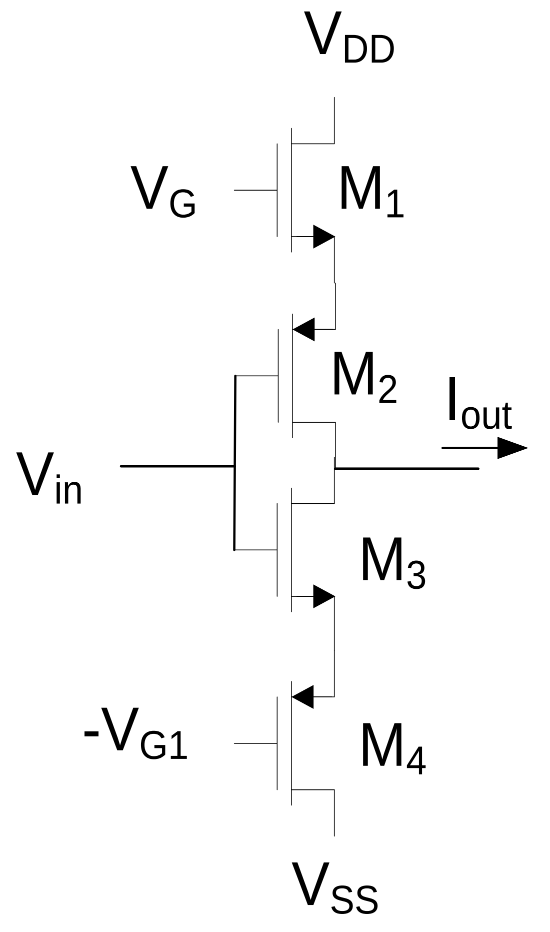
EE 508

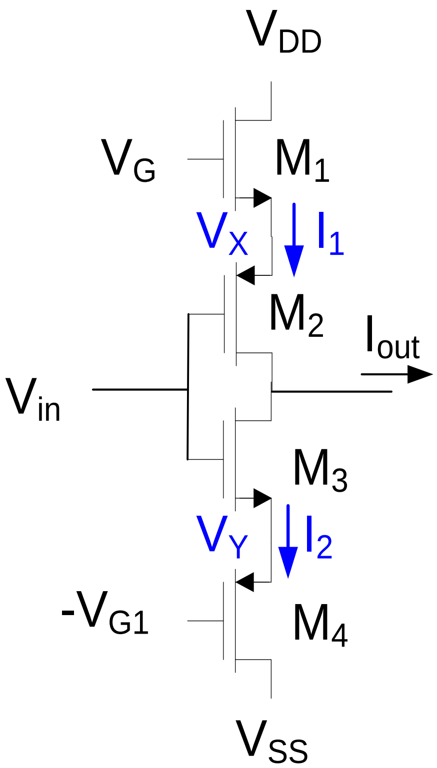
Lecture 35

Transconductor Design and Applications

Simple single-ended OTA



Simple single-ended OTA



$$\left.\begin{aligned} I_0 &= I_1 - I_2 \\ I_1 &= \beta_1\left(V_G - V_X - V_{Tn}\right)^2 \\ I_1 &= \beta_2\left(V_X - V_{in} + V_{Tp}\right)^2 \\ I_2 &= \beta_3\left(V_{in} - V_Y - V_{Tn}\right)^2 \\ I_2 &= \beta_4\left(V_Y + V_{G1} + V_{Tp}\right)^2 \end{aligned}\right\}$$

Taking the square root of the two I_1 equations

$$\sqrt{\frac{1}{\beta_1}}\sqrt{I_1} = \left(V_G - V_X - V_{Tn}\right)$$

$$\sqrt{\frac{1}{\beta_2}}\sqrt{I_1} = \left(V_X - V_{in} + V_{Tp}\right)$$

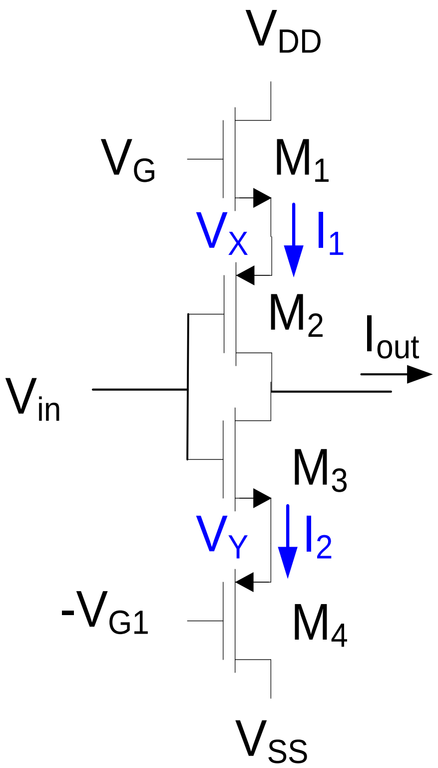
Adding these two equations, we obtain

$$\left(\sqrt{\frac{1}{\beta_2}} + \sqrt{\frac{1}{\beta_1}}\right)\sqrt{I_1} = \left(V_G - V_{in} + V_{Tp} - V_{Tn}\right)$$

Similarly, for the last two equations, obtain

$$\left(\sqrt{\frac{1}{\beta_3}} + \sqrt{\frac{1}{\beta_4}}\right)\sqrt{I_2} = \left(V_{G1} + V_{in} + V_{Tp} - V_{Tn}\right)$$

Simple single-ended OTA



$$\left.\begin{aligned} &I_0 = I_1 - I_2 \\ &\left(\sqrt{\frac{1}{\beta_2}} + \sqrt{\frac{1}{\beta_1}}\right)\sqrt{I_1} = \left(V_G - V_{in} + V_{Tp} - V_{Tn}\right) \\ &\left(\sqrt{\frac{1}{\beta_3}} + \sqrt{\frac{1}{\beta_4}}\right)\sqrt{I_2} = \left(V_{G1} + V_{in} + V_{Tp} - V_{Tn}\right) \end{aligned}\right\}$$

Squaring the last two equations we obtain

$$I_1 = \beta_5\left(V_G - V_{in} + V_{Tp} - V_{Tn}\right)^2$$

$$I_2 = \beta_6\left(V_{G1} + V_{in} + V_{Tp} - V_{Tn}\right)^2$$

Define

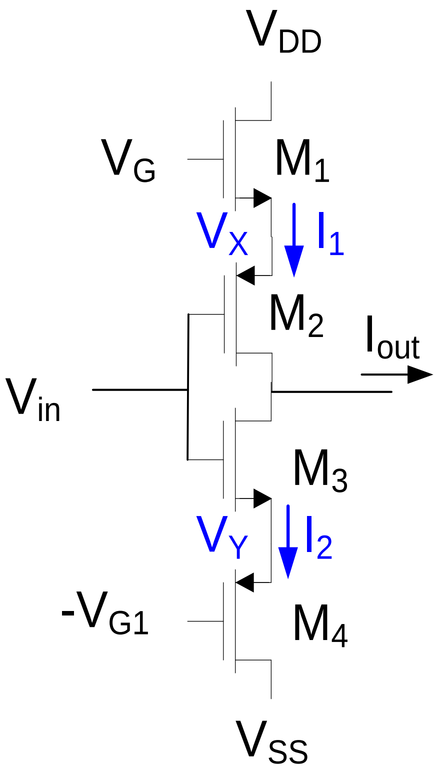
$$\left(\sqrt{\frac{1}{\beta_2}} + \sqrt{\frac{1}{\beta_1}}\right) = \sqrt{\frac{1}{\beta_5}}$$

$$\left(\sqrt{\frac{1}{\beta_3}} + \sqrt{\frac{1}{\beta_4}}\right) = \sqrt{\frac{1}{\beta_6}}$$

Equating the difference to I_0, we obtain

$$\begin{aligned} I_0 = &\left(\beta_5 - \beta_6\right)V_{in}^2 \\ &+ V_{in}\left(2\beta_5\left[V_{Tn} - V_{Tp} - V_G\right] + 2\beta_6\left[V_{Tn} - V_{Tp} + V_{G1}\right]\right) \\ &+ \beta_5\left[V_{Tp} - V_{Tn} + V_G\right]^2 - \beta_6\left[V_{Tp} - V_{Tn} + V_{G1}\right]^2 \end{aligned}$$

Simple single-ended OTA



$$I_0 = (\beta_5 - \beta_6) V_{in}^2 + V_{in}\left(2\beta_5\left[V_{Tn} - V_{Tp} - V_G\right] + 2\beta_6\left[V_{Tn} - V_{Tp} + V_{G1}\right]\right) + \beta_5\left[V_{Tp} - V_{Tn} + V_G\right]^2 - \beta_6\left[V_{Tp} - V_{Tn} + V_{G1}\right]^2$$

