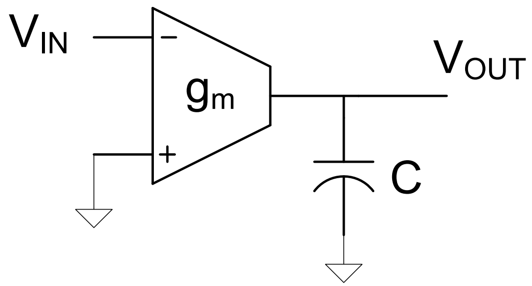
EE 508

Lecture 33

Transconductor Design

Review from Last Lecture

Transconductor Design



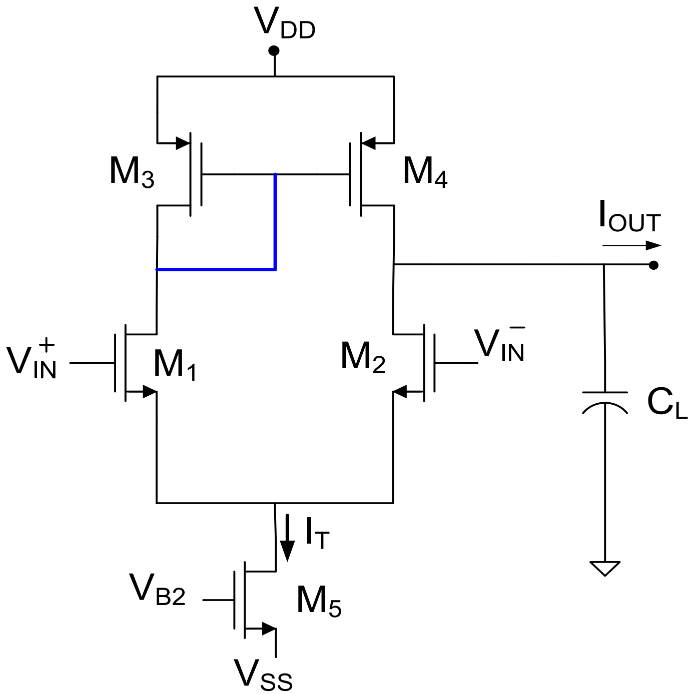
Transconductor-based filters depend directly on the g_m of the transconductor

Feedback is not used to make the filter performance insensitive to the transconductance gain

Linearity and spectral performance of the filter strongly dependent upon the linearity of the transconductor

Often can not justify elegant linearization strategies in the transconductors because of speed, area, and power penalties

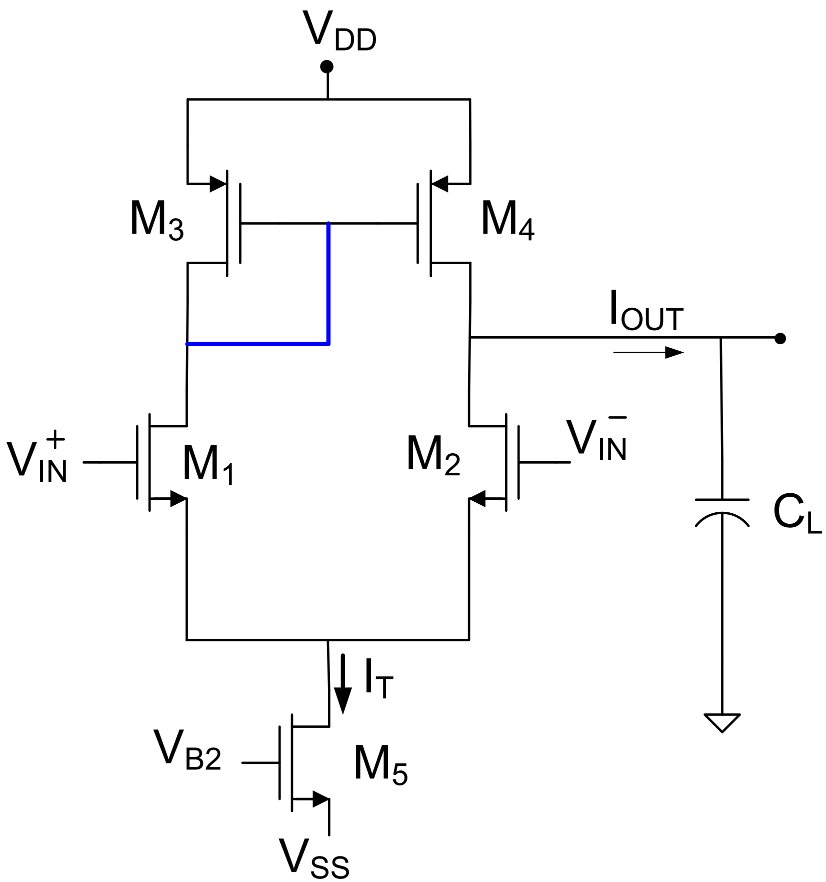
Linearity of Amplifiers

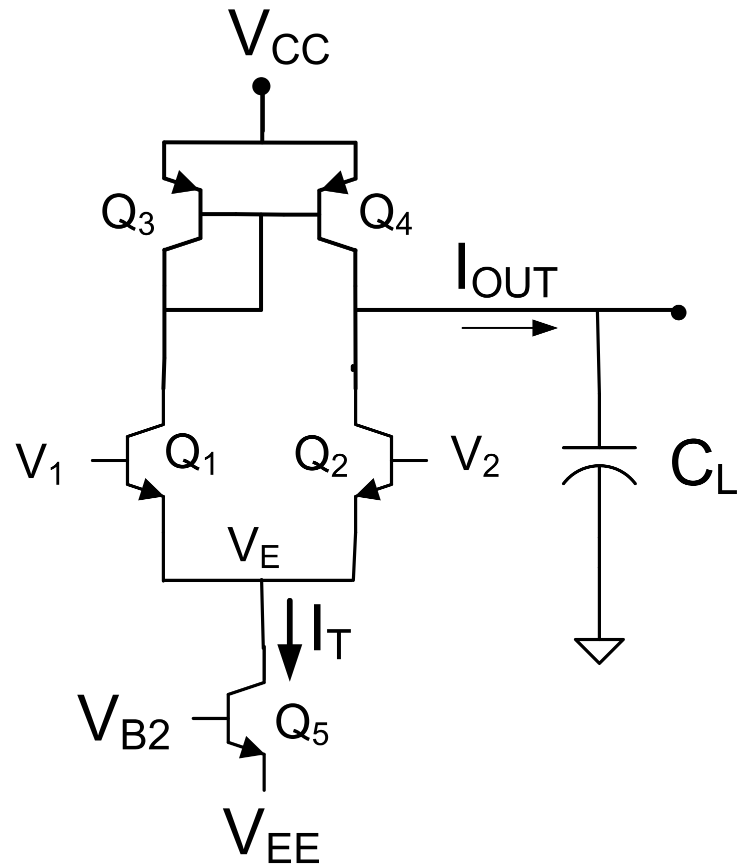


Strongly dependent upon linearity of transconductance of differential pair

Review from Last Lecture

Linearity of Amplifiers

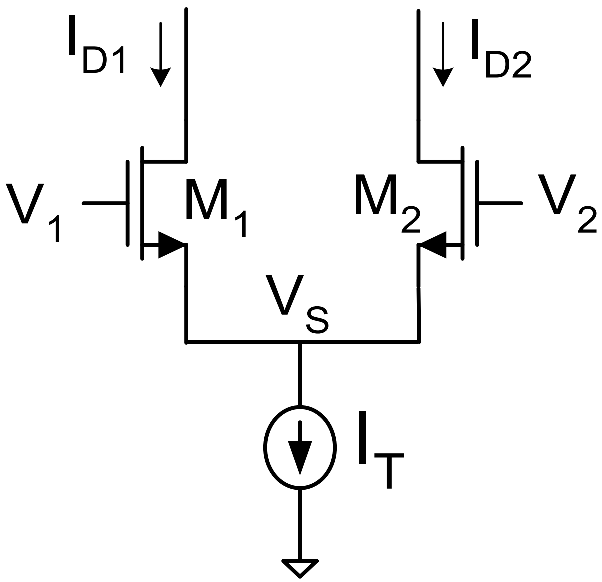




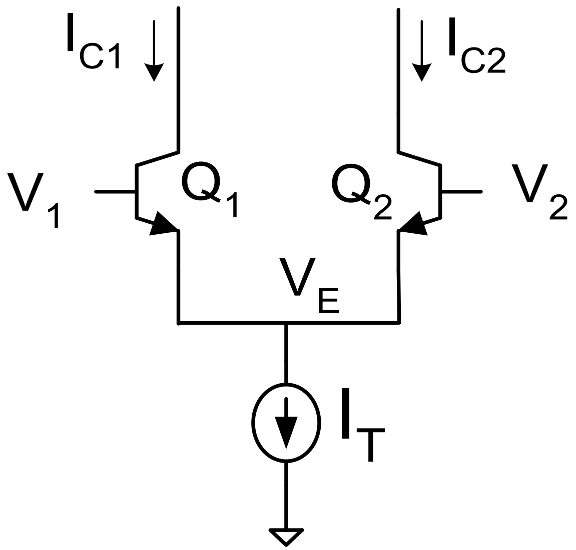
Strongly dependent upon linearity of transconductance of differential pair

