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

- **OTA-C Filter Topologies**
- **OTA-C Filter Non-idealities**
- **Pseudo Differential OTA**
- **OTA-C BP Least Mean Square Tuning Scheme**
- **How to use a conventional OTA as a filter by adding capacitances at the internal nodes.**

Applications for continuous time filters

Top view of a 36 GB, 10,000 RPM, IBM SCSI server hard disk, with its top cover removed.

Read channel of disk drives -- for phase equalization and smoothing the wave form

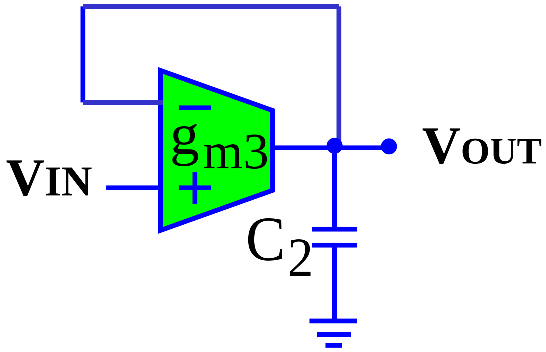
Receivers and Transmitters in wireless applications -- used in PLL and for image rejection

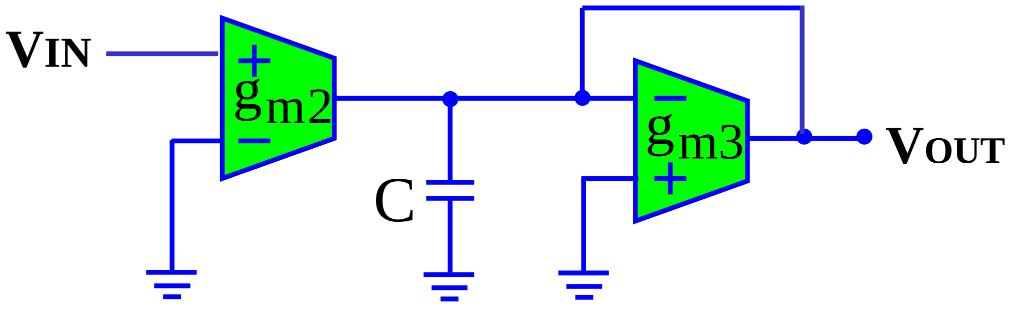
6185i digital cell phone
from Nokia.