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

$$I_0 = V_{in}\left(4\beta_5\left[V_{Tn} - V_{Tp} - V_G\right]\right)$$

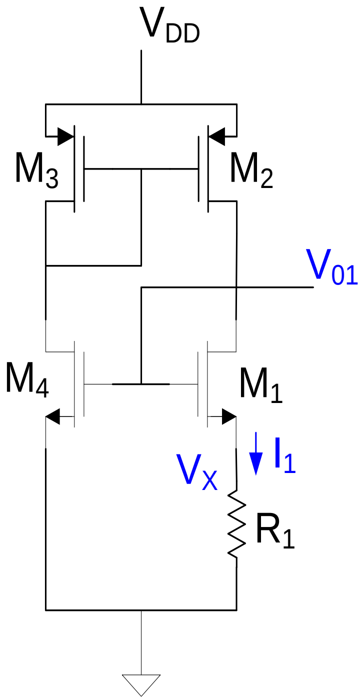
Note this behaves as a linear transconductor !

$$g_m = 4\beta_5\left[V_{Tn} - V_{Tp} - V_G\right]$$

- 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_{DD}

V_{DD} Independent Bias Generators

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



$$\left.\begin{aligned} I_{D2} &= I_{D1} = MI_{D3} \\ I_{D4} &= I_{D3} = \frac{\mu C_{OX} W_4}{2L_4}\left(V_{01} - V_{Tn}\right)^2 \\ I_{D1} &= \frac{\mu C_{OX} W_1}{2L_1}\left(V_{01} - V_X - V_{Tn}\right)^2 \\ V_X &= I_{D1} R_1 \end{aligned}\right\}$$

4 equations and 4 unknowns
$\{I_{D1}, I_{D3}, V_{01}, V_X\}$

Define:

M is the M_3:M_2 mirror gain

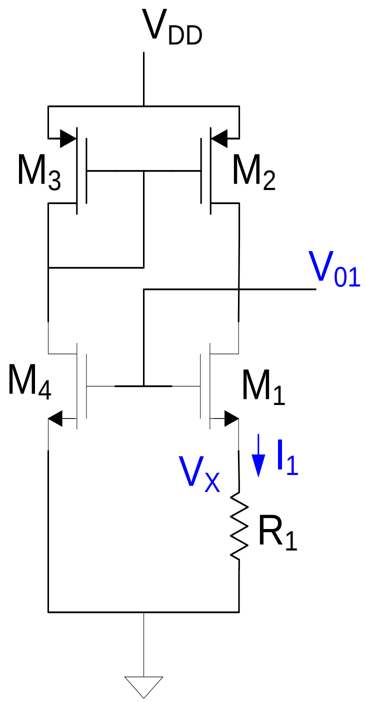
$$\beta_k = \frac{\mu C_{OX} W_k}{2L_k}$$

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

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

V_{DD} Independent Bias Generators

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



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

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

Define:

M is the M_3:M_2 mirror gain

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

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

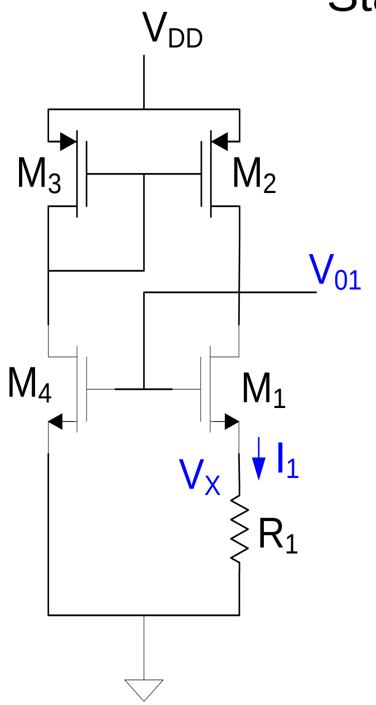
Offers some attractive properties when used as part of a temperature sensor as well

Requires Start-up Circuit

May need compensation for stability

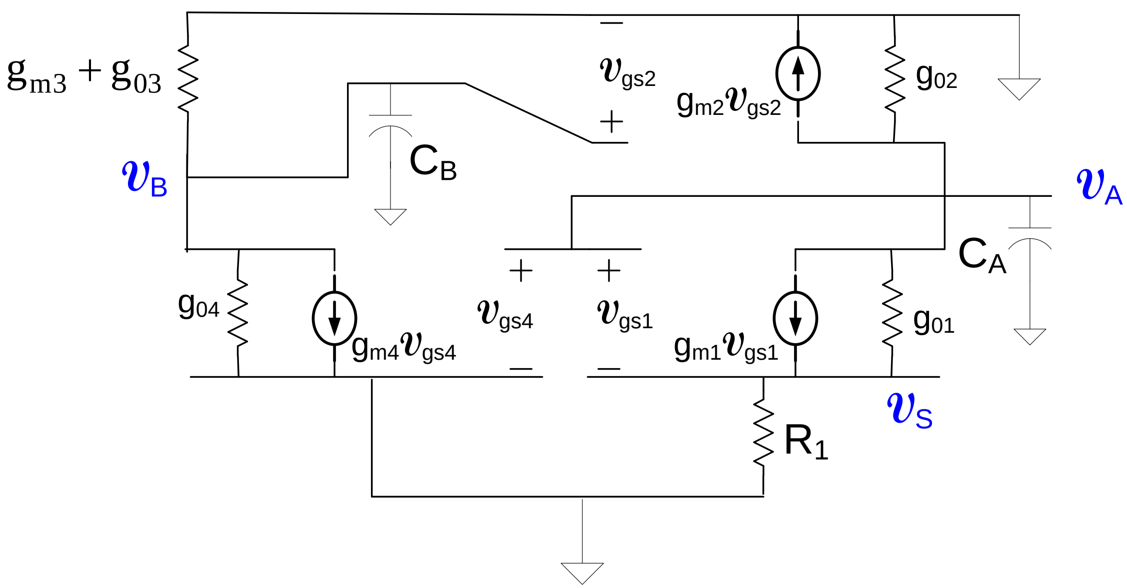
Pseudo-static operation so frequency response of little concern

V_{DD} Independent Bias Generators



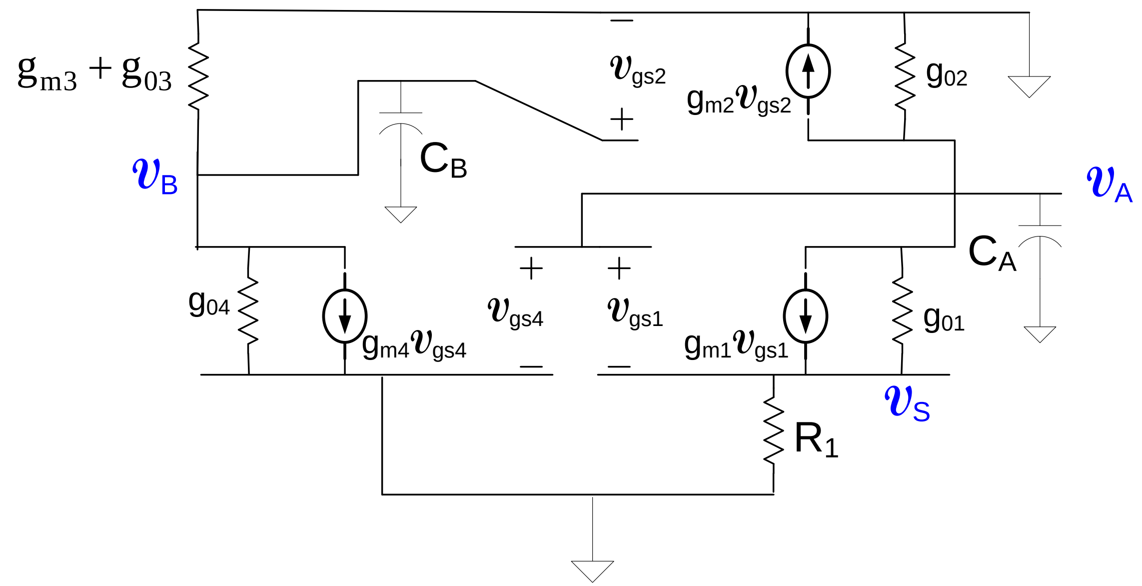
Stability Concerns

Since there is a local feedback loop, the issue of stability must be addressed. To do this, consider the small-signal equivalent circuit



