ELEN 607
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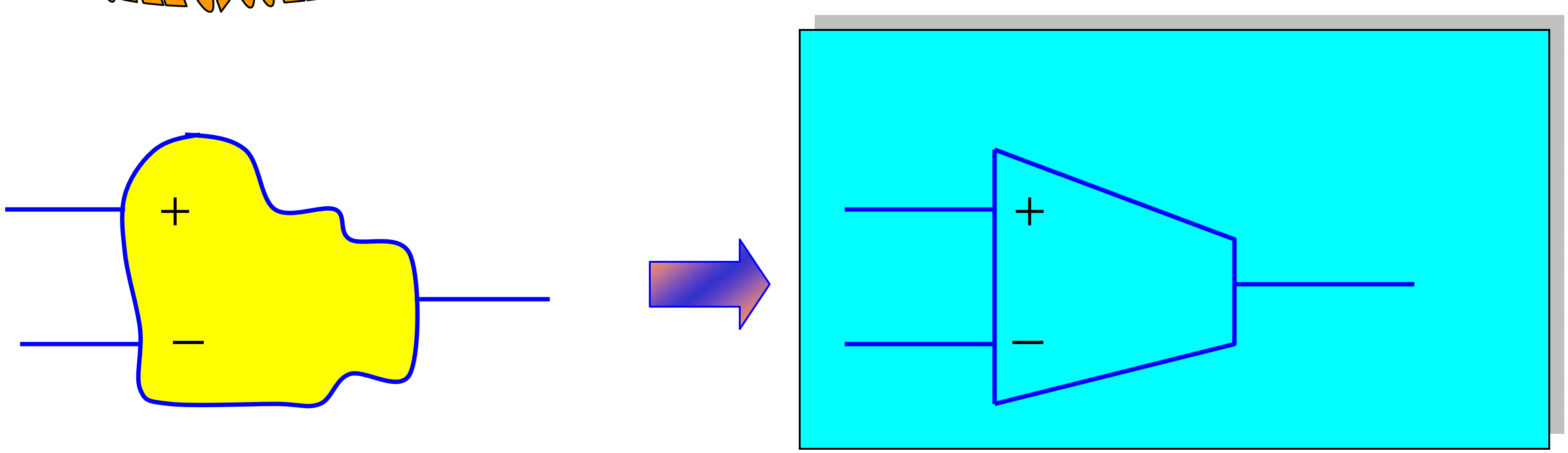
Fundamental techniques for OTA linearization



Fundamental OTA Linearization Techniques

The input-output characteristics of an OTA can be expressed as a non-linear polynomial

$$i_{\mathrm{D}} = I_D + a_1 v_{in} + a_2 v_{in}^2 + \ldots + a_n v_{in}^n$$

For linear applications, the ideal input-output relation for an OTA is:

$$i_{\mathrm{D}} = I_D + i_d = I_D + a_1 v_{in}$$

In actual implementations is not possible to obtain the above expression. Thus, we can only aim to reduce as much as possible the coefficients of the higher-order terms , that is $a_2, a_3, \ldots, a_n$

In what follows, we will discuss different linearization techniques to minimize these higher order terms.For instance differential outputs can ideally cancel all even coefficients.

To be able to make a weakly linear approximation, we can express the polynomial as a Taylor Series, as shown below:

$$i_d = i_d(v_{in})$$

$$i_d = I_D + \frac{\partial i_d}{\partial v_{in}}\Big|_Q v_{in} + \frac{1}{2}\frac{\partial^2 i_d}{\partial {v_{in}}^2}\Big|_Q {v_{in}}^2 + \frac{1}{6}\frac{\partial^3 i_d}{\partial {v_{in}}^3}\Big|_Q {v_{in}}^3 + \bullet\bullet$$

$$\text{for} \cdot v_{in} = V\sin(\omega t)$$

$$i_d \cong I_D + \left.\frac{1}{4}\frac{\partial^2 i_d}{\partial {v_{in}}^2}\right|_Q V^2 + \left.\left(\frac{\partial i_d}{\partial v_{in}}\right|_Q + \left.\frac{\partial^3 i_d}{\partial {v_{in}}^3}\right|_Q \frac{V^2}{8} + \cdots\right) V\sin(\omega t) +$$

$$+ \left(\left.\frac{1}{4}\frac{\partial^2 i_d}{\partial {v_{in}}^2}\right|_Q + \cdots\right) V^2\sin(2\omega t) + \left(\left.\frac{1}{24}\frac{\partial^3 i_d}{\partial {v_{in}}^3}\right|_Q + \cdots\right) V^3\sin(3\omega t) + \bullet\bullet$$

Nonlinear Metrics

Transconductance

$$g_m = \left.\frac{\partial i_d}{\partial v_{in}}\right|_Q + \left.\frac{\partial^3 i_d}{\partial v_{in}^3}\right|_Q \frac{V^2}{8} + \cdots$$

Second Harmonic Distortion

$$HD_2 \cong \frac{1}{4} \frac{\left.\frac{\partial^2 i_d}{\partial v_{in}^2}\right|_Q}{\left.\frac{\partial i_d}{\partial v_{in}}\right|_Q} V$$

Third Harmonic Distortion

$$HD_3 \cong \frac{1}{24} \frac{\left.\frac{\partial^3 i_d}{\partial v_{in}^3}\right|_Q}{\left.\frac{\partial i_d}{\partial v_{in}}\right|_Q} V^2$$

Third Intermodulation Distortion

$$IM_3 = 3 * HD_3$$

Rough Classification of Transconductance Linearization Schemes

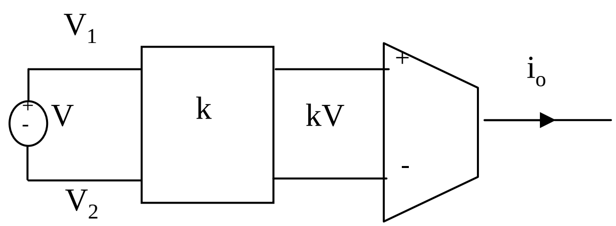
1. By attenuating the input signal.

2. By using negative feedback.

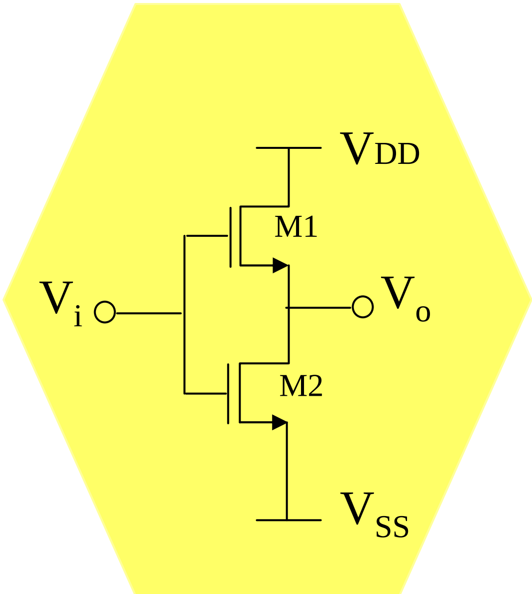
3. By strategically adding several non-linear OTAs.

4. By connecting a nonlinear function and its inverse non-linear function.

OTA Linearization Techniques using attenuation



$$i_o = a_1(kV) + a_2(kV)^2 + a_3(kV)^3 + \cdots$$



Active attenuation based on a self-cascode transistor

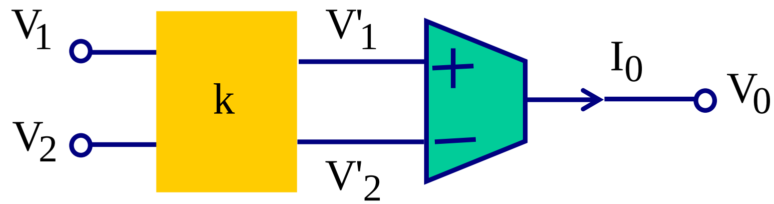
$$k = 1 - 1/\sqrt{1 + W_1L_2/W_2L_1} \text{ for } V_{T_1} = V_{T_2}$$

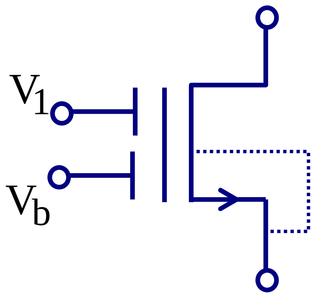
Nonlinearity Reduction

1. Input Attenuation

Potential Implementations :

- Floating Gate





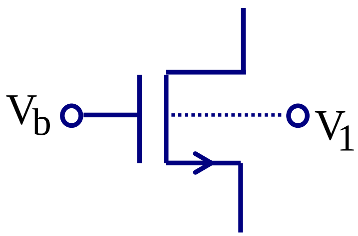
$$\Rightarrow k_1 g_m \cong \frac{C_b}{C_1 + C_b} g_m$$

Similar S/N with and without Floating Gate.

$\uparrow V_{off} = I_{off} / kai$

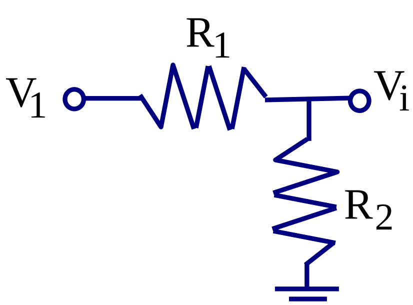
k can be frequency dependent

- Bulk Driven



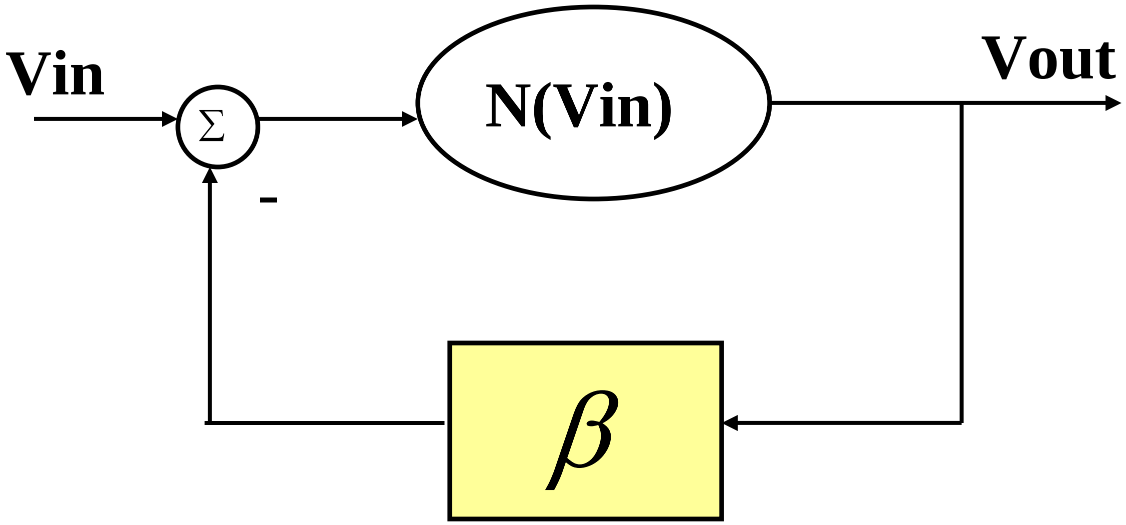
$$\Rightarrow \gamma g_m , \qquad 0.2 < \gamma < 0.4$$

