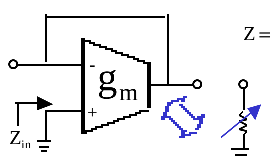
Operational Transconductance – C (OTA-C) and Current-Mode Filter Structures

- **OTA-C Filter Topologies**
- **What is current-mode and how is related to Transconductance mode ?**
- **Current-Mode Filters**
- **How to use a conventional OTA as a filter by adding capacitances at the internal nodes.**

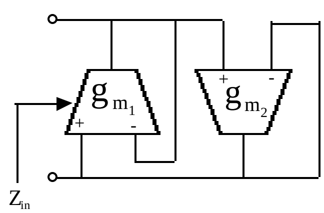
Edgar Sánchez-Sinencio

Controlled Impedance Elements



$Z = \frac{1}{g_m}$

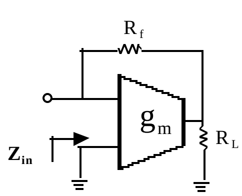
(a) Single-ended voltage variable resistor (VVR).



$g_{m_1} = g_{m_2} = g_m$

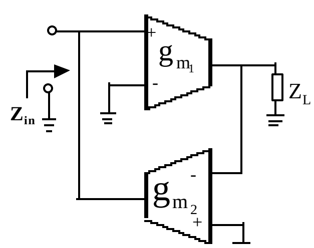
$Z_{in} = \frac{1}{g_m}$

(b) Floating VVR



$Z_{in} = \frac{1 + \frac{R_f}{R_L}}{g_m}$

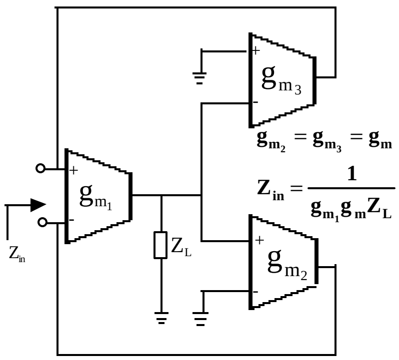
(c) Scaled VVR



$Z_{in} = \frac{1}{g_{m_1} g_{m_2} Z_L}$

(d) Voltage variable impedance inverter

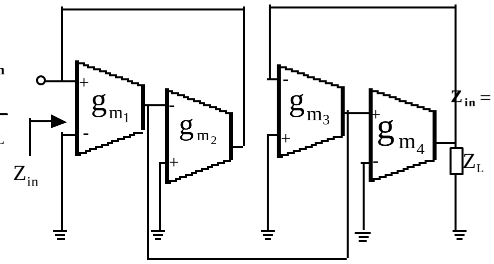
Controlled Impedance Elements



$$g_{m_2} = g_{m_3} = g_m$$

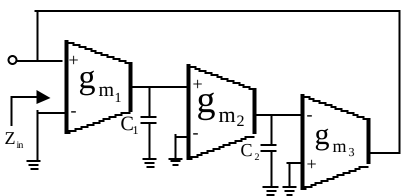
$$Z_{in} = \frac{1}{g_{m_1} g_m Z_L}$$

(e) **Voltage variable floating impedance**



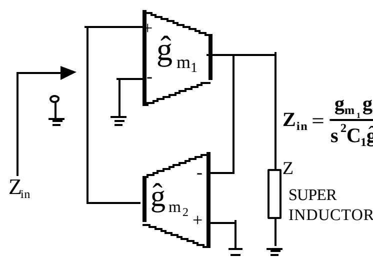
$$Z_{in} = \frac{Z_L g_{m_3} g_{m_4}}{g_{m_1} g_{m_2}}$$

(f) Impedance multiplier



(g) Super inductor

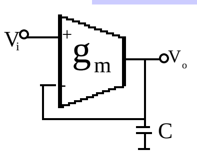
$$Z_{in} = \frac{s^2 C_1 C_2}{g_{m_1} g_{m_2} g_{m_3}}$$



$$Z_{in} = \frac{g_{m_1} g_{m_2} g_{m_3}}{s^2 C_1 \hat{g}_{m_1} \hat{g}_{m_2}}$$

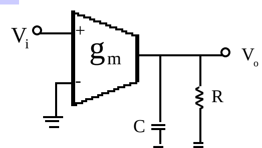
(h) FDNR

First-Order OTA-C Circuits



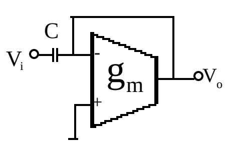
$$\frac{V_o}{V_i} = \frac{g_m}{sC + g_m}$$

(a) Lowpass, fixed dc gain pole adjustable



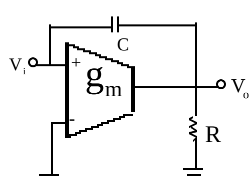
$$\frac{V_o}{V_i} = \frac{g_m}{sC + 1/R}$$

(b) Lowpass fixed pole, adjustable dc gain



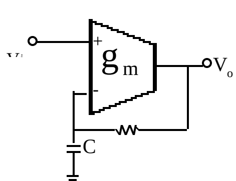
$$\frac{V_o}{V_i} = \frac{sC}{sC + g_m}$$

(c) Highpass, fixed high-frequency gain, adjustable pole



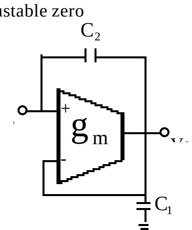
$$\frac{V_o}{V_i} = \frac{R(sC + g_m)}{sRC + 1}$$

(d) Shelving equalizer, fixed hig-frequency gain, fixed pole, adjustable zero



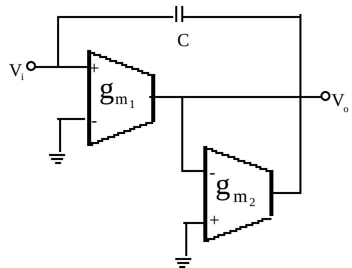
$$\frac{V_o}{V_i} = \frac{g_m (1 + RC)}{sC + g_m}$$

(e) shelving equalizer, fixed high-frequency gain, fixed zero, adjustable pole



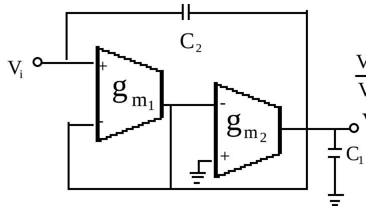
$$\frac{V_o}{V_i} = \frac{sC_2 + g_m}{s(C_1 + C_2) + g_m}$$

(f) Lowpass filter adjustable pole and zero, fixed ration



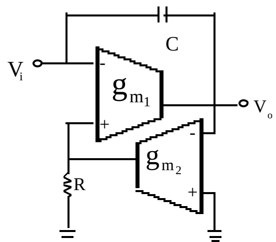
$$\frac{V_o}{V_i} = \frac{sC + g_{m_1}}{sC + g_{m_2}}$$

(g) Shelving equalizer, independently adjustable pole and zero



$$\frac{V_o}{V_i} = \frac{g_{m_1} + sC_2}{s(C_1 + C_2) + g_{m_1} + g_{m_2}}$$

(h) Lowpass or highpass filter, adjustable zero and pole, fixed ratio or independent adjustment

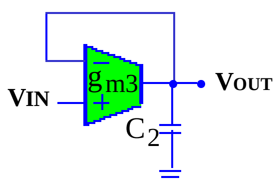


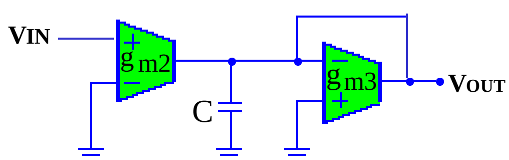
$$\frac{V_o}{V_i} = \frac{sC - g_{m_1}}{sC + g_{m_1} g_{m_2} R}$$