V_{DD} Independent Bias Generators

Stability Concerns



Summing current at the three nodes, we obtain

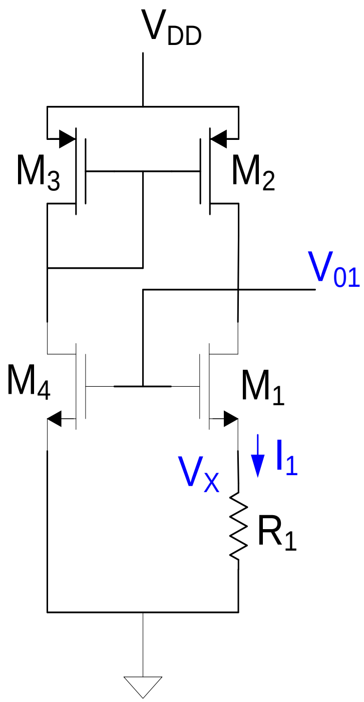
$$\left.\begin{aligned} v_A\left(g_{01}+g_{02}+sC_A\right)+g_{m2}v_B+g_{m1}\left(v_A-v_S\right)&=g_{01}v_S \\ v_B\left(g_{m3}+g_{01}+g_{03}+sC_B\right)+g_{m1}v_A&=0 \\ v_S\left(g_{01}+g_1\right)&=g_{01}v_A+g_{m1}\left(v_A-v_S\right) \end{aligned}\right\}$$

Solving and neglecting g_0 terms compared to g_m terms, we obtain the characteristic polynomial

$$D(s)=s^2C_AC_B+s\left(C_Ag_{m3}-C_B\frac{g_{m1}^2}{g_1+g_{m1}}\right)+g_{m1}\left(g_{m3}-g_{m2}\right)$$

V_{DD} Independent Bias Generators

Stability Concerns



$$D(s) = s^2 C_A C_B + s\left(C_A g_{m3} - C_B \frac{g_{m1}^2}{g_1 + g_{m1}} \right) + g_{m1}\left(g_{m3} - g_{m2}\right)$$

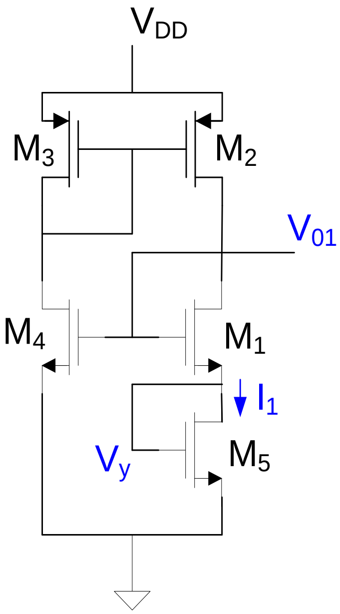
Thus, for stability, must have

$$g_{m3} > g_{m2}$$

$$C_A g_{m3} > C_B \frac{g_{m1}^2}{g_1 + g_{m1}}$$

V_{DD} Independent Bias Generators

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



$$I_{D2} = I_{D1} = I_{D5} = MI_{D3}$$

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

$$I_{D1} = \frac{\mu C_{OX} W_1}{2L_1}\left(V_{01} - V_Y - V_{Tn}\right)^2$$

$$I_{D5} = \frac{\mu C_{OX} W_5}{2L_5}\left(V_Y - V_{Tn}\right)^2$$

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

Define:

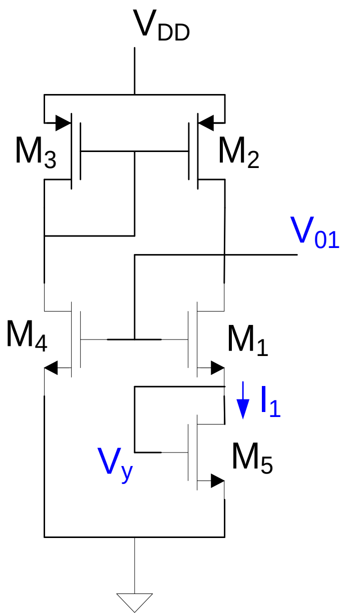
M to be the M_3:M_2 mirror gain

$$V_{01} = V_{Tn}\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right) - 2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right) - 1}\right)$$

$$V_Y = V_{Tn}\left[\left(\sqrt{\frac{W_5L_1}{W_1L_5}} - 1\right) + \left(\frac{1}{1+\sqrt{\frac{W_5L_1}{W_1L_5}}}\right)\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right) - 2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right) - 1}\right)\right]$$

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

$$V_{01} = V_{Tn}\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-1}\right)$$

$$V_Y = V_{Tn}\left[\left(\sqrt{\frac{W_5L_1}{W_1L_5}}-1\right)+\left(\frac{1}{1+\sqrt{\frac{W_5L_1}{W_1L_5}}}\right)\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-1}\right)\right]$$

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

Applications well beyond this biasing requirement

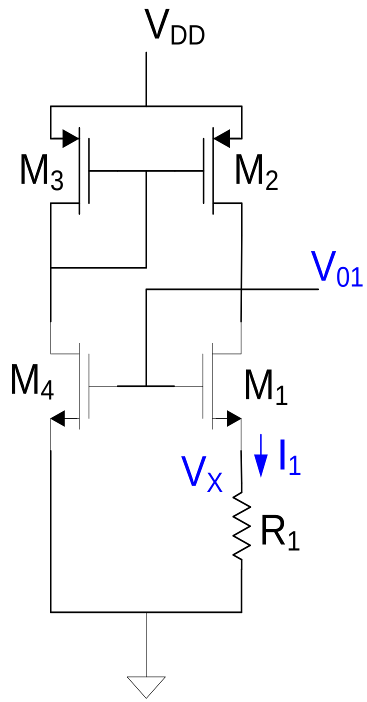
Requires Start-up Circuit

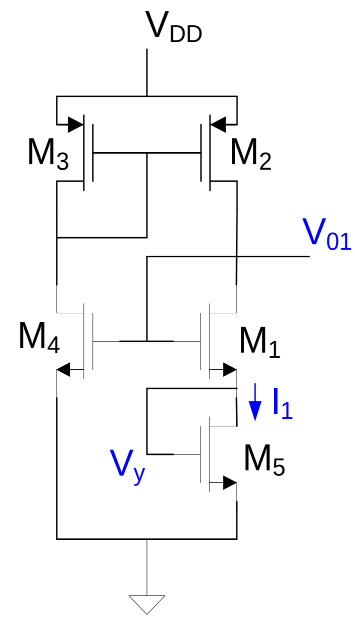
May need compensation for stability

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)





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

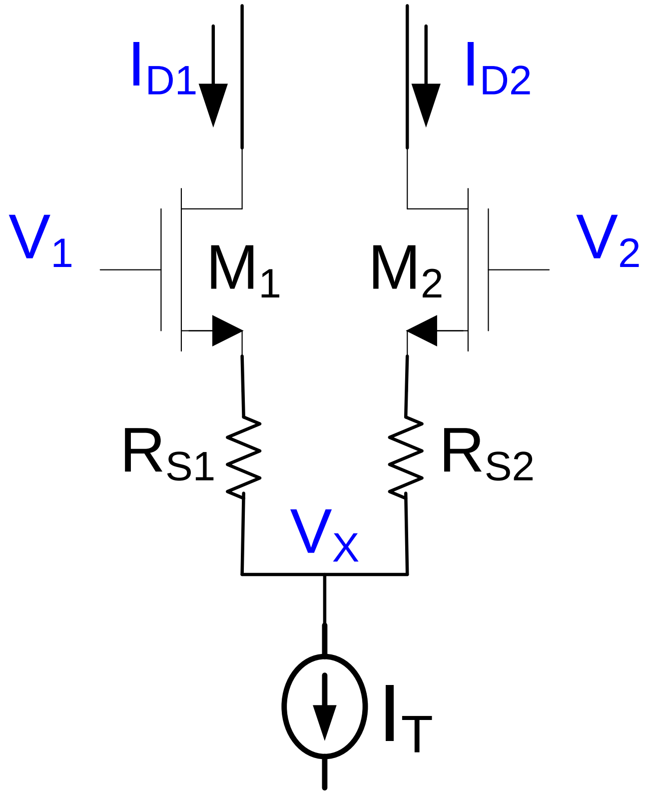
$$V_{01} = V_{Tn} + \left(\frac{1}{MR}\right)\left(\frac{1}{\beta_4} - \frac{1}{\sqrt{M\beta_1\beta_4}}\right)$$

$$V_{01} = V_{Tn}\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-1}\right)$$

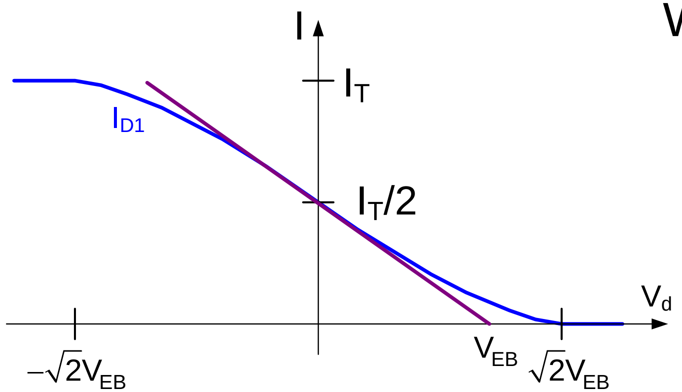
$$V_Y = V_{Tn}\left[\left(\sqrt{\frac{W_5L_1}{W_1L_5}}-1\right)+\left(\frac{1}{1+\sqrt{\frac{W_5L_1}{W_1L_5}}}\right)\left(\frac{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-2}{\sqrt{M}\left(1+\sqrt{\frac{W_1L_5}{W_5L_1}}\right)-1}\right)\right]$$