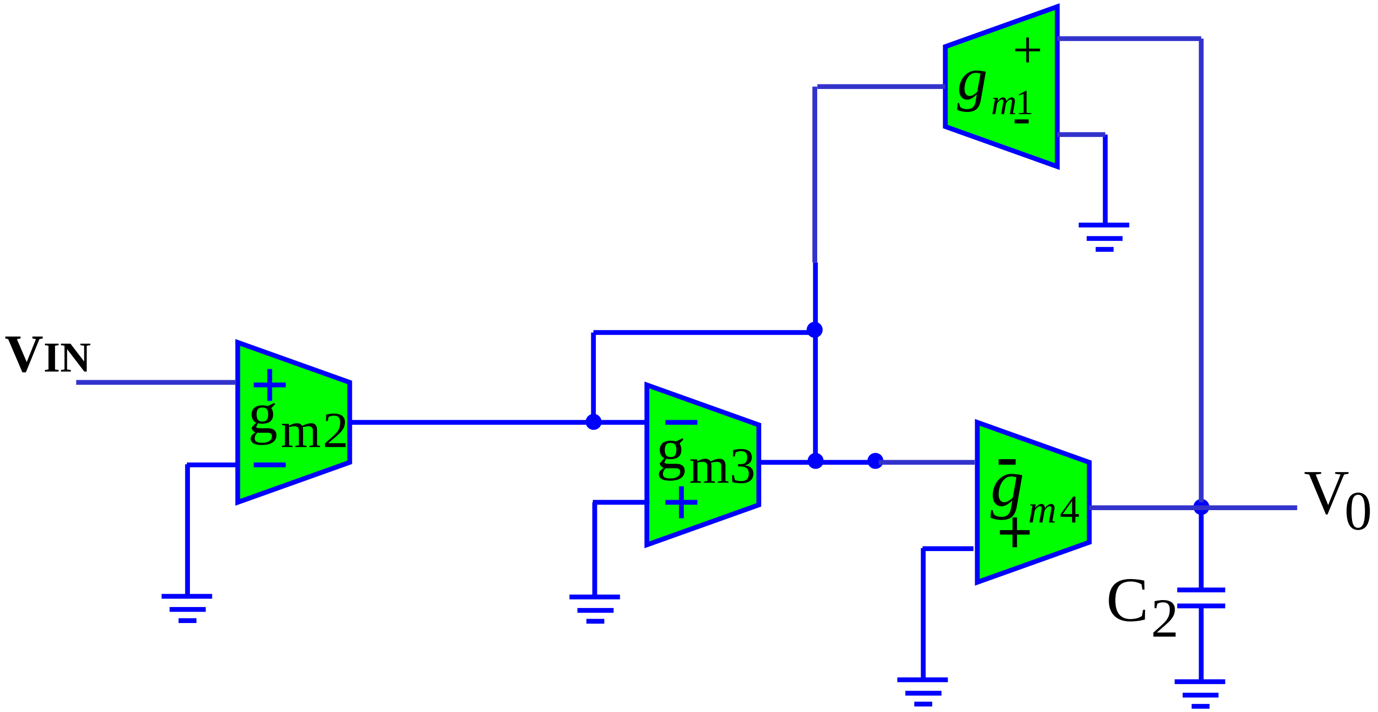
CMP-35 portable MP3 player

All multi media applications --Anti aliasing before ADC and smoothing after DAC

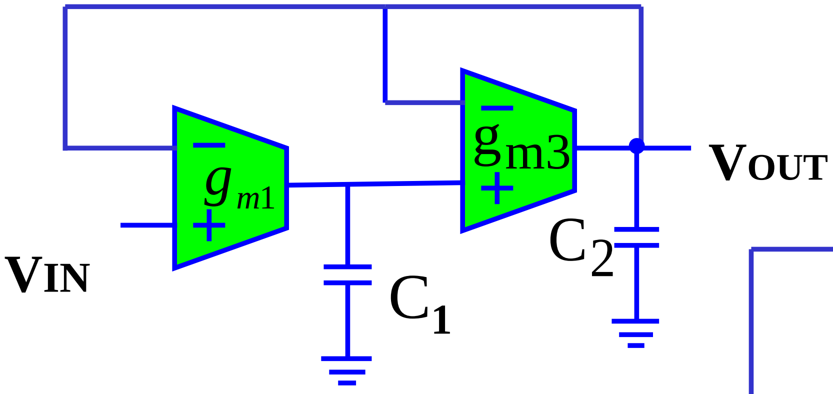
LOSSY OTA-C INTEGRATORS



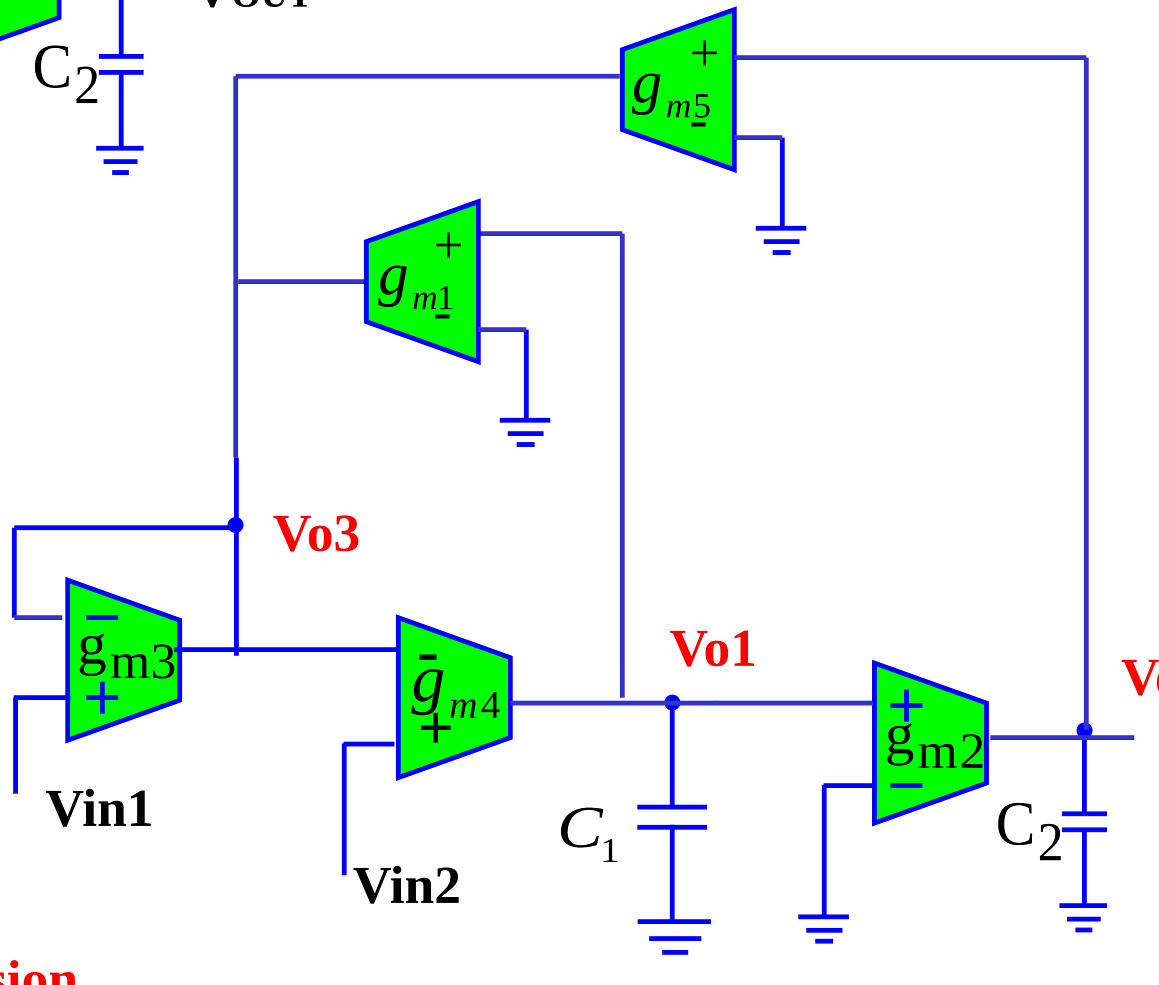




OTA-C Two Integrator Loop Filters



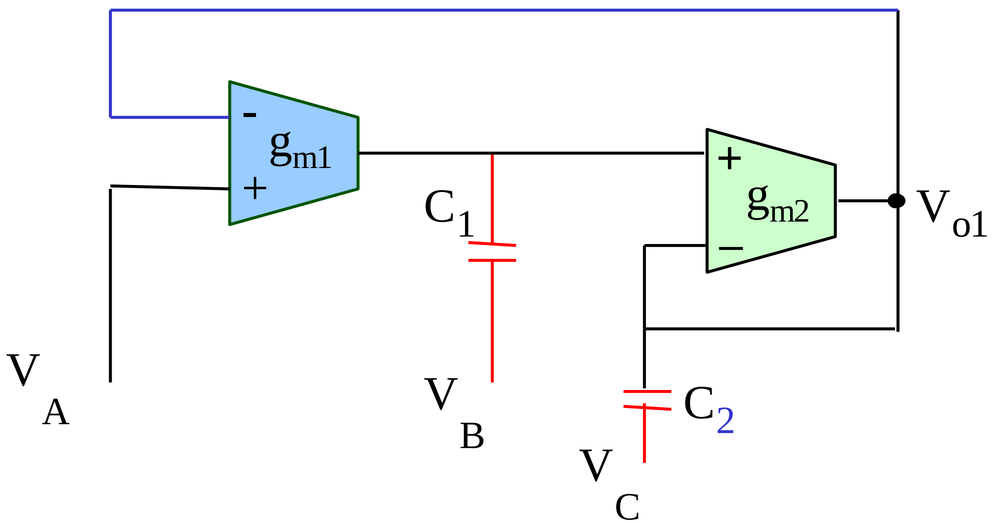
Two OTAs Filter



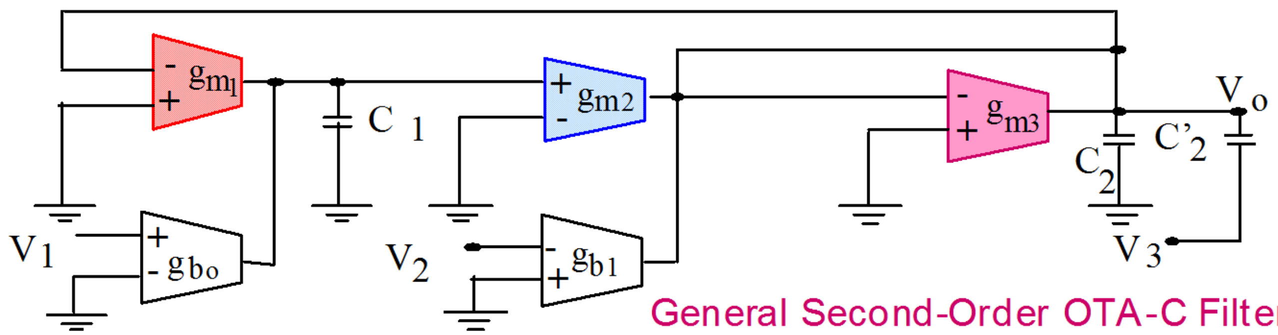
KHN OTA-C Version

How to generate the zeros of the filter ?

Canonic OTA-C Biquad



$$V_{01} = \frac{s^2 C_1 C_2 V_C + sC_1 g_{m2} c V_B + g_{m1} g_{m2} V_A}{s^2 C_1 C_2 + sC_1 g_{m2} + g_{m1} g_{m2}}$$



For $V_1=V_2=V_3$

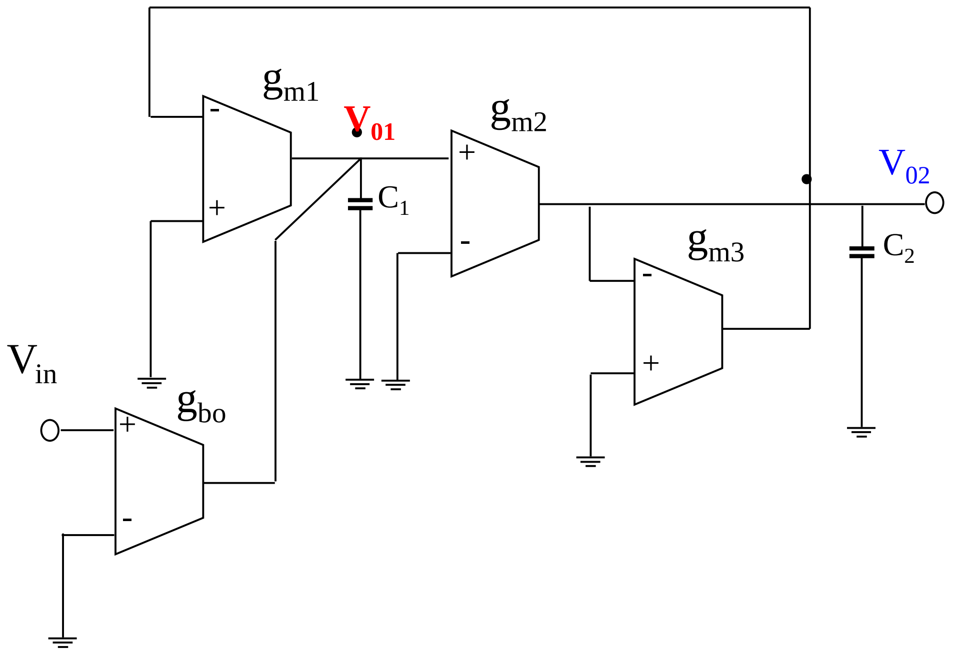
$$H(s) = \frac{C_2}{C_2 + C'_2} \frac{s^2 - s\frac{g_{b1}}{C_2} + \frac{g_{bo} g_{m2}}{C_1 C_2}}{s^2 + s\frac{g_{m3}}{(C_2 + C'_2)} + \frac{g_{m1} g_{m2}}{C_1 \left(C_2 + C'_s\right)}}$$

$$\omega_o^2 = \frac{g_{m1} g_{m2}}{C_1 (C_2 + C'_2)} \quad ; \quad \omega_z^2 = \frac{g_{bo} g_{m2}}{C_1 C_2}$$

$$\frac{\omega_o}{Q} = \frac{g_{m3}}{(C_2 + C'_2)} \quad ; \quad \frac{\omega_z}{Q_z} = \frac{g_{b1}}{C_2}$$

$$|H(0)| = \frac{g_{bo}}{g_{m1}} \quad ; \quad |H(\infty)| = \frac{C_2}{C_2 + C'_2}$$

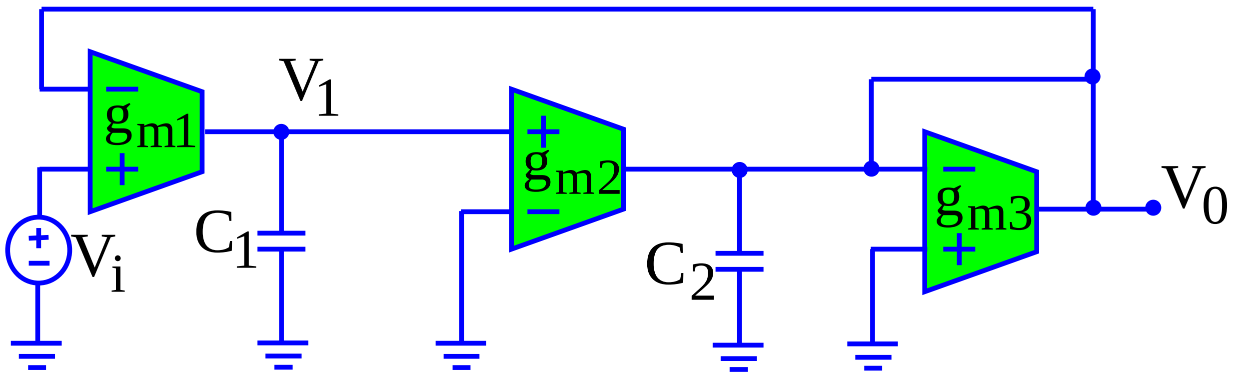
INTERNAL VOLTAGE SCALING



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},

OTA-C Three OTA Filter: Transfer Function Derivation taking into Account the OTA non-idealities.



Assume ideal OTAs first, then :

$$V_1 = \frac{1}{sC_1} g_{m1}(V_i - V_0) \qquad (1)$$

$$V_0 = (g_{m2}V_1 - g_{m3}V_0)\frac{1}{sC_2} \qquad (2)$$

(1) into (2)

$$(sC_2)V_0 = \left[g_{m2} \frac{g_{m1}}{sC_1}(V_i - V_0) - g_{m3}V_0 \right]$$

$$V_0\left[sC_2 + \frac{g_{m1}g_{m2}}{sC_1} + g_{m3}\right] = \frac{g_{m1}g_{m2}}{sC_1}V_i$$

$$H_{LP}(s) = \frac{V_0}{V_i} = \frac{g_{m1}g_{m2}}{s^2C_1C_2 + sC_1g_{m3} + g_{m1}g_{m2}} = \frac{\frac{g_{m1}g_{m2}}{C_1C_2}}{s^2 + s\frac{g_{m3}}{C_2} + \frac{g_{m1}g_{m2}}{C_1C_2}}$$

$$\omega_o^2 = \frac{g_{m1}g_{m2}}{C_1C_2} \quad , \quad BW = \frac{\omega_o}{Q} = \frac{g_{m3}}{C_2}$$

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

Now let's assume the transconductance is characterized by:

$$g_m = g_{mo} e^{-s/\omega_p} \cong g_{mo}\left(1 - s/\omega_p\right) \text{ for } \omega_p << \omega_o.$$

Under this condition the excess phase can be expressed as $\phi_E \cong \omega_o / \omega_p$.