- For discrete components, a voltage divider is possible :



$$V_i = \frac{R_2}{R_1 + R_2} V_1 = \frac{1}{1 + \frac{R_1}{R_2}} V_1 = k V_1$$

2. Linearization by using negative feedback.



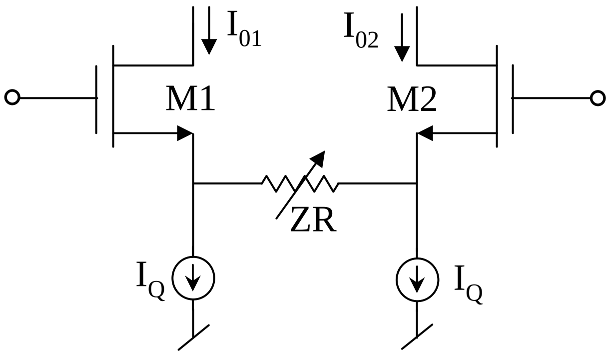
Example: A source degeneration differential pair

$$\beta = R_S \qquad \mathbf{N(Vin) = g_m}$$

$$G_m = \frac{g_m}{1+g_m R_s} = \frac{g_m}{1+N}$$

DEGENERATED DIFFERENTIAL PAIRS

⇨ CONCEPT



$V_c + \frac{1}{2}V_d$

$V_c - \frac{1}{2}V_d$

➢ BASIC PRINCIPLE:

Linearity is determined by the resistor

$$g_m = \frac{I_{01} - I_{02}}{V_d} = \frac{g_m I}{1 + g_{m2} R} \approx \frac{1}{R}$$

➢ CONDITIONS:

M1, M2 are saturated
M1, M2are perfectly matched

$g_{m1} R >> 1$

(g_{m1}, intrinsic transconductance of M1, M2)

V_{GS} OF M1, M2 can be regarded constant

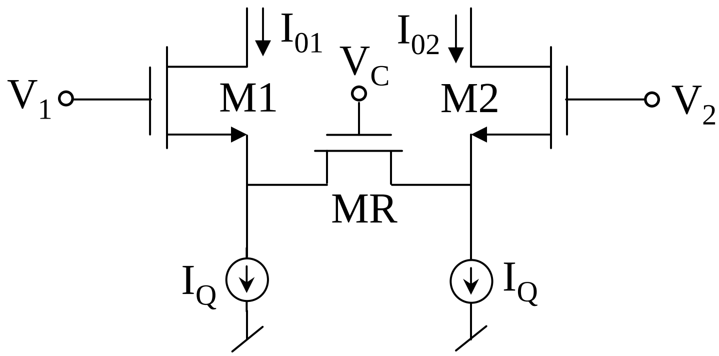
➢ GLOBAL CHARACTERISTICS:

High linearity at the cost of large area occupation.

Low transconductance (less suited for high frequency applications)

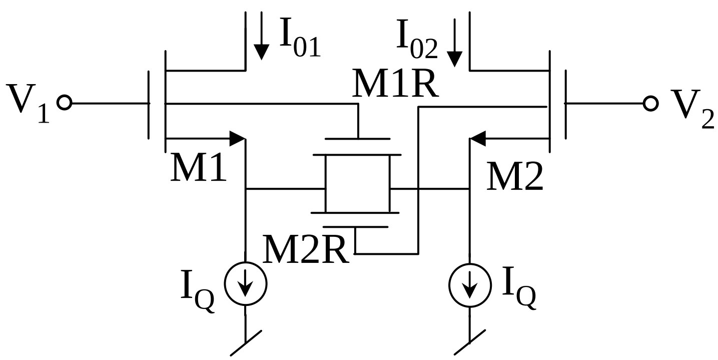
DEGENERATED DIFFERENTIAL PAIRS-IMPLEMENTATIONS

- Single Most in Ohmic Region As a Resistor (TSIV86)



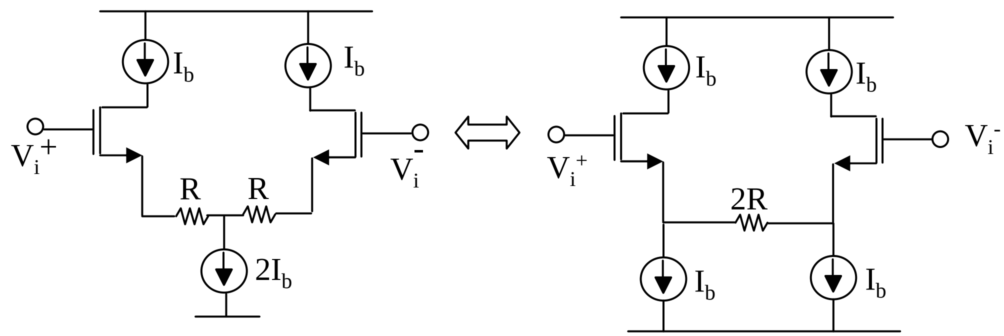
- Tuning via V_C
- $V_C >> \longrightarrow$
  MR in ohmic region
- $\left(\frac{W}{L}\right)_{1,2} >> \rightarrow$
  $V_{GE1,2} >> \sim coust$

- Two Most in Ohmic Region as a Resistor (KRUM88)

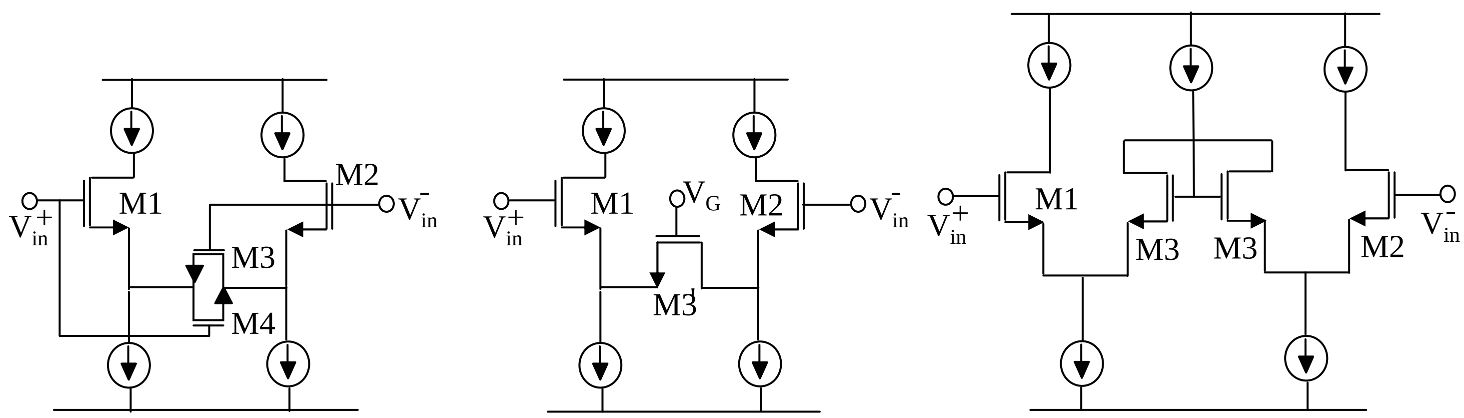


- Tuning via I_Q

Transconductance linearization schemes via source degeneration.



(a) (b)



(a) (b) (c)

Active Source Degeneration topologies; (a) and (b) transistors biased on triode region and (c) with saturated transistors.

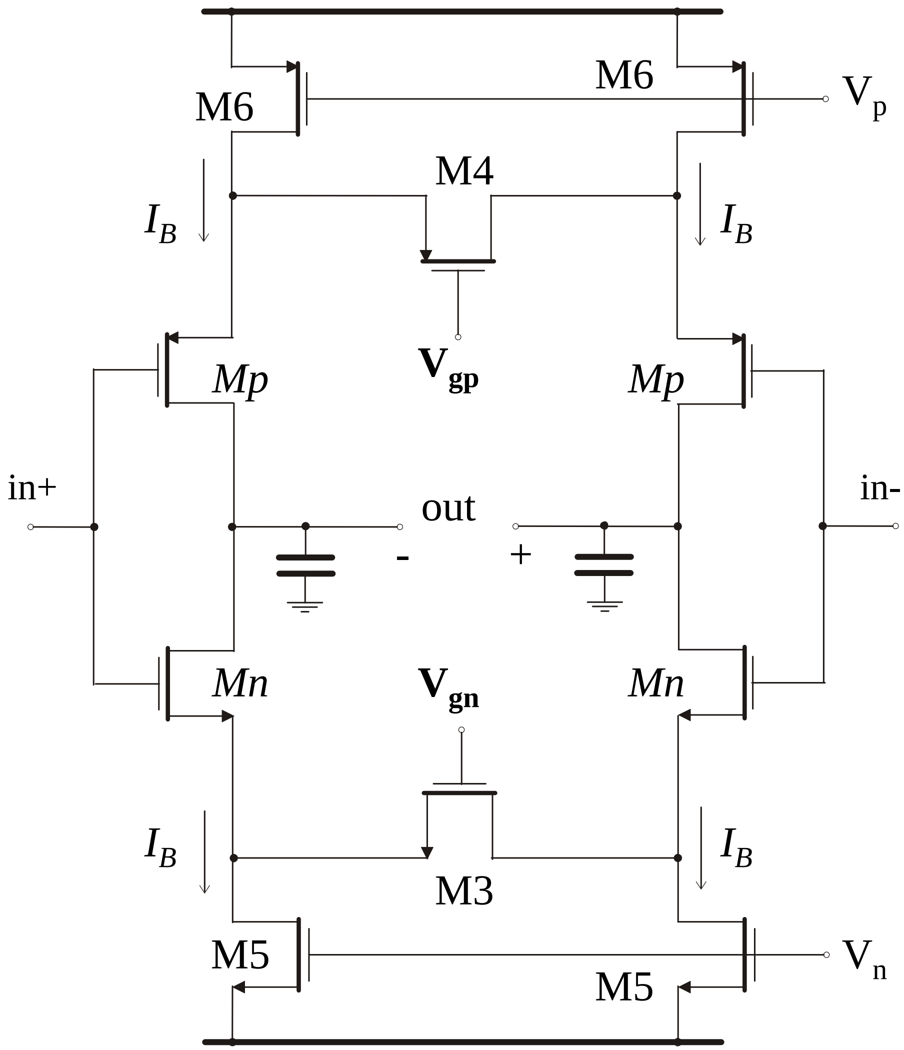
Table 1. Main characteristics for differential pair based OTAs.
N(=G_mR) is the source-degeneration factor.