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

Transconductance Linearization Strategies

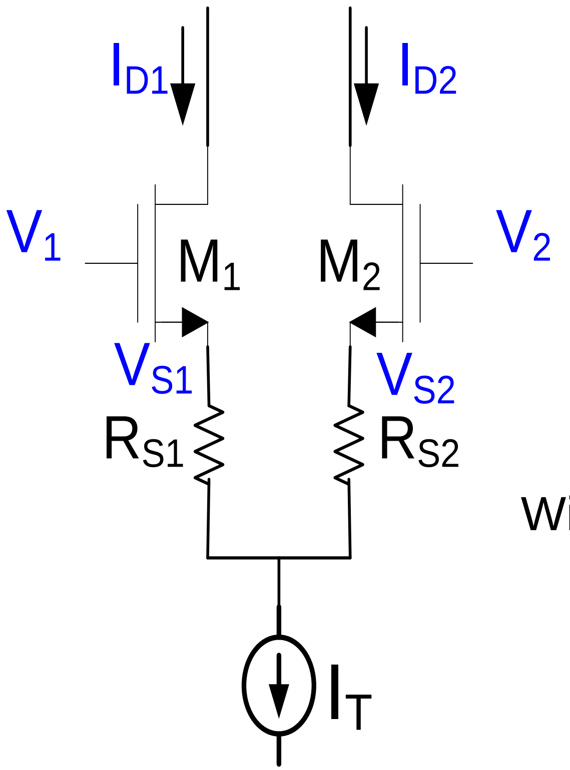


Recall with R_S=0



Widely used source degeneration

Transconductance Linearization Strategies



$$\left.\begin{aligned} I_{D1} &= \beta\left(V_1 - V_{S1} - V_T\right)^2 \\ I_{D2} &= \beta\left(V_2 - V_{S2} - V_T\right)^2 \\ V_{S1} - I_{D1}R_{S1} &= V_{S2} - I_{D2}R_{S2} \\ I_{D1} + I_{D2} &= I_T \end{aligned}\right\}$$

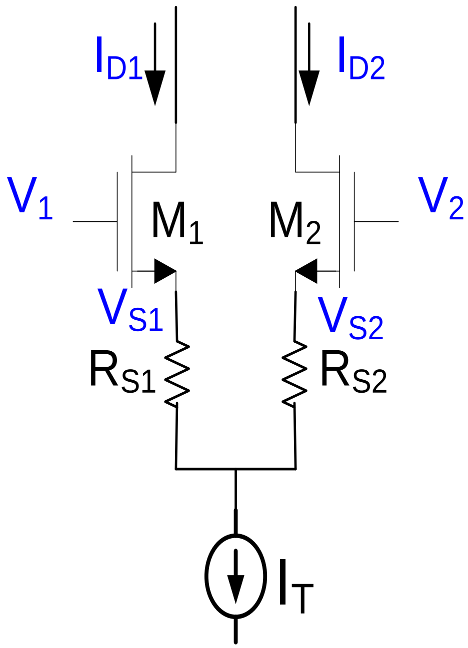
With a straightforward analysis, we obtain the expression

$$\sqrt{\frac{1}{\beta}}\left(\sqrt{I_T - I_{D1}} - \sqrt{I_{D1}}\right) + R_S\left(I_T - 2I_{D1}\right) = V_d$$

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



$$\sqrt{\frac{1}{\beta}}\left(\sqrt{I_T - I_{D1}} - \sqrt{I_{D1}}\right) + R_S\left(I_T - 2I_{D1}\right) = V_d$$

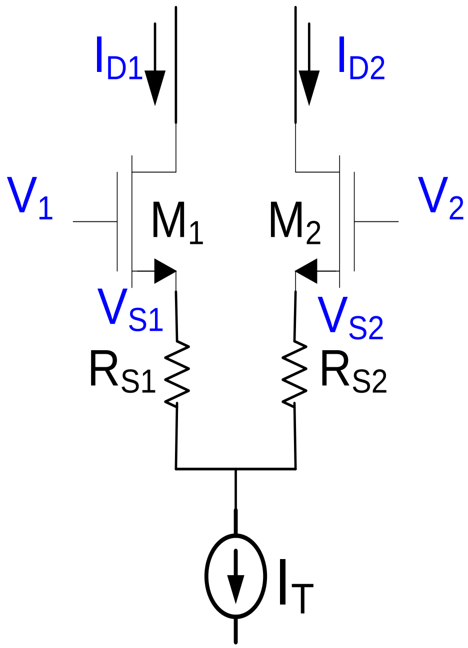
The transconductance of this structure can be readily derived to obtain

$$g_m = \left.\frac{\partial V_d}{\partial I_{D1}}\right|_{Q-pt}^{-1} = \left[\sqrt{\frac{1}{\beta}} \bullet \frac{1}{2}\left(-\left(I_T - I_{D1}\right)^{-1/2} - I_{D1}^{-1/2}\right) - 2R_S\right]\Bigg|_{Q-pt}^{-1}$$

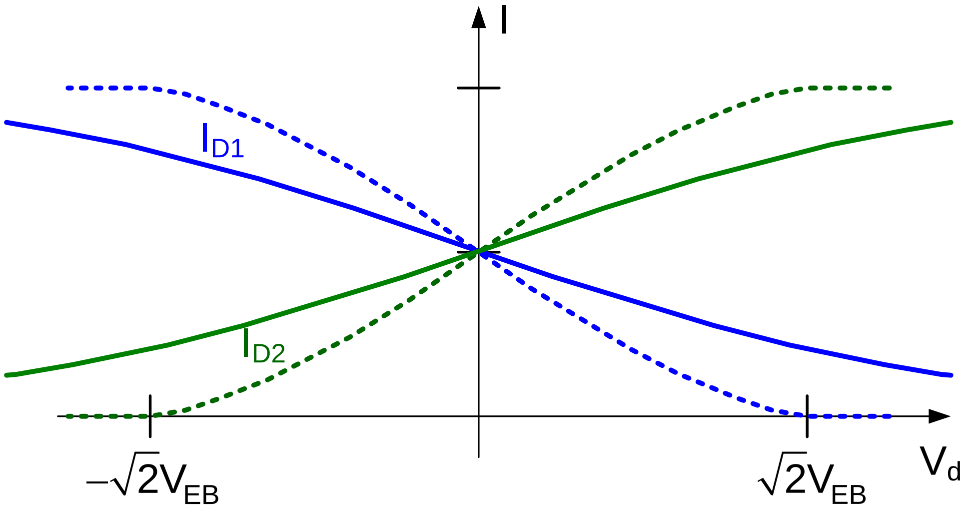
This can be expressed as

$$g_m = \left.\frac{\partial V_d}{\partial I_{D1}}\right|_{Q-pt}^{-1} = -\frac{1}{\left[\sqrt{\frac{2}{\beta I_T}} + 2R_S\right]} = -\frac{\beta V_{EB}}{1 + 2\beta V_{EB} R_S}$$

Transconductance Linearization Strategies



$$\sqrt{\frac{1}{\beta}}\left(\sqrt{I_T - I_{D1}} - \sqrt{I_{D1}}\right) + R_S\left(I_T - 2I_{D1}\right) = V_d$$



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 date 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