Note that ideally $\phi_E = 0^0$.

then,

$$H_{LP}(s) = \frac{g_{mo1} g_{mo2} (1 - s/\omega_{p1})(1 - s/\omega_{p2})}{s^2 C_1 C_2 + s C_1 g_{mo3} (1 - s/\omega_{p3}) + g_{mo1} g_{mo2} (1 - s/\omega_{p1})(1 - s/\omega_{p2})}$$

$$D(s) = s^2 C_1 C_2 + s C_1 g_{mo3} - s^2 \frac{C_1 g_{mo3}}{\omega_{p3}} + g_{mo1} g_{mo2}\left(1 - s\left(\frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}}\right) + \frac{s^2}{\omega_{p1}\omega_{p2}}\right)$$

$$D(s) = s^2 \left\{C_1 C_2 - \frac{C_1 g_{mo3}}{\omega_{p3}} + \frac{g_{mo1} g_{mo2}}{\omega_{p1}\omega_{p2}}\right\} + s\left\{C_1 g_{mo3} - \left(\frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}}\right) g_{mo1} g_{mo2}\right\} + g_{mo1} g_{mo2}$$

Then the actual ω_{oa} and BW_a become

$$\omega_{oa}^2 = \frac{g_{mo1} g_{mo2}}{C_1 C_2 + \frac{g_{mo1} g_{mo2}}{\omega_{p1}\omega_{p2}} - \frac{C_1 g_{mo3}}{\omega_{p3}}}$$

$$BW_a = \frac{\omega_{oa}}{Q_a} = \frac{C_1 g_{mo3} - \left(\frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}}\right) g_{mo1} g_{mo2}}{C_1 C_2 - \frac{C_1 g_{mo3}}{\omega_{p3}} + \frac{1}{\omega_{p1}\omega_{p2}}}$$

Let us also assume that $\omega_{p1} = \omega_{p2} = \omega_p$, then $\omega_{oa} \cong \omega_o$, thus,

$$BW_a = \frac{\omega_{oa}}{Q_a} = \frac{C_1 g_{mo3} - \frac{2}{\omega_{p1}} g_{mo1} g_{mo2}}{C_1 C_2 - \frac{C_1 g_{mo3}}{\omega_p} + \frac{1}{\omega_p^2}} \cong \frac{C_1 g_{mo3} - \frac{2}{\omega_p} g_{mo1} g_{mo2}}{C_1 C_2}$$

$$BW_a \cong \frac{g_{mo3}}{C_2} - \frac{g_{mo1} g_{mo2}}{C_1 C_2} \frac{2}{\omega_{p1}} = BW - \omega_{oa}^2 \cdot \frac{2}{\omega_{p1}} = BW - \frac{2\omega_{oa}^2}{\omega_{p1}}$$

$$Q_a = \frac{\omega_{oa}}{BW_a} \cong \frac{1 \cdot \omega_{oa}}{\frac{g_{mo3}}{C_2} - \omega_{oa}^2 \frac{2}{\omega_{p1}}}$$

$$Q_a = \frac{\frac{C_2 \omega_{oa}}{g_{m3}}}{1 - \frac{C_2 \omega_{oa}}{g_{m3}} \frac{2\omega_{oa}}{\omega_{p1}}} = \frac{Q}{1 - \frac{Q_2 \omega_{oa}}{\omega_{p1}}} = \frac{Q}{1 - \frac{2\omega_{oa}}{\omega_{p1}} Q}$$

Alternatively, Q_a can be expressed in terms of the excess phase $\phi_E = \tan^{-1} \frac{\omega_o}{\omega_p} \cong \frac{\omega_o}{\omega_p}$ then

$$Q_a \cong \frac{Q}{1 - 2\phi_E Q} \cong Q(1 + 2\phi_E Q)$$

$$BW_a = BW - 2\omega_{oa}\phi_E$$

Furthermore, if $A_{vo} = g_m R_o$ is taken into account, then

$$Q_a = \frac{Q}{1 + \dfrac{2Q}{A_{vo}}}$$

If $A_{vo} = 500$

$$Q_a = \frac{Q}{1 + 4 \times 10^{-3} Q}$$

Q	Q_a
1	0.996
5	4.902
10	9.6
50	41.667

Note that :

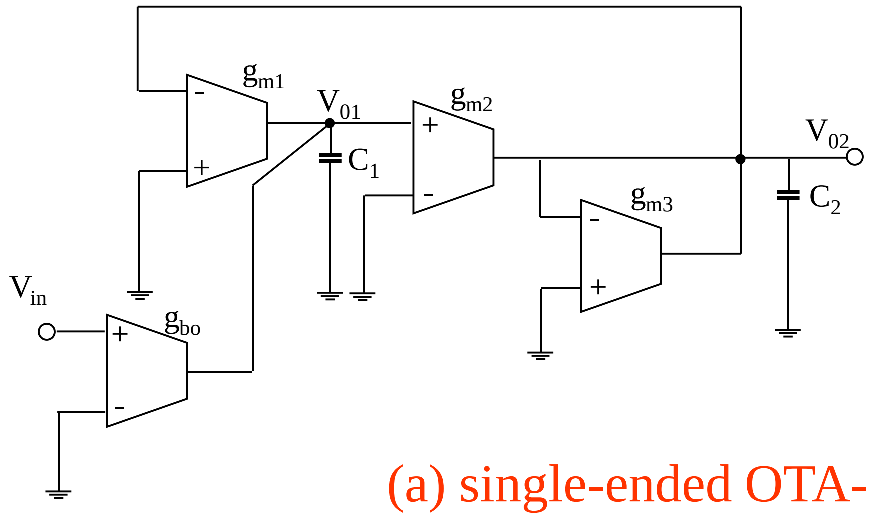
$$Q_a \downarrow \quad \text{when} \quad A_{vo} \downarrow$$

$$BW_a \downarrow \quad Q_a \uparrow \quad \text{when} \quad \phi_E \uparrow$$

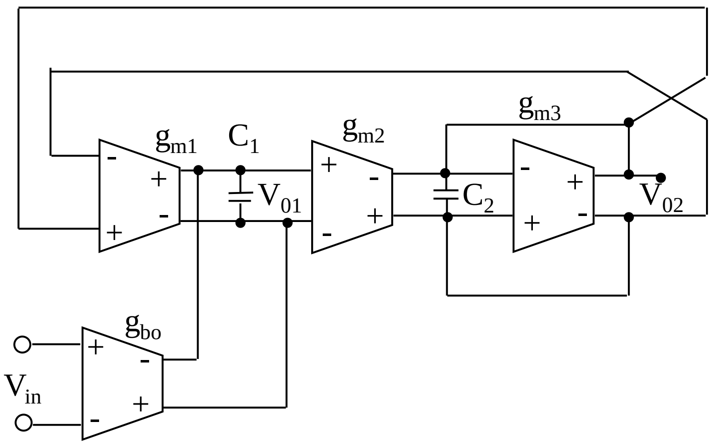
$$Q_a \cong \frac{Q}{1 - 2\phi_E Q} \cong Q(1 + 2\phi_E Q)$$

$$BW_a = BW - 2\omega_{oa}\phi_E$$

Two-integrator biquad with gain control



(a) single-ended OTA-C Biquad with one input



(b)

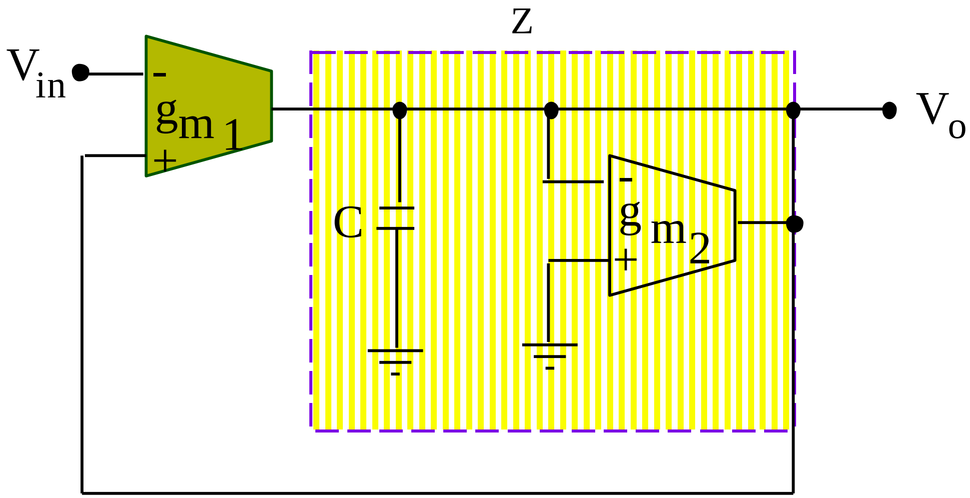
Fully differential OTA-C Biquad

Assuming a one pole OTA model
Table OTA finite parameters effects for biquad on the resonant frequency and bandwidth

Poles frequency*	$\sqrt{\dfrac{g_{m1}g_{m2}}{C_1C_2}}\left[\sqrt{1+\left(\dfrac{g_{o1}}{g_{m1}}\right)\left(\dfrac{g_{m3}+g_{o2}+g_{o3}}{g_{m2}}\right)-\left(\dfrac{\frac{g_{m1}}{C_1}}{\omega_{P1}}\right)\left(\dfrac{\frac{g_{m2}}{C_2}}{\omega_{P2}}\right)}\right]$
Bandwidth*	$BW_{ideal}(1-error) \cong \left(\dfrac{g_{m3}}{C_1}\right)\left(1-\dfrac{g_{o1}+g_{o2}+g_{o3}}{g_{m3}}-2Q\dfrac{\omega_0}{\omega_{P1,2}}\right)$

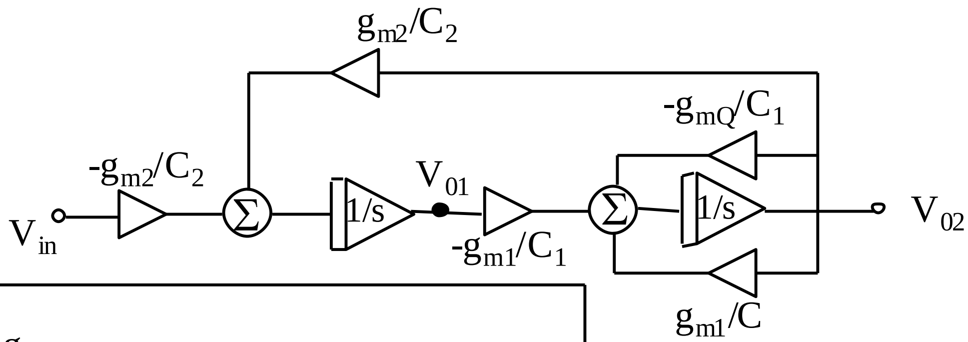
* $\omega_{P1,2}$ and $g_{o1,2}$ are the non-dominant pole and output conductance, respectively.

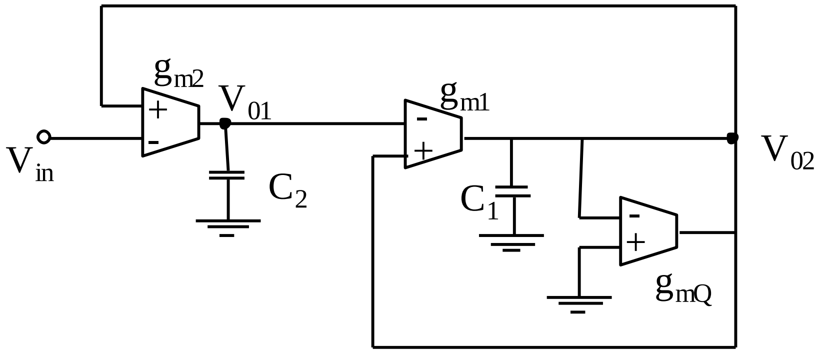
Lossy Integrator With Positive Feedback



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

Low-Frequency, High-Q OTA-C Biquad

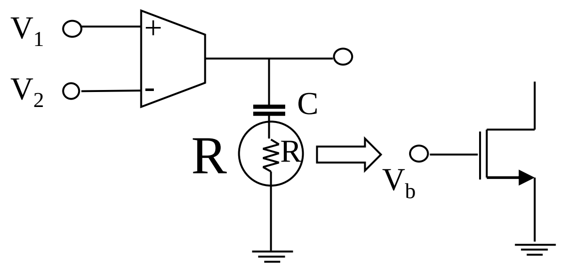




$$\frac{V_{02}}{V_{in}} = \frac{g_{m1}g_{m2}/C_1C_2}{s^2 + s\left(g_{m1} - g_{m2}\right)/C_1 + g_{m1}g_{m2}/C_1C2} = \frac{g_{m1}g_{m2}/C_1C_2}{D(s)}$$ LP

$$\frac{V_{01}}{Vin} = \frac{g_{m1}/C_2\left(s + \left(g_{m2} - g_{mQ}\right)/C_1\right)}{D(s)}$$ Resonator

OTA Excess Phase Compensation



. Phase compensation techniques: passive for integrators.