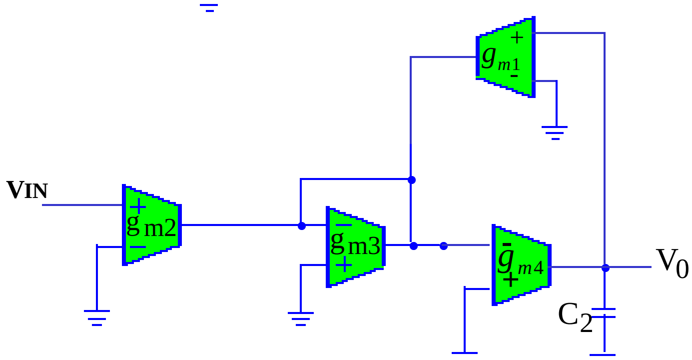
$$g_{m_2} R = 1$$

(i) Phase shifter, adjustable with g_m

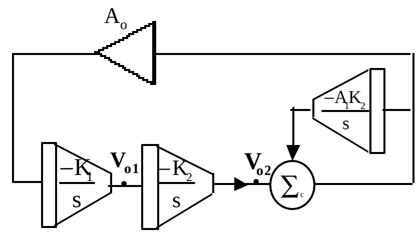
LOSSY OTA-C INTEGRATORS







TWO-INTEGRATOR LOOP OTA-C



(a) Cascade of one lossless and one lossy integrators

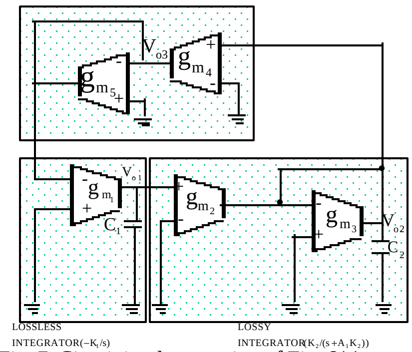
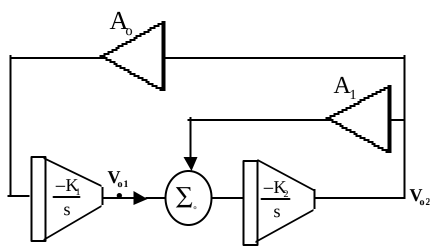


Fig. 7. Circuit implementation of Fig. 6(a)



(b) Cascade of two lossless integrators

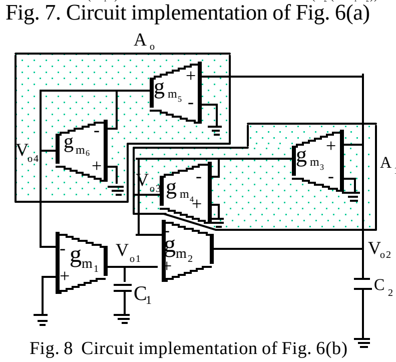
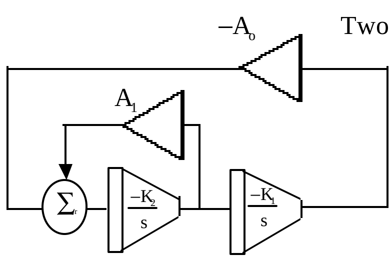
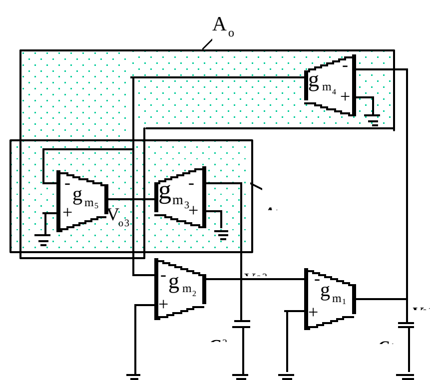


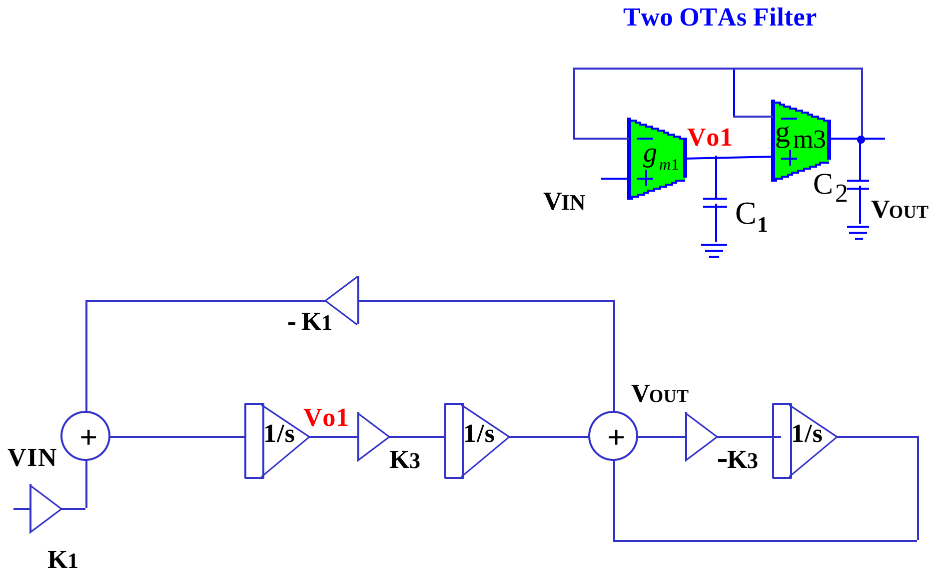
Fig. 8 Circuit implementation of Fig. 6(b)

Two integrator loops with a single feedback summer

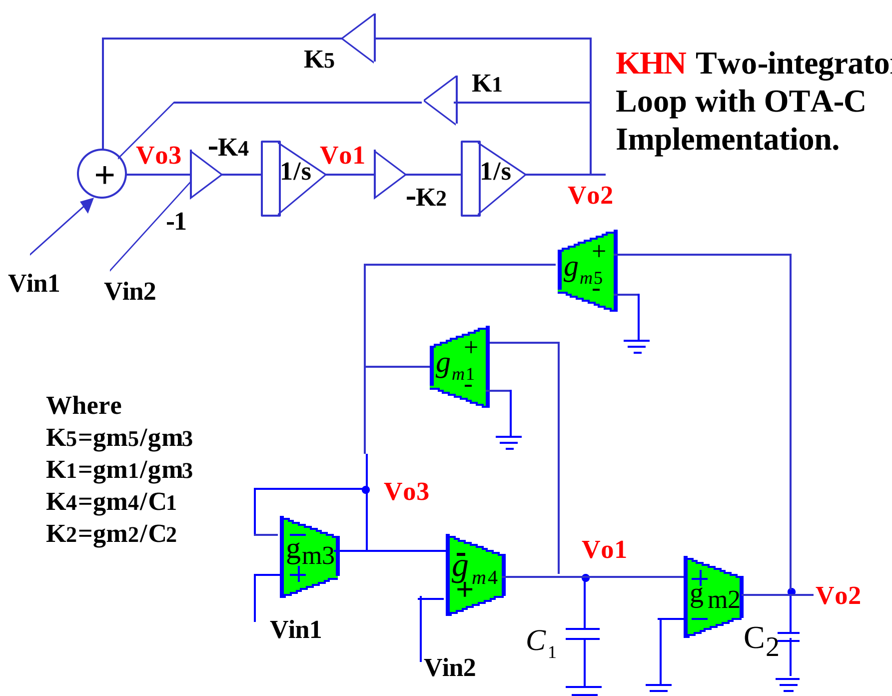




OTA-C Two Integrator Loop Filters

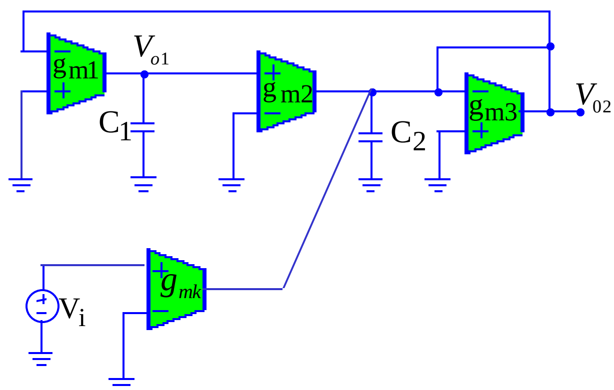


KHN Two-integrator Loop with OTA-C Implementation.