Review from Last Lecture

Differential Input Pairs



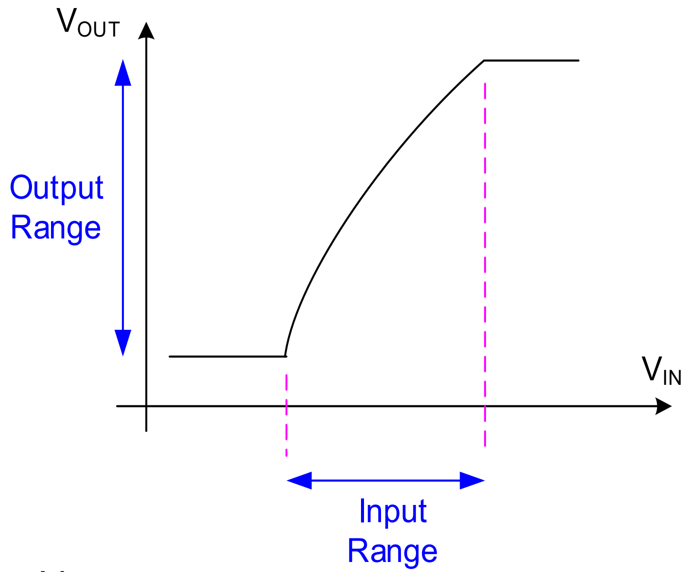
MOS Differential Pair

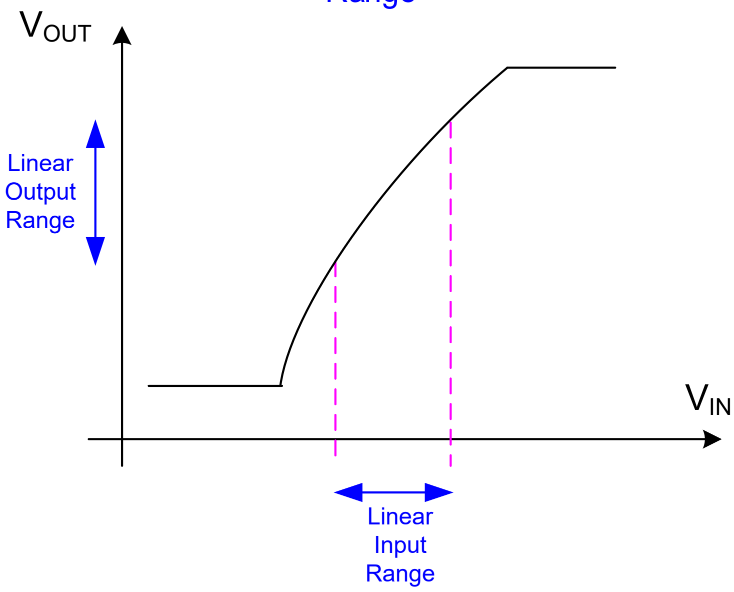


Bipolar Differential Pair

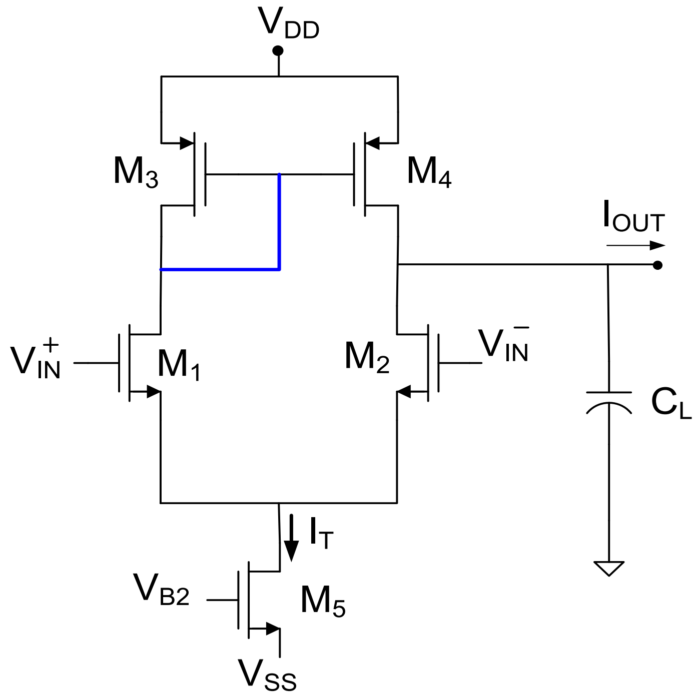
Review from Last Lecture

Signal Swing and Linearity





Linearity of Amplifiers

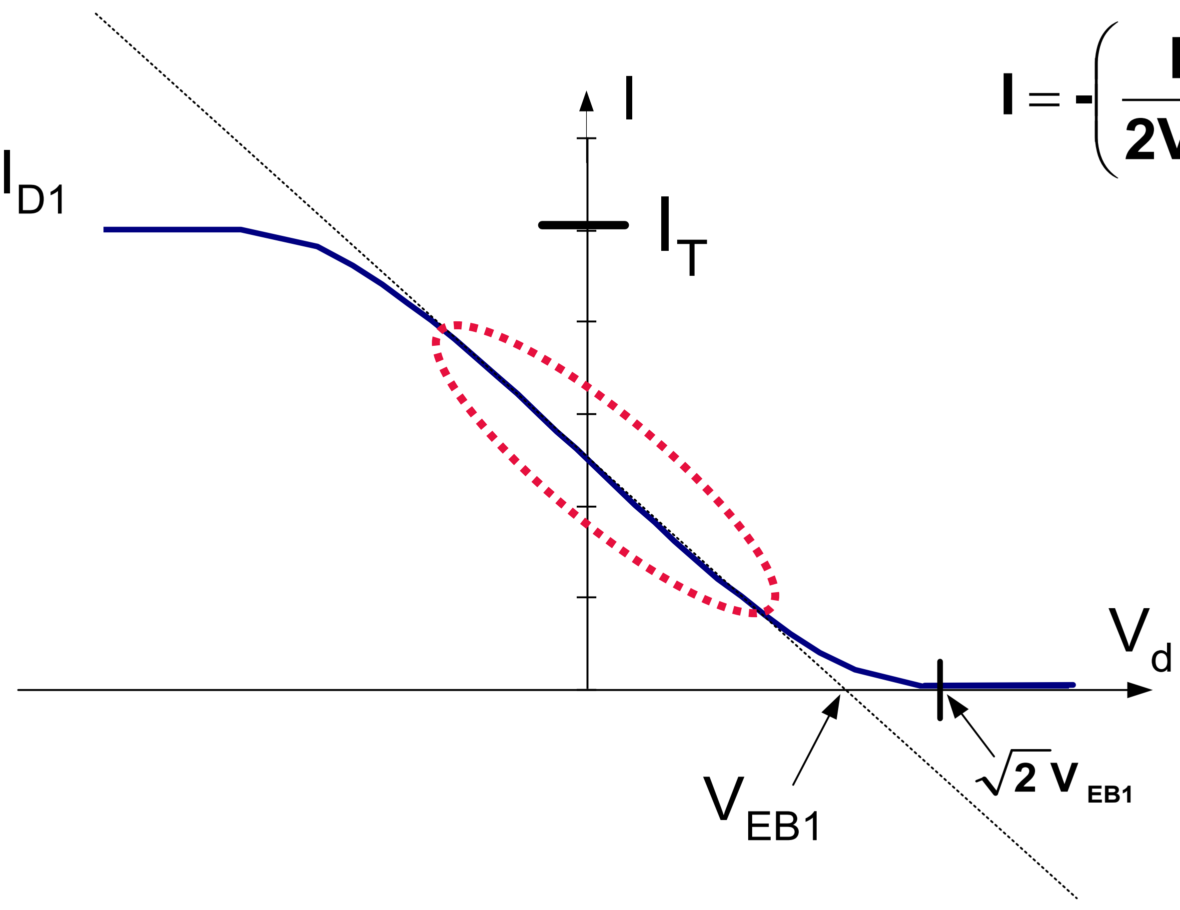


Strongly dependent upon linearity of transconductance of differential pair

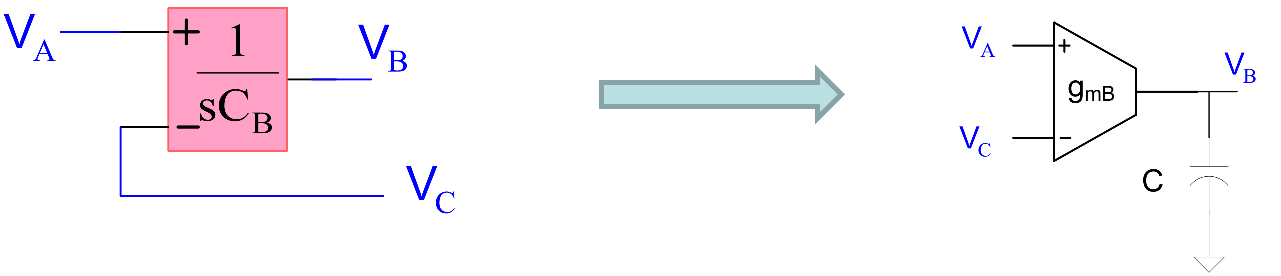
How linear is the amplifier ?

$$V_{dint} = -\frac{h}{m} = -\frac{I_T}{2m} = V_{EB1}$$

$$I = -\left(\frac{I_T}{2V_{EB1}}\right)V_d + \frac{I_T}{2}$$



Programmable Filter Structures



$$\left|\omega_0\right| = \frac{g_m}{C}$$

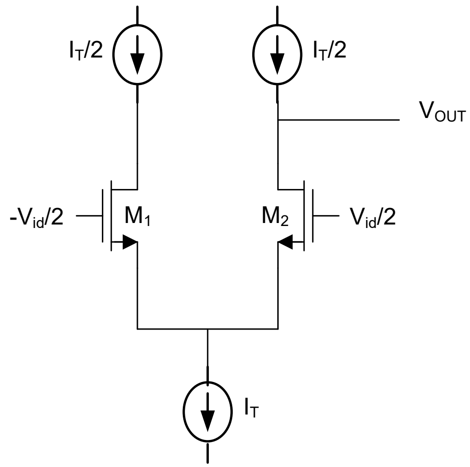
Often want to program or trim filters

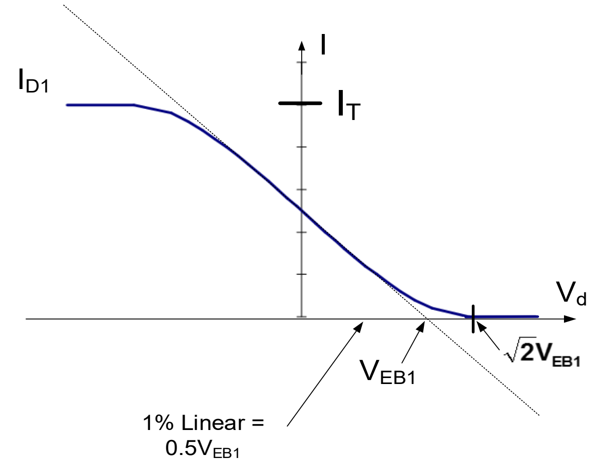
Applicable in wide variety of filter architectures (here showing integrator-based)

Attractive to do this by adjusting g_m , in part, because g_m can be continuously adjustable with some transconductance devices

Review from Last Lecture

What input range is possible when using the tail current to program the OTA (i.e. after W/L fixed)?



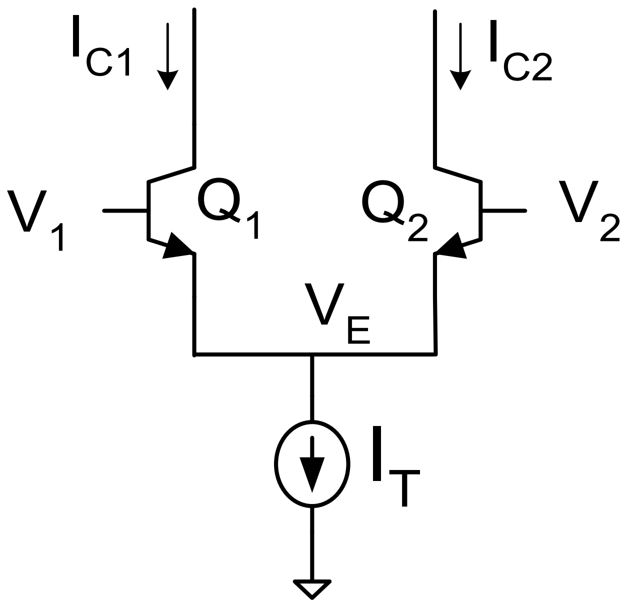


$$g_m = \mu C_{OX}\frac{W}{L}V_{EB} == \sqrt{I_T}\sqrt{\mu C_{OX}\frac{W}{L}}$$

$$\mathbf{V_{dx}} = \pm\sqrt{\frac{\mathbf{2L}}{\mathbf{\mu C_{OX} W}}}\left(\sqrt{\mathbf{I_T}}\right)$$

- Input signal swing decreases linearly with decreases in g_m for fixed W/L
- One decade reduction in g_m results in one decade decrease in signal swing
- One decade reduction in g_m requires two decade decrease in I_T
- Though MOS OTA can have very good single swing with large V_{EB}, very limited tail current programmability with basic MOS OTA
- There are, however, other ways to program MOS OTA without big penalty in signal swing

Bipolar Differential Pair



$$\left.\begin{aligned} I_{C1} &= J_S A_{E1} e^{\frac{V_1 - V_E}{V_t}} \\ I_{C2} &= J_S A_{E2} e^{\frac{V_2 - V_E}{V_t}} \\ I_{C1} + I_{C2} &= I_T \end{aligned}\right\}$$

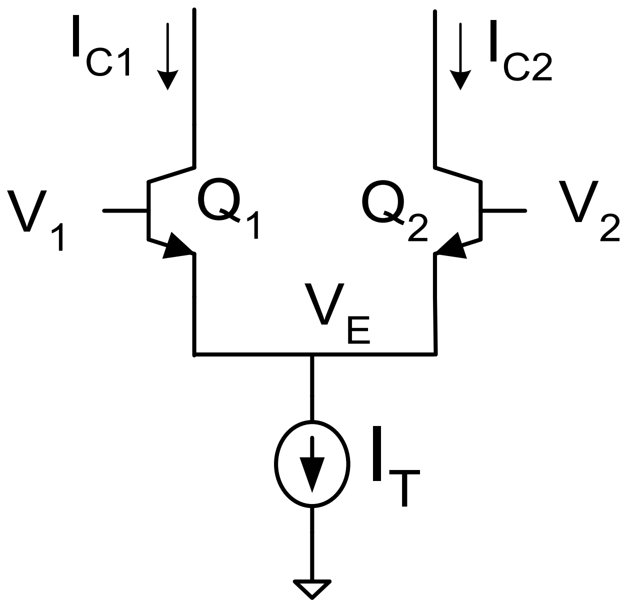
$$V_1 = V_E + V_t \ln\left(\frac{I_{C1}}{J_S A_{E1}}\right)$$

$$V_2 = V_E + V_t \ln\left(\frac{I_{C2}}{J_S A_{E2}}\right)$$

$$V_d = V_2 - V_1$$

$$V_d = V_t\left(\ln\left(\frac{I_{C2}}{J_S A_{E2}}\right) - \ln\left(\frac{I_{C1}}{J_S A_{E1}}\right)\right) \overset{A_{E1}=A_{E2}}{=} V_t \ln\left(\frac{I_{C2}}{I_{C1}}\right)$$

Bipolar Differential Pair



$$\mathbf{V_d = V_2 - V_1}$$

$$\mathbf{V_d = V_t\left(\ln\left(\frac{I_{C2}}{J_S A_{E2}}\right) - \ln\left(\frac{I_{C1}}{J_S A_{E1}}\right)\right) \overset{A_{E1}=A_{E2}}{=} V_t \ln\left(\frac{I_{C2}}{I_{C1}}\right)}$$

$$\mathbf{V_d = V_t \ln\left(\frac{I_T - I_{C1}}{I_{C1}}\right)}$$

$$\mathbf{V_d = V_t \ln\left(\frac{I_{C2}}{I_T - I_{C2}}\right)}$$

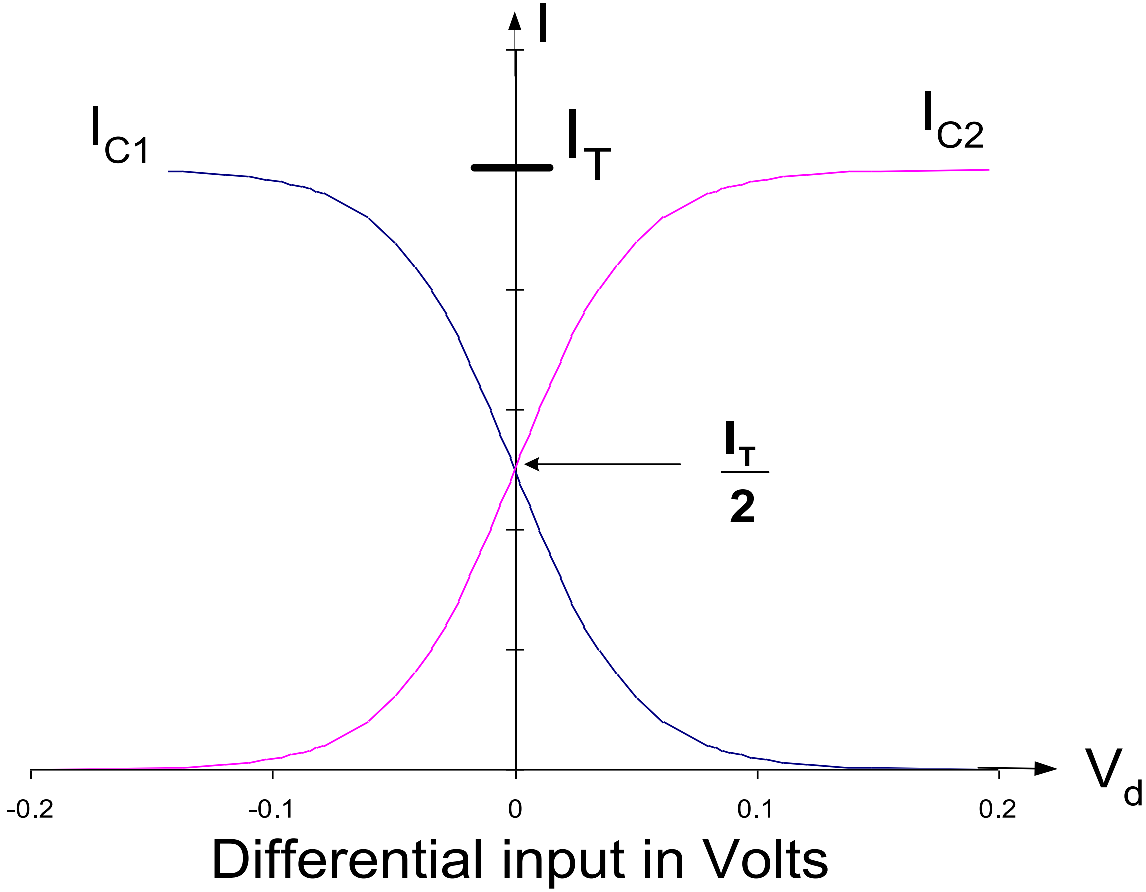
At $I_{C1}=I_{C2}=I_T/2$, $V_d=0$

As I_{C1} approaches 0, V_d approaches infinity

As I_{C1} approaches I_T, V_d approaches minus infinity

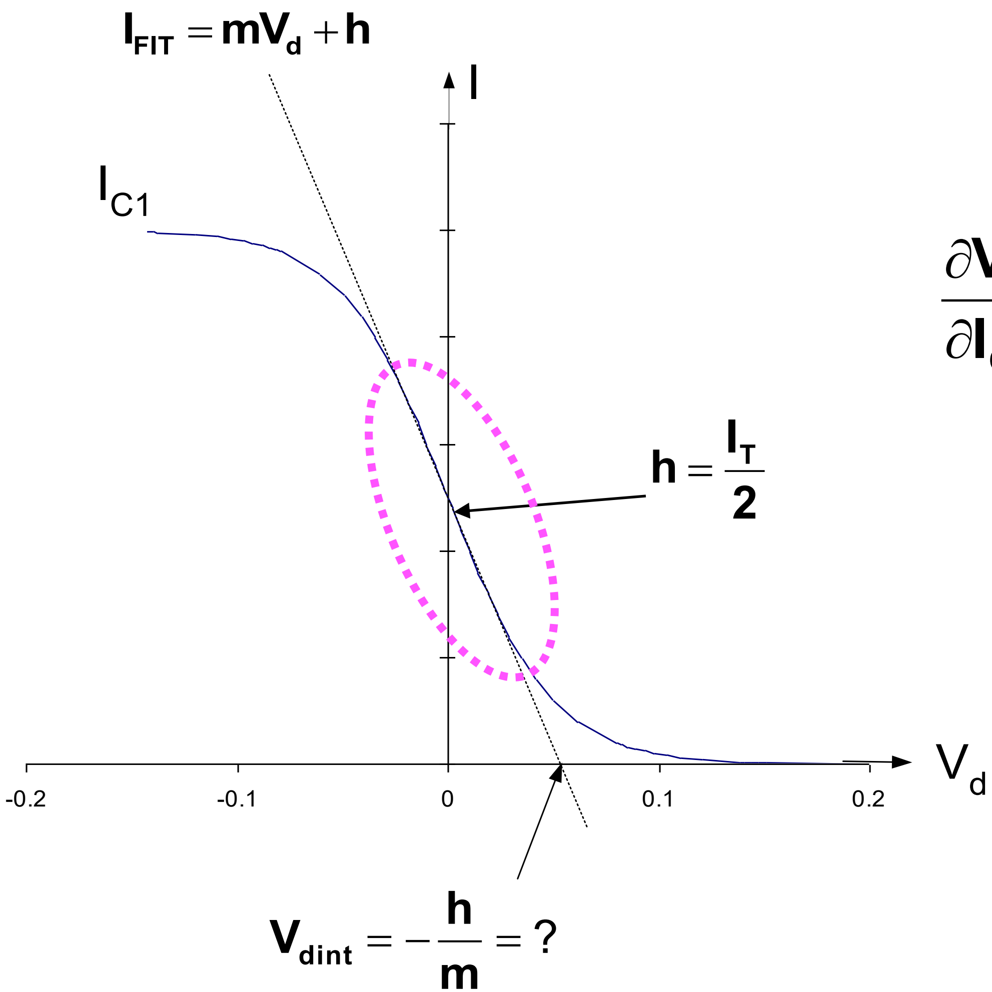
Transition much steeper than for MOS case