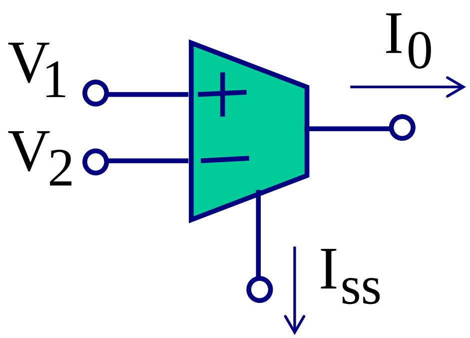
How to determine the value of RC ?

The R is implemented with a transistor operating in the triode (ohmic) region.

The zero generated by the RC should cancel the dominant pole of Gm(s).

Active Frequency Compensation Transconductor

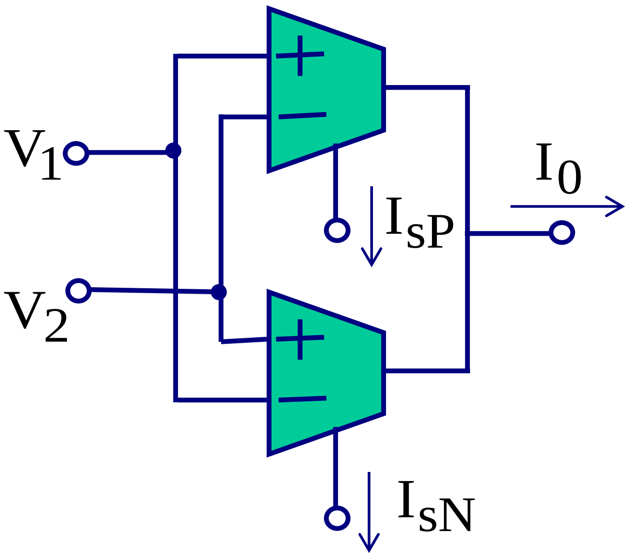
[J. Ramirez-Angulo and E. Sanchez-Sinencio, "Active Compensation of Operational Transconductance Amplifier Filters Using Partial Positive Feedback," IEEE Journal of Solid-State Circuits, vol. 25, No. 4, pp. 1024-1028, August 1990]



$$I_0 = g_m (V_1 - V_2)$$

$$g_m(s) = g_{mo}\left(1 - \frac{s}{\omega}\right)$$

ω depends on I_{ss}



$$I_0 = (g_{mp}(s) - g_{mN}(s))\Delta V = g_{meff}(s)\Delta V$$

$$g_{meff}(s) = g_{meffo}\left[1 - \frac{s}{\omega_{eff}}\right]$$

$$g_{meffo} = g_{mPo} - g_{mNo} \quad , \quad \omega_{eff} = \frac{g_{meffo}}{\frac{g_{mPo}}{\omega_p} - \frac{g_{mNo}}{\omega_N}}$$

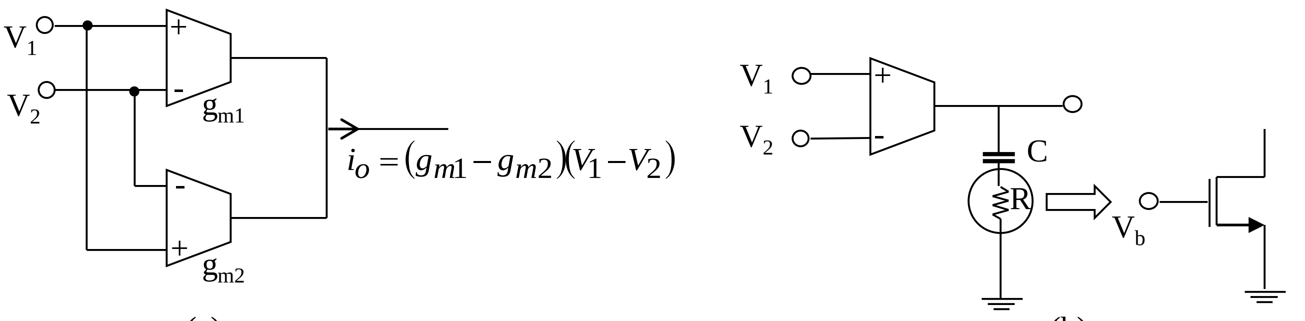
It is possible to make

$$\omega_{eff} >> \omega_p, \omega_N$$

Phase compensation techniques

In a Biquad:

$$Q_a = \frac{Q}{1 + 2\left(\frac{1}{A_{vo}} - \phi_E\right)Q}$$



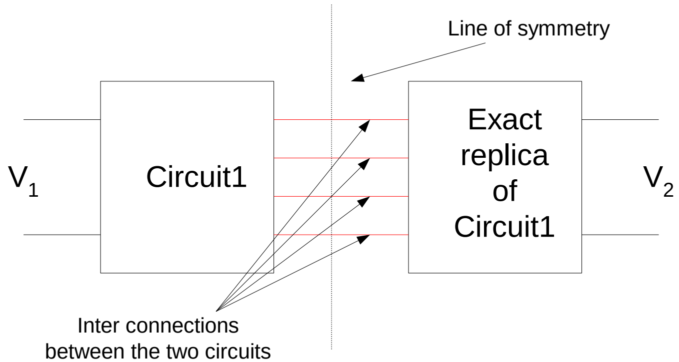
(a) (b)

(a) active; and (b) passive for integrators.

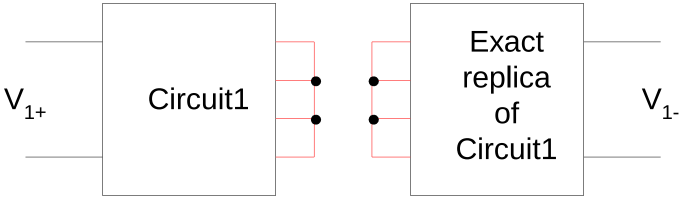
Recall that

$$g_m = \frac{g_{mo}}{1 + s/\omega_p} \cong g_{mo}\left(1 - s/\omega_p\right) \cong g_{mo}\, e^{-s/\omega_p}$$