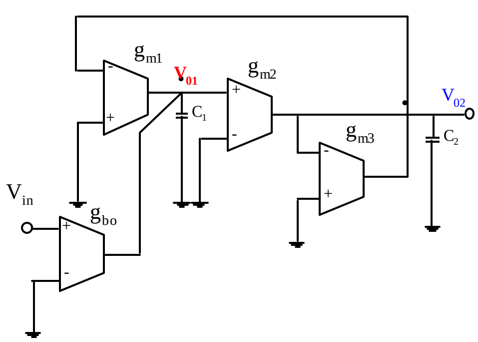


Where
K5=gm5/gm3
K1=gm1/gm3
K4=gm4/C1
K2=gm2/C2

OTA-C Biquadratic Filter based on a Two-integrator Loop



INTERNAL VOLTAGE SCALING

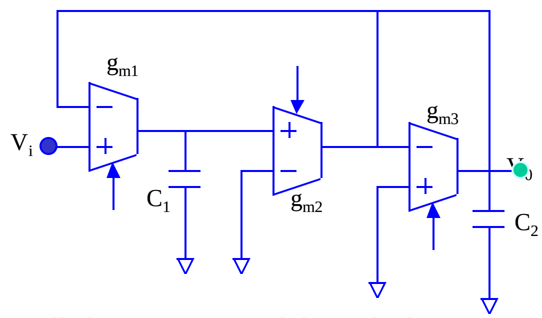


Assume the voltage V_{01} needs to be scaled by a factor "a" without changing the other node voltages:

1. The impedance at the node under consideration must be increased by "a". In this case C1 becomes C1/a.
2. Multiply all the transconductances leaving that node by the factor "a". In this case g_{m2} becomes ag_{m2},

Real example - Low Pass/Low Frequency Filter

FILTER TOPOLOGY **with g_m = 11.8nA/V**



DESIGN EQNS

$$\left(\frac{V_0}{V_i}\right) = \frac{w_0^2}{s^2 + \frac{w_0}{Q}s + w_0^2}$$

$$w_0 = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} = \frac{g_m}{C}$$

$$Q = \frac{1}{g_{m3}}\sqrt{g_{m1}g_{m2}\frac{C_2}{C_1}}$$

$$Q = \frac{g_m}{g_{m3}}$$

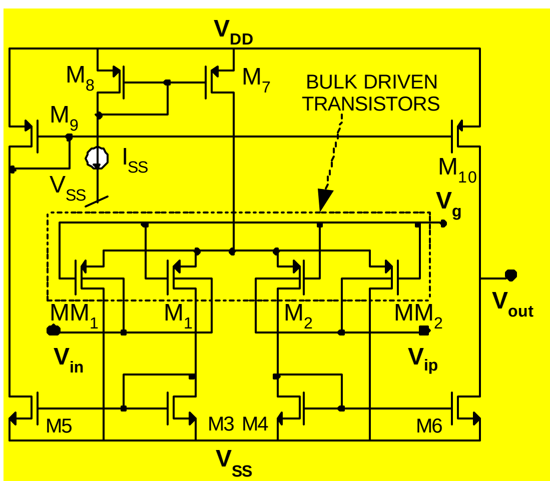
$$g_{m1} = g_{m2} = g_m$$

$$C_1 = C_2 = C$$

- **Bulk driven OTAs used due to higher input range**
- **Vi can be applied to the input of OTA 1 because the input parasitic capacitance does not affect the performance.**

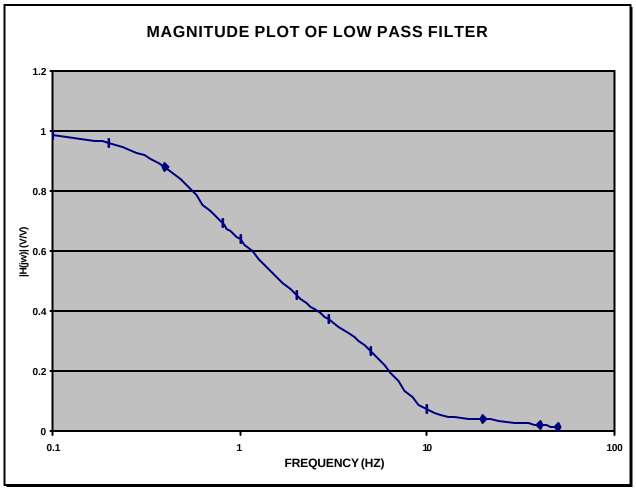
Reference.- A. Veeravalli, E. Sanchez-Sinencio and J. Silva-Martinez," Transconductance Amplifiers with Very Small Transconductance: A Comparative Design Approach" to appear in J. of Solid-State Circuits

Bulk-Driven OTA used in the filter



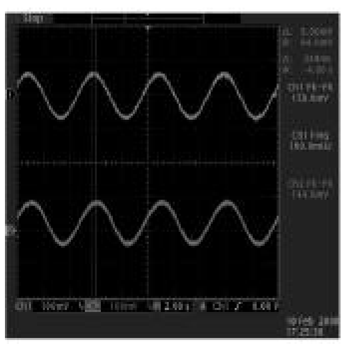
$$SNR = 20\log\left(\frac{V_{in\_rms}}{V_{n\_rms}}\right) = 20\log\left[\left(\frac{\frac{G_m}{K_P}\sqrt{32HD_3}}{\sqrt{\frac{2K_F}{C_{ox}^2}\ln\left(\frac{f_2}{f_1}\right)}}\right)\left(\frac{L^{1.5}W^{0.5}}{\alpha}\right)\right]$$

MEASURED RESULTS

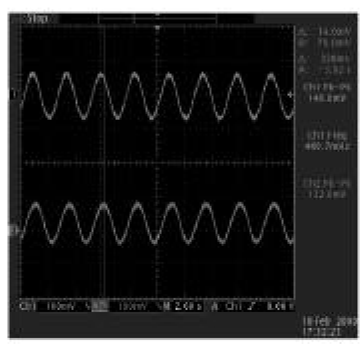


- f_{3dB} ~ 0.7 Hz with C_L = 2.7nF
- gm = $2\pi C_L f_{3dB}$ ~ 11.8nA/V

EXPERIMENTAL RESULTS 1.2 micron technology

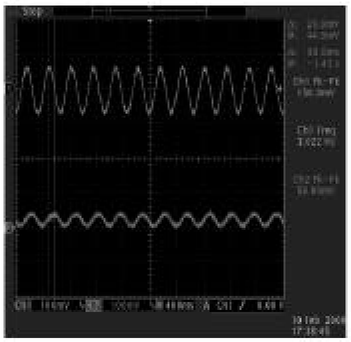


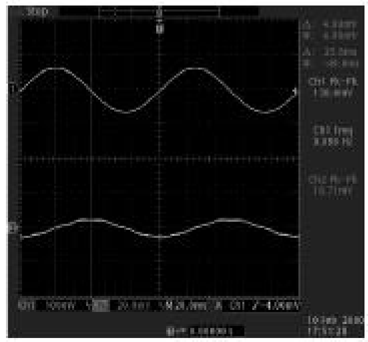
Input Ch1 150mVpp @ 0.2 Hz
Output Ch2 144mVpp



Input Ch1 150mVpp @ 0.4 Hz
Output Ch2 132mVpp

WAVEFORMS...





Input Ch1 150mVpp @ 3 Hz
Output Ch2 56mVpp

Input Ch1 150mVpp @ 10 Hz
Output Ch2 11mVpp

Summary of Experimental results

PARAMETER	VALUE
Filter order	2
-3dB Bandwidth (Hz)	0.3
HD_3 @ V_{in}=150mVpp (dB)	-45
Total input noise (µVrms)	15.6
SNR (dB)@ HD_3 = -45dB	70.5
Power consumption (µW)	8.18
Power supply (V)	±1.35
Total filter area (mm²)	0.06

• HD2 ~ -33dBm ~ 2.2 %

- **FULLY CURRENT-MODE**

Input Signal: Current

Output Signal: Current

Basic Building Blocks are:

Inverting Integrators

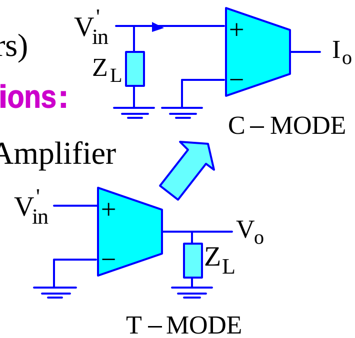
Inverting (Current Amplifiers)

Primitive Circuit Implementations:

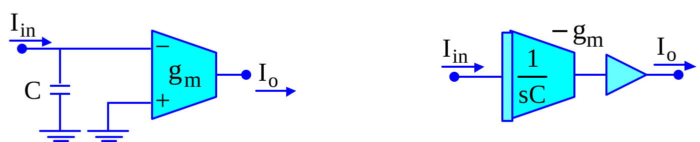
Single Transistor Inverting Amplifier

Simple Current Mirror

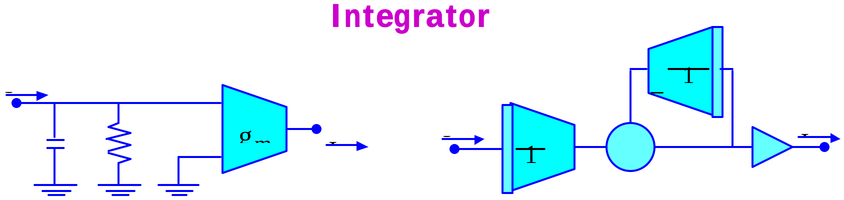
Capacitor



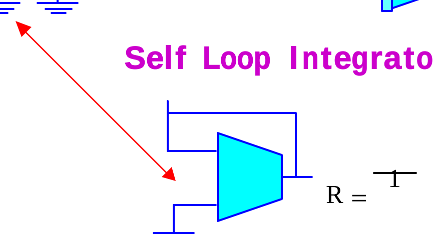
Current-Mode Implementation using OTA's



Integrator

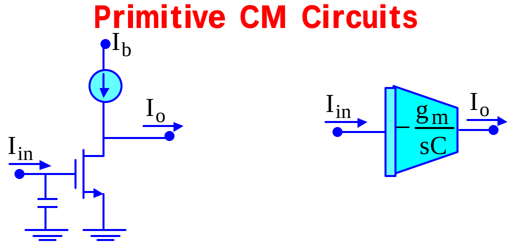


Self Loop Integrator

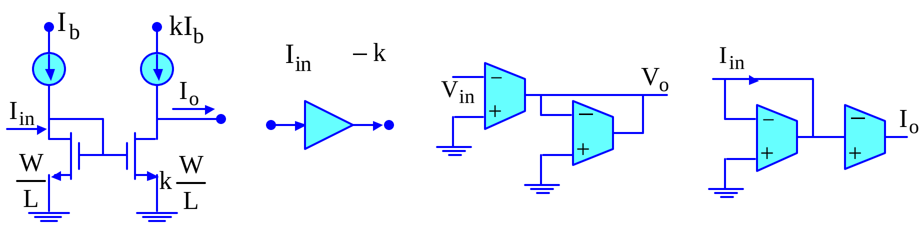


In order to fully obtain the benefits of current-mode techniques simpler circuits with reduced parasitics are desireable.

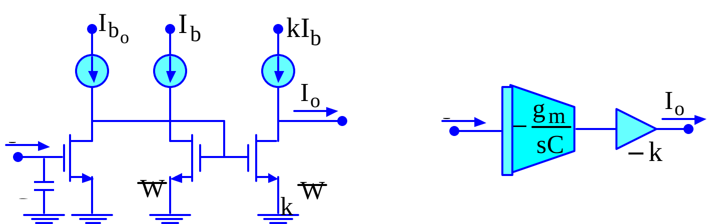
Primitive CM Circuits



Inverting Integrator

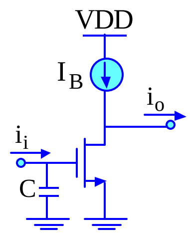


Amplifier (Multiplier by a constant)

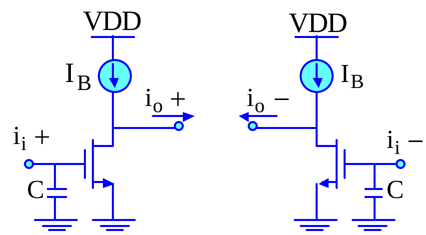


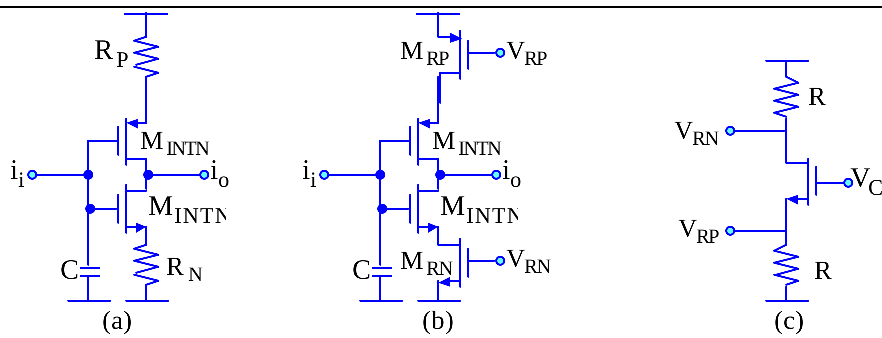
Non-Inverting Integrator

3.3V Power supply
High frequency
Low area
Suitable for digital process
Good PSR
Poor linearity, efficiency (1% THD => η < 4%)
Poor voltage gain



Low power supply (3.3V)
High frequency
Low area
Suitable for digital process
Very good PSR
Good Linearity (differential)
Excellent efficiency (≈ 100%)
Poor common mode rejection





(a) Tunable CMOS class AB integrator (b) Transistor Implementation with Mrn and Mrp operating in triode region (c) Bias implementation (diffusion or poly resistors).