Transfer Characteristics of Bipolar Differential Pair



Transition much steeper than for MOS case
Asymptotic Convergence to 0 and I_T

Signal Swing and Linearity of Bipolar Differential Pair



$$m = \left.\frac{\partial I_{C1}}{\partial V_d}\right|_{Q-point}$$

$$\left.\frac{\partial V_d}{\partial I_{C1}}\right|_{Q-point} = -V_t \left.\frac{I_T}{I_{C1}\left(I_T - I_{C1}\right)}\right|_{I_{C1}=\frac{I_T}{2}}$$

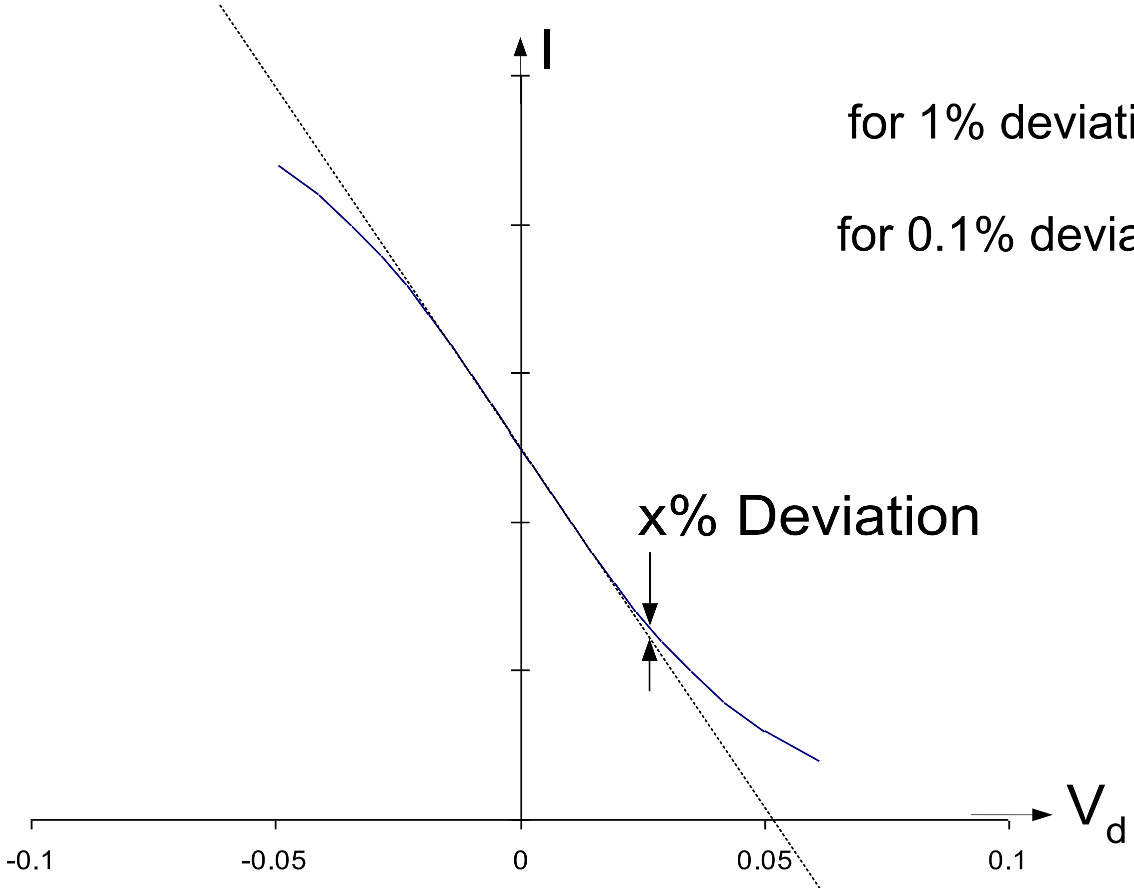
$$\left.\frac{\partial V_d}{\partial I_{C1}}\right|_{Q-point} = -\frac{4V_t}{I_T}$$

$$I_{FIT} = -\frac{I_T}{4V_t}V_d + \frac{I_T}{2}$$

$$V_{dint} = -\frac{h}{m} = 2V_t$$

Note V_{dint} is independent of I_T in contract to what we saw for MOS differential pairs

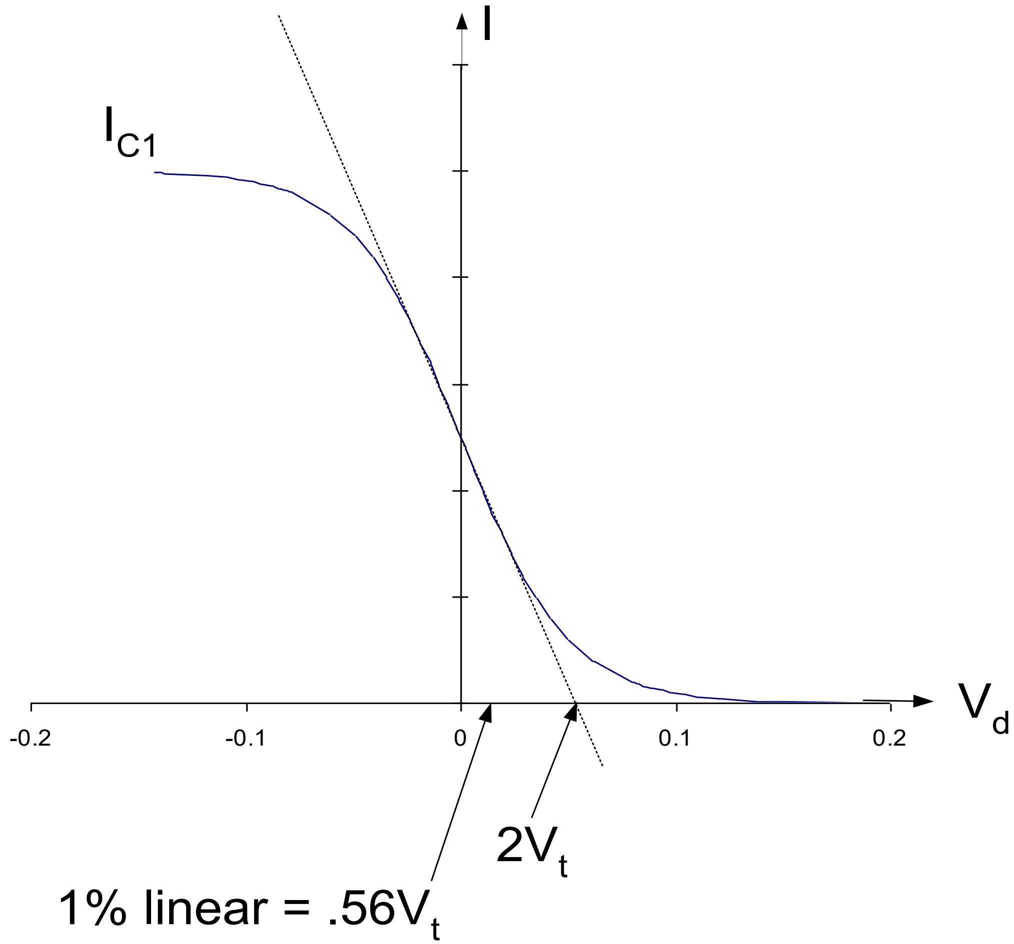
Signal Swing and Linearity of Bipolar Differential Pair



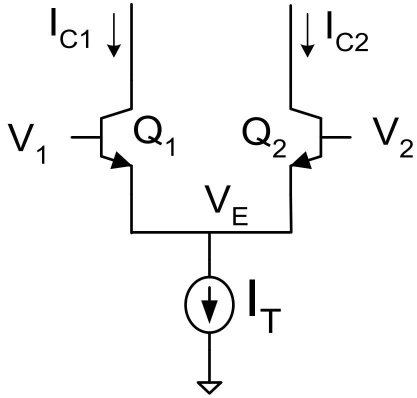
for 1% deviation, V_d=.56V_t

for 0.1% deviation, V_d=.27V_t

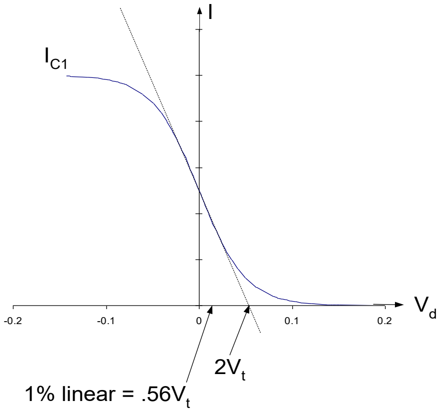
Signal Swing and Linearity of Bipolar Differential Pair



What input range is possible when using the tail current to program the OTA ?

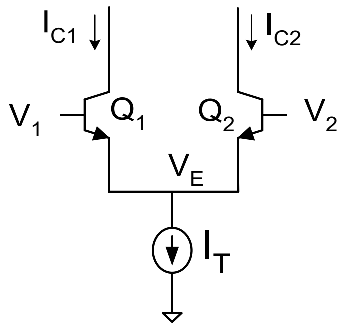


$$g_m = \frac{I_T}{2V_t}$$

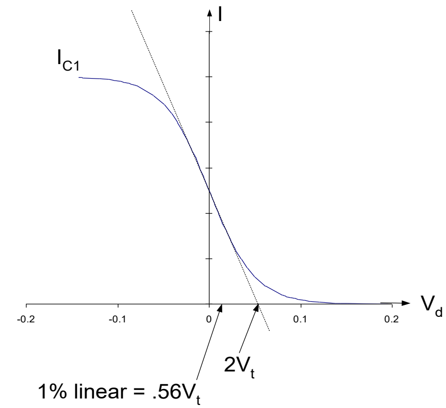


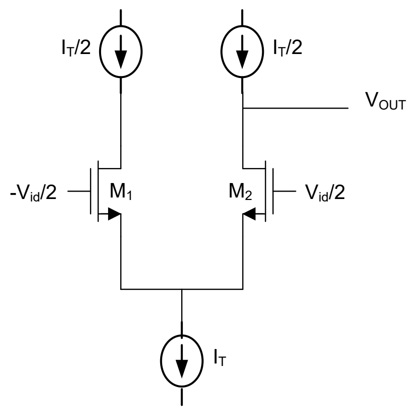
Since input signal swing not affected by I_T, Multi-decade adjustment of g_m with I_T without degrading signal swing

Signal Swing for basic MOS and BJT transconductors

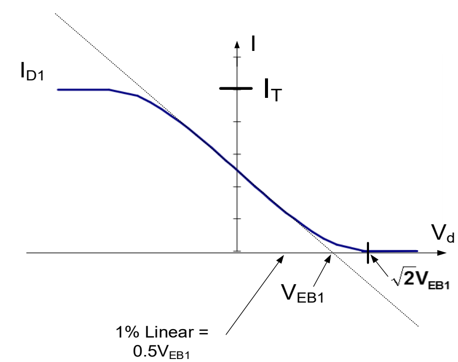


$$g_m = \frac{I_T}{2V_t}$$





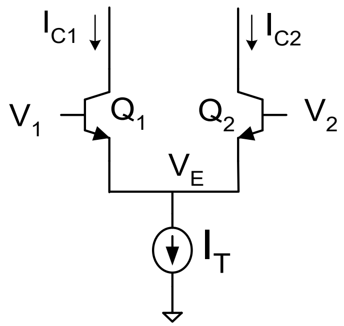
$$g_m = \mu C_{OX} \frac{W}{L} V_{EB} == \sqrt{I_T}\sqrt{\mu C_{OX} \frac{W}{L}}$$



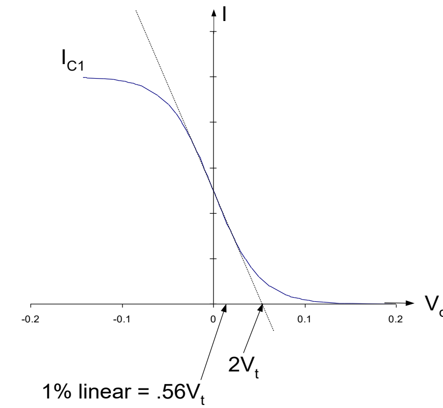
Signal Swing and Linearity Summary

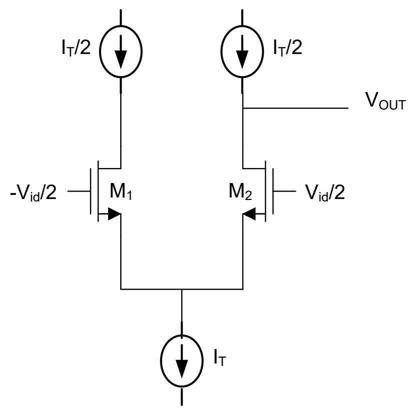
- Signal swing of MOSFET can be rather large if V_{EB} is large but this limits gain
- Signal swing of MOSFET degrades significantly if V_{EB} is changed for fixed W/L
- Bipolar swing is very small but independent of g_m
- Multiple-decade adjustment of bipolar g_m is practical
- Even though bipolar input swing is small, since gain is often very large, this small swing does usually not limit performance in feedback applications

Does the MOS or BJT transconductor have larger input signal swing?