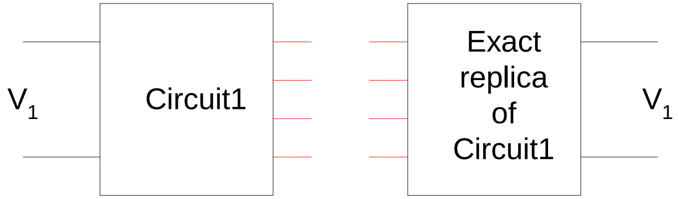
Behavior of symmetric circuits



An example of fully symmetric circuit

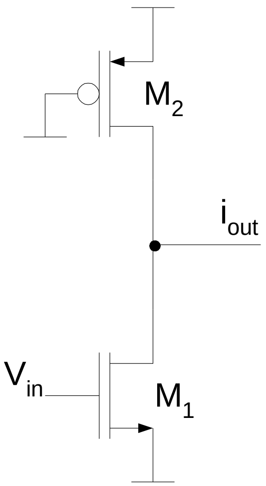


Equivalent circuit for fully differential input

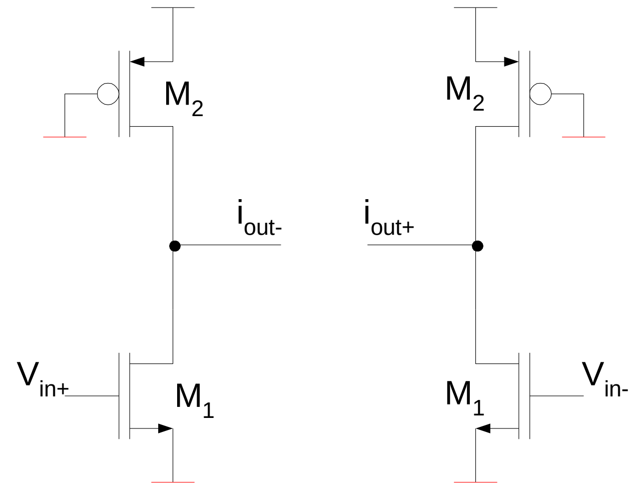


Equivalent circuit for common mode input

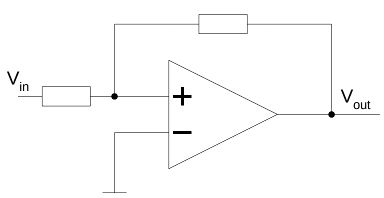
Derivation of CMFF OTA

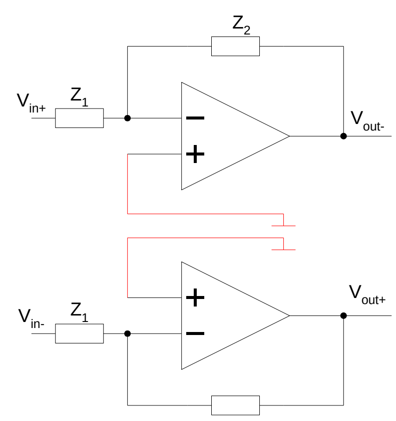


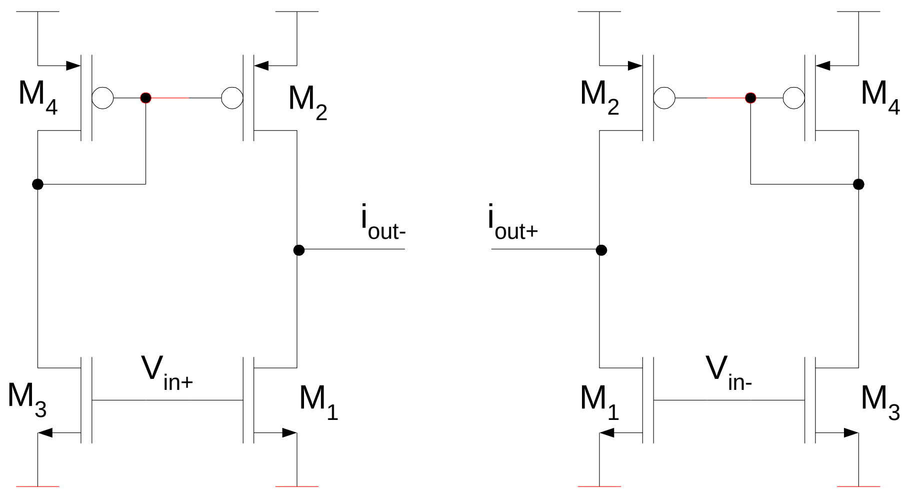
Single ended OTA circuit



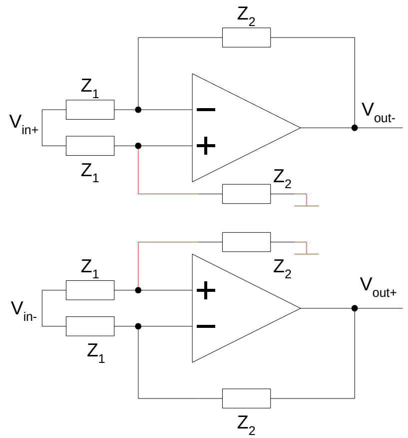
Circuit of OTA for differential input





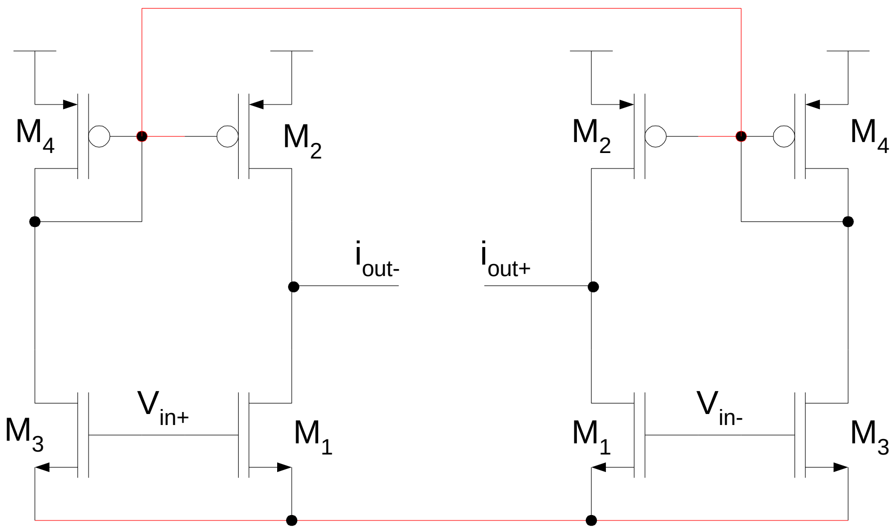


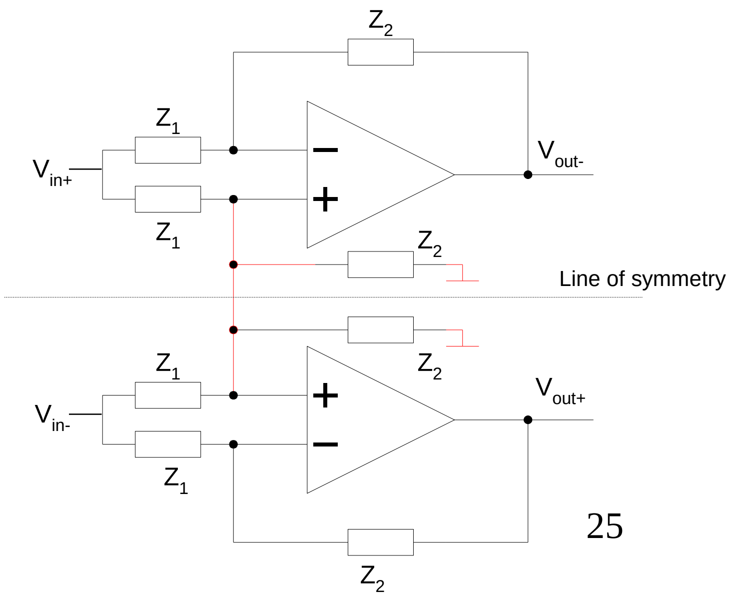
Circuit of OTA for common mode signals



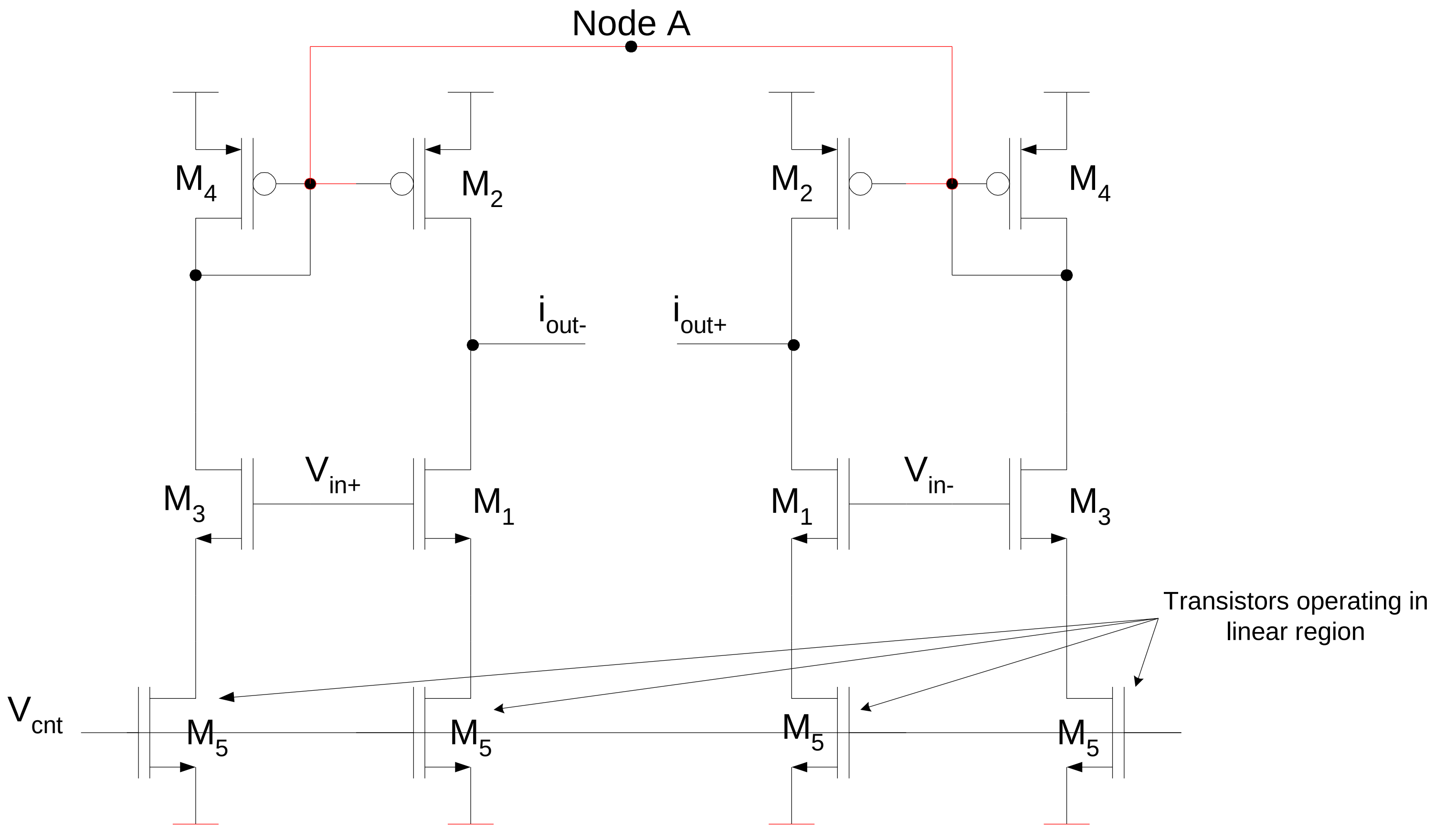
Note.- Independent trajectories, poor CMRR

