}} \right)^2 \]

\[ V_{01} = V_{Tn} + \left( \frac{1}{MR} \right) \left( \frac{1}{\beta_4} - \frac{1}{\sqrt{M\beta_1\beta_4}} \right) \]
$V_{\text{DD}}$ Independent Bias Generators

Two widely-used $V_{\text{DD}}$ independent bias generators (start-up ckts not shown)

\[ V_X = \frac{1}{R_1} \left( \sqrt{\frac{1}{M \beta_4}} - \sqrt{\frac{1}{\beta_1}} \right)^2 \]

\[ V_{01} = V_{Tn} + \left( \frac{1}{MR} \right) \left( \frac{1}{\beta_4} - \frac{1}{\sqrt{M \beta_1 \beta_4}} \right) \]

Define:

$M$ is the $M_3:M_2$ mirror gain

$\beta_k = \frac{\mu C_{\text{OX}} W_k}{2L_k}$

Observe $V_X$ is independent of both $V_T$ and $V_{DD}$

Offers some attractive properties when used as part of a temperature sensor as well

Requires Start-up Circuit

May need compensation for stability

Pseudo-static operation so frequency response of little concern
**$V_{DD}$ Independent Bias Generators**

Two widely-used $V_{DD}$ independent bias generators (start-up ckts not shown)

\[
\begin{align*}
I_{D2} &= I_{D1} = I_{D5} = M I_{D3} \\
I_{D4} &= I_{D3} = \frac{\mu C_{OX} W_4}{2L_4} (V_{01} - V_{Tn})^2 \\
I_{D1} &= \frac{\mu C_{OX} W_1}{2L_1} (V_{01} - V_Y - V_{Tn})^2 \\
I_{D5} &= \frac{\mu C_{OX} W_5}{2L_5} (V_Y - V_{Tn})^2
\end{align*}
\]

4 equations and 4 unknowns \{\(I_{D1}, I_{D3}, V_{01}, V_Y\)\}

Define:

\[
V_{01} = V_{Tn} \left( \sqrt{M} \left( 1 + \frac{W_1 L_5}{W_5 L_4} \right)^{-2} \right) \\
\frac{1}{\sqrt{M} \left( 1 + \frac{W_1 L_5}{W_5 L_4} \right)^{-1}}
\]

\[
V_Y = V_{Tn} \left[ \left( \frac{W_5 L_1}{W_1 L_5} - 1 \right) + \frac{1}{\left( 1 + \frac{W_1 L_1}{W_5 L_4} \right)} \right] \left( \frac{W_5 L_1}{W_1 L_5} - 1 \right)
\]

Define:

\[
M \text{ to be the } M_3:M_2 \text{ mirror gain}
\]
$V_{DD}$ Independent Bias Generators

Two widely-used $V_{DD}$ independent bias generators (start-up ckts not shown)

Define:
M to be the $M_3:M_2$ mirror gain

Note $V_{01}$ and $V_Y$ are dependent only upon $V_T$

Applications well beyond this biasing requirement

Requires Start-up Circuit

May need compensation for stability

Pseudo-static operation so frequency response of little concern
$V_{DD}$ Independent Bias Generators

Two widely-used $V_{DD}$ independent bias generators (start-up ckts not shown)

$V_X = \frac{1}{R_1} \left( \sqrt{\frac{1}{M\beta_4}} - \sqrt{\frac{1}{\beta_1}} \right)^2$

$V_{01} = V_{Tn} + \left( \frac{1}{MR} \right) \left( \frac{1}{\beta_4} - \frac{1}{\sqrt{M\beta_1\beta_4}} \right)$

where $\beta_k = \frac{\mu C_{OX} W_k}{2L_k}$ and M is the $M_3:M_2$ mirror gain.
Transconductance Linearization Strategies

\[
\begin{align*}
\text{Recall with } R_S &= 0 \\
\text{Widely used source degeneration}
\end{align*}
\]
Transconductance Linearization Strategies

\[ I_{D1} = \beta (V_1 - V_{S1} - V_T)^2 \]
\[ I_{D2} = \beta (V_2 - V_{S2} - V_T)^2 \]
\[ V_{S1} - I_{D1} R_{S1} = V_{S2} - I_{D2} R_{S2} \]
\[ I_{D1} + I_{D2} = I_T \]