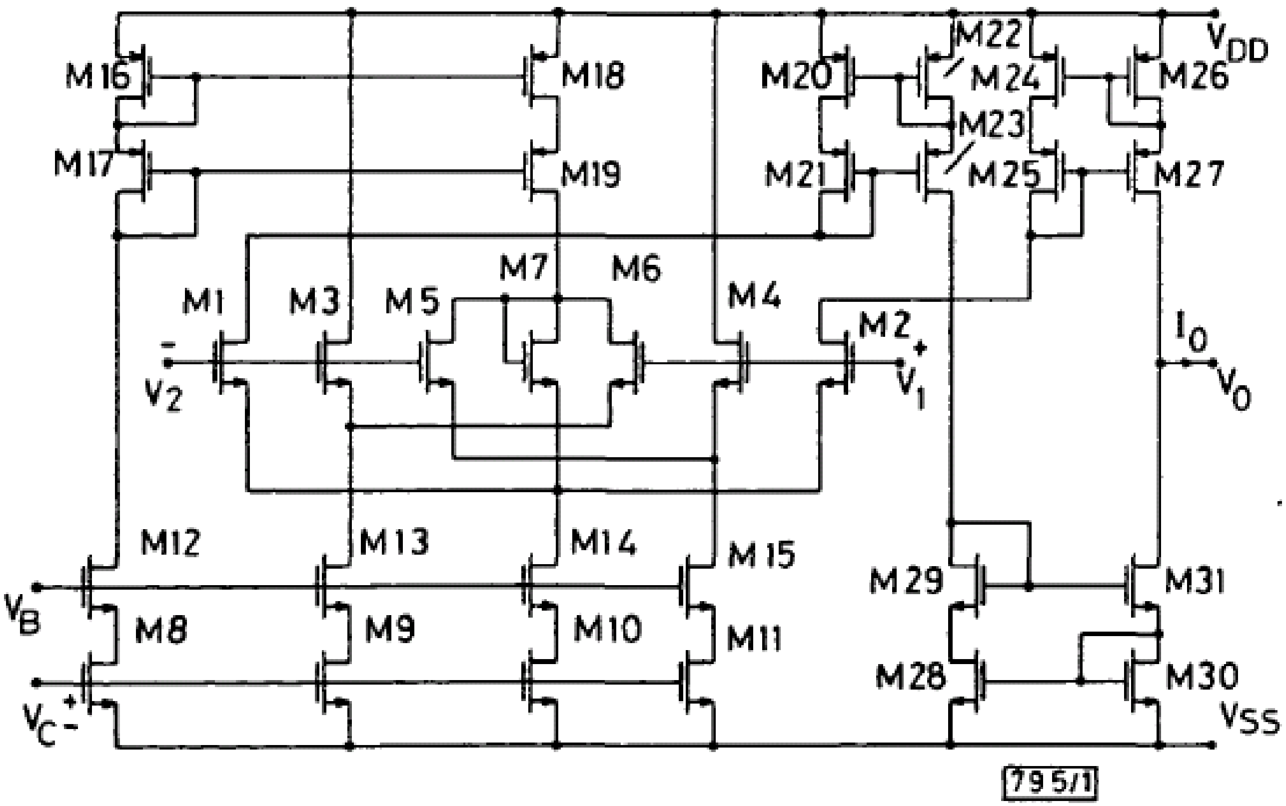
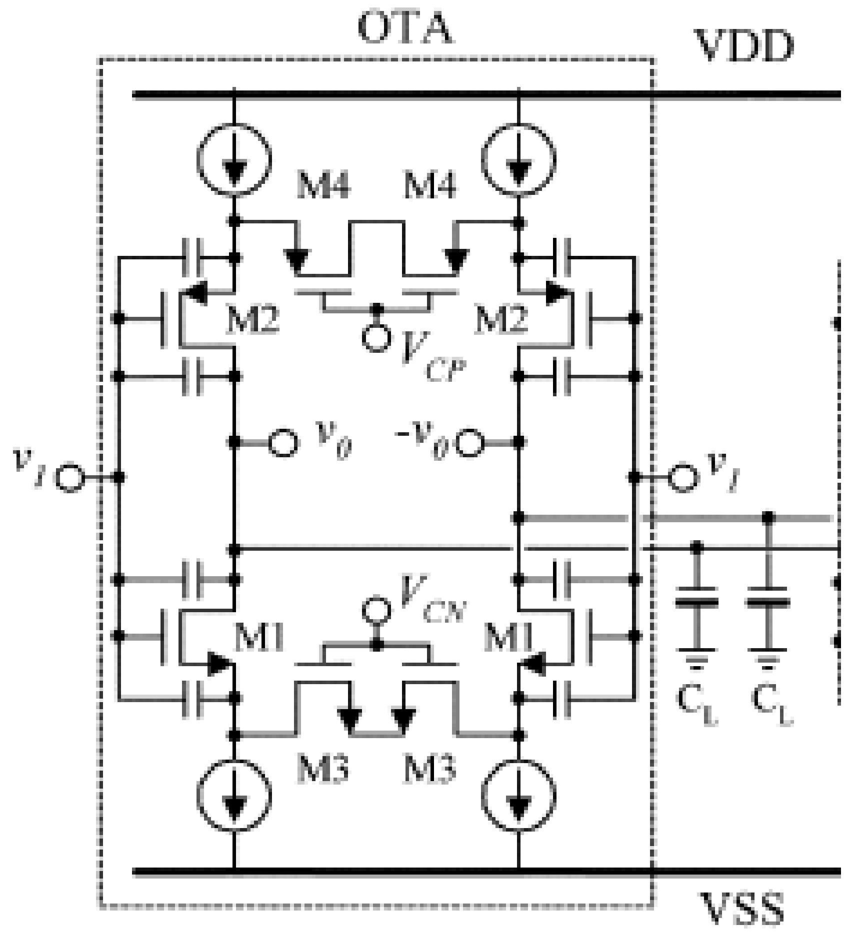


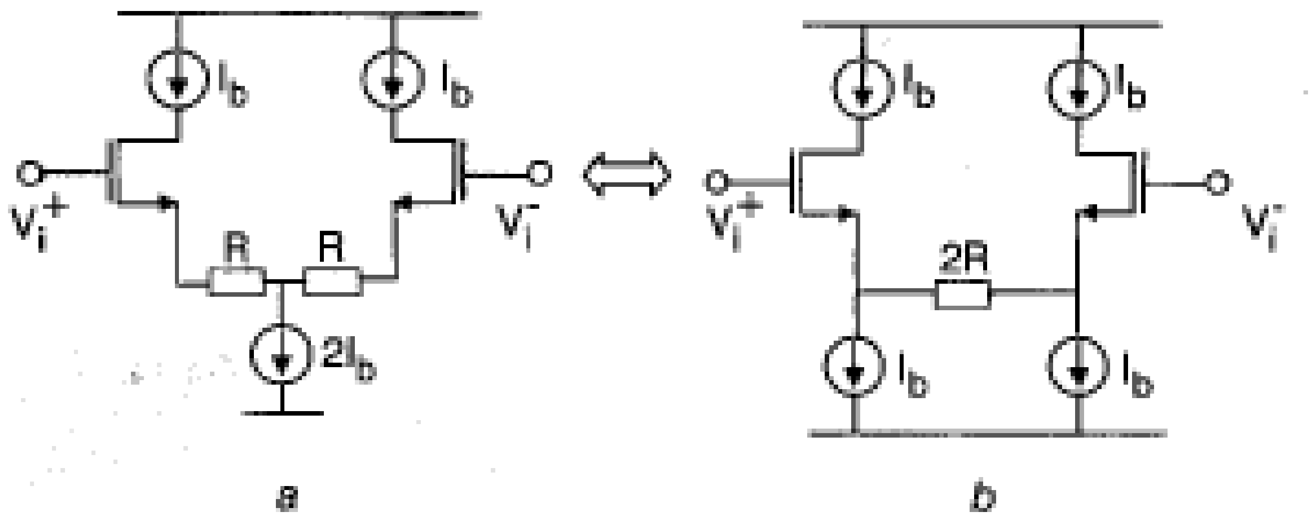
Fig. 1 *Linearised CMOS transconductance circuit*

Transconductance Linearization Strategies



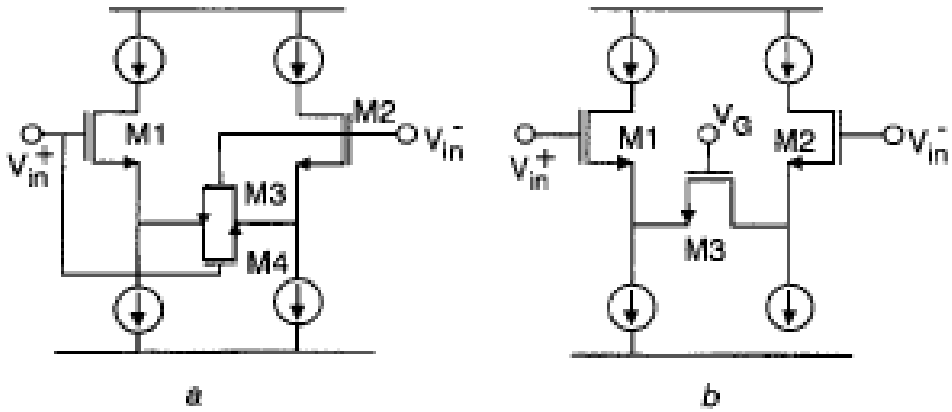
From CAS 2006 P 811 Jose Silva

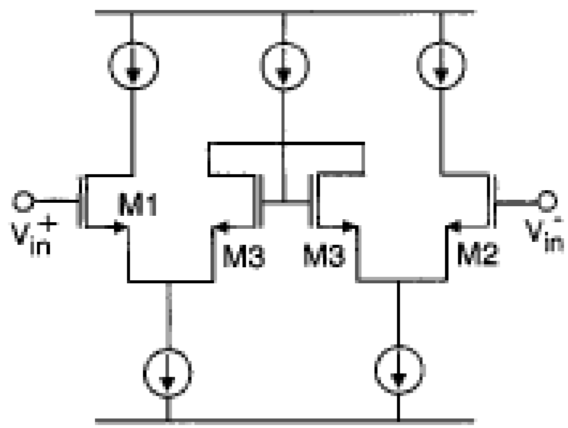
Transconductance Linearization Strategies