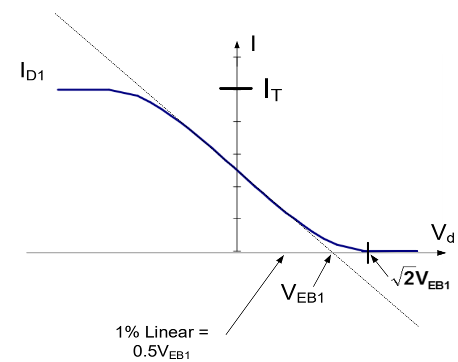


$$g_m = \frac{I_T}{2V_t}$$



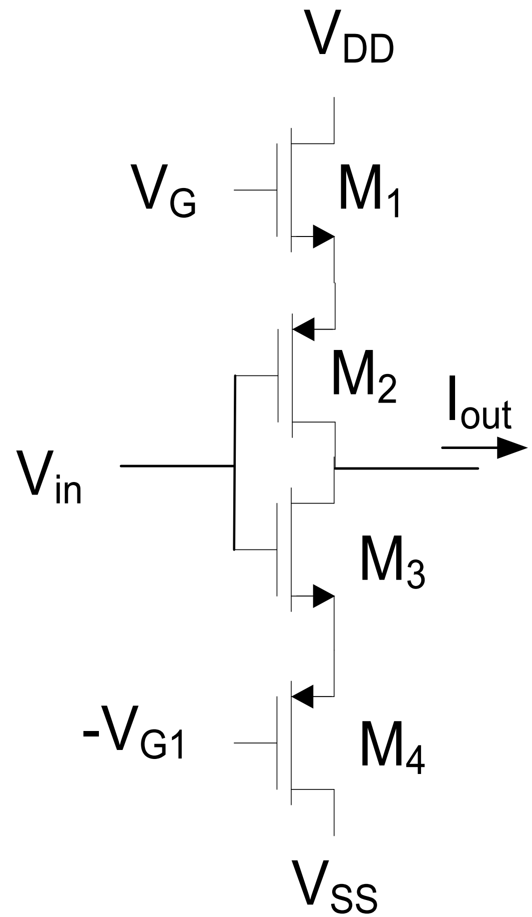


$$g_m = \mu C_{OX} \frac{W}{L} V_{EB}$$

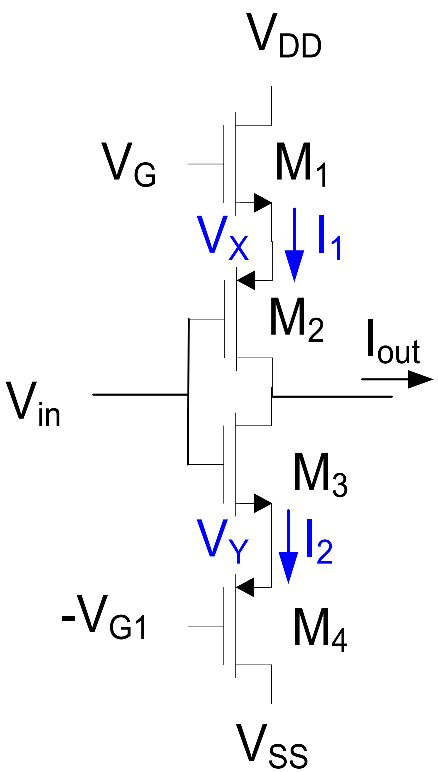


Depends upon how much adjustment range is desired

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

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

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

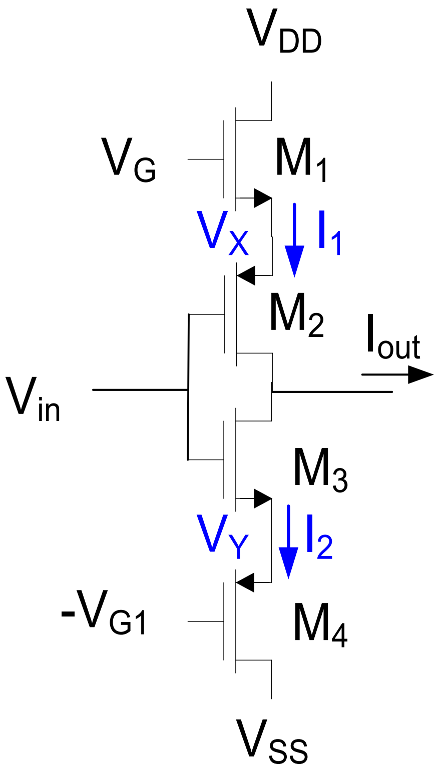
Adding these two equations, we obtain

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

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

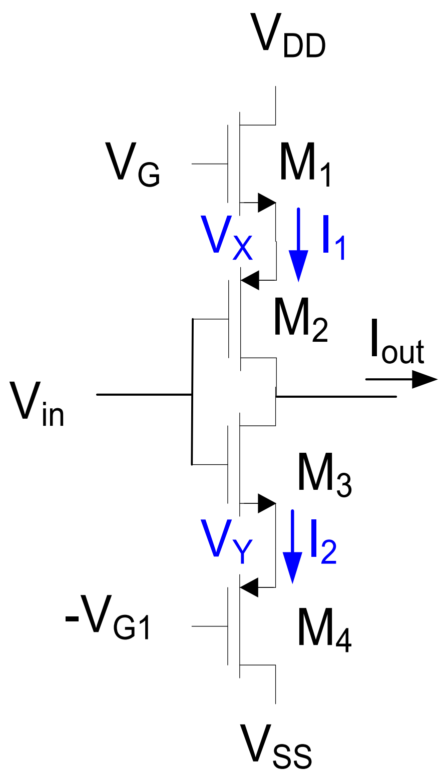
$$\left(\sqrt{\frac{1}{\beta_2}} + \sqrt{\frac{1}{\beta_1}}\right) = \sqrt{\frac{1}{\beta_5}}$$

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

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



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

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}

Bias Generators

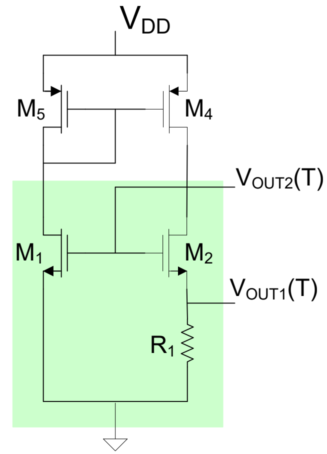
Bias voltage generators are widely used to bias cascode devices and other transistors in an IC

Key goal is often to have bias voltages independent of V_{DD} to avoid coupling supply noise into linear circuits

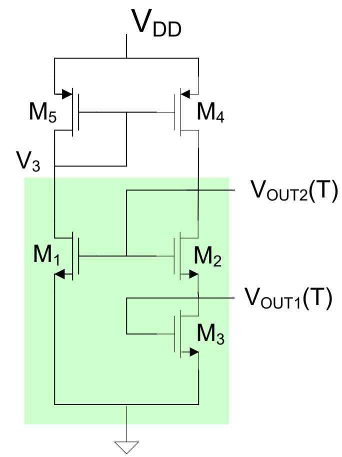
Potential Bias Generators

Consider the following four circuits:

Inverse Widlar



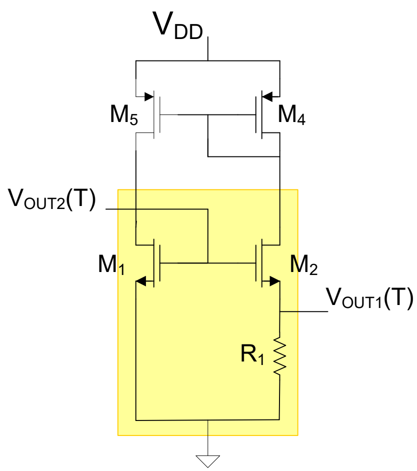
(a)



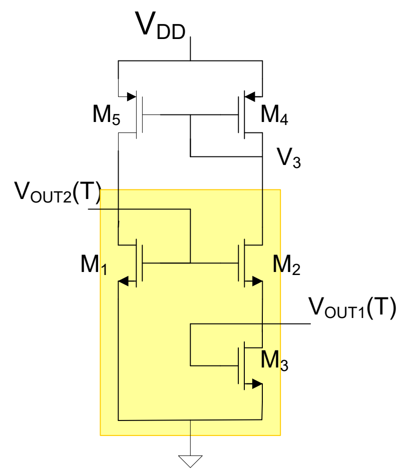
(b)

Inverse Widlar

Widlar



(c)



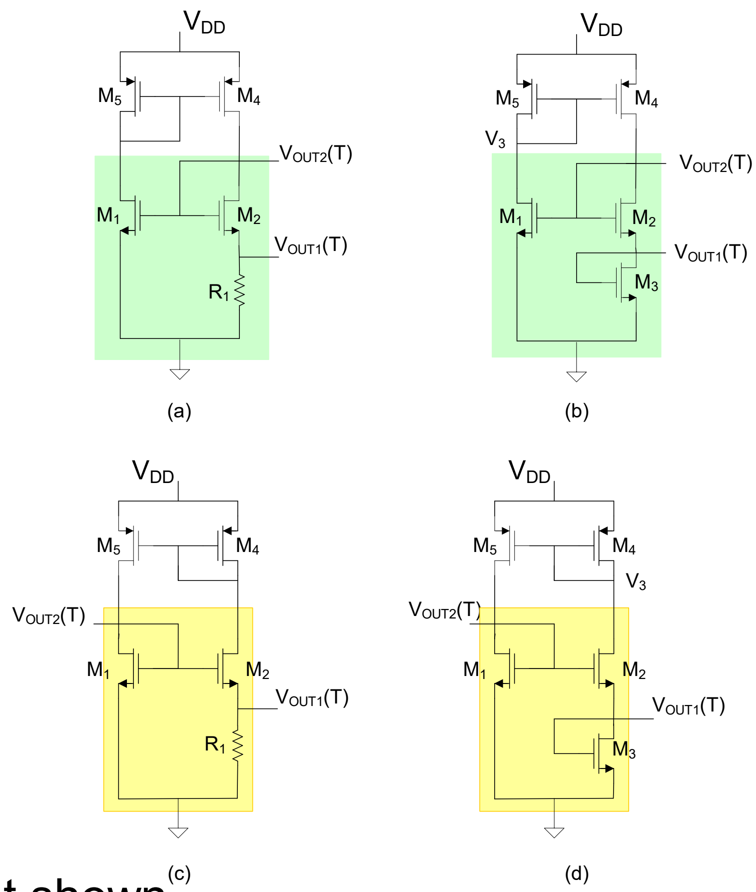
(d)

Widlar

Potential Bias Generators

Inverse Widlar

Inverse Widlar



Widlar

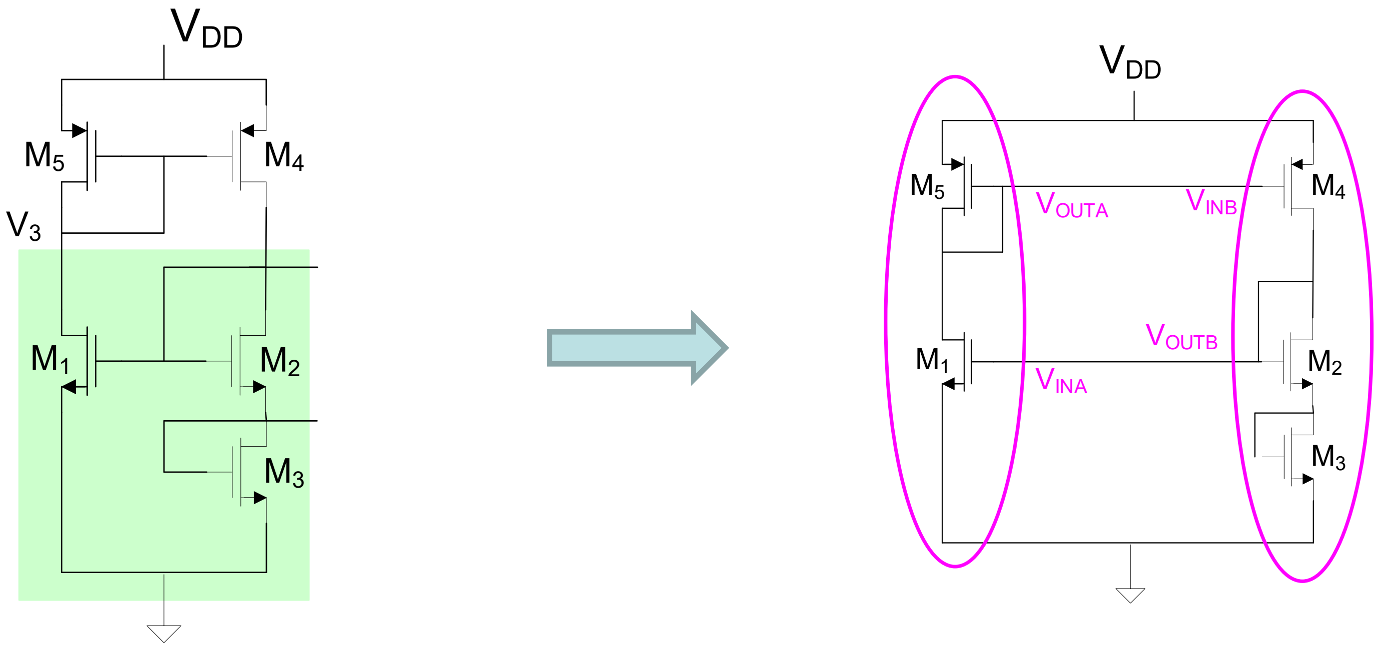
Widlar

- Start up circuits not shown
- If g_o is neglected, it can be shown that all devices are operating in the saturation region, the output voltages are independent of V_{DD}
- Note all have a positive feedback loop !

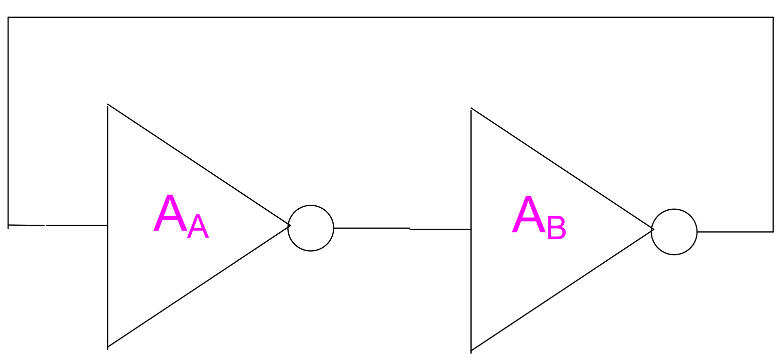
Regenerative Feedback Loops Can Provide Some Very Useful Properties but Can Also Offer Some Surprises !!

Theorem: If the small signal loop gain of the positive feedback loop is less than unity at an equilibrium point of the return map, then the equilibrium point is a stable equilibrium point and if the loop gain is larger than unity at an equilibrium point the equilibrium point is an unstable equilibrium point.