Parameter	Differential Pair	Source Degeneraton
Small-signal transconductance	$G_m = \sqrt{2\mu_n C_{OX}} \sqrt{\frac{W_n}{L_n}} \sqrt{I_B}$	$G_{m,sd} = \frac{G_m}{1+N}$
Third Harmonic Distortion (HD3)	$HD3 = \frac{1}{32}\left(\frac{v_{id}}{V_{DSAT}}\right)^2$	$\left(\frac{1}{1+N}\right)^2 HD3$
Input referred thermal noise density	$\frac{16}{3}\frac{kT}{G_m}\left(1+\frac{g_{mp}}{G_m}\right)$	$\frac{16}{3}\frac{kT}{G_{m,sd}}\left(1+\frac{g_{mp}+\left(\frac{N}{1+N}\right)^2 g_{mn}}{G_{m,sd}}\right)$
Dynamic Range	$DR = \sqrt{\frac{(HD3)(C_L)}{NF} 10^{11} V_{DSAT}}$	$\sqrt{\frac{NF}{NF_{sd}}(1+N)DR}$
Current consumption*	$2I_B$	$2(1+N)I_B$
Transistor dimensions*	$\frac{W}{L}$	$(1+N)\frac{W}{L}$

* For comparison, same transconductance and same V_{DSAT} are fixed for both structures.
NF= NF_{sd} for N=0

Table 2. Properties of OTAs using source degeneration

Reference/Figure	Transconductance	Properties
Fig. (a)	$\dfrac{g_{m1}}{1+\dfrac{\beta_1}{4\beta_3}}$	Low sensitive to common-mode input signals. The linear range is limited to $V_{in}<V_{DSAT}$, and THD=-50 dB. M1=M2, M3=M4
Fig. (b)	$\dfrac{g_{m1}}{1+g_{m1}R}$ $R=1/\mu_o C_{ox}\left(V_{gs}-V_T\right)$	Highly sensitive to common-mode input signals. For better linearity large V_{GS3} voltages are required. Large tuning range if V_G is used.
Fig. (c)	$\dfrac{g_{m1}}{1+g_{m1}/g_{m3}}$ M1=M2	Low sensitive to common-mode input signals. Limited linearity improvement, HD3 reduces by -12 dB. More silicon area is required.



$$i = i_n + i_p$$

$$i = \frac{g_{mn} v}{N_n + 1}\sqrt{1 - \left(\frac{v}{2(N_n+1)V_{DSsatn}}\right)^2} +$$

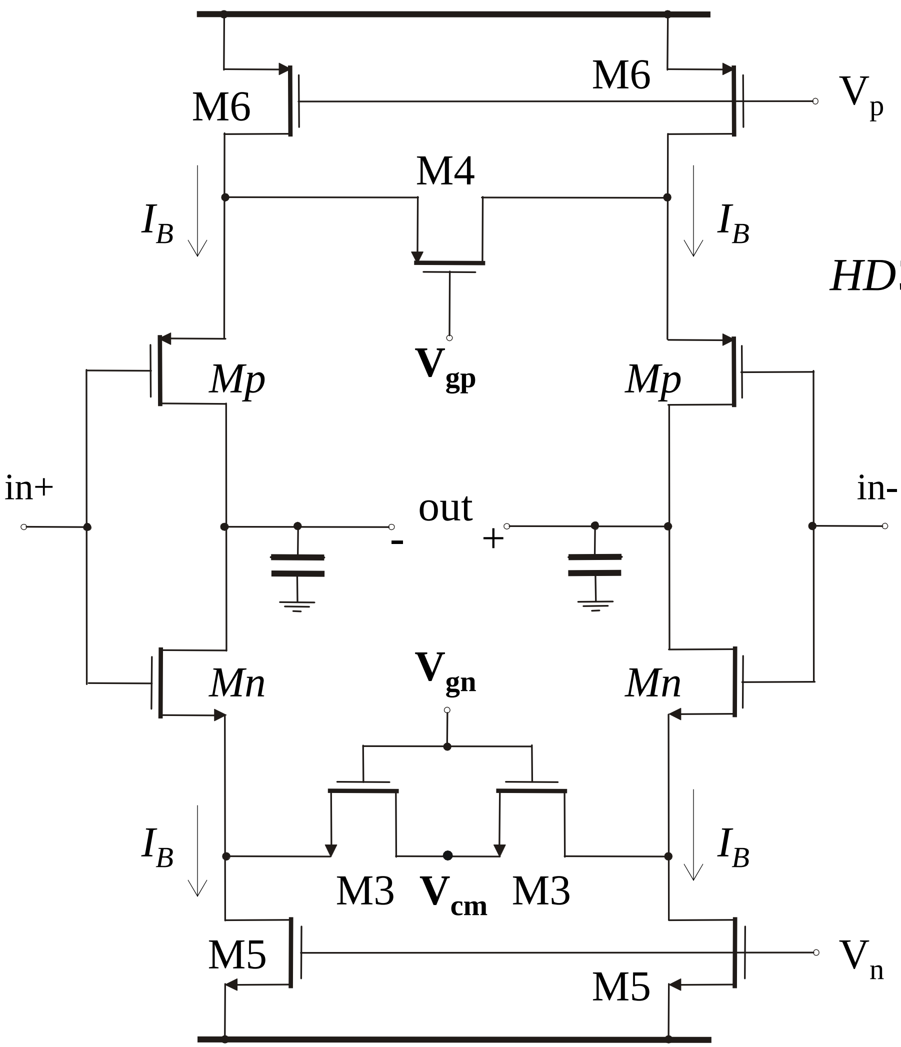
$$+ \frac{g_{mp} v}{N_p + 1}\sqrt{1 - \left(\frac{v}{2(N_p+1)V_{DSsatp}}\right)^2}$$

$$G_m = \frac{g_{mn}}{N_n + 1} + \frac{g_{mp}}{N_p + 1}$$

$$R_{DS} = \frac{1}{\beta(V_{GS} - V_T)}$$

$$N_n = g_{mn} R_n$$

$$N_p = g_{mp} R_p$$

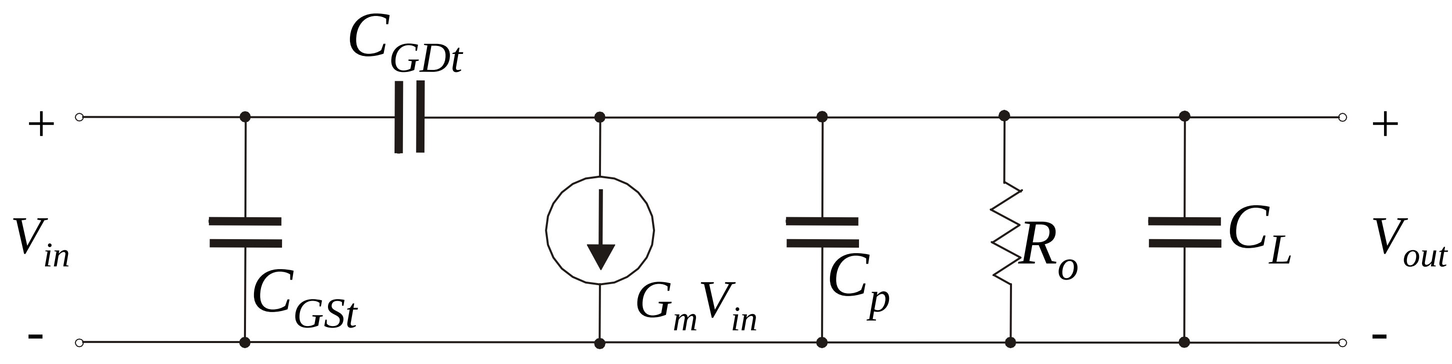


$$HD2 = 0$$

$$HD3 = \frac{1}{32}\left(\frac{1}{N_n+1}\right)^2\left(\frac{v}{V_{DSsatn}}\right)^2 \frac{g_{mn}(N_p+1)}{g_{mn}(N_p+1)+g_{mp}(N_n+1)} + \frac{1}{32}\left(\frac{1}{N_p+1}\right)^2\left(\frac{v}{V_{DSsatp}}\right)^2 \frac{g_{mp}(N_n+1)}{g_{mn}(N_p+1)+g_{mp}(N_n+1)}$$

$$HD3 \propto \frac{a_1}{(N_n+1)^2} + \frac{a_2}{(N_p+1)^2}$$

$$G_m \propto \frac{b_1}{N_n+1} + \frac{b_2}{N_p+1}$$



DC Gain: $$A_o = \frac{V_{out}}{V_{in}} = \frac{(G_m V_{in}) R_o}{V_{in}} = G_m R_o$$

Dominant Pole Frequency: $$f_p = \frac{1}{2\pi\, R_o (C_L + C_p)}$$

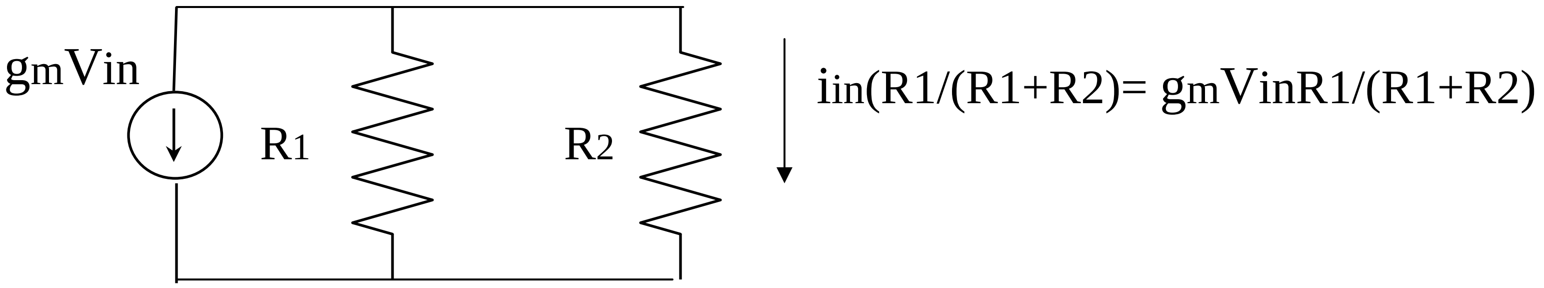
Gain-Bandwidth Product: $$GBW = \frac{G_m}{2\pi\,(C_L + C_p)}$$

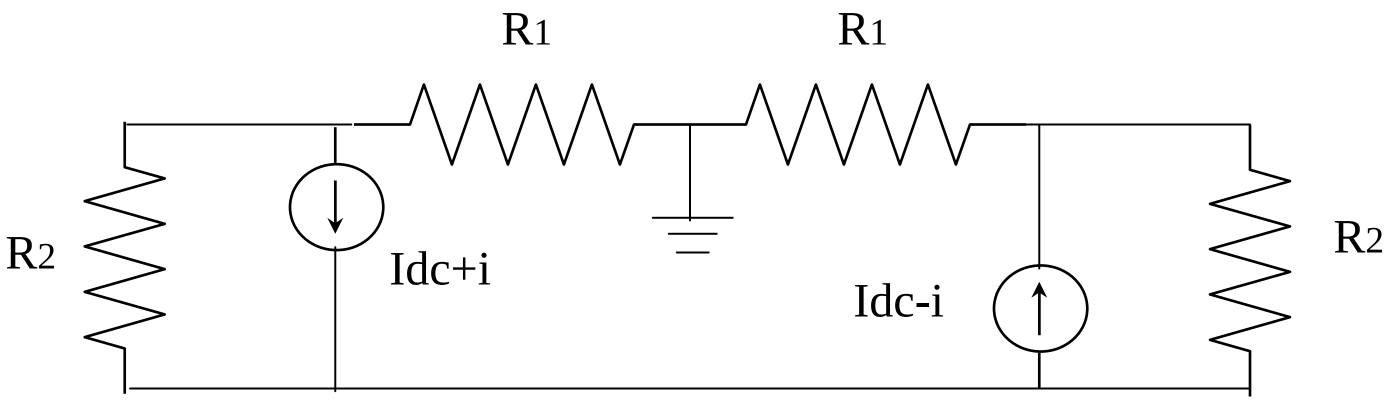
C_{GDt} effect:

$$f_z = \frac{G_m}{2\pi\, C_{GDt}}$$

Ref.-Antonio López/Silva-Martinez

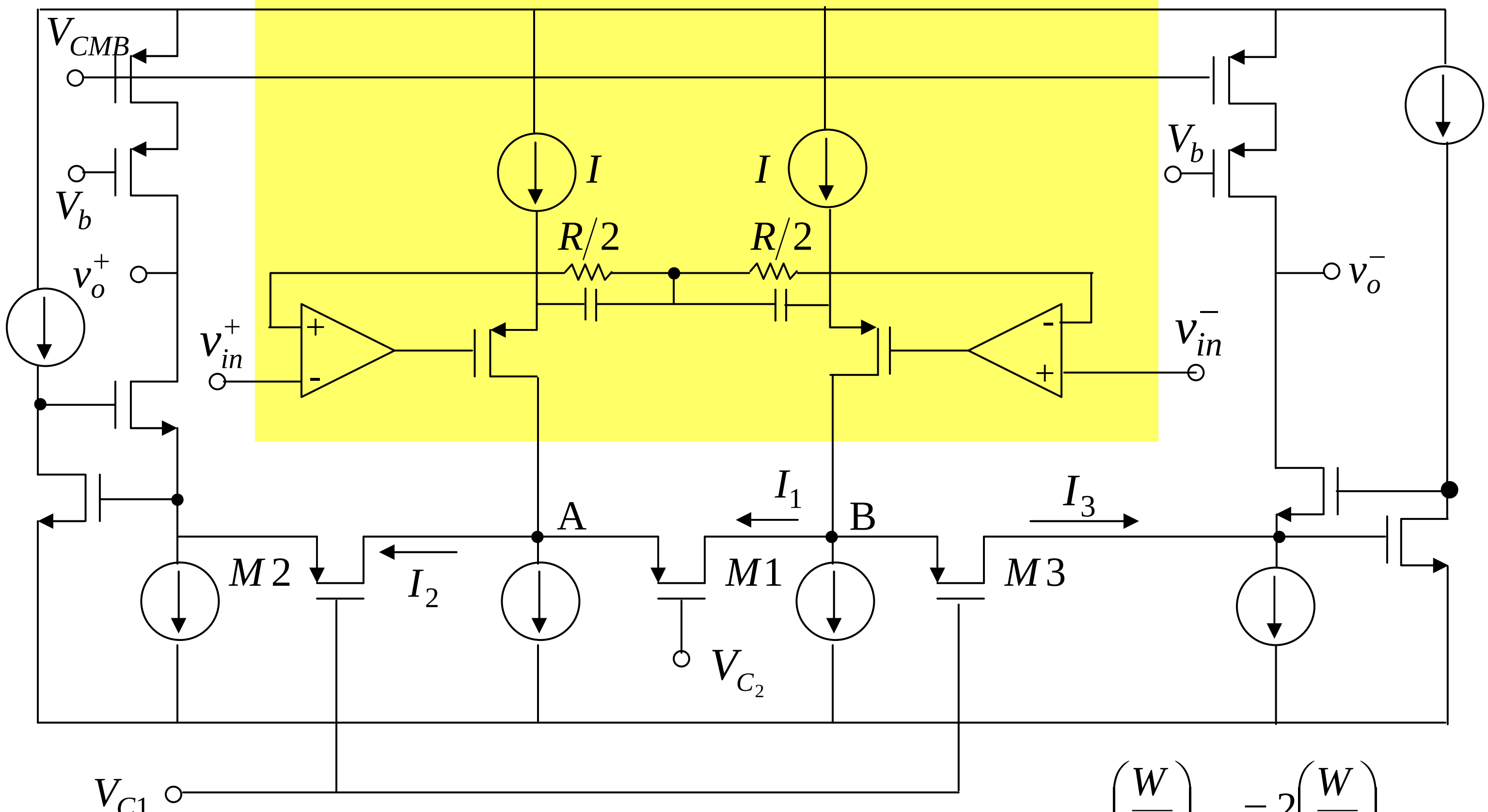
Current Divider Concept to tune Transconductance Amplifier





Symmetric Current Divider for Differential Structures

Source Degeneration Transconductance with Current Divider

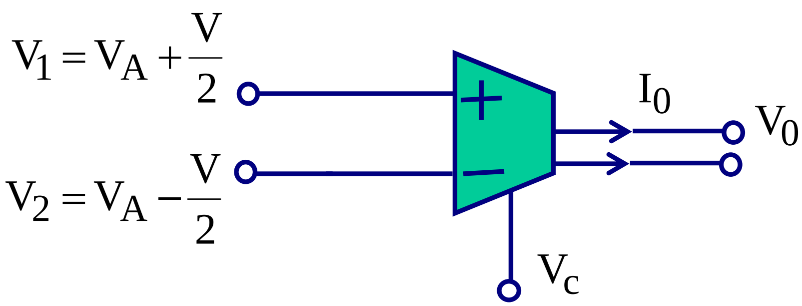


$$\left(\frac{W}{L}\right)_{2,3} = 2\left(\frac{W}{L}\right)_1$$

$$G_m = \frac{1}{R}\left(\frac{1}{2} + \frac{V_{C1} - V_{C2}}{\frac{V_{C1}+V_{C2}}{2} - V_T - V_A} \qquad ; \quad \text{i.e. } 3V_A = V_{CM}\right)$$

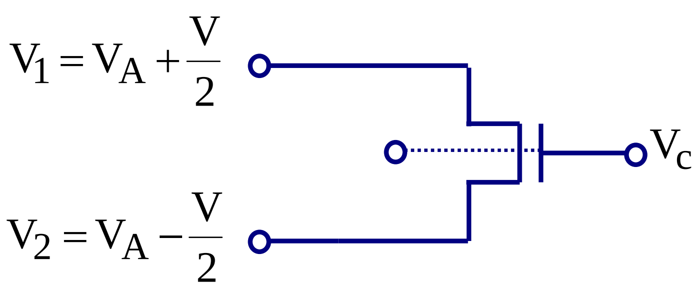
Ref.- Z.Y. Chang, D. Haspeslagh, J. Boxho and D. Macq, “A Highly Linear CMOS Gm-C Bandpass Filter for Video Applications,” IEEE CICC, 1996.

Balanced Nonlinearity Cancellation (no even power terms)

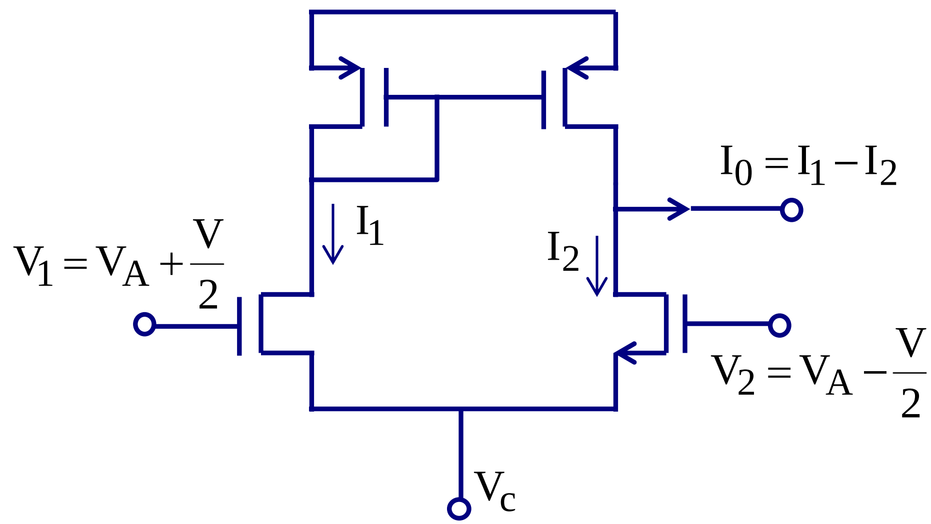


Implementations

$$I_0 = V(a_1 + 2a_2 V_A + \cdots) + \frac{1}{4}V^3\left(a_3 + 4a_4 V_A + 10a_5 V_A^2 + \cdots\right) + \frac{1}{16}V^5(a_5 + 6a_6 V_A + \cdots)$$



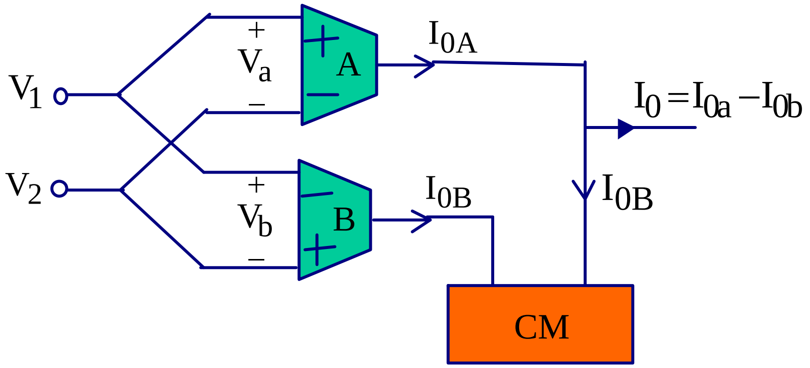
Single Device in Ohmic Region



Two Matched Devices in Saturation

- CMR is poor
- The technique requires perfect matching of components and voltages.

3. Subtraction of Polynomials. This is Another Balanced Nonlinearity Cancellation.

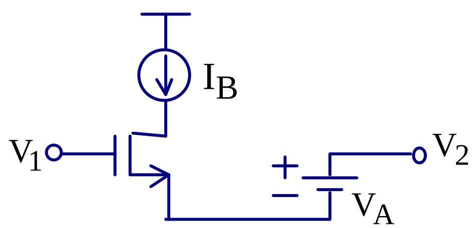


V_a and V_b are level shifting voltages that keep the transconductor active even for $V_1 = V_2 = 0$

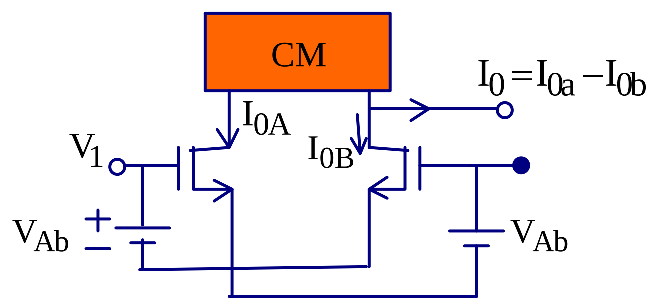
$$I_0 = 2\left(a_1 V + a_3 V^3 + a_5 V^5 + \cdots\right)$$

$$V = V_1 - V_2$$

Implementations:

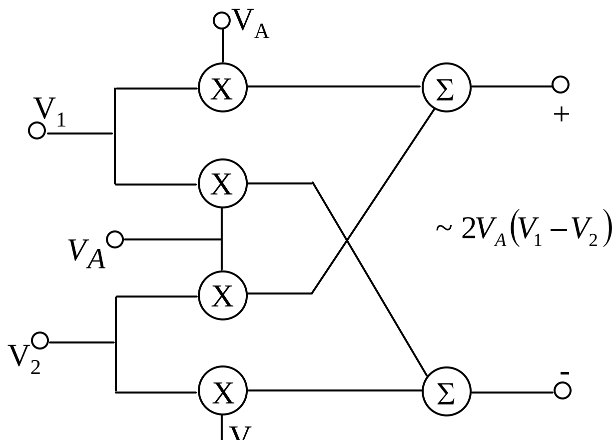


Single Device. V_A is a level shifting voltage keeping the transistor in saturation