Linearity Enhancement with Source Degeneration

Transconductance Linearization Strategies



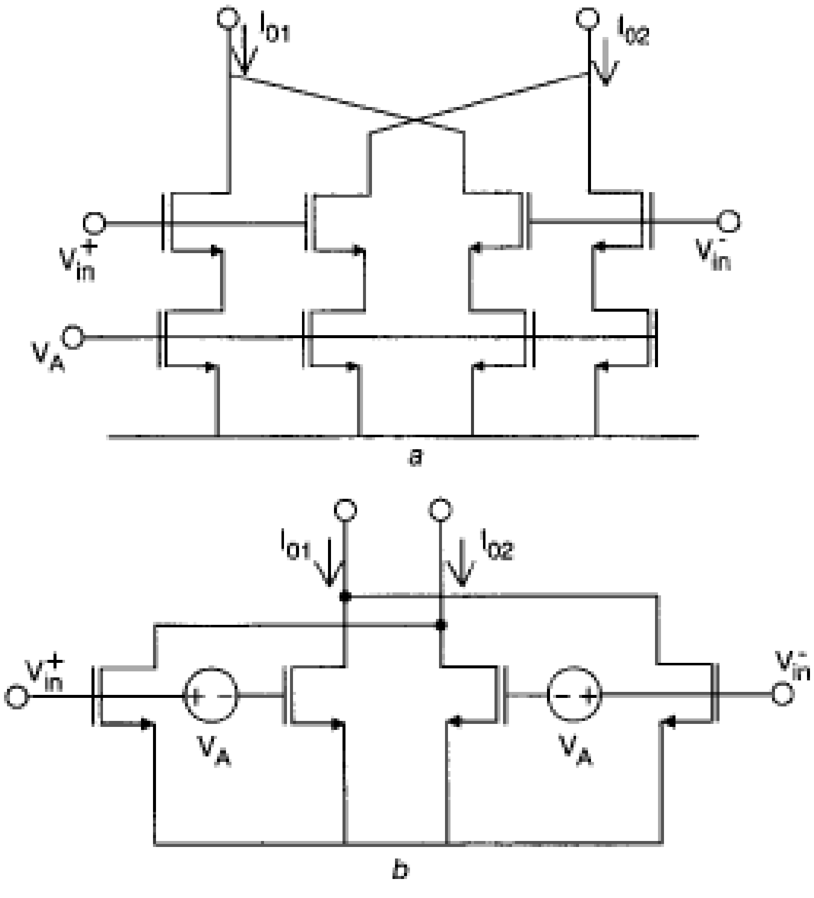


Linearization with active source degeneration

CMOS transconductance amplifiers, architectures and active filters: a tutorial

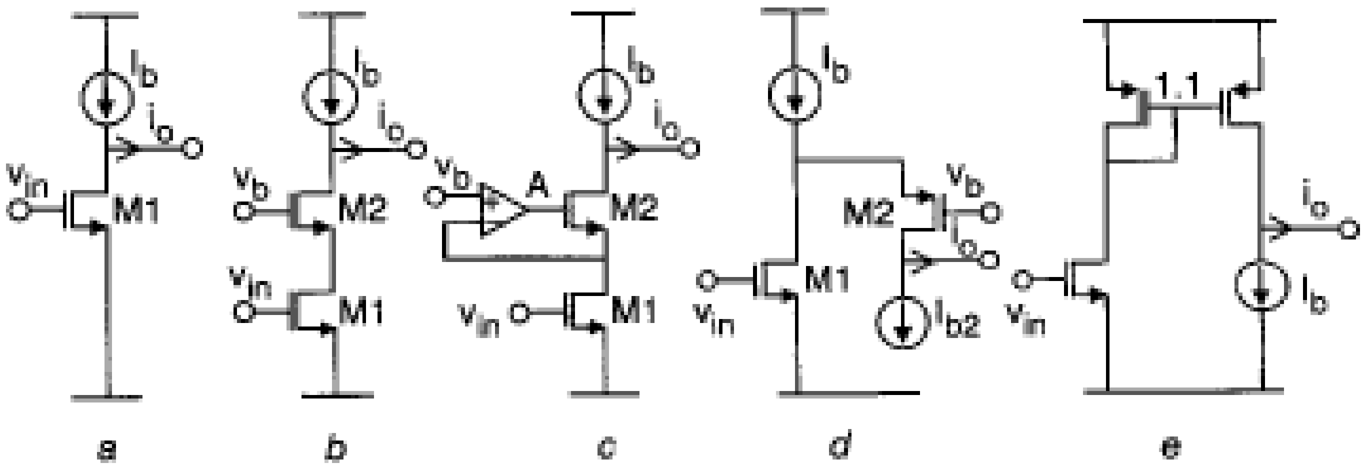
E.Sánchez-Sinencio and J.Silva-Martínez

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, nonlinearised 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.