Consider the Inverse Widlar Bias Generator



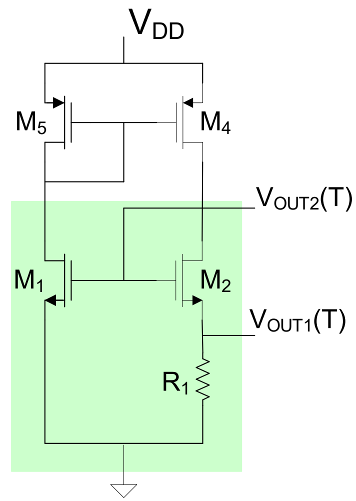
Can be viewed as two common-source amplifiers in a loop

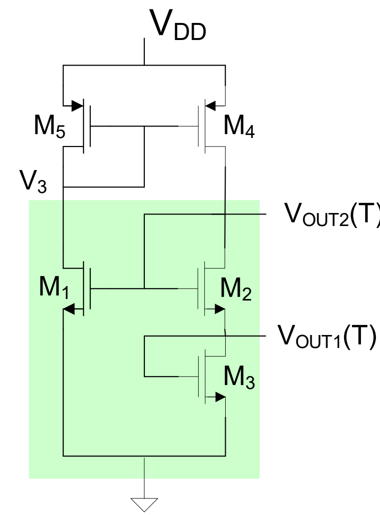


Same observation about the other 3 structures

V_{DD} Independent Bias Generators

Consider the two Inverse Widlar bias generators (start-up ckts not shown)





Assuming all devices in saturation,

$$V_{02} = V_{Tn} + \frac{\theta}{2} \pm \sqrt{\frac{\theta V_{Tn}}{2} + \left(\frac{\theta}{2}\right)^2}$$

$$V_{01} = \frac{\theta}{2} \pm \sqrt{\frac{\theta V_{Tn}}{2} + \left(\frac{\theta}{2}\right)^2}$$

$$-\sqrt{\frac{2L_2}{\mu_n C_{OX} W_2 R_1}}\sqrt{V_{Tn} + \frac{\theta}{2} \pm \sqrt{\frac{\theta V_{Tn}}{2} + \left(\frac{\theta}{2}\right)^2}}$$

$$V_{01} = V_{Tn}\left(\frac{1-\sqrt{\frac{W_2 L_1}{M_{54} W_1 L_2}}}{1+\sqrt{\frac{W_2 L_3}{W_3 L_2}} - \sqrt{\frac{W_2 L_1}{M_{54} W_1 L_2}}}\right)$$

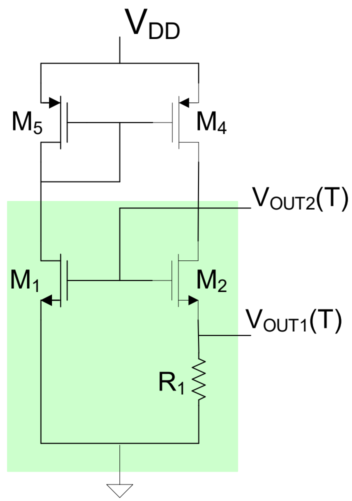
$$V_{02} = V_{Tn}\left(\frac{1+\sqrt{\frac{W_2 L_3}{W_3 L_2}} - 2\sqrt{\frac{W_2 L_1}{M_{54} W_1 L_2}}}{1+\sqrt{\frac{W_2 L_3}{W_3 L_2}} - \sqrt{\frac{W_2 L_1}{M_{54} W_1 L_2}}}\right)$$

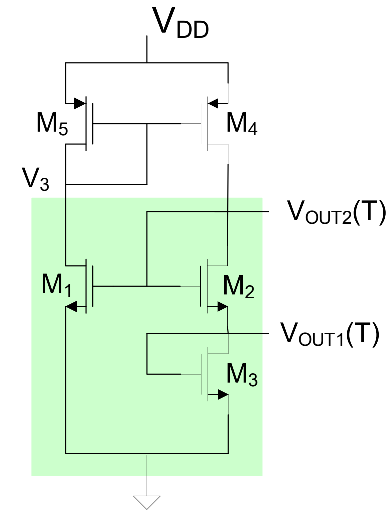
where $\theta = \frac{2L_1}{M_{54} R_1 \mu_n C_{OX} W_1}$ and M_{54} is the M_5:M_4 mirror gain

Note: Outputs V_{DD} independent !

V_{DD} Independent Bias Generators

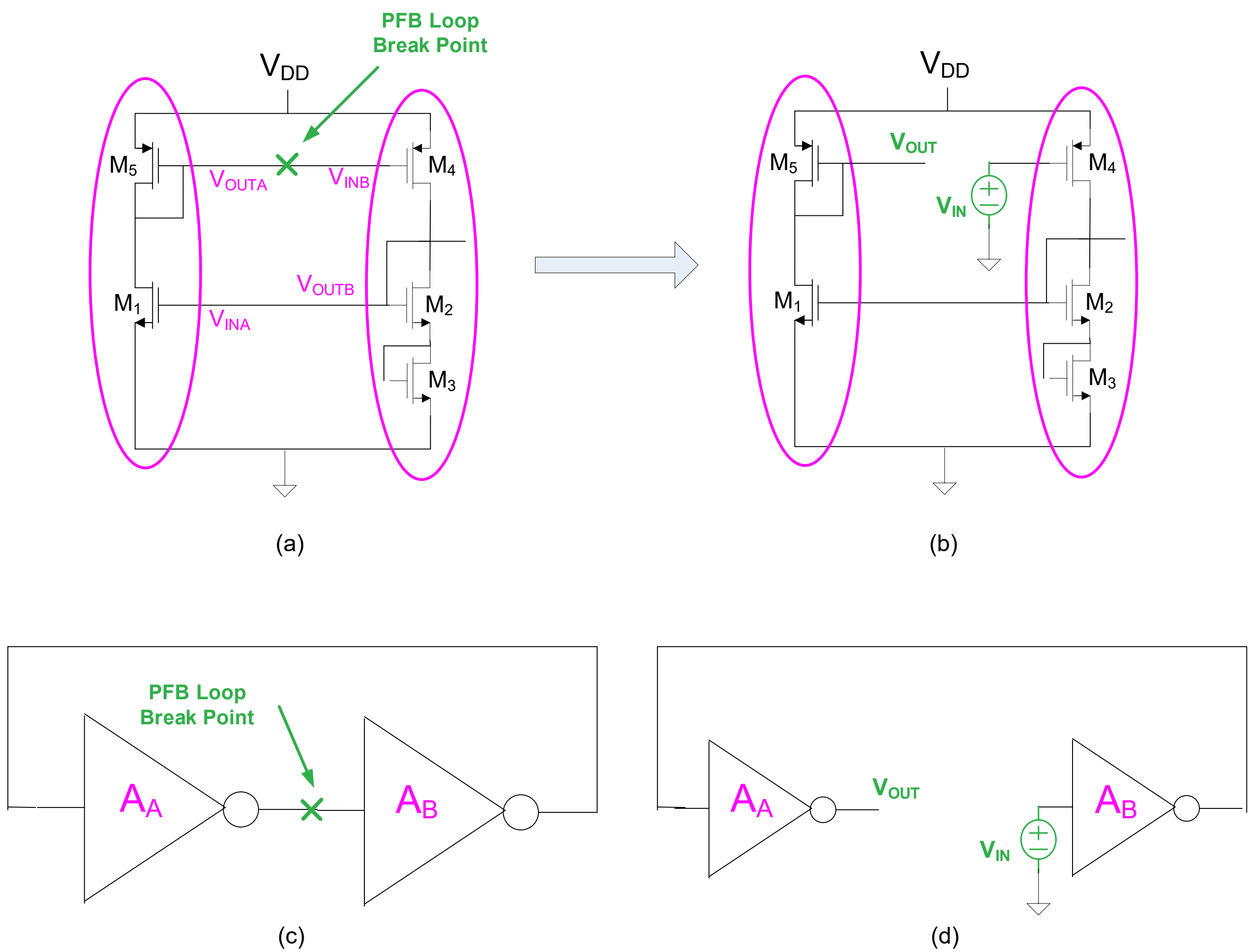
Consider the two Inverse Widlar bias generators (start-up ckts not shown)





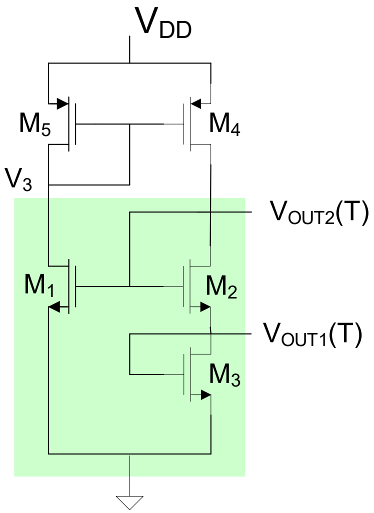
Must still check for stationarity of operating point, stability, and start-up

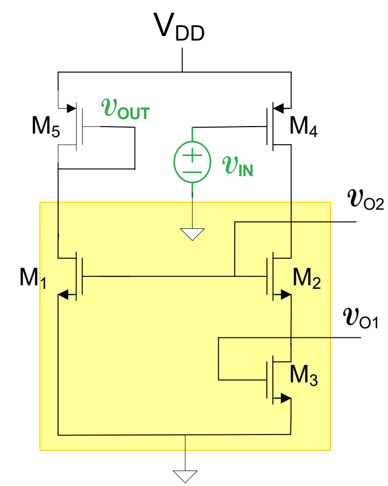
Consider Inverse Widlar with Transistor M_3 first



V_{DD} Independent Bias Generators

Check for stationarity of operating point





$$A_{LOOP} = \frac{g_{m1}}{g_{m5}} \bullet g_{m4}\left(\frac{1}{g_{m2}} + \frac{1}{g_{m3}}\right)$$

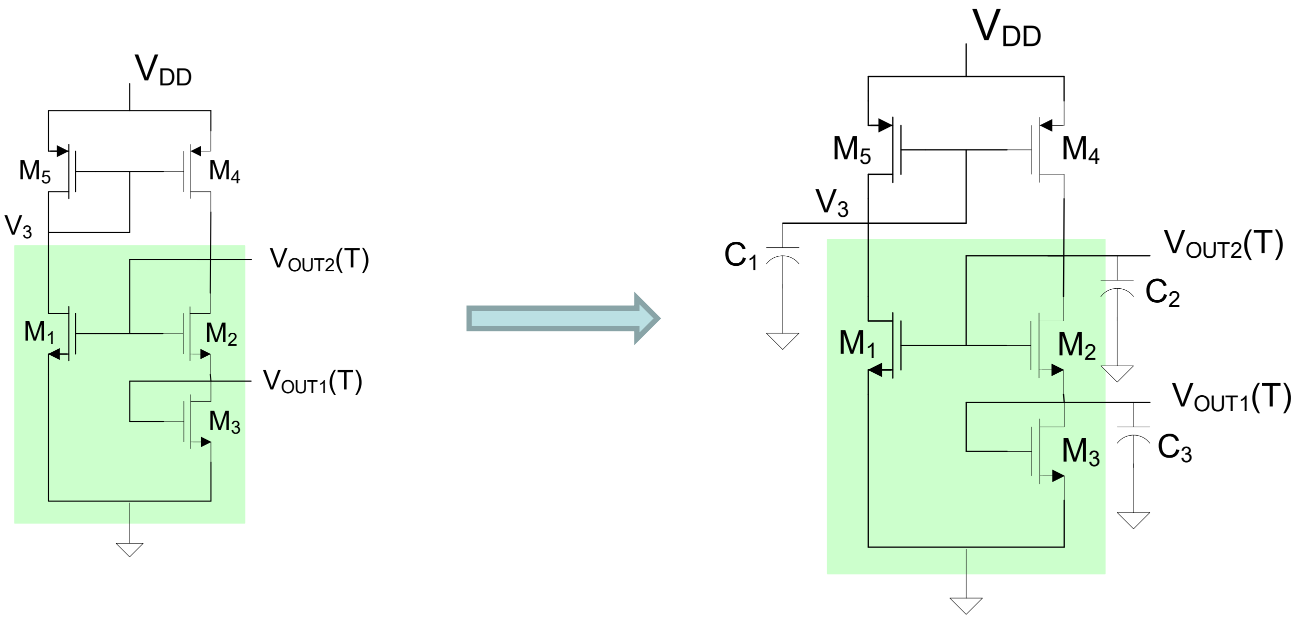
$$g_m = \frac{2I_Q}{V_{EB}}$$

$$A_{LOOP} = \frac{V_{EB2} + V_{EB3}}{V_{EB2} + V_{EB3} + V_{Tn}}$$

- Observe loop gain is always less than 1
- So it is a viable circuit for a bias generator

V_{DD} Independent Bias Generators

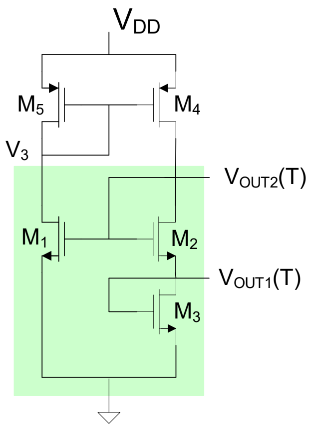
Check for stability

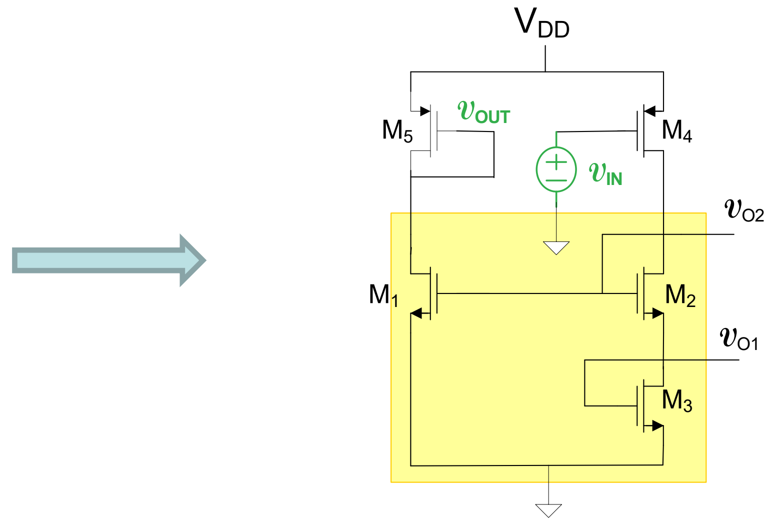


- Circuit has 3 poles
- May use RH criteria
- If unstable, adjust one of the capacitors