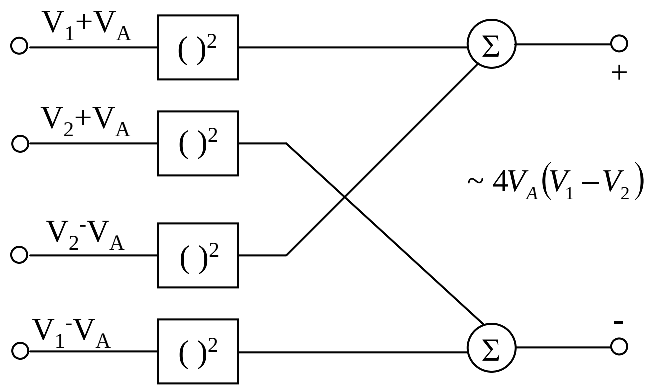


Balanced Configuration

3. General transconductance linearization by non-linear terms sum cancellation techniques

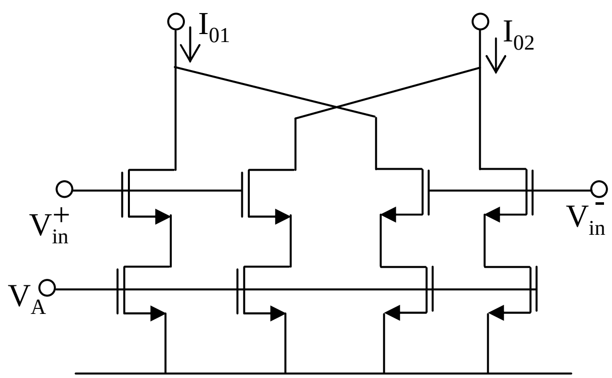


(a)

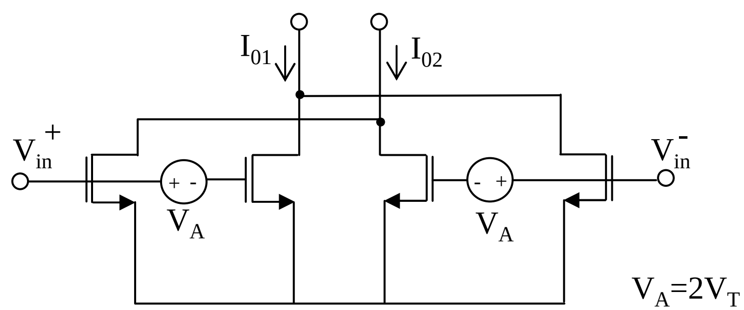


(b)

(a) using single multipliers by a constant (V_A), (b) using single-quadrant devices. $V_1 = - V_2$.



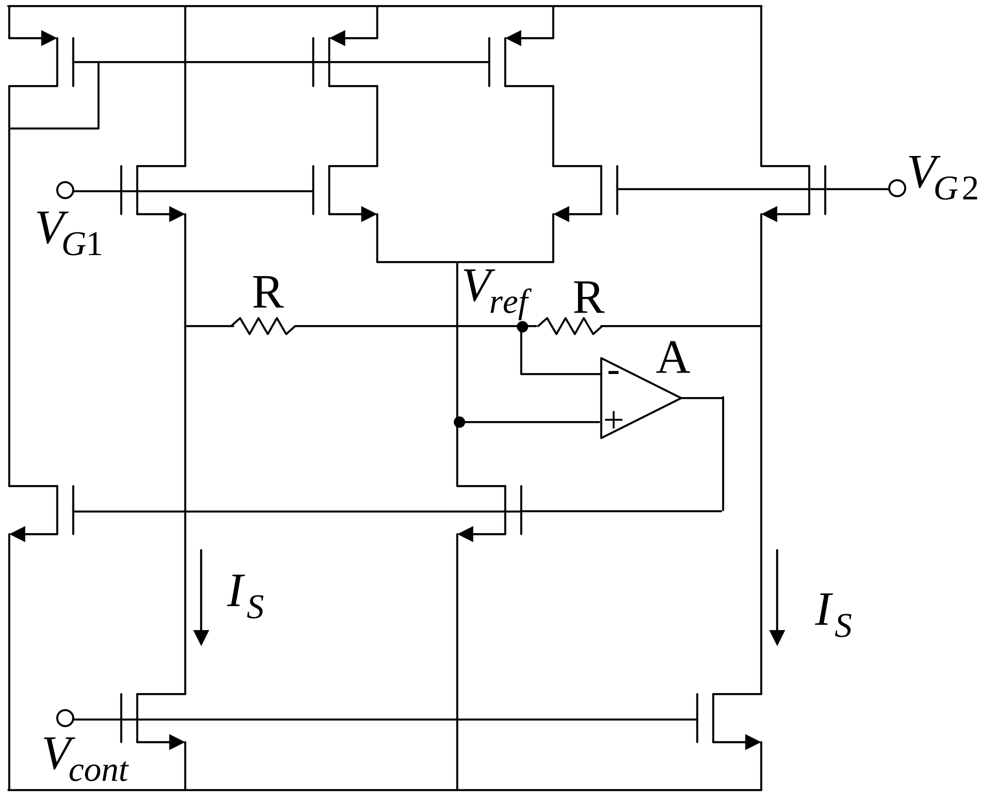
(A)



(B)

Transconductance (A) based on Fig (a); (B) based on Fig. (b). Note that in (A) the input signal might need a DC bias.

LINEAR OTA USING FEEDBACK CONTROL ON COMMON SOURCE NODE



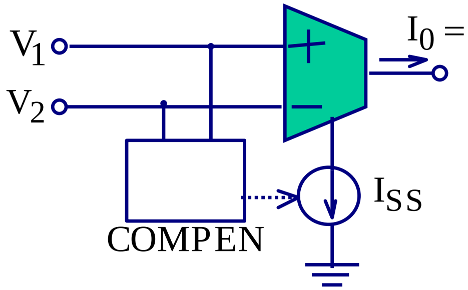
$$G_m = 2k\left(\frac{V_{G1}+V_{G2}}{2} - V_{ref} - V_T\right)$$

$$V_G = \frac{V_{G1}+V_{G2}}{2} - V_{ref} = k'\sqrt{I_S} = k''V_{cont}$$

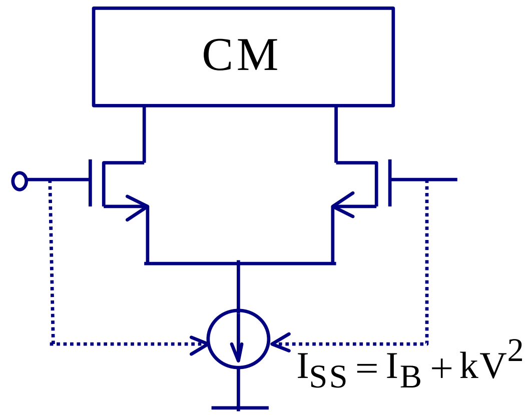
$$G_m = 2k\left(k''V_{cont} - V_T\right)$$

Ref.- J. Sevenhans and M. Van Paemel, “Novel CMOS Linear OTA Using Feedback Control on Common Source Node, “Elect. Letters, pp. 1873-1874, 26th Sept. 1996.

Nonlinearity Cancellation Using Function Compensation



$V = V_1 - V_2$



$I_0 = F(V, I_{SS})$

Make

$I_{SS} = I_B + F_1(V)$

Where $F_1(V)$ is chosen such that,

$$I_0 = F\left(V, I_B + F_1(V)\right) = VF_2(I_B) = G_m V$$

Example 1. Linearization via bias current modulation of a differential pair operating in strong inversion.

$$I_{out} = K\Delta V\sqrt{2I_{SS}/K - \Delta V^2} \quad ; K = \mu_o C_{ox}\frac{W}{2L} \tag{1}$$

Goal is to add $I_M = K\Delta V^2/2$ to I_{SS} , thus

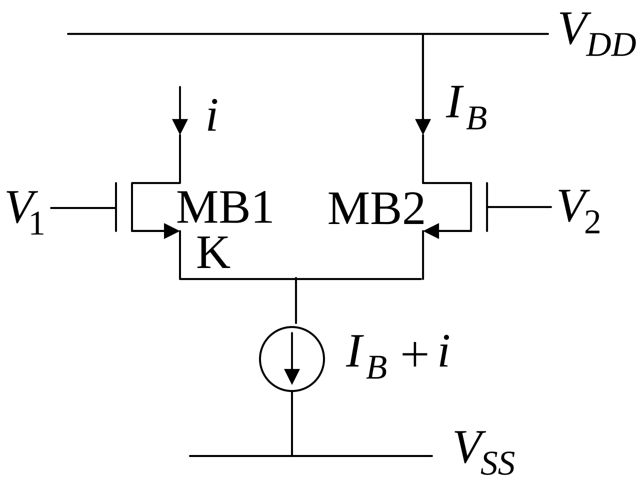
$$I_m = K\Delta V^2/2$$
$$\frac{2(I_{SS} + I_M)}{K} = \frac{2I_{SS}}{K} + \frac{K\Delta V^2}{K} - \Delta V^2 = \frac{2I_{SS}}{K} \tag{2}$$

(2) into (1)

$$I_{out} = K\Delta V\sqrt{2I_{SS}/K} = \sqrt{2KI_{SS}}\ \Delta V = G_m\Delta V \tag{3}$$

Where G_m is the linearized transconductance, and ΔV is the input differential voltage.

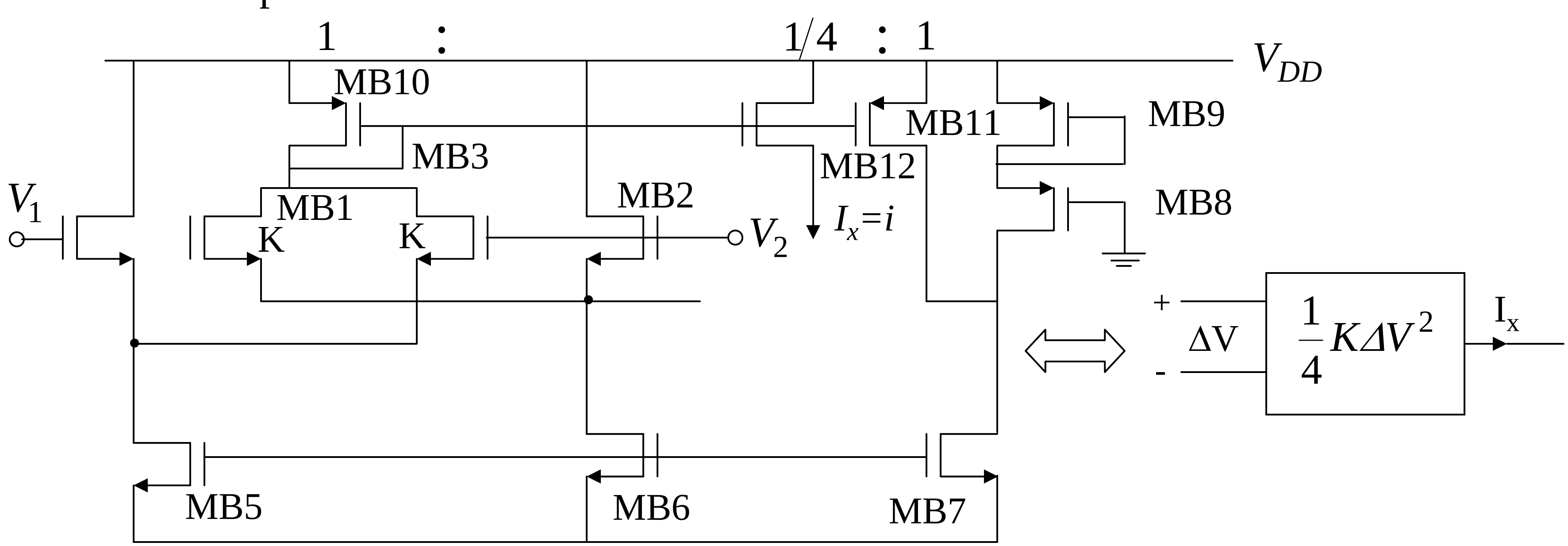
How to generate $I_M = K\Delta V^2/2$?