With a straightforward analysis, we obtain the expression

\[ \sqrt{\frac{1}{\beta}} (\sqrt{I_T - I_{D1}} - \sqrt{I_{D1}}) + R_S (I_T - 2I_{D1}) = V_d \]

The first term on the right is the nonlinear term of the original source coupled pair and the second is linear in \( I_{D1} \)

The larger the second term becomes, the more linear the transfer characteristics are
Transconductance Linearization Strategies

There are a host of transconductance linearization strategies that have been discussed in the literature.

Some are shown below.

Many are strongly dependent upon a precise square-law model of the MOS devices and do not provide practical solutions when the devices are not square-law devices.

Analysis or simulation with a more realistic model is necessary to validate linearity and practical applications of these structures.
Transconductance Linearization Strategies

How good is the square-law model that we have been using for predicting filter performance?

It is reasonably good when analyzing structures whose linearity characteristics are not strongly dependent upon the device model.

The circuits considered to data are not particularly linear so the square-law model probably does a pretty good job of predicting their performance.

More accurate models are usually unwieldy for hand analysis.
Transconductance Linearization Strategies

Fig. 1 Linearised CMOS transconductance circuit
Transconductance Linearization Strategies

From CAS 2006 P 811  Jose Silva
Transconductance Linearization Strategies

Linearity Enhancement with Source Degeneration
Transconductance Linearization Strategies

Linearization with active source degeneration
CMOS transconductance amplifiers, architectures and active filters: a tutorial

E. Sánchez-Sinencio and J. Silva-Martínez

Abstract: An updated version of a 1985 tutorial paper on active filters using operational transconductance amplifiers (OTAs) is presented. The integrated circuit issues involved in active filters (using CMOS transconductance amplifiers) and the progress in this field in the last 15 years is addressed. CMOS transconductance amplifiers, non-linearised and linearised, as well as frequency limitations and dynamic range considerations are reviewed. OTA-C filter architectures, current-mode filters, and other potential applications of transconductance amplifiers are discussed.
Linearity compensation with cross-coupled feedback
Single-ended input TAs
Differential input OTAs

Differential input and output OTAs
Parasitic Capacitances and Floating Nodes

There is invariably a parasitic capacitance associated with every terminal of every element in a filter.

These parasitic capacitances can be significant in integrated filters.

These can be combined into a single parasitic capacitance on each node.
Parasitic Capacitances and Floating Nodes

A floating node is a node that is not connected to either a zero-impedance element or across a null-port.

Floating nodes are generally avoided in integrated filters because the parasitic capacitances on the floating nodes usually degrades filter performance and often increases the order of the filter.

Some filter architectures inherently have no floating nodes, specifically, most of the basic integrator-based filters have no floating nodes.
Parasitic Capacitances and Floating Nodes
Parasitic Capacitances and Floating Nodes

No floating nodes!
Signal Swing in OTA Circuits

The signal swing for the basic bipolar OTA is limited to a few mV for reasonably linear operation.

This limited signal swing limits the use of the OTS.

The following circuit (with maybe a 100:1 or more attenuation) can be used to increase the input signal swing to the volt range and although it involves quite a few more components, the functionality can be most significant.

Program range is not affected by adding the attenuators.

\[ g_{meq} = \theta g_m \]

*Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial*

Randall L. Geiger and Edgar Sánchez-Sinencio
Programmable Filter Structures

It will be assumed that the transconductance gain can be programmed with either a dc current or a dc voltage.

\[ T(s) = \frac{g_m}{g_m + sC} \]

Programmable First-Order Low-Pass Filter
Programmable Filter Structures

Programmable First-Order High-Pass Filter

\[ T(s) = \frac{sC}{g_m + sC} \]