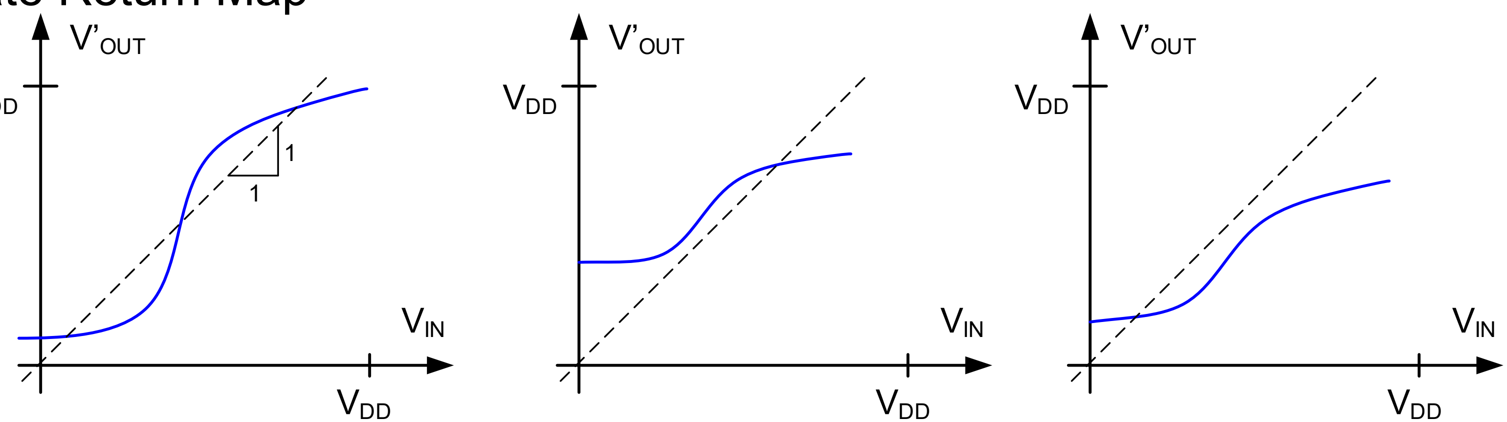
V_{DD} Independent Bias Generators

Check for startup





Create Return Map

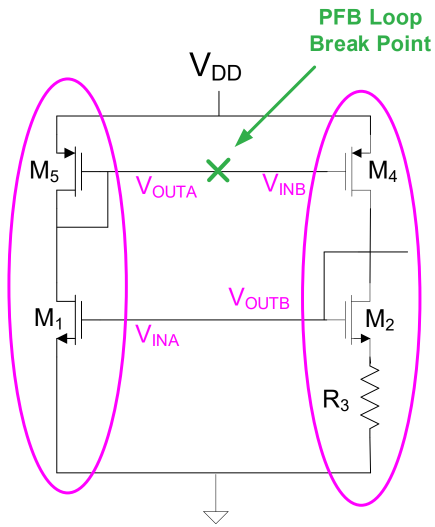


(a) (b) (c)

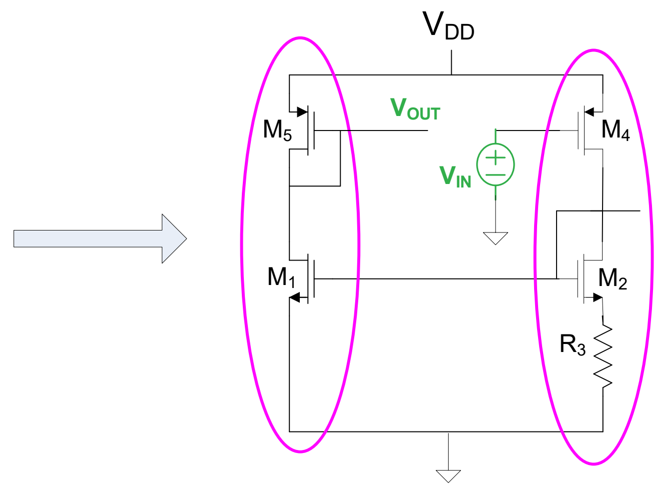
Must have single intersection point (desired point) with slope at unity gain crossing less than 1 over PVT variations

Add/modify startup circuit if necessary (usually necessary with this structure)

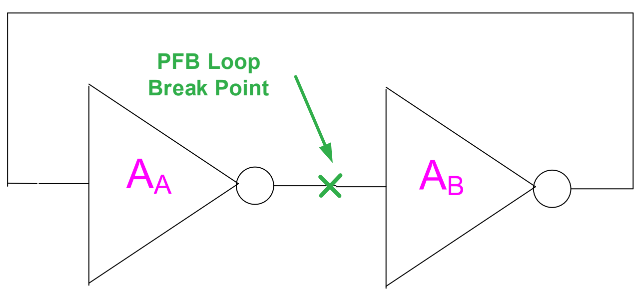
Consider Inverse Widlar with Transistor Resistor



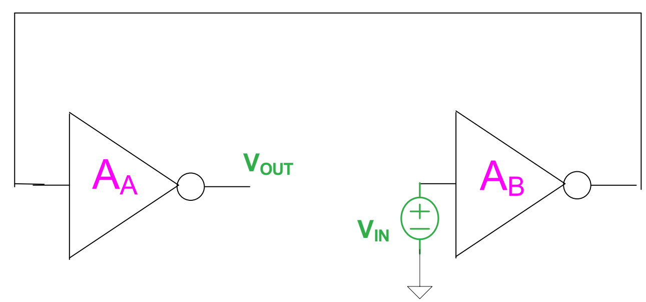
(a)



(b)



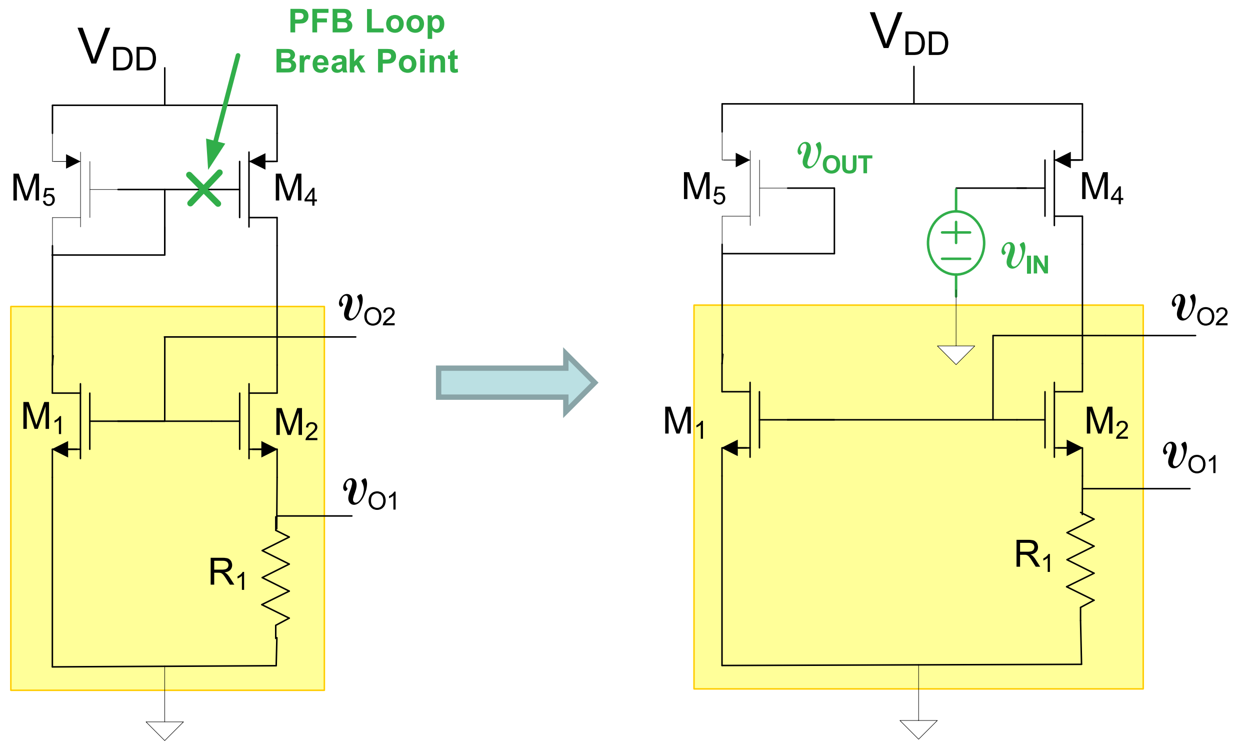
(c)



(d)

V_{DD} Independent Bias Generators

Check for stationarity of operating point



$$A_{LOOP} = \frac{g_{m1}}{g_{m5}} g_{m4} \left(\frac{1}{g_{m2}} + R_1 \right)$$

$$A_{LOOP} = \left(1 + \frac{V_{01}}{V_{02} - V_{Tn}} \right) > 1$$

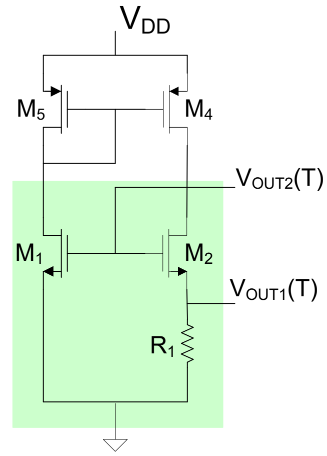
- Observe loop gain is always larger than 1
- So it is not a viable circuit for a bias generator

Basic Bias Generator Circuits

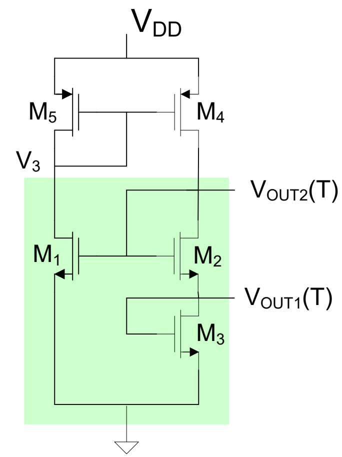
Only two of these circuits are useful directly as bias generators!

Inverse Widlar

Not stationary equilibrium point !



(a)



(b)

Inverse Widlar

$$V_{01} = V_{Tn}\left(\frac{1-\sqrt{\frac{W_2L_1}{M_{IW}W_1L_2}}}{1+\sqrt{\frac{W_2L_3}{W_3L_2}}-\sqrt{\frac{W_2L_1}{M_{IW}W_1L_2}}}\right)$$

$$V_{02} = V_{Tn}\left(\frac{1+\sqrt{\frac{W_2L_3}{W_3L_2}}-2\sqrt{\frac{W_2L_1}{M_{IW}W_1L_2}}}{1+\sqrt{\frac{W_2L_3}{W_3L_2}}-\sqrt{\frac{W_2L_1}{M_{IW}W_1L_2}}}\right)$$

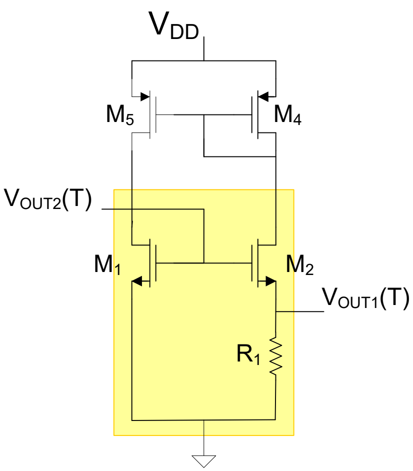
Widlar

$$V_{01} = \left(\frac{\theta_1}{2} \pm \sqrt{\frac{\theta_1 V_{Tn}}{2} + \left(\frac{\theta_1}{2}\right)^2}\right)\left(1-\sqrt{\frac{W_1L_2}{M_W W_2 L_1}}\right)$$

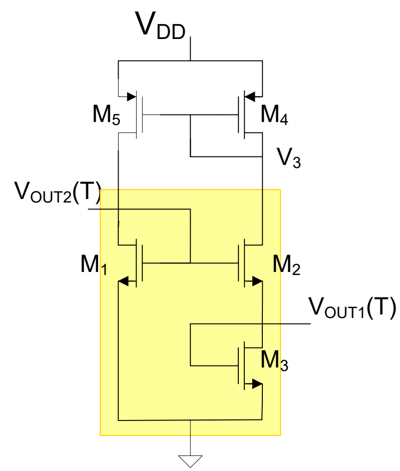
$$V_{02} = V_{Tn} + \frac{\theta_1}{2} \pm \sqrt{\frac{\theta_1 V_{Tn}}{2} + \left(\frac{\theta_1}{2}\right)^2}$$

$$I_{D1} = M_W I_{D2}$$

$$\theta_1 = \frac{M_W 2L_1}{R\mu_n C_{OX} W_1}$$



(c)

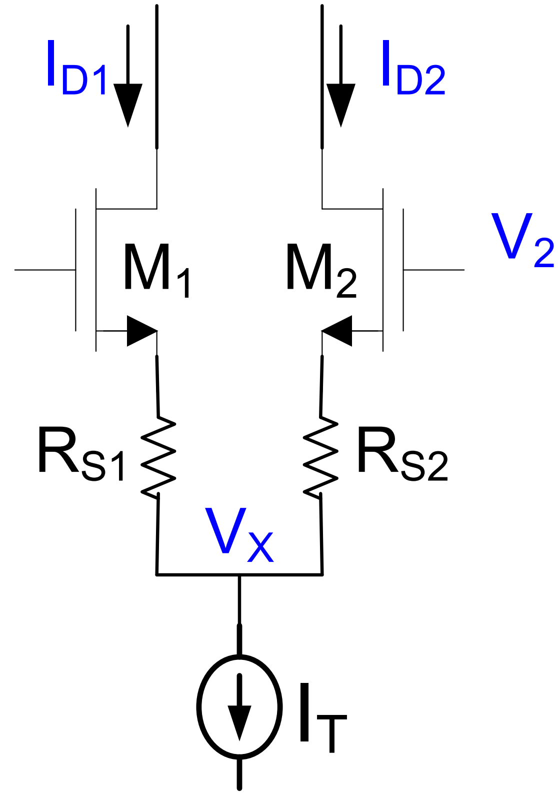


(d)

Widlar

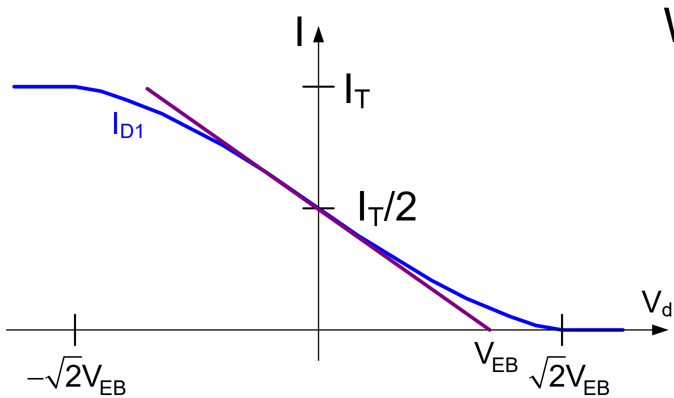
Not stationary equilibrium point !