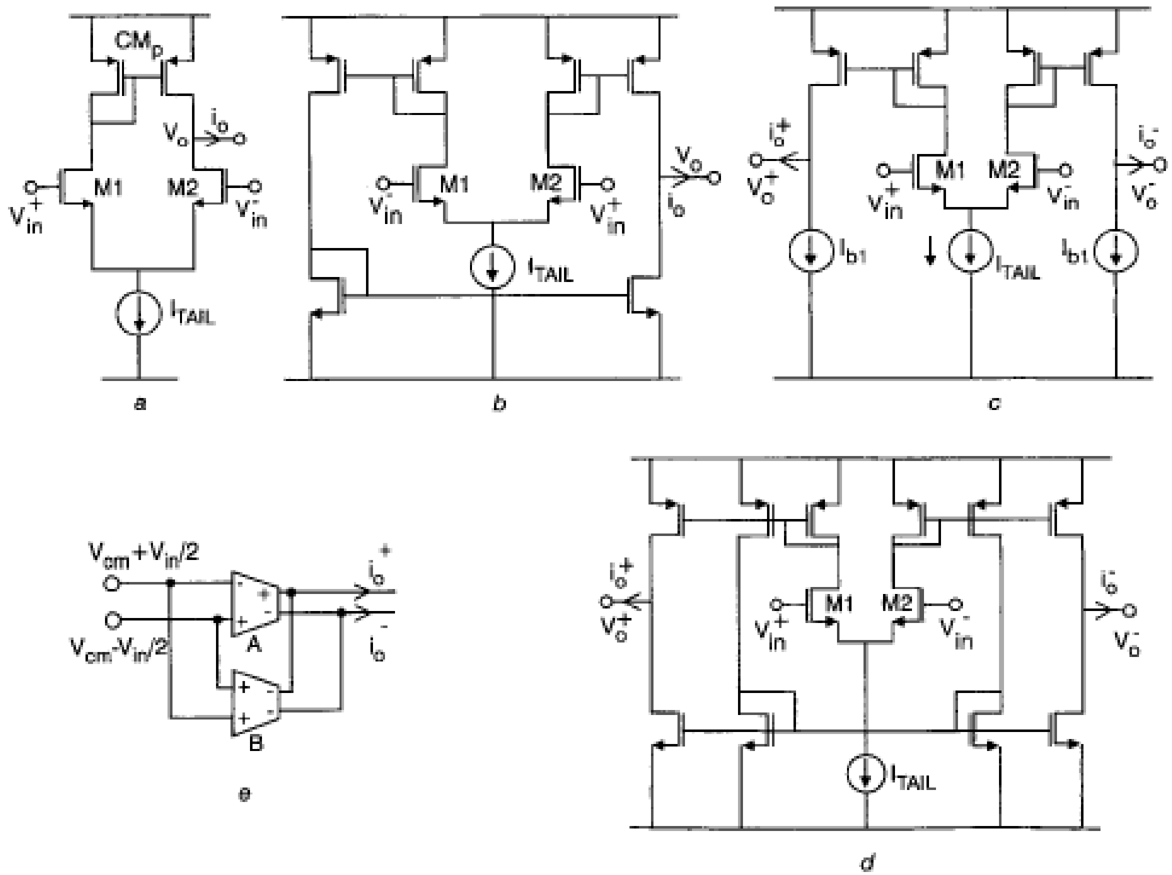


Linearity compensation with cross-coupled feedback

Single-ended input TAs

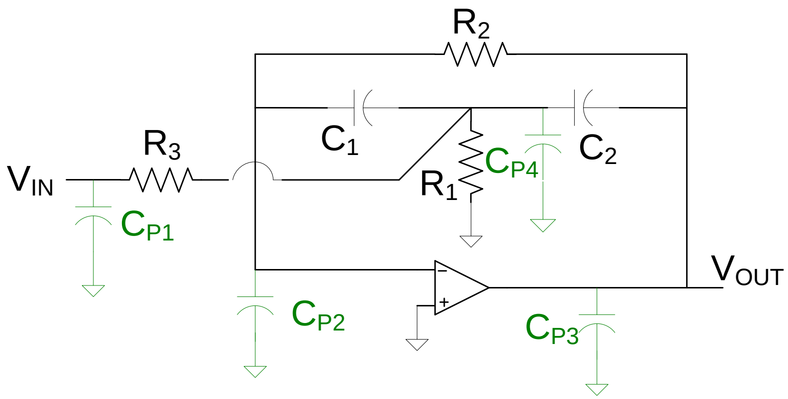


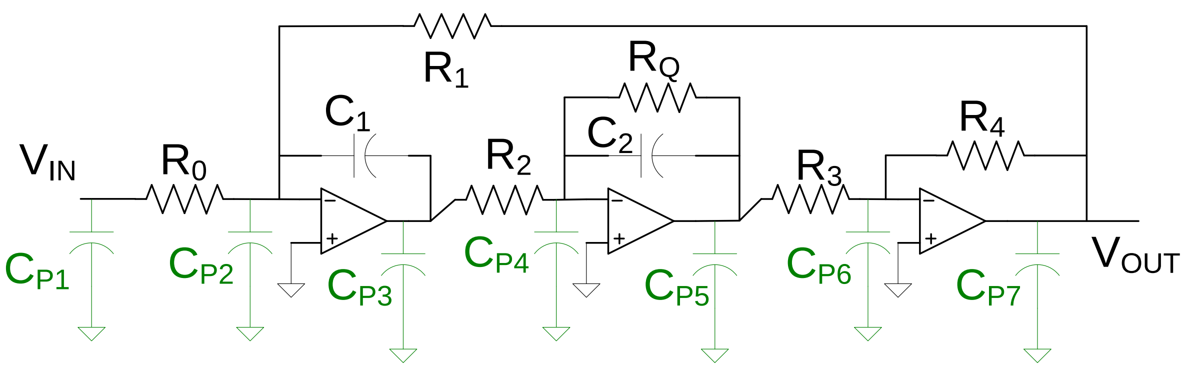
Differential input OTAs

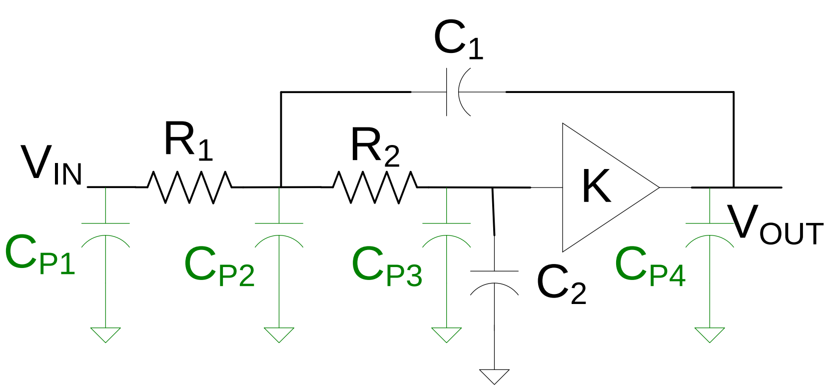


Differential input and output OTAs

Parasitic Capacitances and Floating Nodes





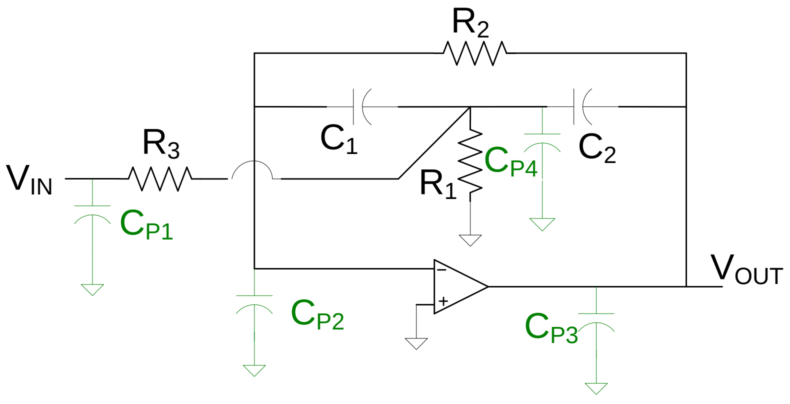


There is invariably a parasitic capacitance associated with every terminal of every element in a filter

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

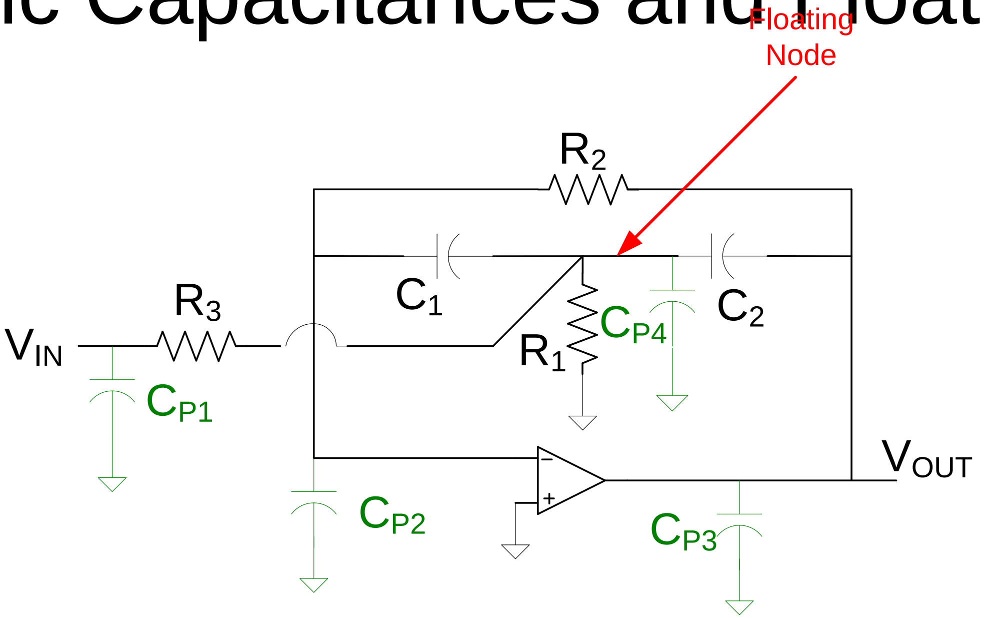


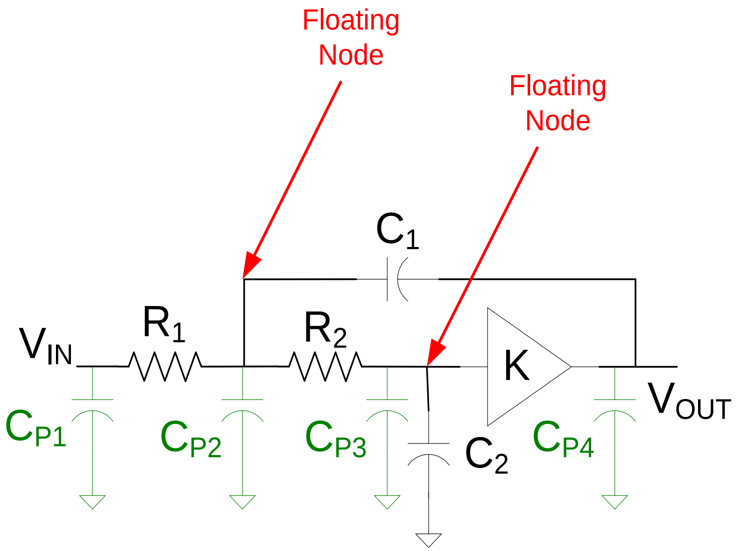
A floating node is a node that is not connected to either a zero-impedance element or across a null-port