Linearity sufficient

Very high efficiency (> 100%) => AB, low power

Very high frequency

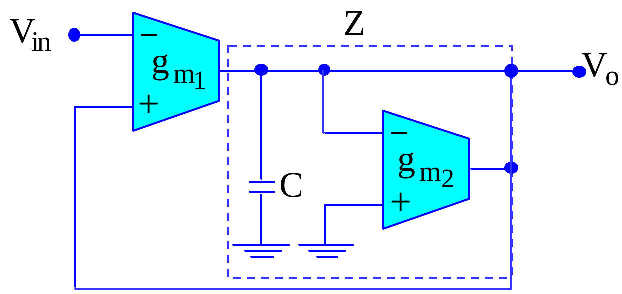
Small area

Low Power Supply

Linearity dep. on process variations

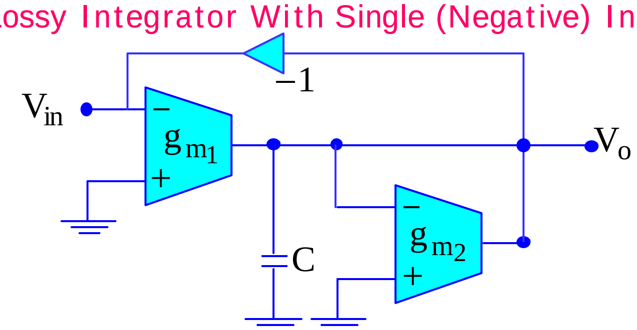
PSR poor

Lossy Integrator With Positive Feedback

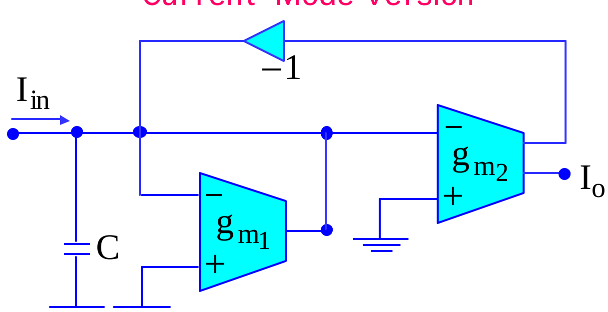


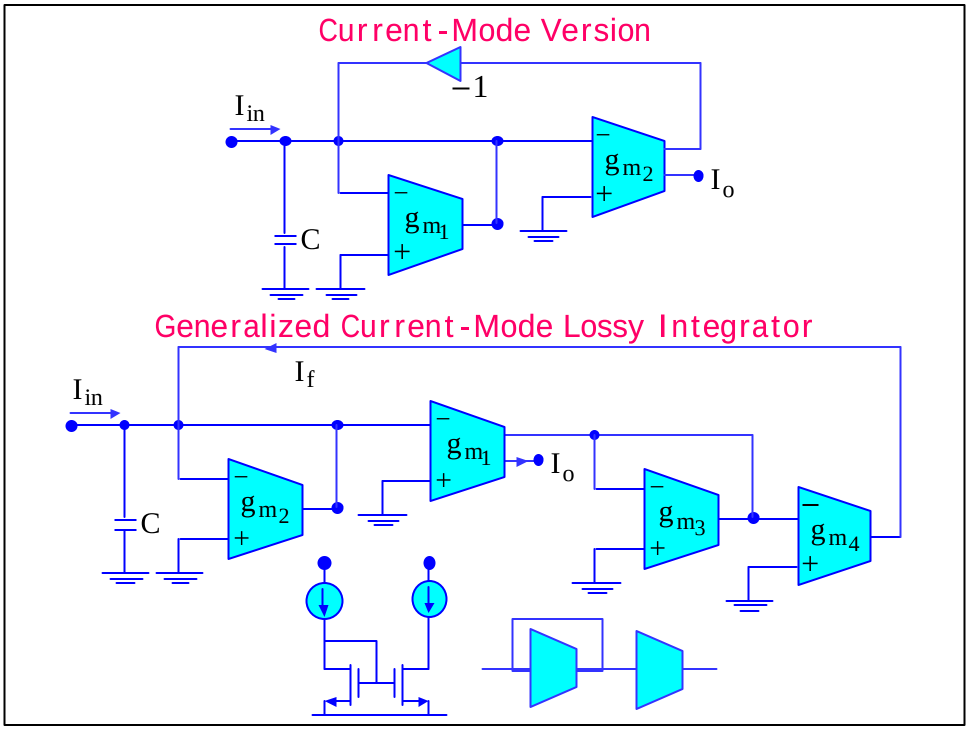
$$\frac{V_o}{V_{in}} = \frac{-g_{m_1}Z}{1-g_{m_1}Z} = -\frac{g_{m_1}}{sC_2+(g_{m_2}-g_{m_1})}$$

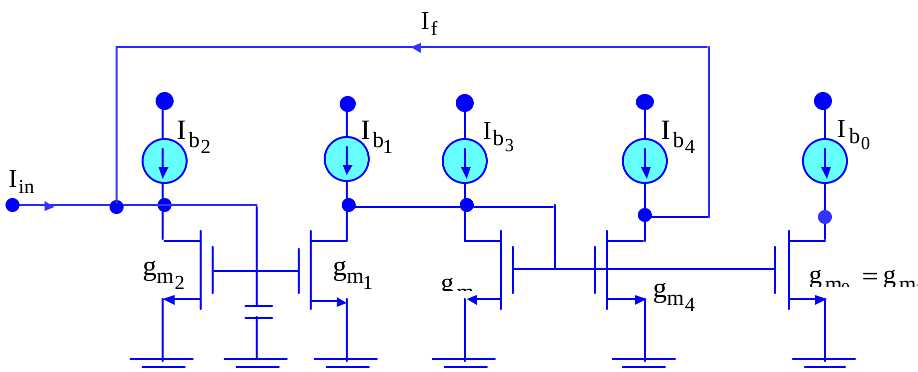
OTA-C Lossy Integrator With Single (Negative) Input OTA's



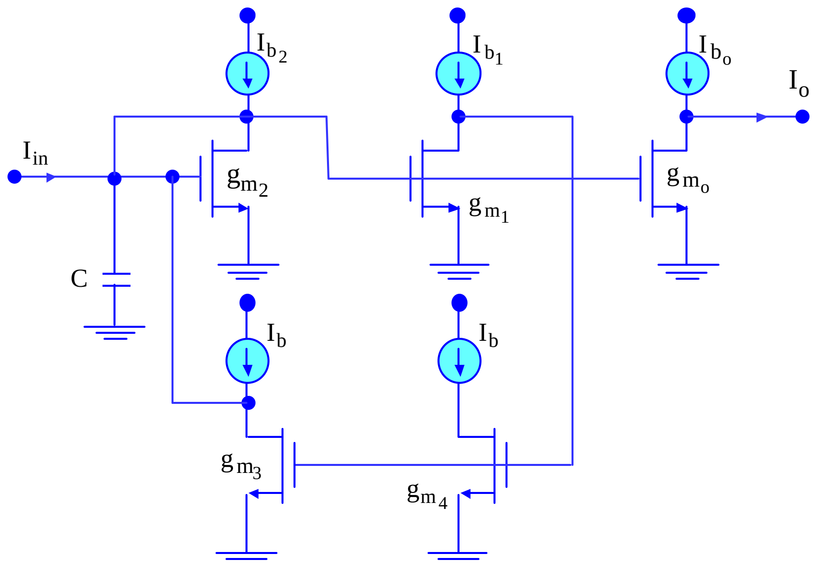
Current-Mode Version





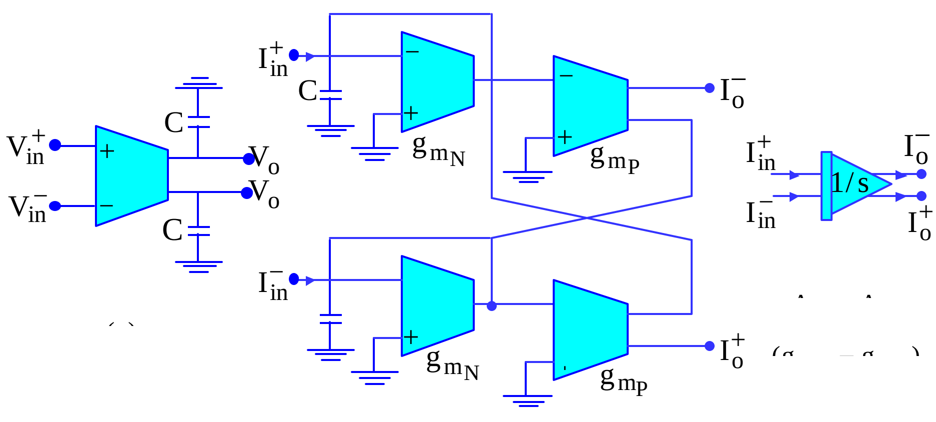


Transistor Level Implementation
CM Lossy (Lossless) Integrator



Redrawing the CM Integrator

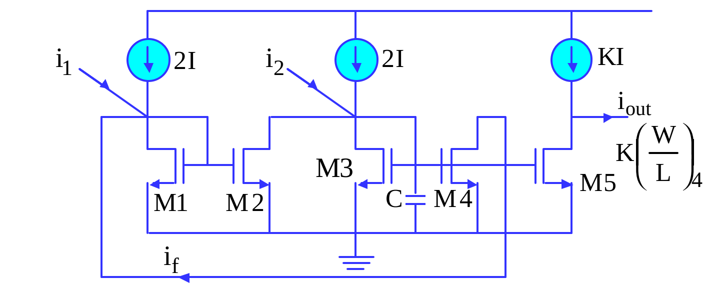
Fully Pseudo Differential Integrator

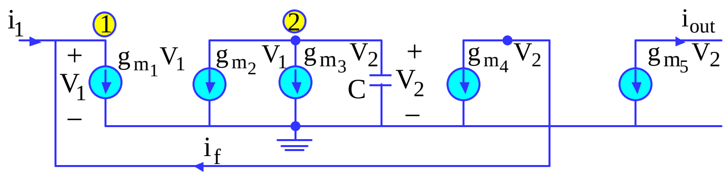


$$A_{dm} = \frac{I_o^+ - I_o^-}{I_{in}^+ - I_{in}^-} = \frac{-\omega_u}{[illegible]}$$

$$A_{cm} = \frac{I_o^+ + I_o^-}{I_{in}^+ + I_{in}^-} = \frac{-\omega_u}{[illegible]}$$

Continuous - Time Current-Mode Integrator Based On Current-Mirrors.





$$i_f = \frac{i_1 \dfrac{g_{m_2}}{g_{m_1}} - i_2}{g_{m_1}\left(g_{m_3} + sC\right) - g_{m_2} g_{m_4}} \bullet g_{m_1} g_{m_4}$$

$$i_f = g_{m_1} g_{m_4} \frac{i_1 \dfrac{g_{m_2}}{g_{m_1}} - i_2}{g_{m_1} g_{m_3} - g_{m_2} g_{m_4} + g_{m_1} sC}$$

a) Lossless Integrator

$g \;=\; g \quad \text{and} \quad g \;=\; g$

$$i_f = \frac{g_{m4}}{} (i_1 - i_2)$$

$$i_{out} = K \frac{g_{m4}}{} (i_1 - i_2)$$

b) Lossy Integrator

$$g_{m_1}g_{m_3} > g_{m_2}g_{m_4} \quad , \quad g_{m_1} = kg_{m_2} \ , \ g_{m_3} = kg_{m_4}$$

$$i_f = \frac{k}{k^2-1}\frac{ki_1 - i_2}{1+\frac{sC}{g_{m_4}}\frac{k^2-1}{k}} \quad , \quad k > 1$$

i.e. $k = 2$

$$i_f = \frac{2}{3}\frac{2i_1 - i_2}{1+\frac{sC}{g_{m_4}}\frac{3}{2}}$$

If the parasitic capacitances and the output conductances are considered, then

$$i_f = \frac{-k_1(s-z_1)i_1}{(s+p_1)(s+p_2)} - \frac{k_2(s+z_2)}{(s+p_1)(s+p_2)}$$