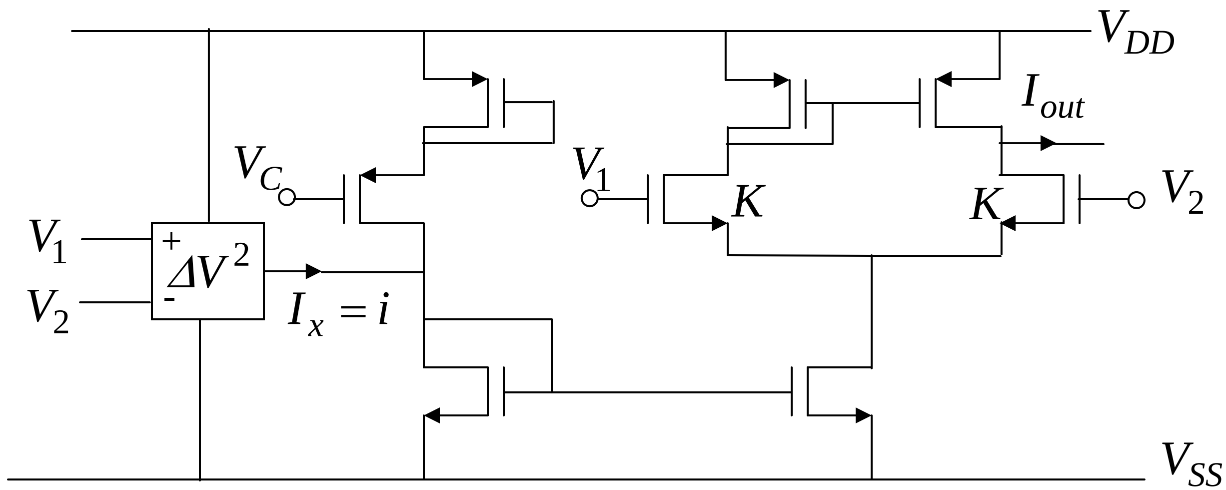
Conceptual Circuit

$$\Delta V = V_1 - V_2$$

$$i = \begin{cases} K\Delta V^2 & \text{for } \Delta V \geq 0 \\ 0 & \text{for } \Delta V < 0 \end{cases}$$



Complete ΔV^2 circuit structure



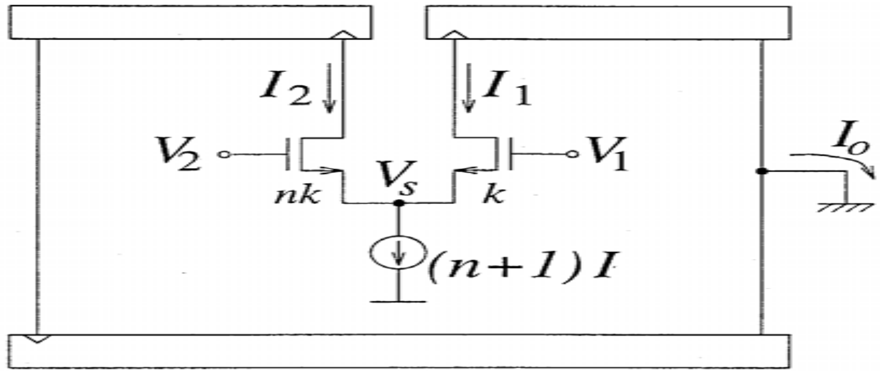
Single-Ended OTA with the Tail Current and KΔ² Circuit Generator

Ref. – T. Inoue, F. Ueno, Y. Aramaki, O. Matsumoto and M. Suefuji, “ A Design of CMOS OTA’s Using Simple Linearizing Techniques and Their Application to High-Frequency Continuous-Time Filters,” 2000 IEEE.

Nedungadi Transconductor

[A. Nedungadi and T. R. Viswanathan, "Design of Linear CMOS Transconductor Elements," *IEEE Transactions on Circuits and Systems*, vol. CAS-31, No. 10, pp. 891-894, October 1984]

Unbalanced Differential Pair



$$\left.\begin{aligned} I_1 &= k(V_1 - V_s - V_T)^2 \\ I_2 &= nk(V_2 - V_s - V_T)^2 \\ I_1 + I_2 &= (n+1)I \end{aligned}\right\} \Rightarrow V_s^2 - 2V_s\left(\frac{V_1 + nV_2}{n+1} - V_T\right) + \frac{(V_1 - V_T)^2 + n(V_2 - V_T)^2}{n+1} = \frac{I}{k}$$

$$V_s = \frac{V_1 + nV_2 - (n+1)V_T}{n+1} - \sqrt{\frac{I}{k} - \frac{n\Delta V^2}{(n+1)^2}} \Rightarrow$$

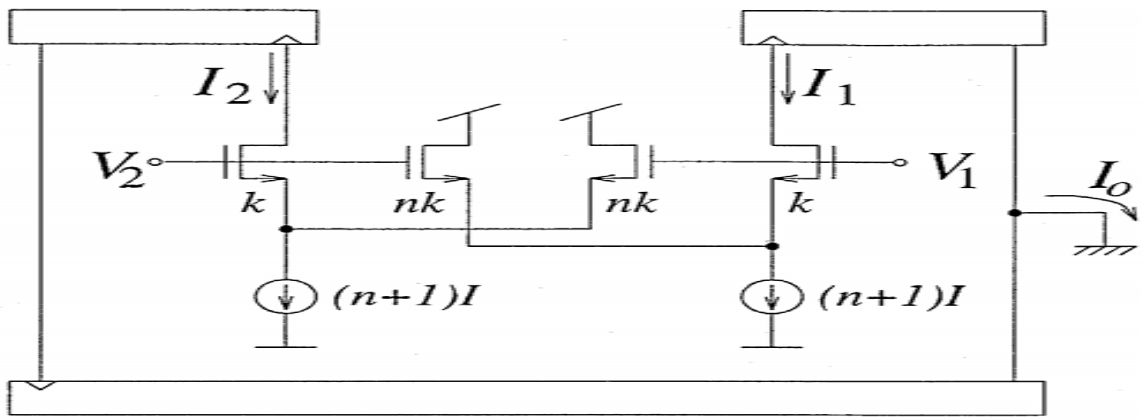
$$\Rightarrow \begin{cases} I_1 = k\left[\dfrac{n\Delta V}{n+1} + \sqrt{\dfrac{I}{k}}\sqrt{1 - \dfrac{kn\Delta V^2}{(n+1)^2 I}}\right]^2 \\ I_2 = nk\left[\dfrac{-\Delta V}{n+1} + \sqrt{\dfrac{I}{k}}\sqrt{1 - \dfrac{kn\Delta V^2}{(n+1)^2 I}}\right]^2 \end{cases}$$

Range:

$$\left.\begin{aligned} I_1 &= (n+1)I \\ I_2 &= 0 \end{aligned}\right\} \Rightarrow \Delta V = \sqrt{\frac{I}{k}(n+1)}$$

$$\left.\begin{aligned} I_1 &= 0 \\ I_2 &= (n+1)I \end{aligned}\right\} \Rightarrow \Delta V = -\sqrt{\frac{I}{k}\left(1 + \frac{1}{n}\right)}$$

The Cross-Coupled Quad Transconductor



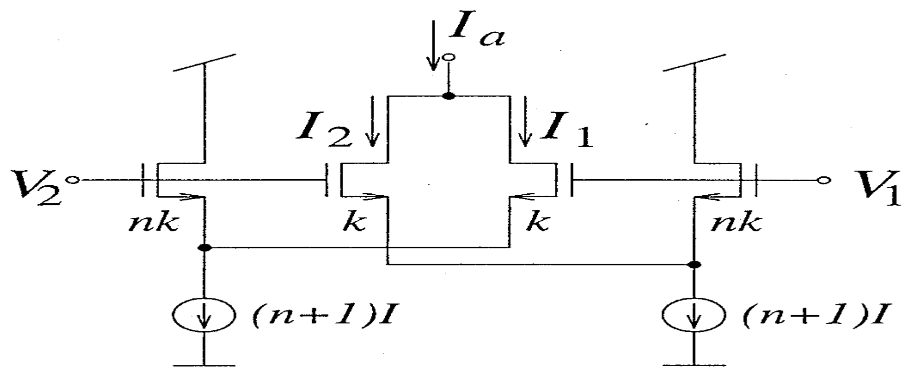
$$\left.\begin{aligned} I_1 &= k\left[\frac{n\Delta V}{n+1} + \sqrt{\frac{I}{k}}\sqrt{1 - \frac{kn\Delta V^2}{(n+1)^2 I}}\right]^2 \\ I_2 &= k\left[\frac{-n\Delta V}{n+1} + \sqrt{\frac{I}{k}}\sqrt{1 - \frac{kn\Delta V^2}{(n+1)^2 I}}\right]^2 \end{aligned}\right\} \Rightarrow$$

$$\Rightarrow I_o = I_1 - I_2 \Rightarrow \begin{cases} I_o = g_m \Delta V \sqrt{1 - \beta \Delta V^2} \\ g_m = \dfrac{4n}{n+1}\sqrt{Ik} \quad , \quad \beta = \dfrac{k}{I}\dfrac{n}{(n+1)^2} \end{cases}$$

Range: $|\Delta V| \le \sqrt{\frac{I}{k}\left(1 + \frac{1}{n}\right)}$

For $\Delta V = \sqrt{\frac{I}{k}\left(1 + \frac{1}{n}\right)} \Rightarrow I_o = \frac{4n}{n+1} I$,