Transconductance Linearization Strategies

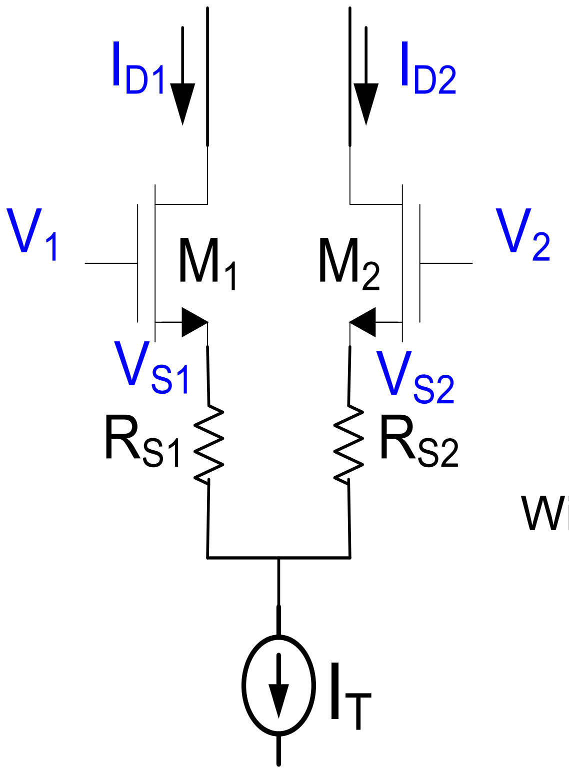


Widely used source degeneration

Recall with R_S=0



Transconductance Linearization Strategies



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

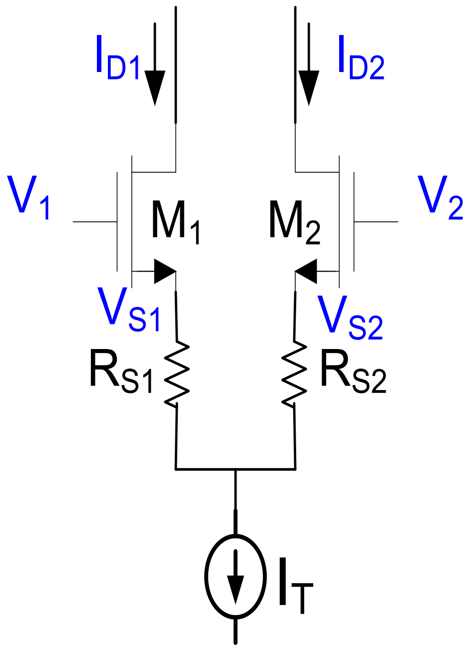
With a straightforward analysis, we obtain the expression

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

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



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

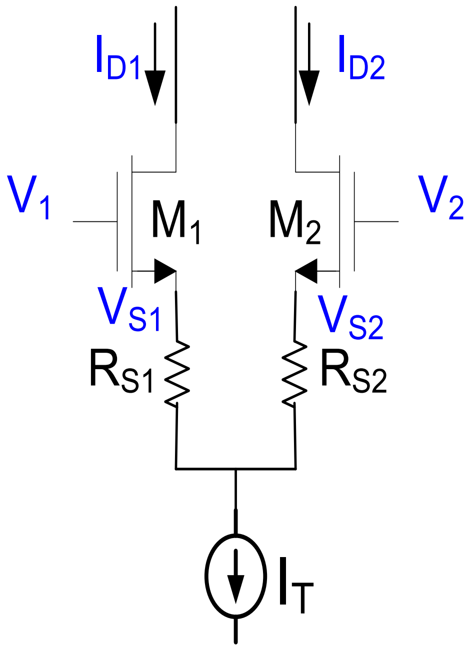
The transconductance of this structure can be readily derived to obtain

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

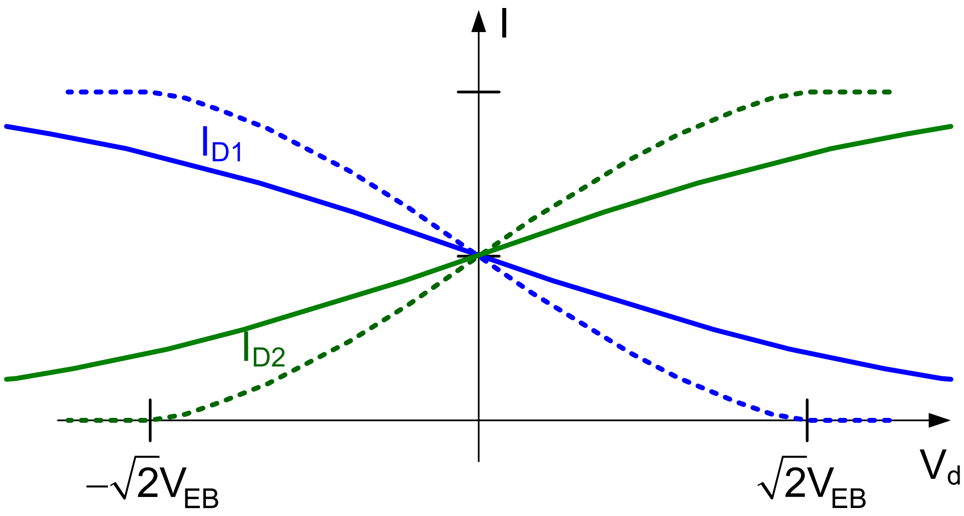
This can be expressed as

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

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

Transconductance Linearization Strategies

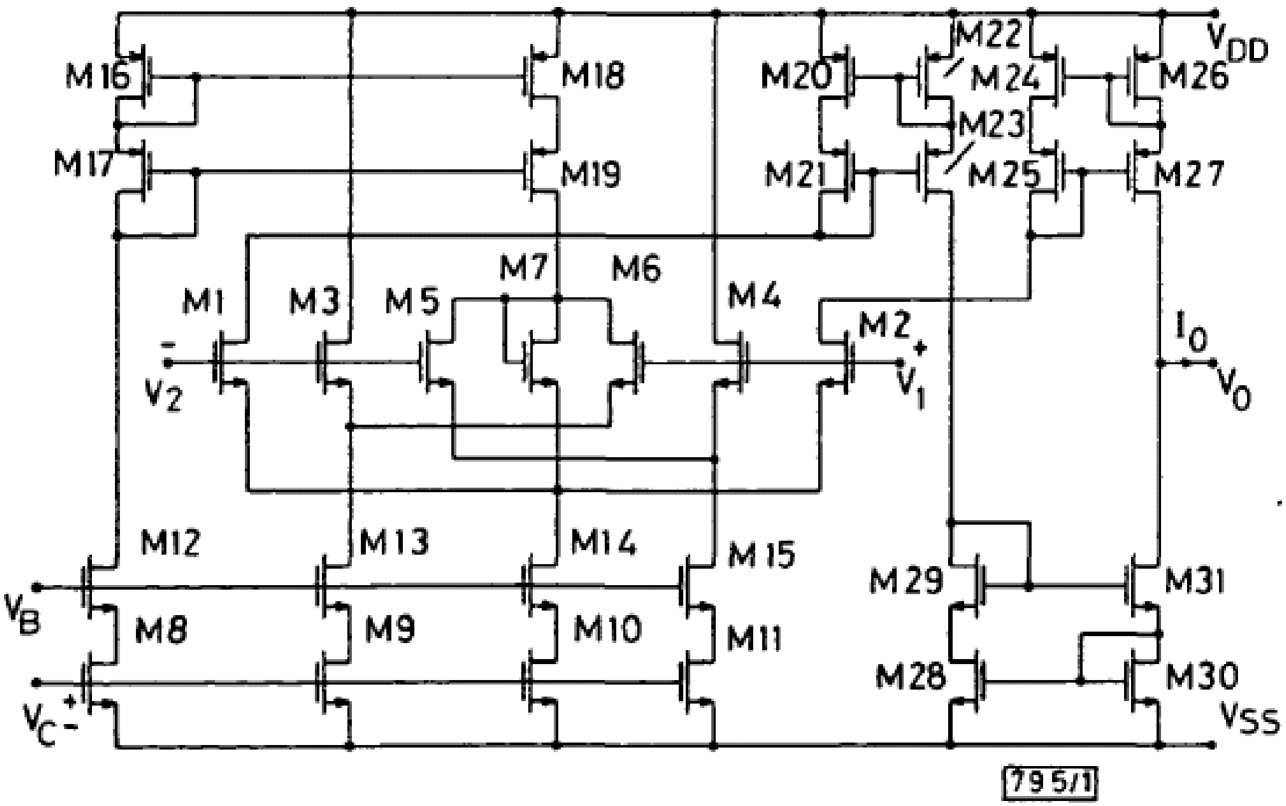
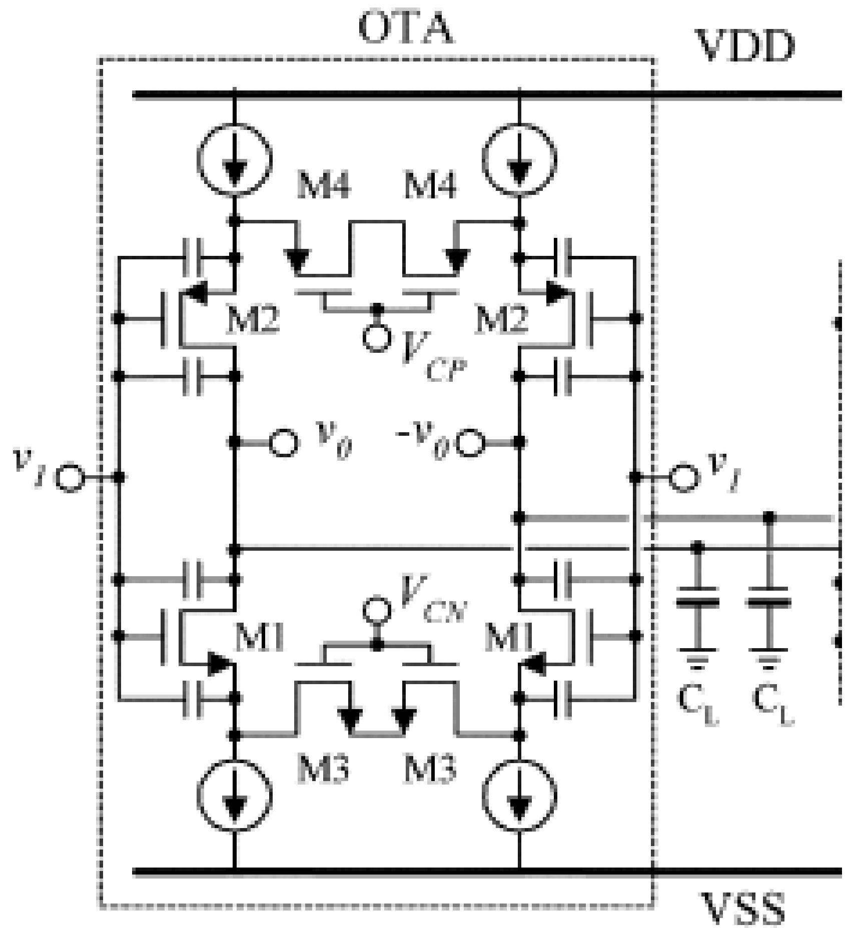


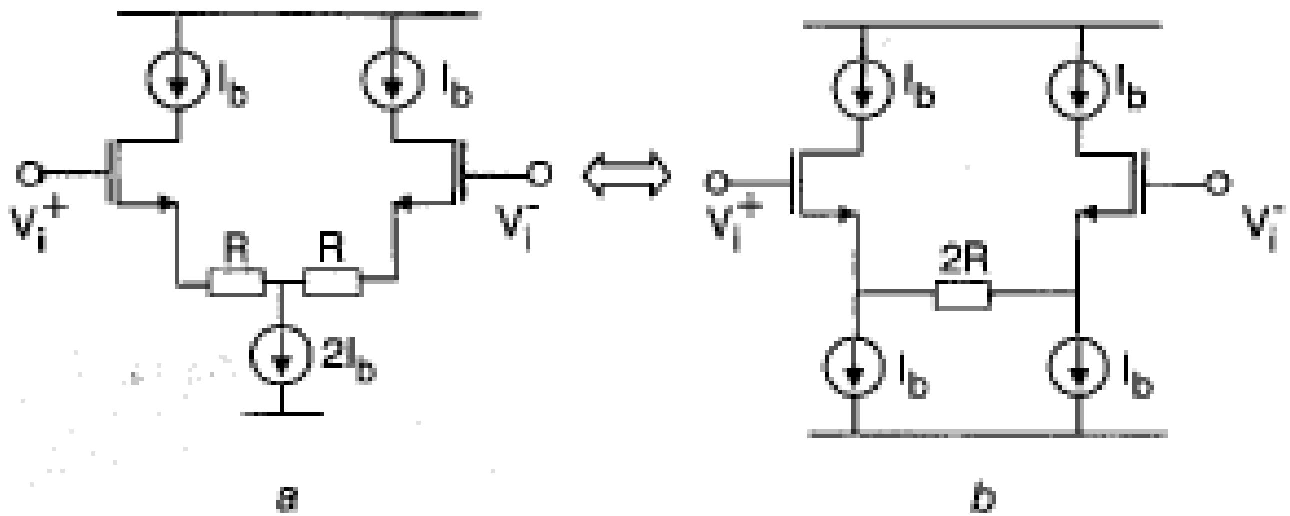
Fig. 1 *Linearised CMOS transconductance circuit*

Transconductance Linearization Strategies



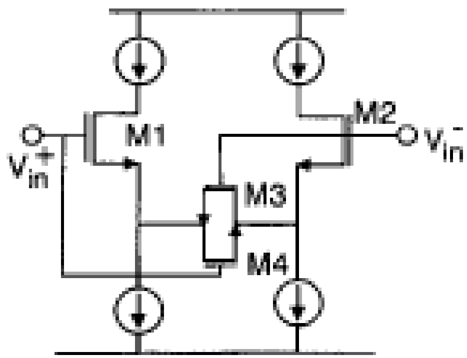
From CAS 2006 P 811 Jose Silva

Transconductance Linearization Strategies

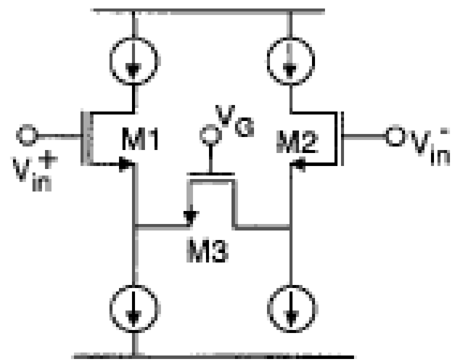


Linearity Enhancement with Source Degeneration

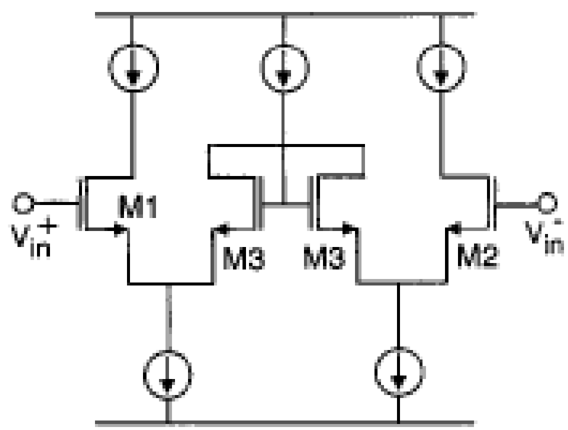
Transconductance Linearization Strategies