Fully-balanced, fully-symmetric CMFF OTA



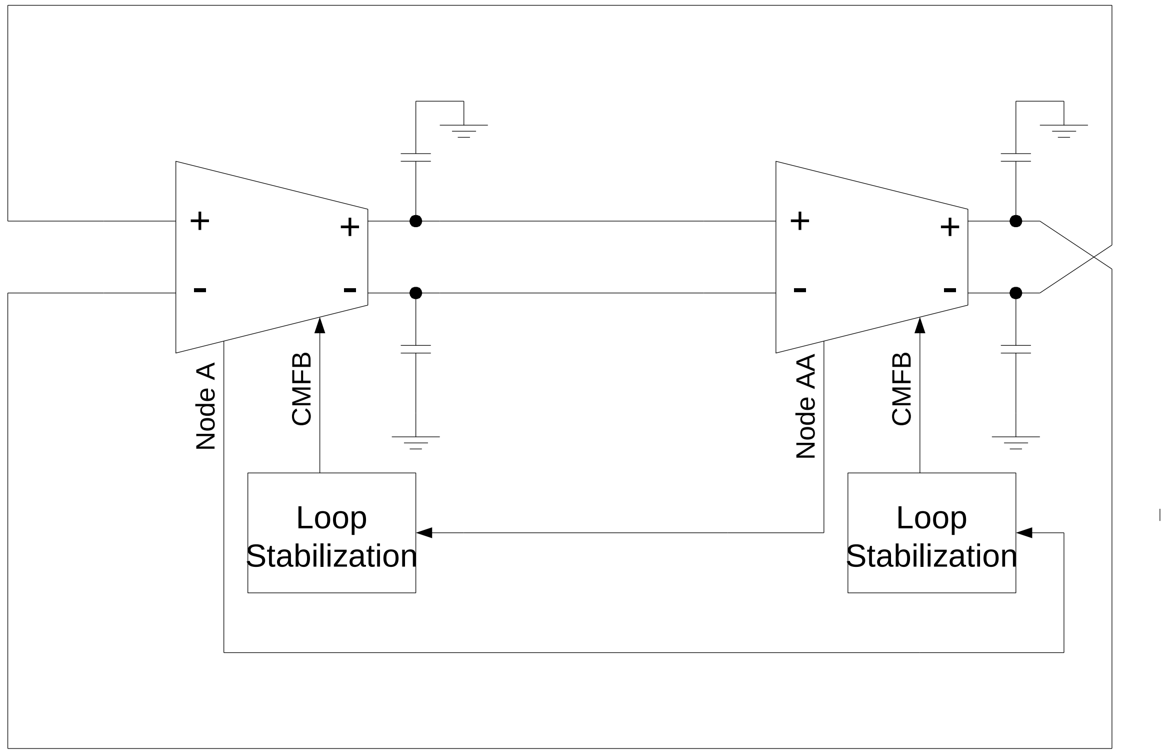


OTA with improved flexibility

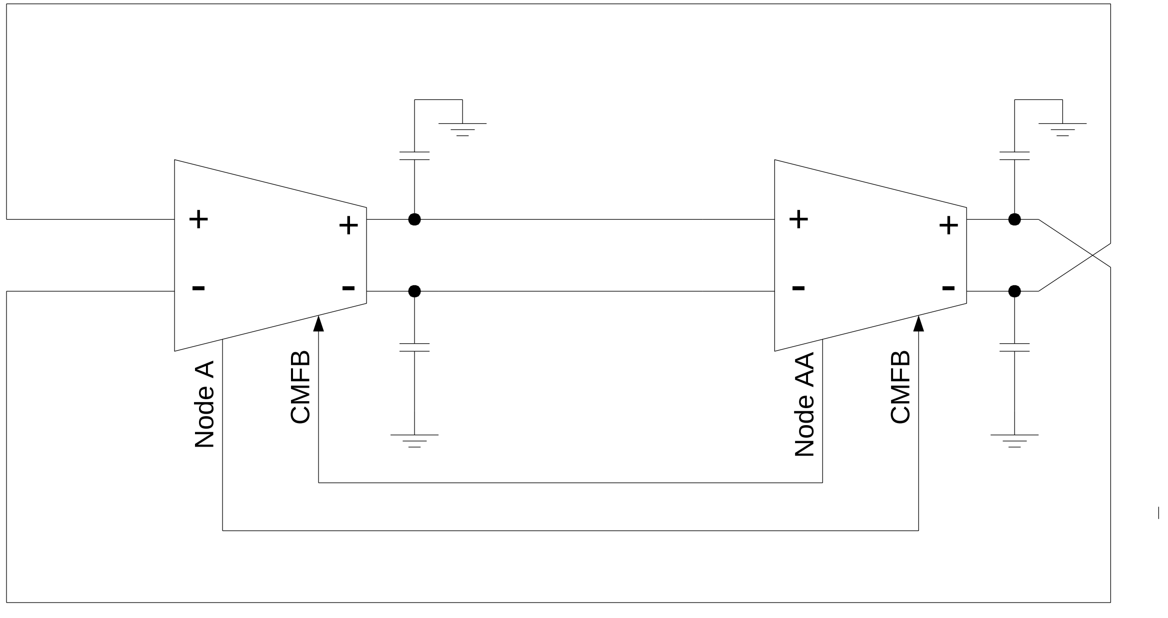


Fully-balanced, fully-symmetric, pseudo differential CMFF OTA

Two integrator loop

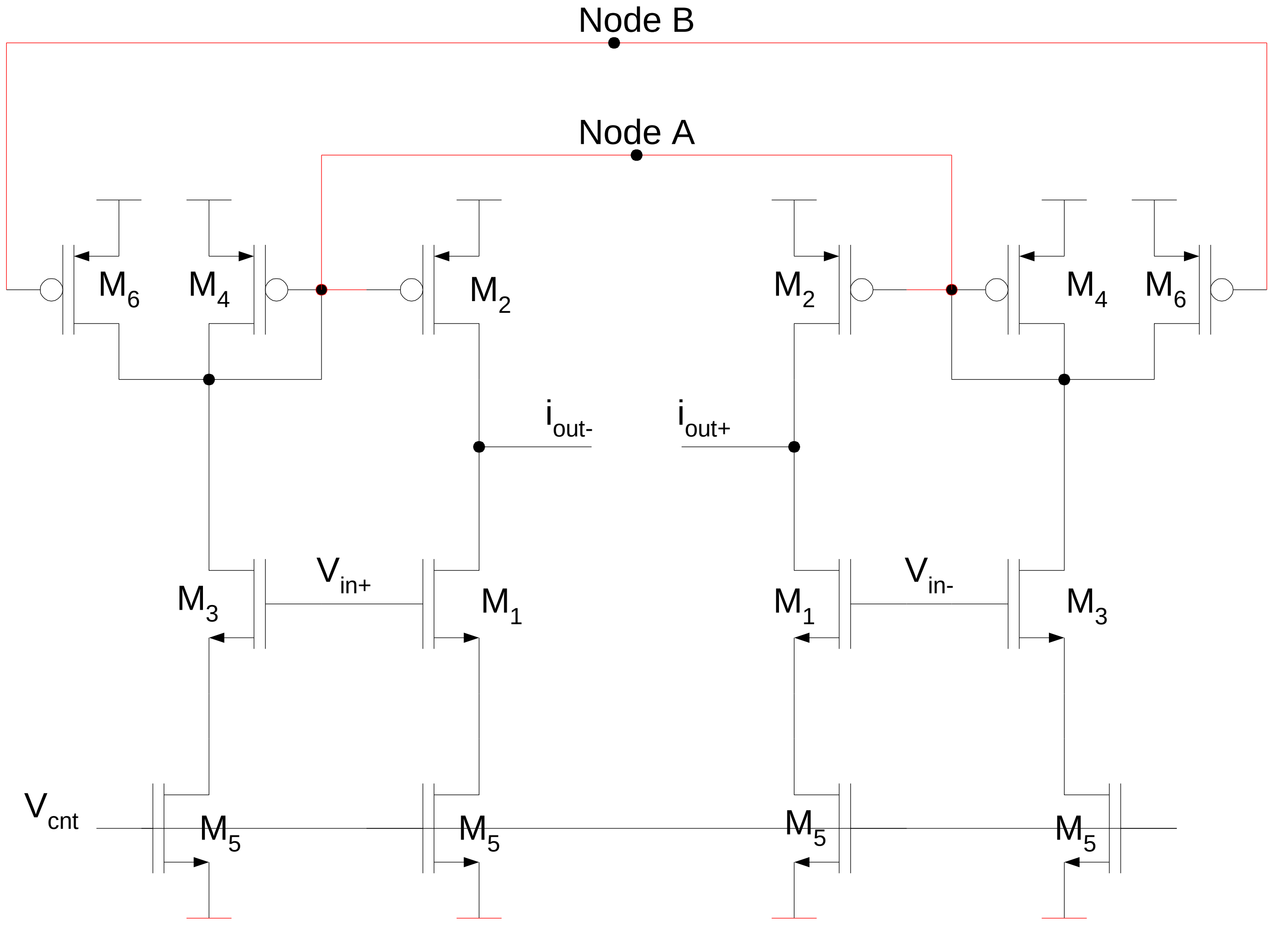


Two integrator loop using CMFF OTA



Two integrator loop using CMFF+CMFB OTA

(CMFF + CMFB) OTA



Fully-balanced, fully-symmetric, pseudo differential (CMFF+CMFB) OTA

Characteristics of the OTA

- Let the total capacitance at 'node A' be C_{int}
- Let the capacitance used in two integrator loop be C_{ext}
- Effective transconductance= $$g_{m1eff} = \frac{g_{m1}}{1 + \frac{g_{m1}}{g_{ds5}}}$$
- CMFB loop gain = $$\frac{g_{m1eff} \cdot g_{m6} \cdot g_{m2}}{g_{m4} \cdot g_{m4} \cdot g_{ds2}} \cdot \frac{1}{\left(1 + \frac{sC_{int}}{g_{m4}}\right)^2 \left(1 + \frac{sC_{ext}}{g_{ds2}}\right)}$$

- Gain (I_o/V_i) = $$g_{ds5}\left\{2-\left(1+\frac{4\beta_1}{g_{ds5}}\left(V_G-V_T\right)\right)^{-\frac{1}{2}}\right\}$$

- Gain(I_o/V_i^3) = $$\frac{12\beta_1^2}{g_{ds5}}\left[1+\frac{4\beta_1}{g_{ds5}}\left(V_G-V_T\right)\right]^{-\frac{5}{2}}=\frac{12\beta_1^2 g_{ds5}^{3/2}}{\left[g_{ds5}+4\beta_1\left(V_G-V_T\right)\right]^{5/2}}$$

- To improve linearity, use larger resistor for source degeneration.

- Differential gain= $\dfrac{g_{m1,eff}}{g_{ds2}}$

- Common mode gain = $\left[1-\dfrac{1}{1+\dfrac{sC_{\text{int}}}{g_{m4}}}\right]\times\dfrac{g_{m1,eff}}{g_{ds2}}+\dfrac{g_{m1,eff}}{g_{m4}}$

- $\text{CMRR}_{(\text{DC})}$ = $\dfrac{g_{m4}}{g_{ds2}}$

- Gain from +ve supply=1
- Gain from -ve supply= $\left[1-\frac{1}{1+\frac{sC_{\text{int}}}{g_{m4}}}\right]\times\frac{g_{m1,eff}}{g_{ds2}}+\frac{g_{m1,eff}}{g_{m4}}$

- Gain from VSS is less than gain from VDD. So, output should be measured wrt VDD
- PSRR is same as CMRR

Output noise current=

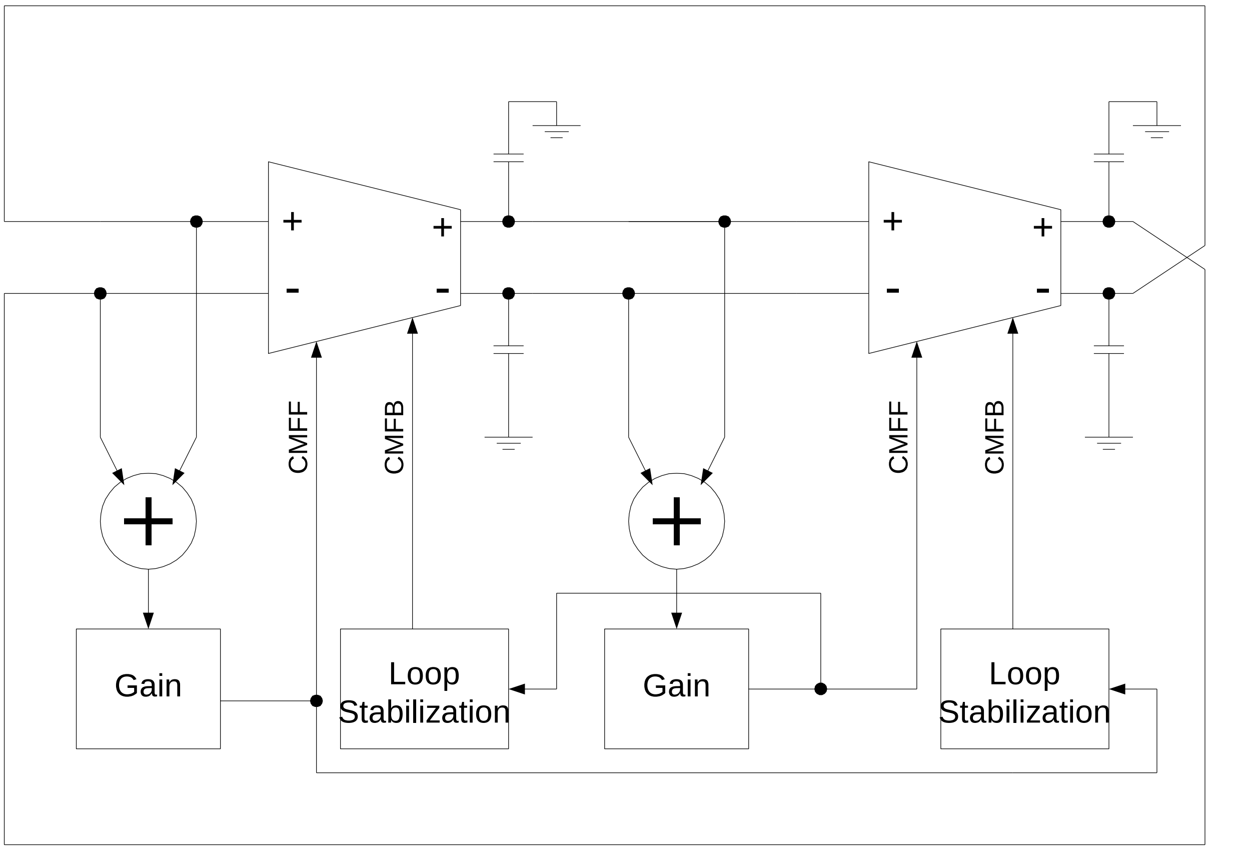
$$i_{n1}^2 \times \left(\frac{1}{1+\frac{g_{m1}}{g_{ds5}}} \right)^2 + 4KTg_{ds5} + 4KTg_{ds5} \times \left(\frac{g_{m2}}{g_{m4}\left(1+\frac{sC_{\text{int}}}{g_{m4}}\right)} \right)^2$$