Where

$$k_1 = g_o/C_1 \quad , \quad k_2 = g_m/C_2$$

$$p_1 = 4g_o/C_2 \quad , \quad p_2 = g_m/C_1$$

$$z_1 = \frac{g_m}{C_2}\frac{g_m}{g_o} \quad , \quad z_2 = \frac{g_m + g_o}{C_1}$$

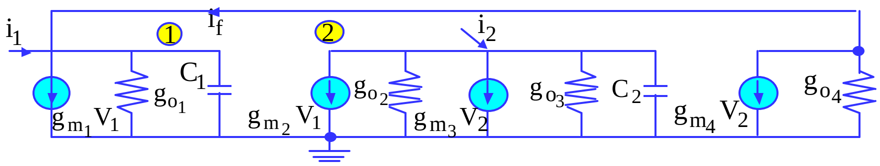
All transistors are equal, and C_1 and C_2 are the lumped nodal capacitances associated with nodes 1 and 2. Note that p_1 moves from the origin to

$$p_1 \rightarrow \frac{\omega_o}{\alpha} = \frac{\frac{g_{m3}}{C_2}}{g_m} = \frac{g_{m3}}{g_m}\frac{4g_o}{C_2}$$

And

$$Q = -\frac{g_{m1}}{\alpha}\frac{C_2}{C_1}$$

Let's consider the input and output impedance,



$$z_{in} = \left.\frac{V_1}{i_1}\right|_{i_2=0} = \frac{g_{m_3} + g_{o_2} + g_{o_3} + sC_2}{-g_{m_4}g_{m_2} + (g_{m_1} + g_{o_1} + g_{o_4} + sC_1)(g_{m_3} + g_{o_2} + g_{o_3} + sC_2)}$$

$$z_{in} = \left.\frac{V_1}{i_1}\right|_{i_2=0} \cong \frac{g_{m_3}(1 + sC_2/g_{m_3})}{-g_{m_4}g_{m_2} + g_{m_1}g_{m_3} + s(C_2 g_{m_1} + C_1 g_{m_3}) + s^2 C_1 C_2}$$

a) Lossless Integrator

$$z_{in}(0) \cong \left.\frac{g_{m_3}}{-g_{m_4}g_{m_2} + g_{m_1}g_{m_3}}\right|_{\substack{g_{m_1}=g_{m_2} \\ g_{m_3}=g_{m_4}}} \rightarrow \infty$$

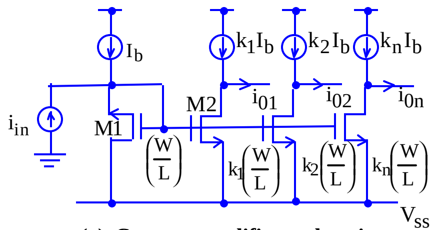
b) Lossy Integrator

$$z_{in}(0) \cong \frac{k^2}{\ldots}\,\frac{1}{\ldots}, \quad k > 0$$

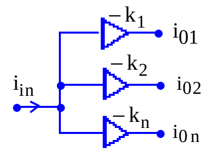
$$z_o(0) \cong \frac{1}{\ldots}$$

Current-Mode: Single and Double-Ended

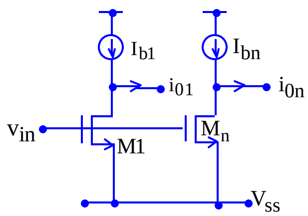
Basic Cell



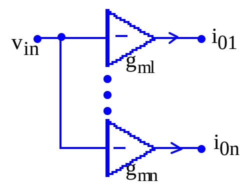
(a) Current amplifier and copies



(b) Representation

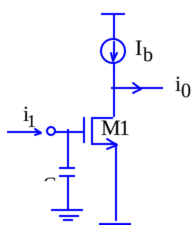


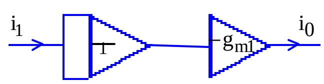
(c) Transconductance amplifier



(d)

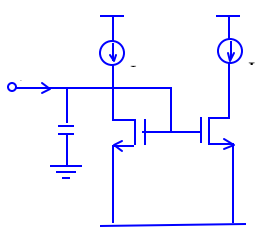
Current-Mode Integrators Single-Ended

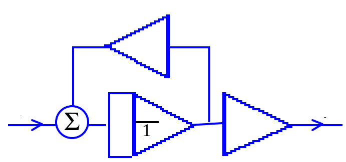




$$\frac{I_0(s)}{I_1(s)} = -\frac{g_{m1}}{sC}$$

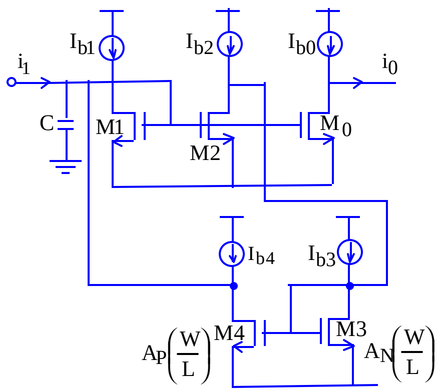
(a) Lossless Integrator

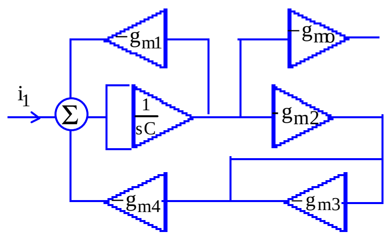




$$\frac{I_0(s)}{I_1(s)} = \frac{-\frac{g_{m2}}{g_{m1}}}{1+\frac{sC}{g_{m1}}}$$

(b) Lossy Integrator



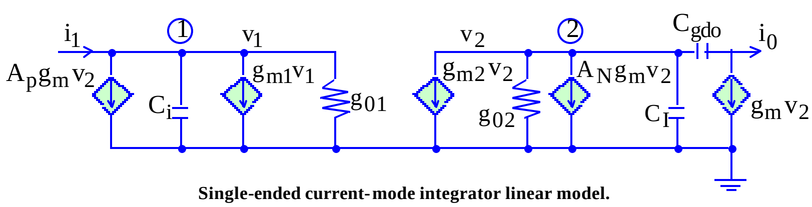


$$\frac{I_0(s)}{I_1(s)} = \frac{-g_{m0}}{g_{m1} - \frac{g_{m4} g_{m2}}{g_{m3}} + sC}$$

(c) Lossy Integrator with additional degrees of freedom

What happened with $g_{m1} = \frac{g_{m4} g_{m2}}{g_{m3}}$?