Squarer

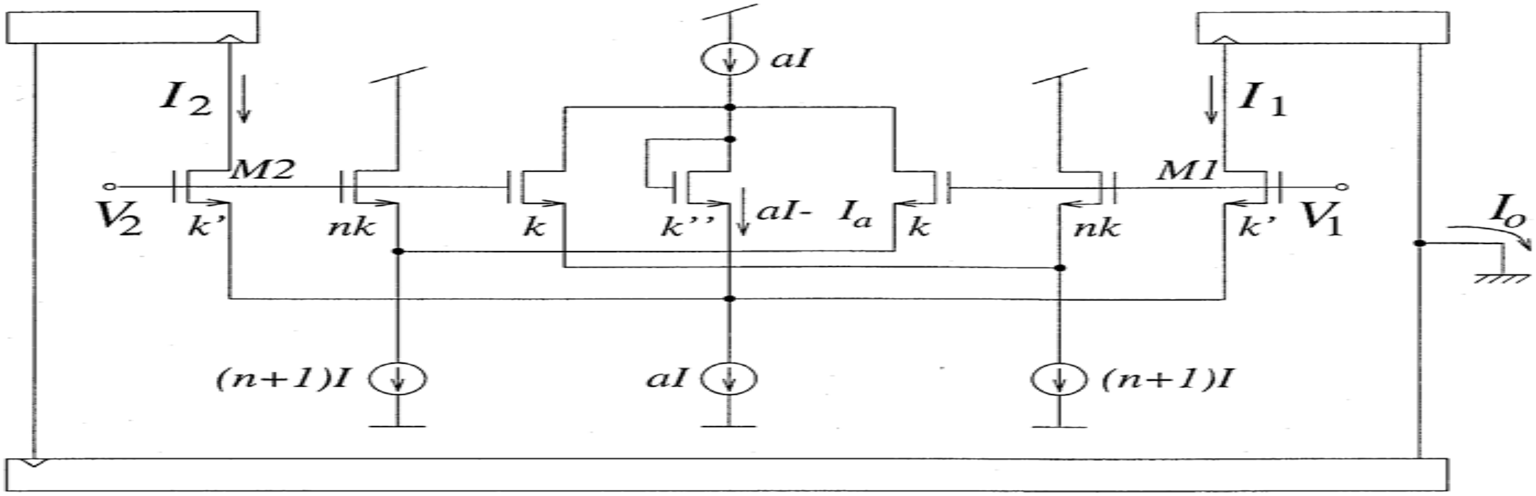


$$I_1 = k\left[\frac{n\Delta V}{n+1} + \sqrt{\frac{I}{k}}\sqrt{1 - \frac{kn\Delta V^2}{(n+1)^2 I}}\right]^2$$

$$I_2 = k\left[\frac{-n\Delta V}{n+1} + \sqrt{\frac{I}{k}}\sqrt{1 - \frac{kn\Delta V^2}{(n+1)^2 I}}\right]^2 \quad \Bigg\} \Longrightarrow$$

$$\Longrightarrow \quad I_a = I_1 + I_2 = 2k\frac{n(n-1)}{(n+1)^2}\Delta V^2 + 2I \quad , \quad \Delta V \le \sqrt{\frac{I}{k}\left(1 + \frac{1}{n}\right)}$$

Nedungadi Transconductor



Transistors M1 and M2 form a simple differential pair biased by

$$I_{ss} = aI - (aI - I_a) = I_a = 2k\frac{n(n-1)}{(n+1)^2}\Delta V^2 + 2I$$

Consequently,

$$I_o = \Delta V\sqrt{2k'I_{ss} - k'^2\Delta V^2} = \Delta V\sqrt{k'^2\Delta V^2\left(4\gamma\frac{n(n-1)}{(n+1)^2}\right) + 4k'I} \quad , \quad \gamma = \frac{k}{k'}$$

If $4\gamma n(n-1) - (n+1)^2 = 0 \quad \Rightarrow \quad I_o = 2\Delta V\sqrt{k'I}$ $\quad \begin{cases} g_m = 2\sqrt{k'I} \\ |\Delta V| \le min\left(\sqrt{\frac{I}{k}\left(1 + \frac{1}{n}\right)}, 2\sqrt{\frac{I}{k'}}\right) \end{cases}$

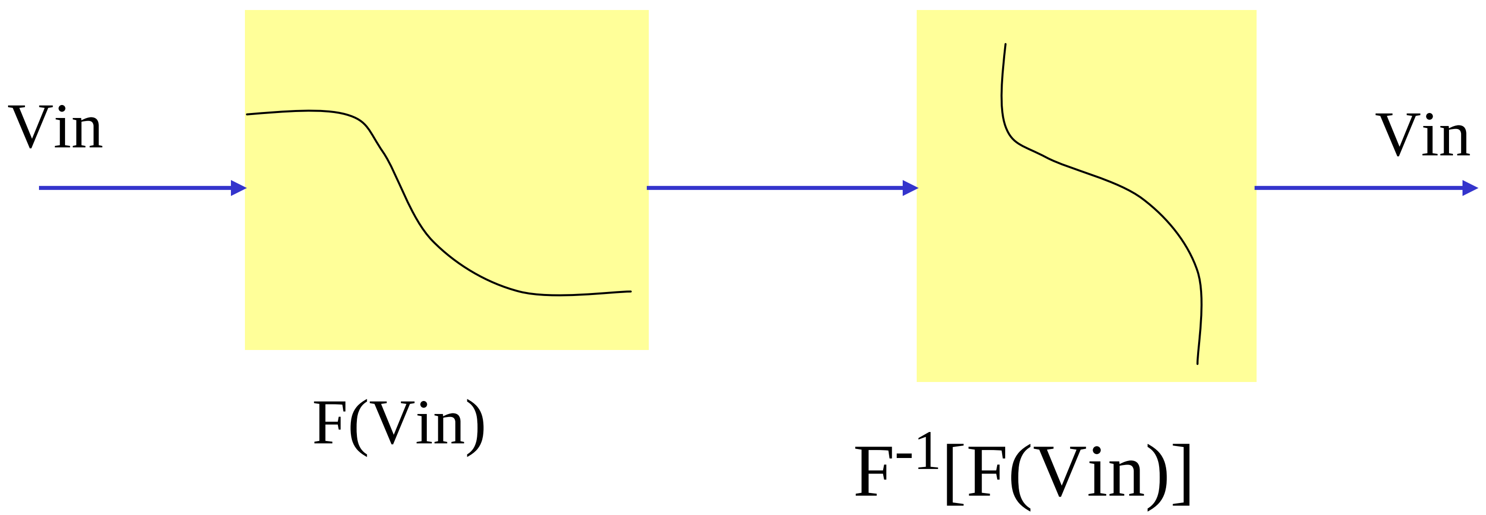
Examples:

$$\gamma = 1 \quad \Rightarrow \quad n = 1 + \frac{2\sqrt{3}}{3} = 2.155$$

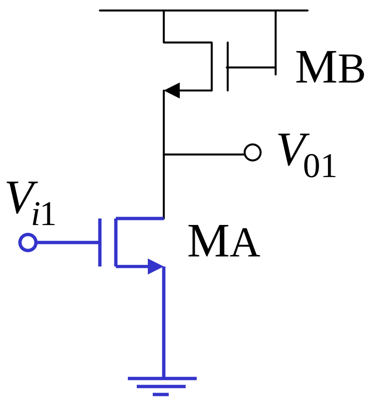
$$n = 3 \quad \Rightarrow \quad \gamma = \frac{(n+1)^2}{4n(n-1)} = \frac{2}{3}$$

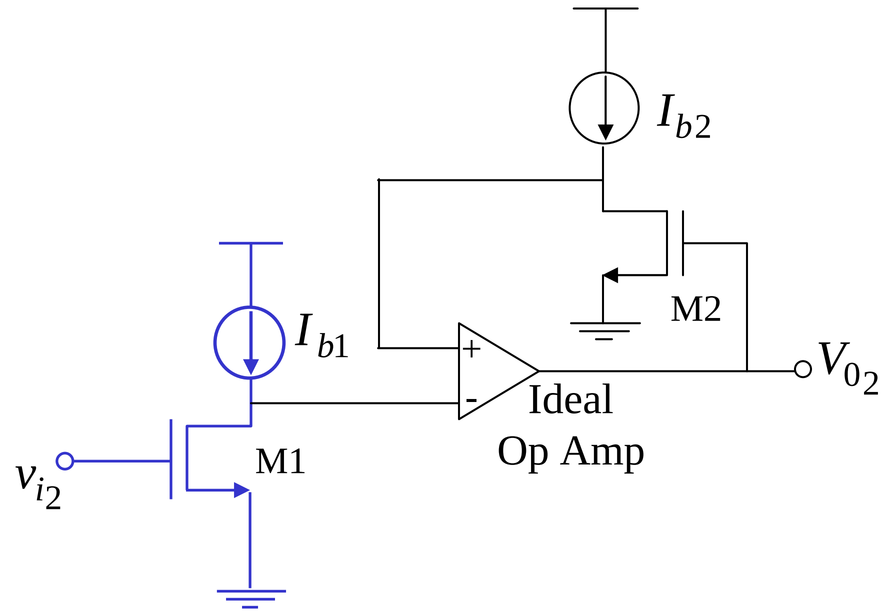
$$aI - I_a \ge 0 \quad \Leftrightarrow \quad \begin{cases} \Delta V^2_{max} = \frac{I}{k}\left(\frac{n+1}{n}\right) \quad \rightarrow \quad a \ge \frac{4n}{n+1} \\ \Delta V^2_{max} = \frac{I_a}{k'} \quad \rightarrow \quad a \ge 2\left[\frac{(1+4\gamma)n^2 + 2n(1-2\gamma) + 1}{(n+1)^2}\right] \end{cases}$$

4. By cascading a nonlinear function and its inverse non-linear function



Nonlinear Function Combined with its Inverse Nonlinear Function

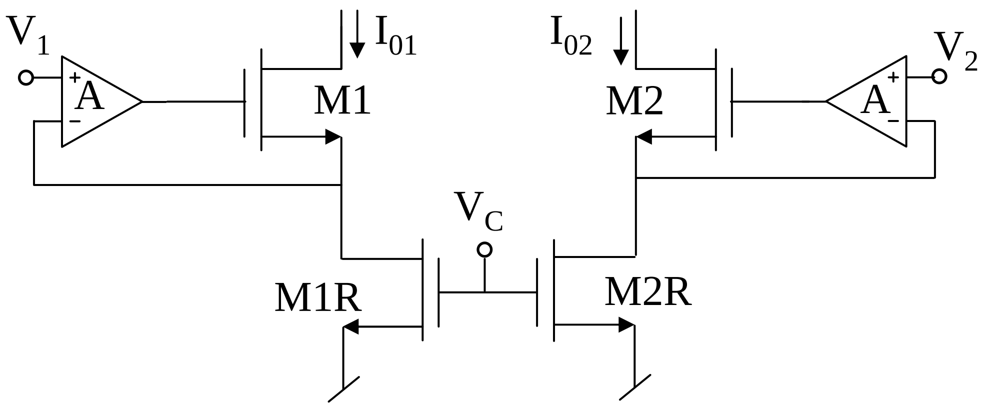




If bulk – and lambda effects are neglected, it can be shown that $HD_2 = 0$.

$$H_1 = \frac{V_{o1}}{V_{i1}} \cong -\frac{g_{mA}}{g_{mB}} \qquad H_2 = \frac{V_{02}}{V_{i2}} = \frac{-g_{m1}g_{m2}}{g_{01}g_{02}}$$

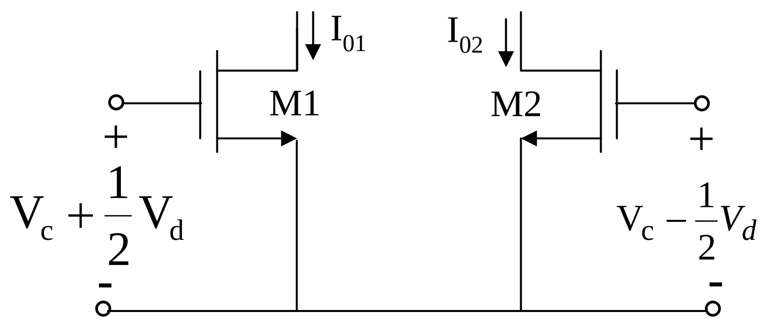
- Improved Transconductor via Series-Shunt Feedback (DUPU89)