$$+ i_{n2}^2 + i_{n4}^2 \times \left(\frac{g_{m2}}{g_{m4}} \right)^2 + i_{n6}^2 \times \left(\frac{g_{m2}}{g_{m4}\left(1+\frac{sC_{\text{int}}}{g_{m4}}\right)} \right)^2$$

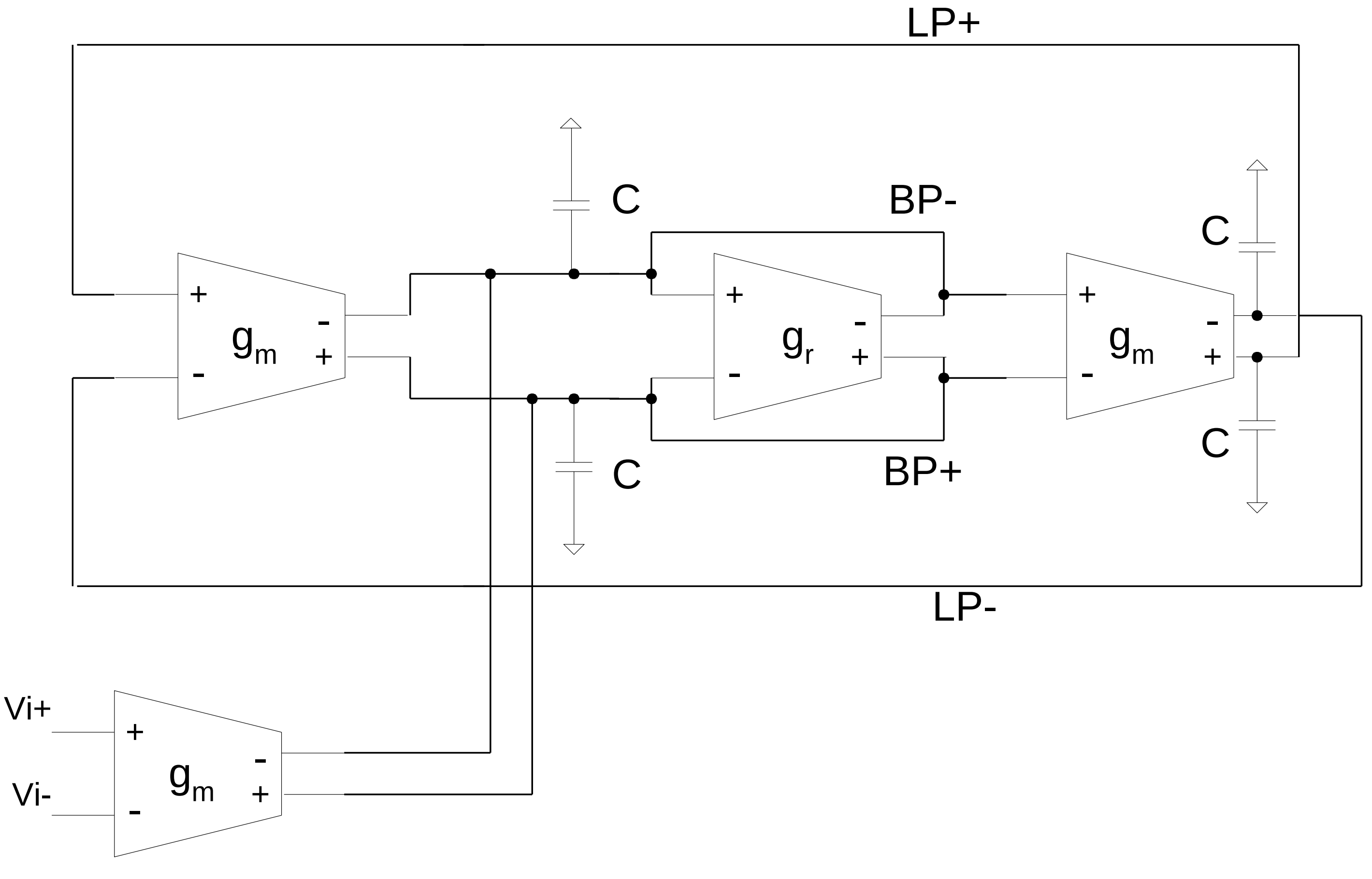
Simplified noise expression

$$KT\left\{\frac{8g_{m1}}{3\left(1+\frac{g_{m1}}{g_{ds5}}\right)^2}+4g_{ds5}+\frac{4g_{ds5}}{\left(1+\frac{sC_{\mathrm{int}}}{g_{m4}}\right)^2}+\frac{8g_{m2}}{3}+\frac{8g_{m2}}{3\left(1+\frac{sC_{\mathrm{int}}}{g_{m4}}\right)^2}\right\}$$

Two integrator loop



Band pass filter



Design of a new high frequency OTA and a Filter Tuning Scheme

Praveen Kallam

Advisor: Dr. E. Sanchez Sinencio

How to build a filter

- OpAmps - Low frequency, high linearity
- OTAs - Medium high frequencies, medium linearity
- Passive components - High frequency
- Transmission lines - Extremely high frequency

NMOS *VS* PMOS

	NMOS	PMOS
Speed	Faster	Slower
Device noise	Low thermal noise	Low flicker noise
Linearity	Bulk effect degrades linearity	No bulk effect
Substrate noise	Higher due to common substrate	Can be better shielded

Advantages of differential Circuits

- Double the signal swings
- Better power supply and substrate noise rejection
- Higher output impedance with conductance cancellation schemes
- Better linearity due to cancellation of even harmonics
- Partial cancellation of systematic errors using layout techniques
- Availability of already inverted signals

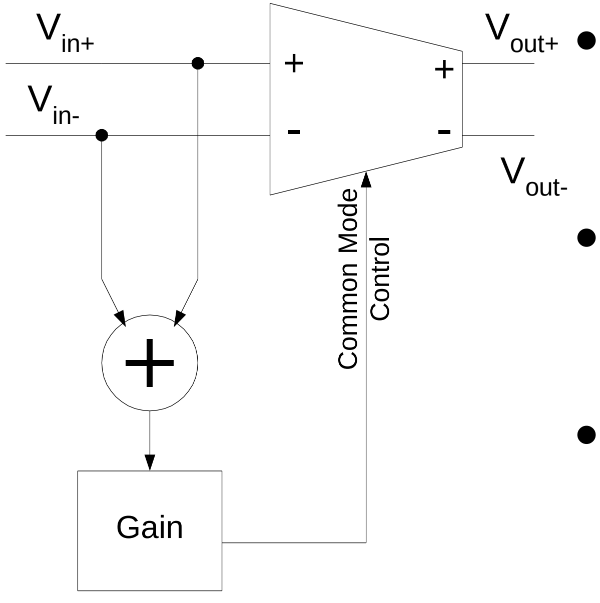
Disadvantages of differential Circuits

- Duplication of circuit requires double the area and power
- Additional circuitry to tackle common mode issues

Common mode issues

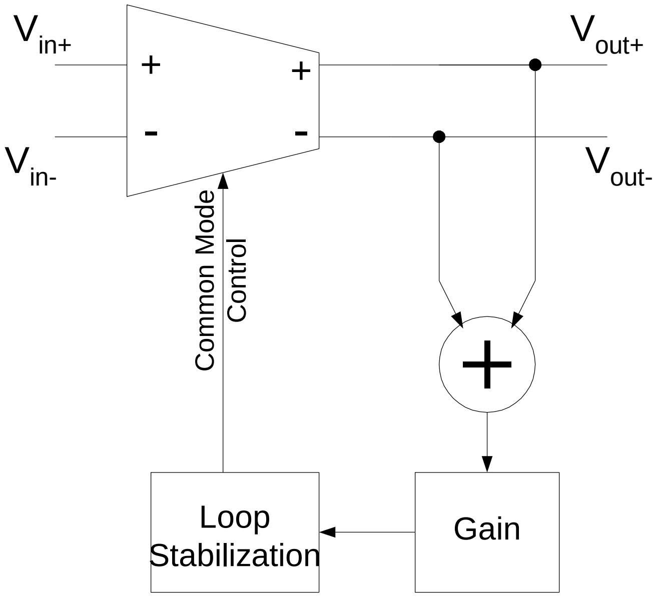
- Output DC common mode voltage should be stabilized (otherwise, the voltage may hit the rails)
- Common mode gain should be small (otherwise, positive feedback in a two integrator loop becomes stronger)