Second-Order Effects on a Single-Ended Current-Mode Integrator (C):



Single-ended current-mode integrator linear model.

$$I_0(S) \approx \frac{g_m\left(1+\frac{S}{\omega_{z2}}\right)}{SC_I\left(1+\frac{\omega_I}{S}\right)}\left[I_1 \frac{1-\delta_1}{\left(1+\frac{S}{\omega_{p1}}\right)}\right]$$

where

$$\delta_1 \equiv g_{01}/g_{m1}, \quad \omega_{z2} = \frac{g_m}{c_{gdo}}$$

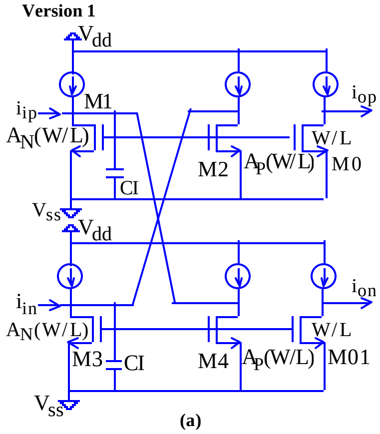
$$\omega_I = \frac{g_{01}+g_{02}}{C_I}, \quad \omega_{p1} = \frac{g_{m1}}{C_i}$$

g_{01} and g_{02} are the sum of the transistor output conductances at nodes 1 and 2 respectively, C_i is the total capacitance at node 1, and C_I the integrating capacitor, is the total capacitance at node 2. The integrator quality factor will be limited by the ratio of the parasitic pole to the integrator unity gain frequencies,

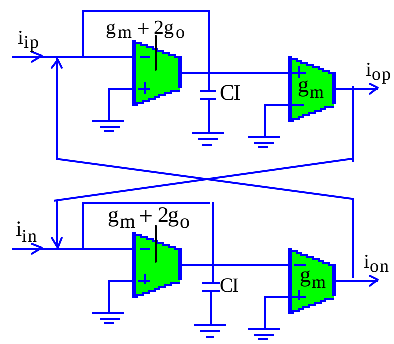
$Q \approx g_{m1}\ C_I$

Improved Balanced Integrators (Double-Ended)

Using the previous single-ended integrator, a double-ended integrator is obtained:
Version 1



(a)



(b)

$$A_{cm1} = \frac{i_{op}+i_{on}}{i_{ip}+i_{in}} = \frac{-g_m}{sC_I + (A_N + A_P)(g_m + 2g_0)}$$

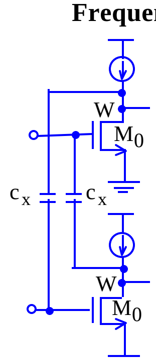
$$A_{dm1} = \frac{i_{op}-i_{on}}{i_{ip}-i_{in}} = \frac{g_m}{sC_I + (A_N - A_P)(g_m + 2g_0)}$$

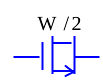
$$IM_3 \approx \frac{3}{32}\left[\frac{i_{0d}}{2I_b}\right]^2$$

and that the harmonic distortion can be approximat ed by

$$THD_3 \approx \frac{1}{32}\left[\frac{i_{0d}}{2I_b}\right]^2$$

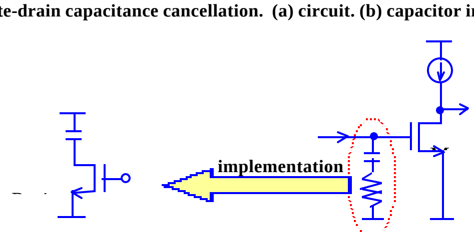
Frequency Compensation





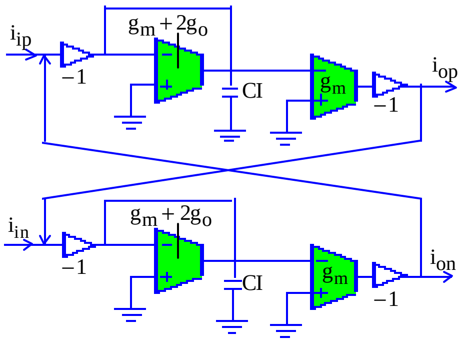
(a) **(b)**

Gate-drain capacitance cancellation. (a) circuit. (b) capacitor impl ementation.

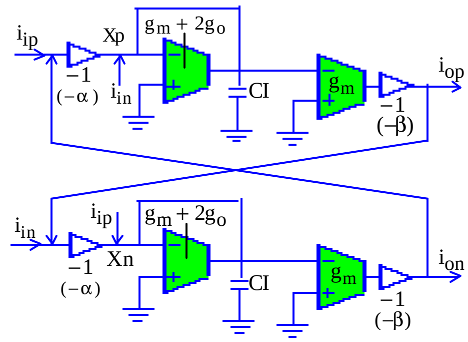


Passive phase compensation of current integrators.

Version 2



Version 3



$$A_{dm2} \cong \frac{-g_m}{sC_I + 6g_0}$$

$$A_{cm2} \cong \frac{-g_m}{sC_I + 2g_m}$$

$$A_{dm3} \cong \frac{-(sC_{px} + 2g_m)}{2(sC_I + 6g_0)}$$

$$A_{cm3} \cong \frac{-(sC_{px} + 2g_0)}{2(sC_I + 2g_m)}$$

A Comparative Table

	Case 1	**Case 2**	**Case 3**
A_{dm}(0)	$\frac{g_m}{2g_o}$	$\frac{g_m}{6g_o}$	$\frac{g_m}{6g_o}$
A_{cm}(0)	$\frac{1}{2}$	$\frac{1}{2}$	$\frac{g_o}{2g_m}$
CMRR(0)	$\frac{g_m}{g_o}$	$\frac{g_m}{3g_o}$	$\frac{g_m^2}{g_o^2}$
CMRR(∞)			

Improved Sensitivity

$$A_{dm} \cong \frac{-(sC_{px} + (\alpha\beta + \beta)g_m}{sC_I + 6g_0 + (1-\varepsilon-\alpha\beta)g_m}$$

$$A_{dm}(0) = \frac{\alpha\beta + \beta}{6g_0 / g_m + (1-\varepsilon-\alpha\beta)}$$

What are the values of α and β?

$$\alpha\beta < 1$$

say $\beta = 9$ and $\alpha = 0.1$

Can you suggest a better common-mode circuit?

Reference:
X. Quan, S. Embabi and E. Sánchez-Sinencio, “Improved Fully Balanced Current-Mirror Integrator”, Electronic Letters, vol. 34, No. 1, pp. 1-3, January 1998.

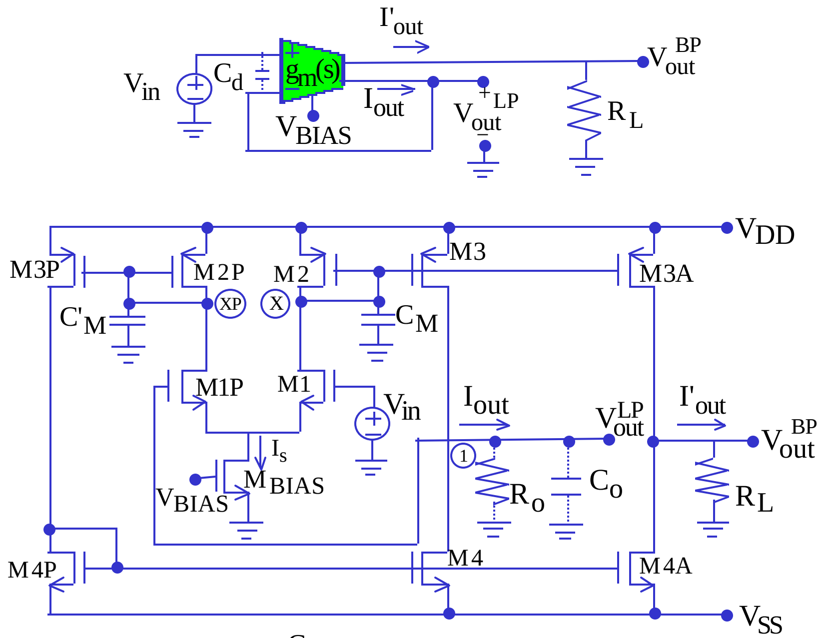
How can you use a conventional OTA as a filter?

How high can the ω_o or the $\omega_o Q$ of a second-order filter be obtained?

Can we reach the limits of the technology?

How can we use OTA multiple outputs?

Reference: J. Ramírez-Angulo, E. Sánchez-Sinencio and M. Howe, “Large foQ Second-Order Filters Using Multiple Outputs OTAS”, Trans. Circuits and Systems II, vol. 41, No. 9, pp. 587-592, September 1994.