a



b

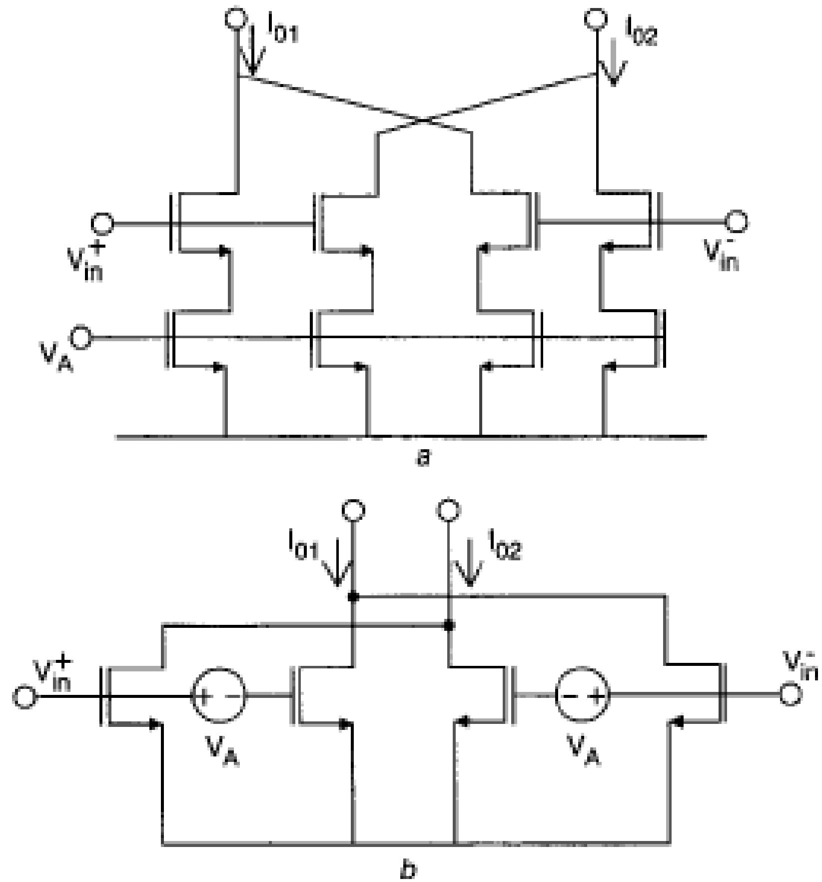


Linearization with active source degeneration

CMOS transconductance amplifiers, architectures and active filters: a tutorial

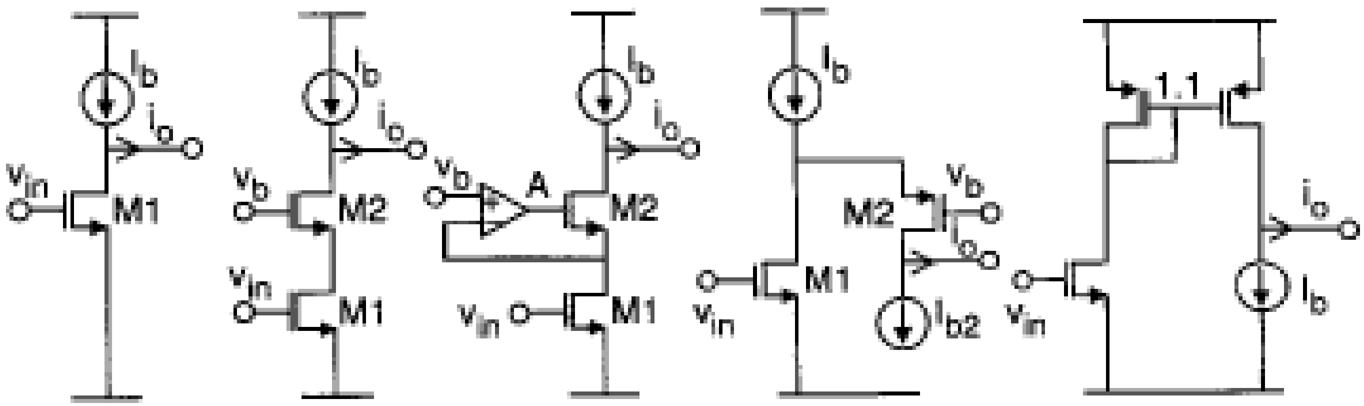
E.Sánchez-Sinencio and J.Silva-Martínez

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



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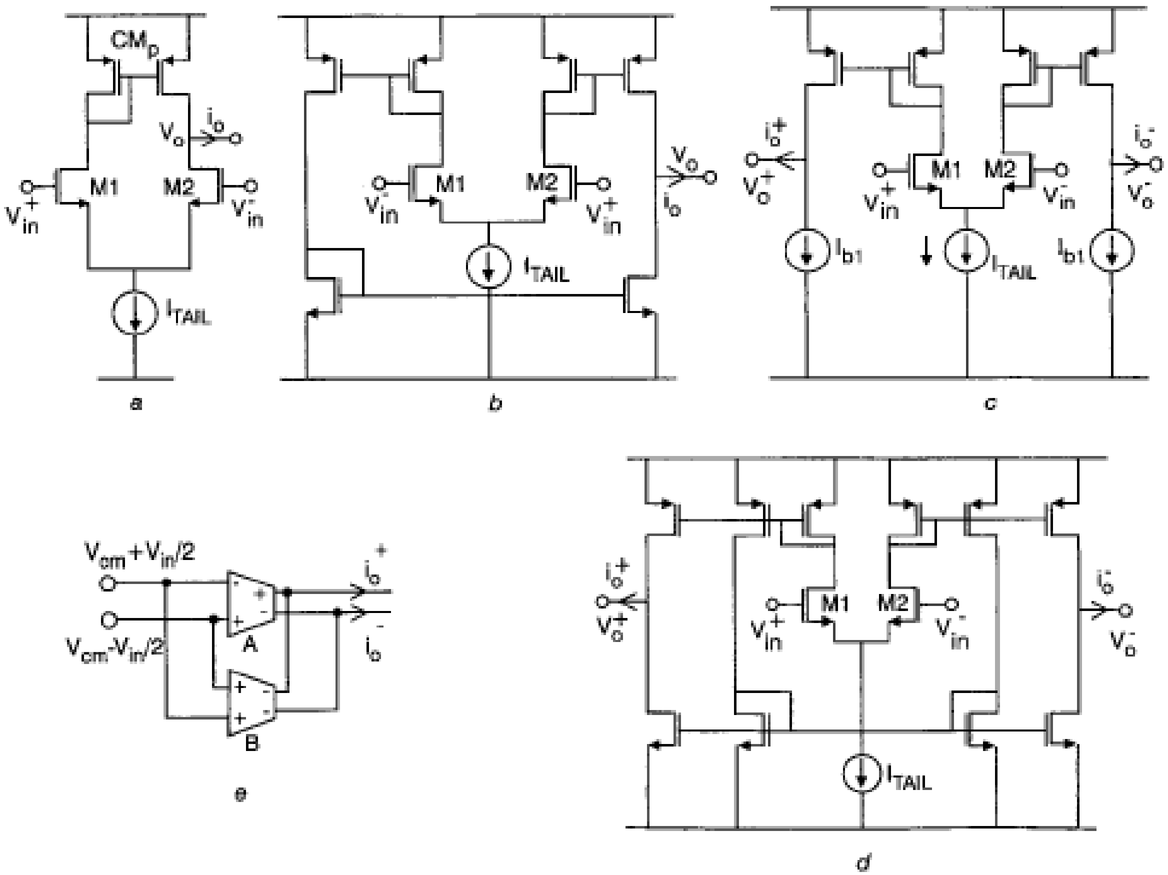
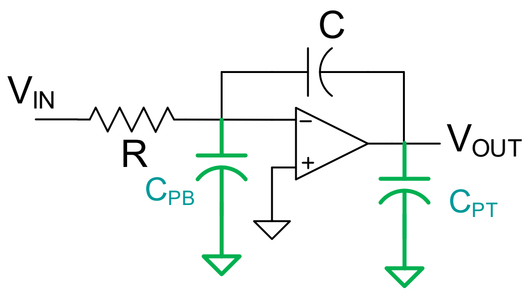
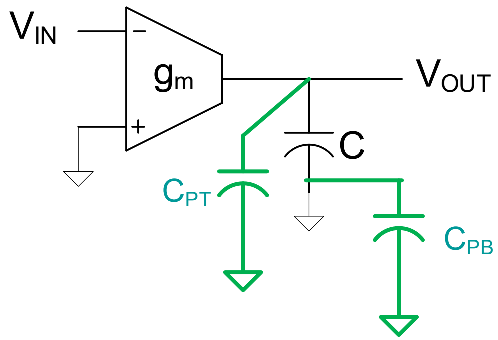
Differential input OTAs



Differential input and output OTAs

Parasitic Capacitances and Floating Nodes





Recall: 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 active RC filters have no floating nodes

Invariably the OTA-C integrators have floating nodes so are sensitive to parasitic capacitances

When filters are programmable or calibrated, floating nodes are less problematic but may add nonlinearity

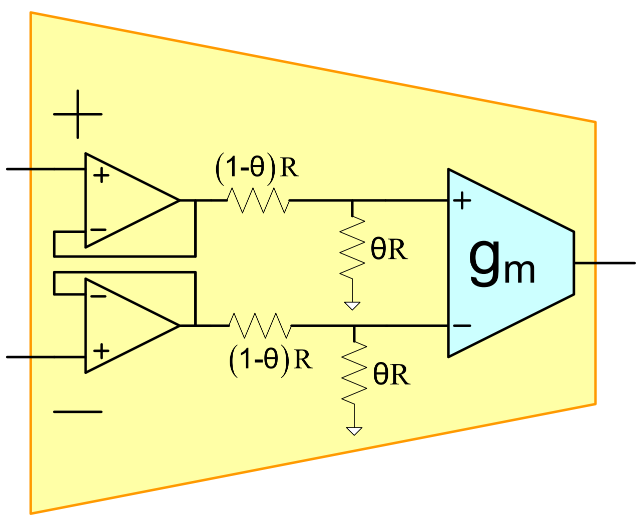
Signal Swing in OTA Circuits

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 OTA

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



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

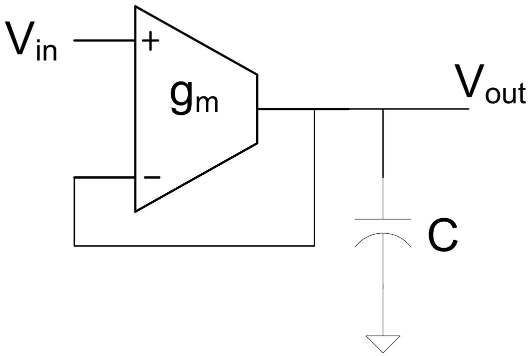
R. L. Geiger and E. Sánchez-Sinencio, "Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial," *IEEE Circuits and Devices Magazine*, Vol. 1, pp.20-32, March 1985.

Active Filter Design Using Operational Transconductance Amplifiers: A Tutorial

Randall L. Geiger and Edgar Sánchez-Sinencio

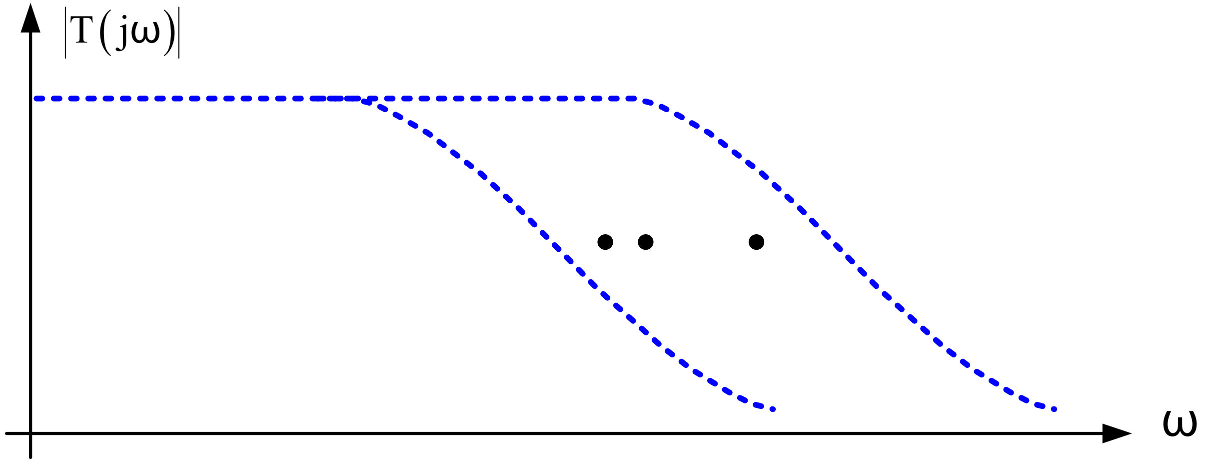
Programmable Filter Structures

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

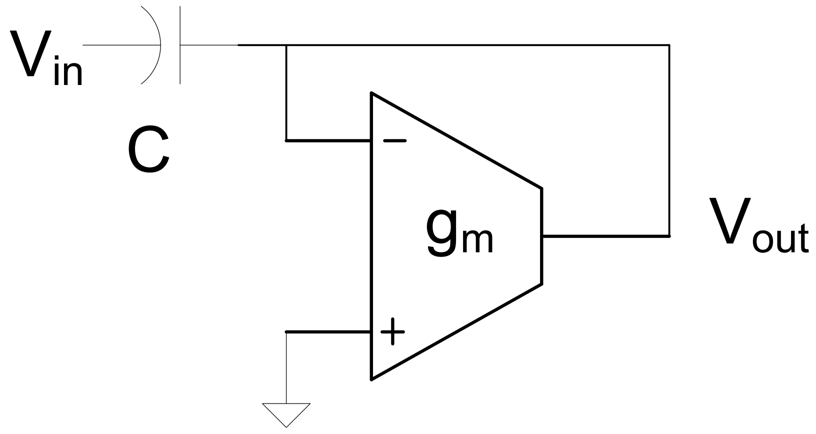


$$T(s) = \frac{g_m}{g_m + sC}$$

Programmable First-Order Low-Pass Filter

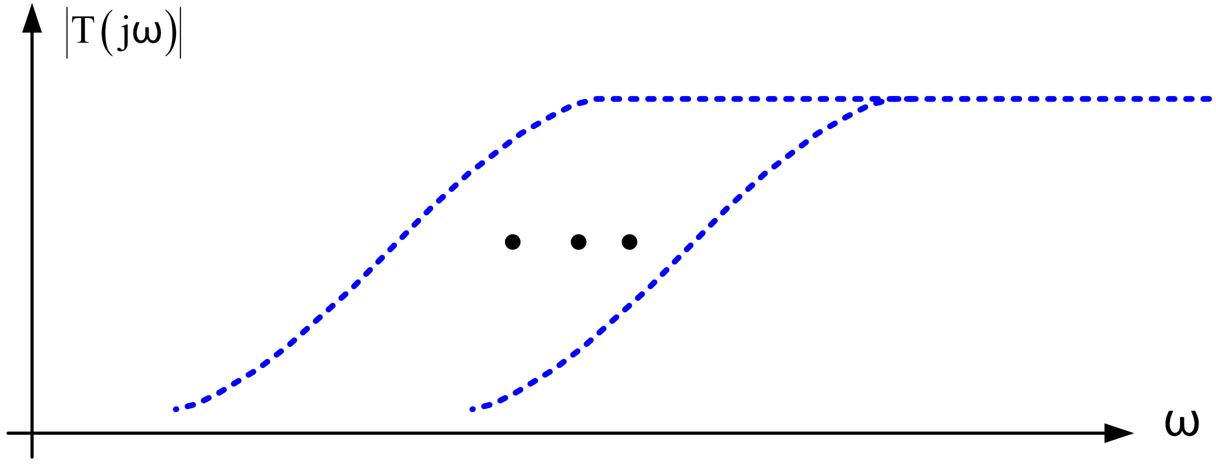


Programmable Filter Structures



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

Programmable First-Order High-Pass Filter



Stay Safe and Stay Healthy !

End of Lecture 33