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

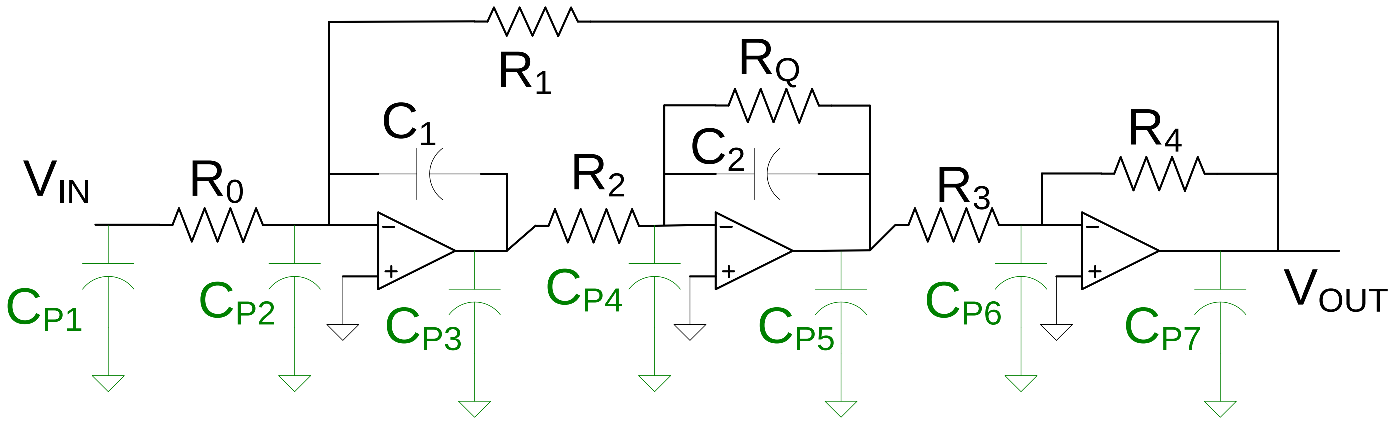
Some filter architectures inherently have no floating nodes, specifically, most of the basic integrator-based filters have no floating nodes

Parasitic Capacitances and Floating Nodes





Parasitic Capacitances and Floating Nodes



No floating nodes !

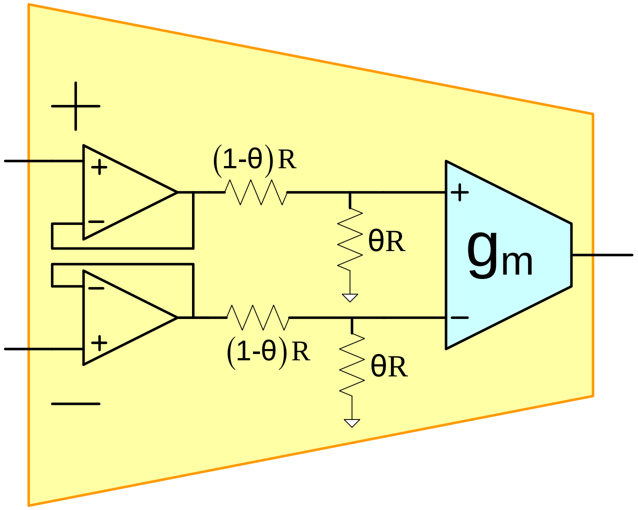
Signal Swing in OTA Circuits

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



$$g_{meq} = \theta g_m$$

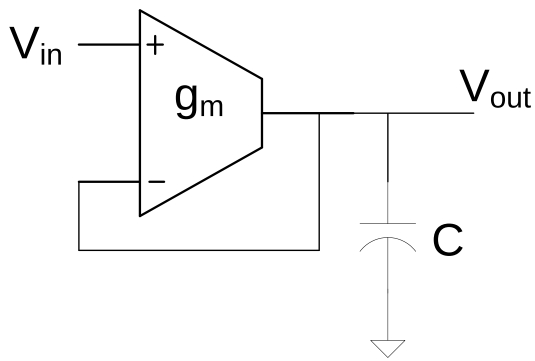
R. L. Geiger and E. Sánchez-Sinencio, "Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial," *IEEE Circuits and Devices Magazine*, Vol. 1, pp.20-32, March 1985.

Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial

Randall L. Geiger and Edgar Sánchez-Sinencio

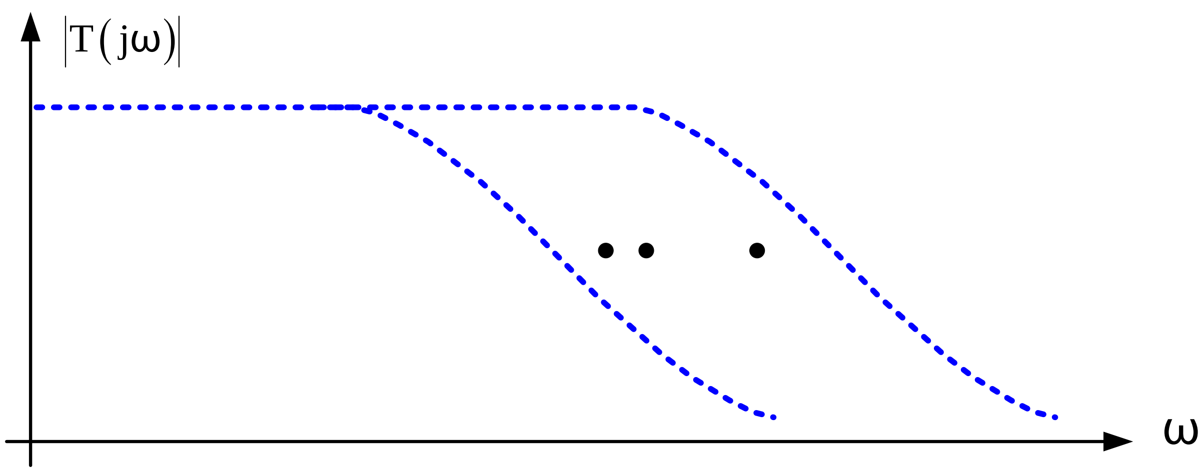
Programmable Filter Structures

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

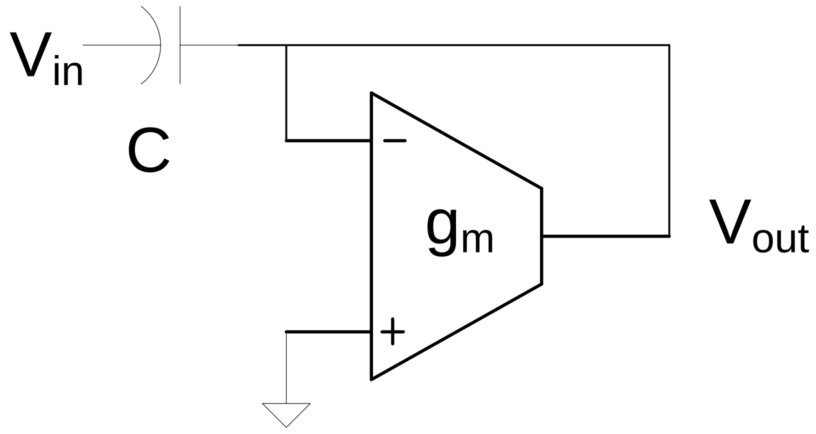


$$T(s) = \frac{g_m}{g_m + sC}$$

Programmable First-Order Low-Pass Filter

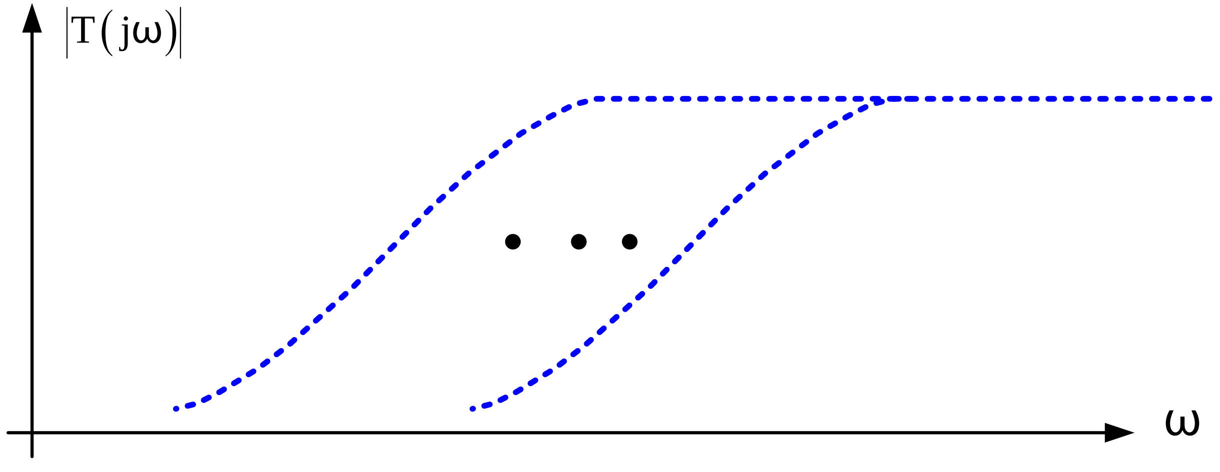


Programmable Filter Structures



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

Programmable First-Order High-Pass Filter



End of Lecture 35