C'_M & C_M ARE REAL CAPACITORS

The corresponding transfer functions are:

$$H_{LP}(s) = \frac{V_{out}^{LP}(s)}{V_{inp}(s)} = \frac{\omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$$

$$H_{BP}(s) = \frac{V_{out}^{BP}(s)}{V_{inp}(s)} = \frac{K_{BP}\frac{\omega_0}{Q}s}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}$$

Where

$$\omega_0 = \sqrt{\frac{g_{m1}}{C_0}\frac{g_{m2}}{C_m}} \qquad Q = \sqrt{\frac{C_m}{C_0}\frac{g_{m1}}{g_{m2}}}$$

And

For $\omega / \omega_p << 1$ and $g_m(s) \cong g_{mo}^{o^{-s/\omega p}}$

$$\omega_p = \frac{g_{m2}}{C_m}$$

$$GB = A_{vo}BW = g_{m1}R_0 \frac{1}{R_0C_0} = \frac{g_{m1}}{C_0}$$

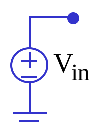
Where $R_0 = 1/(g_{03}+g_{04})$

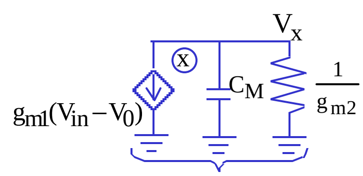
$$\omega_0 = \sqrt{\frac{g_m^0 \omega_p}{C_0}} = \sqrt{GB\omega_p}$$

$$Q = \sqrt{\frac{g_m^0}{C_0\omega_p}} = \sqrt{\frac{GB}{\omega_p}}$$

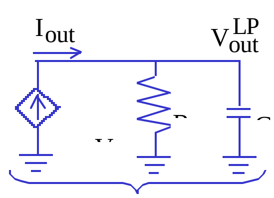
$$K_{BP} = g_m^0 R_L$$

$$f_0 Q = GB / 2\pi$$

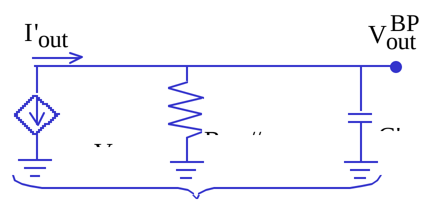




Lossy Integrator



Integrator and First Output

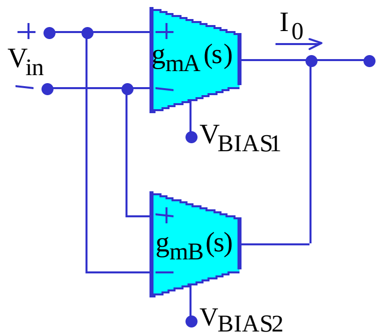


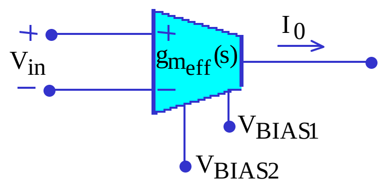
Second Output

$$GB = \frac{g_{m1}}{C_{0PAR}} = \frac{\mu_n C_{ox} \frac{W_1}{L_1}(V_{GS_1} - V_T)}{\frac{2}{3} C_{ox}(W_1L_1)} \cong \frac{\mu_n (V_{GS_1} - V_T)}{L_1^2}\frac{3}{2}$$

i.e., 2μm technology $f_0 Q \sim 2GHz$

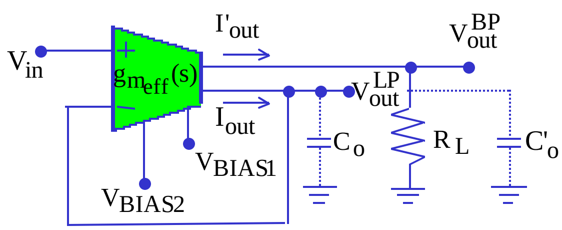
How can we obtain Q and f_0 independently tuned?
A possible solution is shown next.





$$g^o_{meff} = g^o_{mA} - g^o_{mB} = g_{mA}(1 - 1/N)$$

Where $N = \dfrac{g_{mA}}{g_{mB}}$



Analog and Mixed -Signal Center, TAMU

Summary of the Results

$$\omega_0 = \sqrt{\frac{g_{m_A}}{C_O}\omega_{p_A}}\sqrt{1 - \frac{1}{N^2}}$$

$$Q = \frac{1}{\dfrac{\omega_{p_A}}{\omega_o}\left[1 + \dfrac{1}{N} - \dfrac{1}{N+1}\right] - \dfrac{\omega_o}{2\omega_{p_H}}\left[1 - \dfrac{1}{N}\right]}$$

DEPENDENCE OF Q AND f_0 WITH N.

N	f_0 / MHz	Q
∞	16.45	56.66
15	16.41	40.90
7.0	16.28	27.72
3.0	15.50	13.83

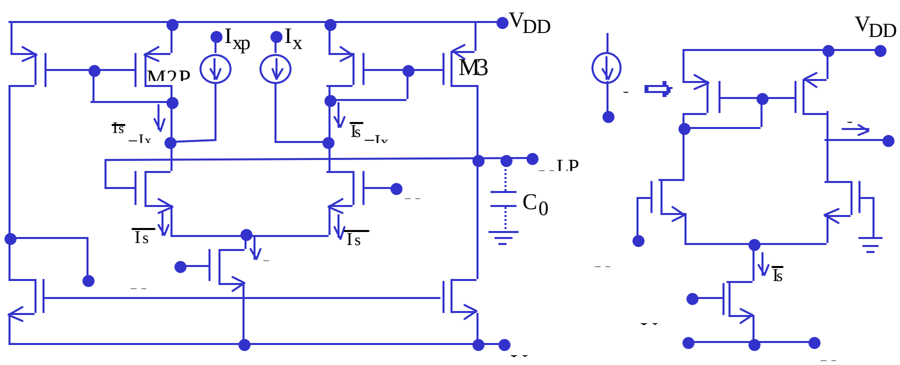
$C_m = 20pF$, $C_0 = 1pF$, $g_{mA} = 850\mu A/V$, $f_{PH} = 75MHz$ and $f_{PA} = 2MHz$

Where ω_{PH} is the high frequency pole of mirror M4, M4P.

Implementations of g_{meff} :

- Two equal OTAs with different bias to satisfy the $N = g_{mA} / g_{mB}$ or different (W/L)s.
- One OTA: To inject a current I_x to the drains of the differential pair. This yields different currents through the load transistor M2 and M2P. Consequently, we can have independent Q tuning.

The actual implementation is discussed next.



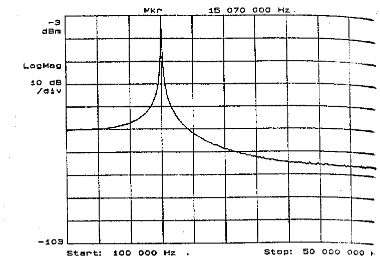
$$g_{m1} = \sqrt{2\mu_n C_{ox} \frac{W_1}{L_1} \frac{I_s}{2}}$$

$$g_{m2} = \sqrt{2\mu_p C_{ox} \frac{W_2}{L_2} \left(\frac{I_s}{2} - I_x \right)}$$

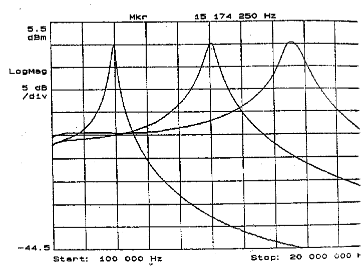
Recall

$$\omega_0^2 = \frac{g_{m1} g_{m2}}{(C_0 C_m)} \quad \& \quad Q^2 = \frac{C_m}{C_0} \frac{g_{m1}}{g_{m2}}$$

Experimental Results



Experimental frequency response of low pass filter tuned for f_0 = MHz, $Q = 200$.



Experimental low pass response showing variable f_0.

Appendix: Derivation of Expressions

$$N = \frac{g^o_{m_A}}{g^o_{m_B}} = \frac{\omega_{p_A}}{\omega_{p_B}}$$

$$\omega_0 = \sqrt{\frac{g_{m_A}}{C_0}\omega_{p_A}}\sqrt{\left(1-\frac{1}{N^2}\right)}$$

$$\tan^{-1} x = \frac{\pi}{2} - \tan^{-1}\frac{1}{x}$$

$$\tan^{-1} x = x - \frac{x^3}{3} + \frac{x^5}{5} - \frac{x^7}{7} + \ldots$$

$$\Phi(\omega) = \frac{\omega_{p_A}}{\omega}\left[1+\frac{1}{N}-\frac{1}{N+1}\right] - \frac{\omega}{2\omega_{PH}}\left[1-\frac{1}{N}\right]$$

$$Q = \frac{1}{\frac{\omega_{p_A}}{\omega_o}\left[1+\frac{1}{N}-\frac{1}{N+1}\right] - \frac{\omega_o}{2\omega_{PH}}\left[1-\frac{1}{N}\right]}$$

$$Q(N \to \infty) = \frac{1}{\frac{\omega_{p_A}}{\omega_o} - \frac{\omega_o}{2\omega_{PH}}}$$

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