- g_{mI} increased by a + A

TRIODE TRANSCONDUCTORS

CONCEPT



BASIC PRINCIPLE:

Linearity is based on the approximately
Linear model of MOSFETs in Ohmic Region

$$I_{oi} = \frac{\beta_i}{Z}\left[Z\left(V_{GS_i} - V_r\right)V_{DS_i} - V_{DS_i}\right]; i = 1, Z$$

$$g_m = \beta V_{OS}$$

CONDITIONS:

M1, m2 are perfectly matched

V_{DS} of M1, M2 must be held constant and below $V_{DS,SAT}$

$$V_{DS1} = V_{DS2}$$

GLOBAL CHARACTERISTICS:

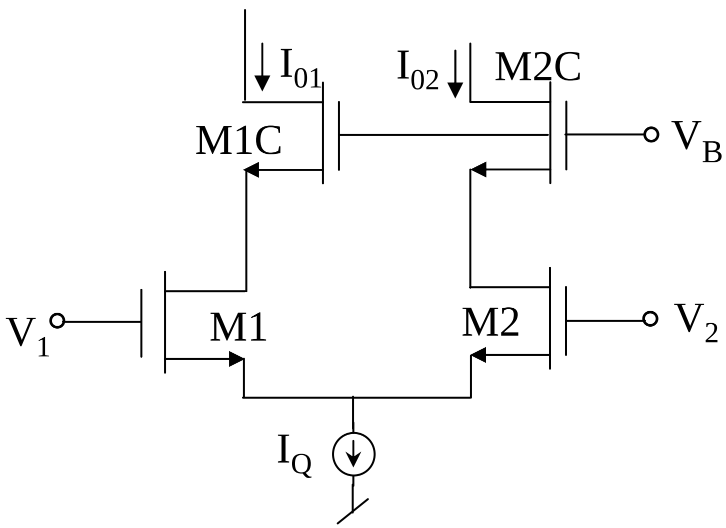
Nonlinearity is mainly due to mobility reduction

$$HD_Z = 0 \qquad HD_S \approx \frac{Q^2V_d^2}{16\left(1 + Q\left(V_C - V_T\right)\right)^2}$$

Well suited for high frequency applications

TRIODE TRANSCONDUCTORS - IMPLEMENTATIONS

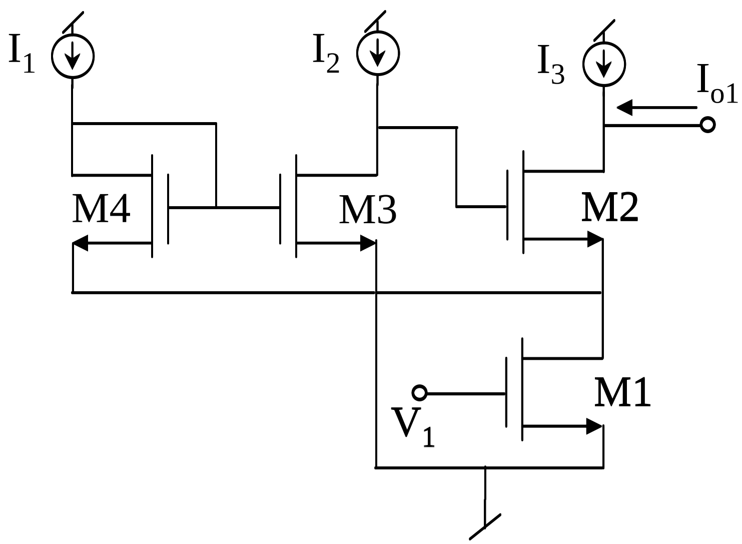
➢ Cascoding (STEF90)



- M1C M2C in saturation
- $\left(\frac{W}{L}\right)_{ic}$ >> →

M1C, M2C act as per-feet source-followers

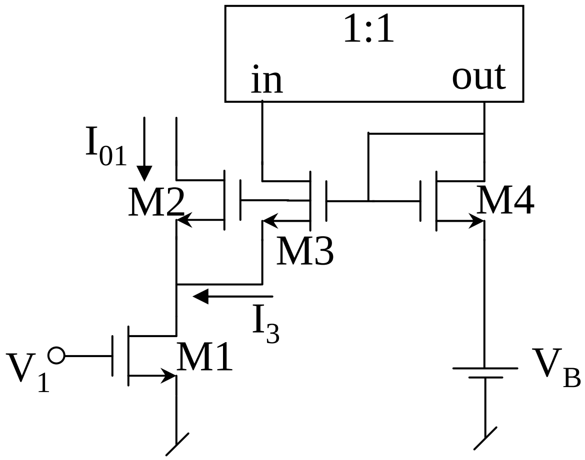
➢ Feedback (GATT90)



- Tuning via I_2
- M2, M3, M4 constitute 2 feedback loop, which keeps V_{og}, constant when V_1 is applied.

Dual

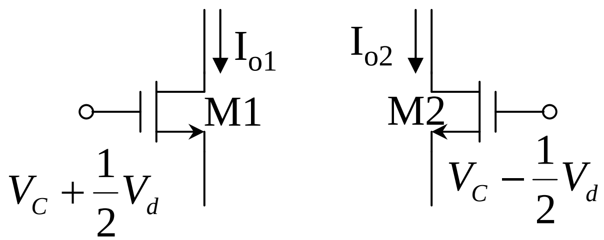
➢ Compensation (NAUT91)



M2 = M3 = M4

• V_{GS2} is modulated, so that, V_{DS1} remain constant and equal to V_B.

- Concept



M1, M2 in strong inversion and saturation

- Basic Principle:

Most of the proposed techniques are based on the simple algebraic identity:

$$(a+b)^2 - (a-b)^2 = 4ab$$

$$g_m = \beta\left(V_{GS_1} + V_{GS_2} - Z\,V_T\right)$$

- Conditions:

  M1, M2 are perfectly matched

  $V_{GS1} + V_{GS2}$ must be held constant

  V_{DS} of M1, M2 above $V_{DA,SAT}$

- Global Characteristics:

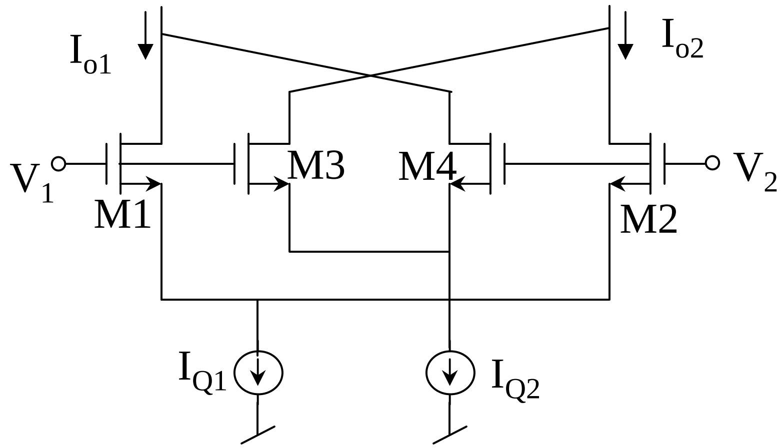
  Also, nonlinearity is mainly due to mobility reduction

$$HD_Z = \phi \quad HD_3 \approx \frac{\theta V_d^2}{16V_o\left(Z + \theta V_o\right)\left(1 + \theta V_o\right)^2}$$

  Well suited for high frequency applications

  Careful layout required

- Cross-Coupling (KHOR84)

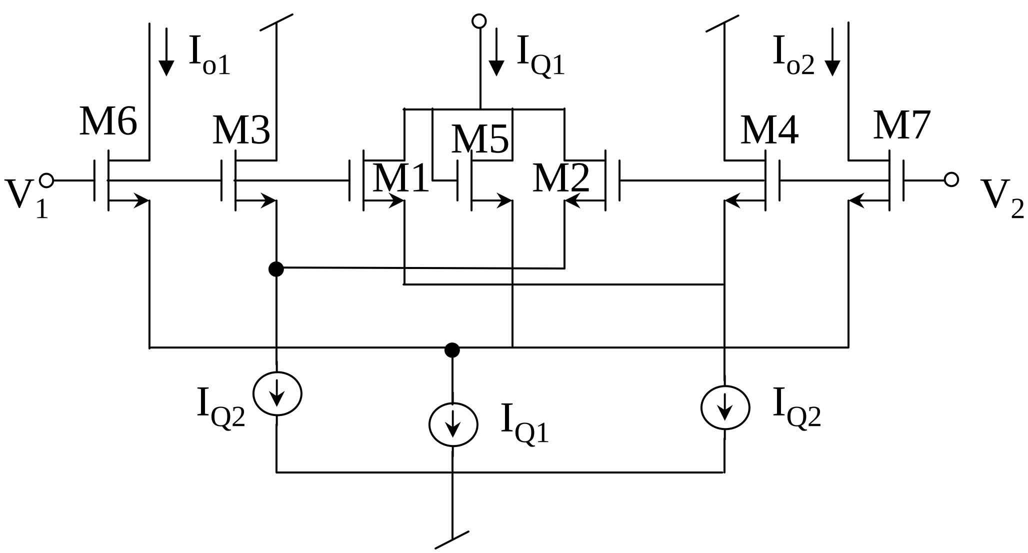


- M1 = M2, M3 = M4
- $\left[\frac{\beta_1}{\beta_3}\right]^{3/2} = \left[\frac{I_{Q1}}{I_{Q2}}\right]^{1/2}$

$\rightarrow$

$$g_m = \sqrt{\beta \cdot I_{Q1}} - \sqrt{\beta_3 I_{Q2}}$$

- Adaptive Biasing (NEDU84)



- $M1 = M2 = M6 = M\beta_1 7$

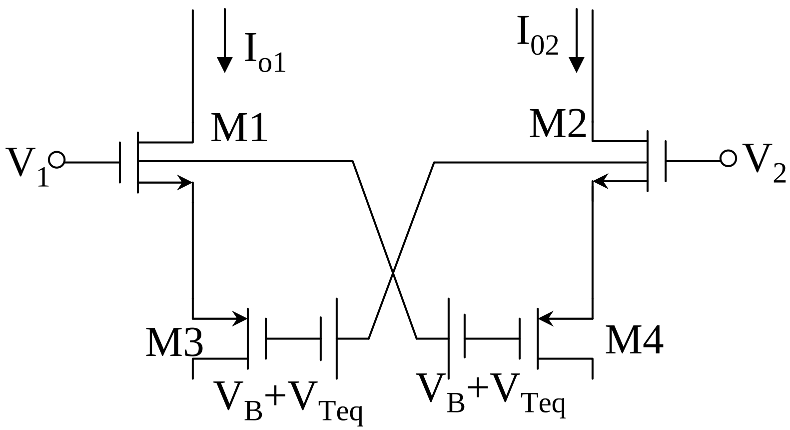
$M3 = M4$

$\beta_3 = n\,\beta_1$

- $\frac{I_{Q1}}{I_{Q2}} = \frac{2}{n+1}$
- $2 > \frac{4n}{n+1}$

$$g_m = \sqrt{\frac{2\beta_1 I_{Q1}}{2}}$$

➢ Class AB (SEEV87)



M1 = M2 = MN

M3 = M4 = MP

- $V_{Teq} = V_{TN} + V_{TF}$
- $$\beta_{eq} = \frac{\beta_N \beta_F}{\left(V_{\beta_N}^{-} + V_{\beta_F}^{-}\right)^2}$$

$$g_m = \beta_{eq} V_B$$