Common Mode Feed Forward



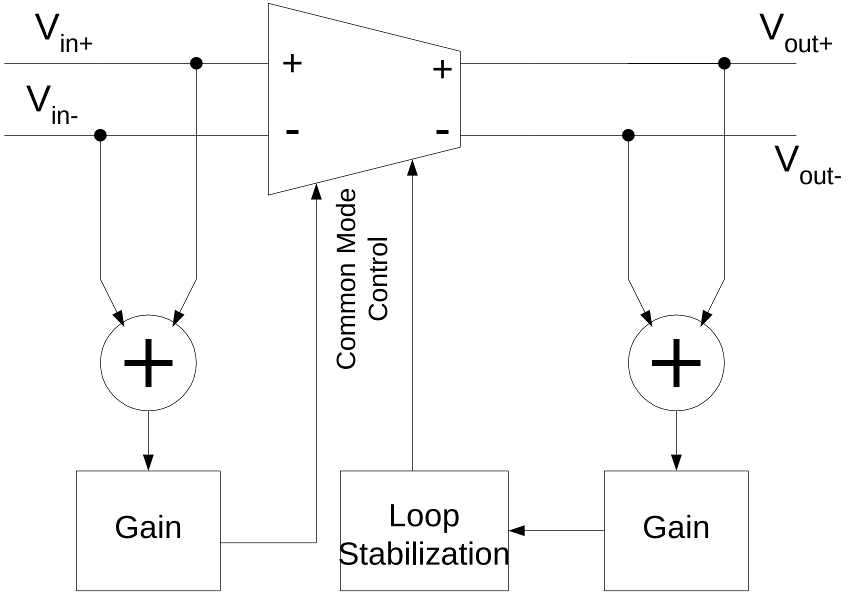
- Can decrease common mode gain even at higher frequencies
- Does not have stability problems
- Cannot stabilize the output DC voltage

Common Mode Feed Back



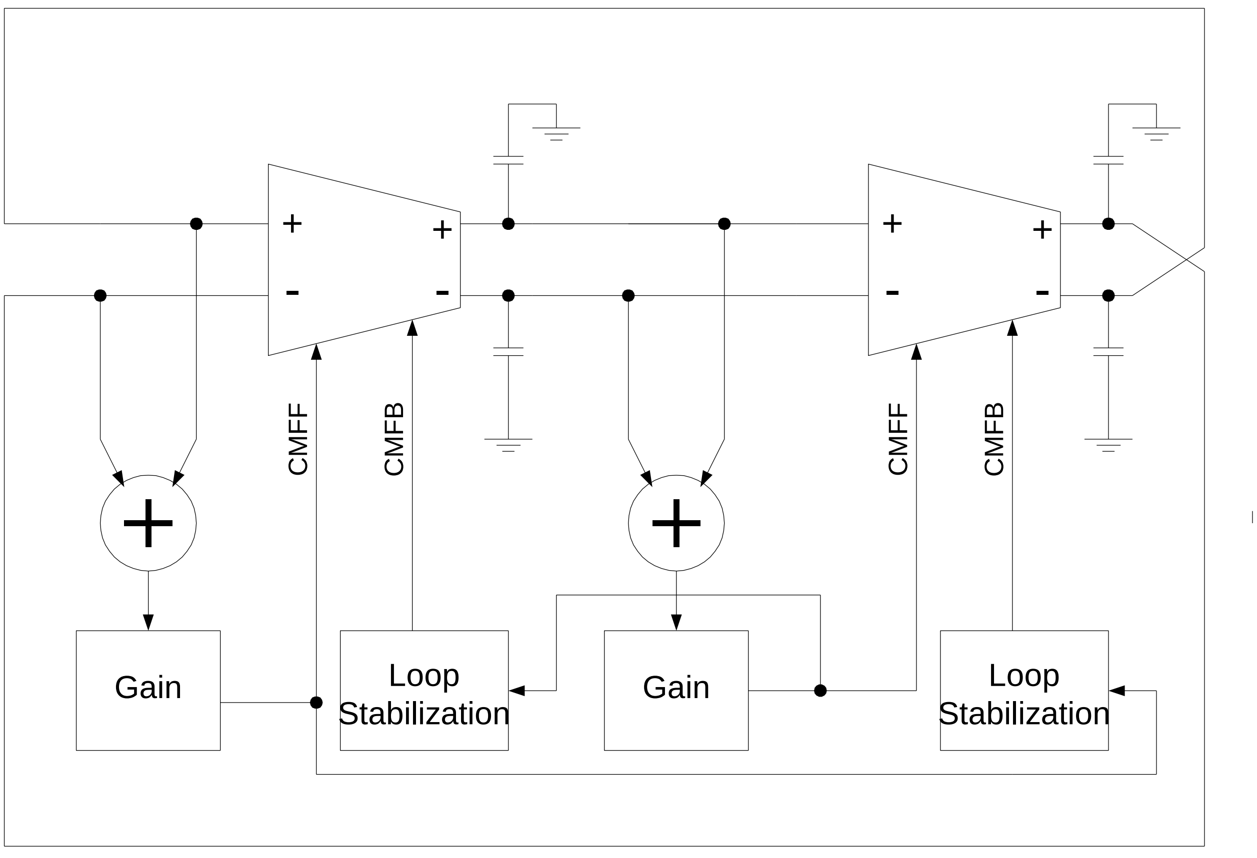
- Stabilizes output DC voltage
- Feedback stability issues make the circuit slow and bulky

CMFF + CMFB

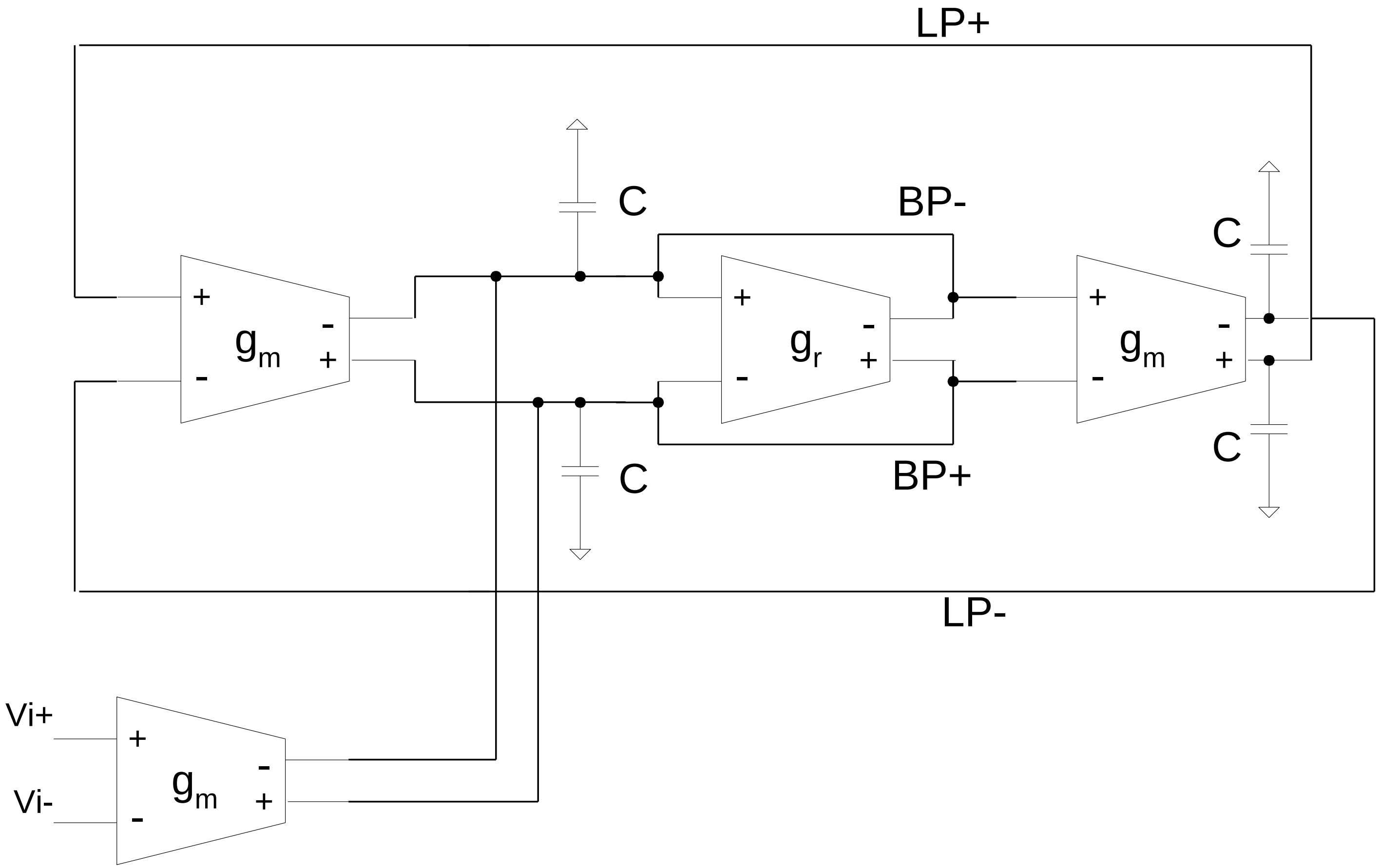


V in+
V in-
+
-
+
-
V out+
V out-
Common Mode Control
Gain
Loop Stabilization
Gain

Two integrator loop



Band pass filter



Need for tuning

- Process parameters can change by 10%
- Parameters also change with temperature and time(aging)

- Another solution for low-frequency is using Switch Capacitor filters

Methods of tuning

- Master-Slave
- Pre-tuning
- Burst tuning
- Switching between two filters

Frequency Tuning

PLL

- Most widely used scheme
- Accurate (less than 1% error is reported)
- **Square wave input reference**
- Only XOR and LPF are the additional components

- Usually used only for filters with Q>10
- Large area overhead

VCF, VCO, Single OTA, Peak detect, adaptive….

Q tuning

Modified LMS

- Accurate
- **Square wave input**
- Independent of frequency tuning

- Not very robust
- Large area overhead

The most accurate scheme so far

- Stevenson, J.M.; Sanchez-Sinencio, E "An accurate quality factor tuning scheme for IF and high-Q continuous-time filters". Solid-State Circuits, IEEE Journal of Volume: 33 12 , Dec. 1998 , Page(s): 1970 -1978

- Combines Master-Slave, PLL and modified LMS
- Less than 1% error in both f-tuning and Q-tuning

LMS Algorithm Derivation.- The mean square error (MSE) is defined as $E(t)=0.5[e(t)]^2 = 0.5[d(t)-y(t)]^2$

where d(t) is the desired output signal, and y(t) is the actual output signal. The steepest descent algorithm is defined as:

$$\frac{dW}{dt} = -\mu \frac{\partial E}{\partial W}$$

$$\frac{dW}{dt} = -\mu \frac{\partial E}{\partial y} \frac{\partial y}{\partial W}$$

$$\frac{dW}{dt} = -\mu \frac{\partial[0.5\{d(t) - y(t)\}^2]}{\partial y} \frac{\partial y}{\partial W}$$

$$\frac{dW}{dt} = \mu[d(t) - y(t)] \frac{\partial y(t)}{\partial W}$$

$$\dot{W} = \mu[d(t) - y(t)]G(t) = \mu e(t)G(t)$$

Linear System case.

$$y(t) = \sum_{i=0}^{n} w_i x_i ,$$

where:

x_i is the input signal.

Therefore:

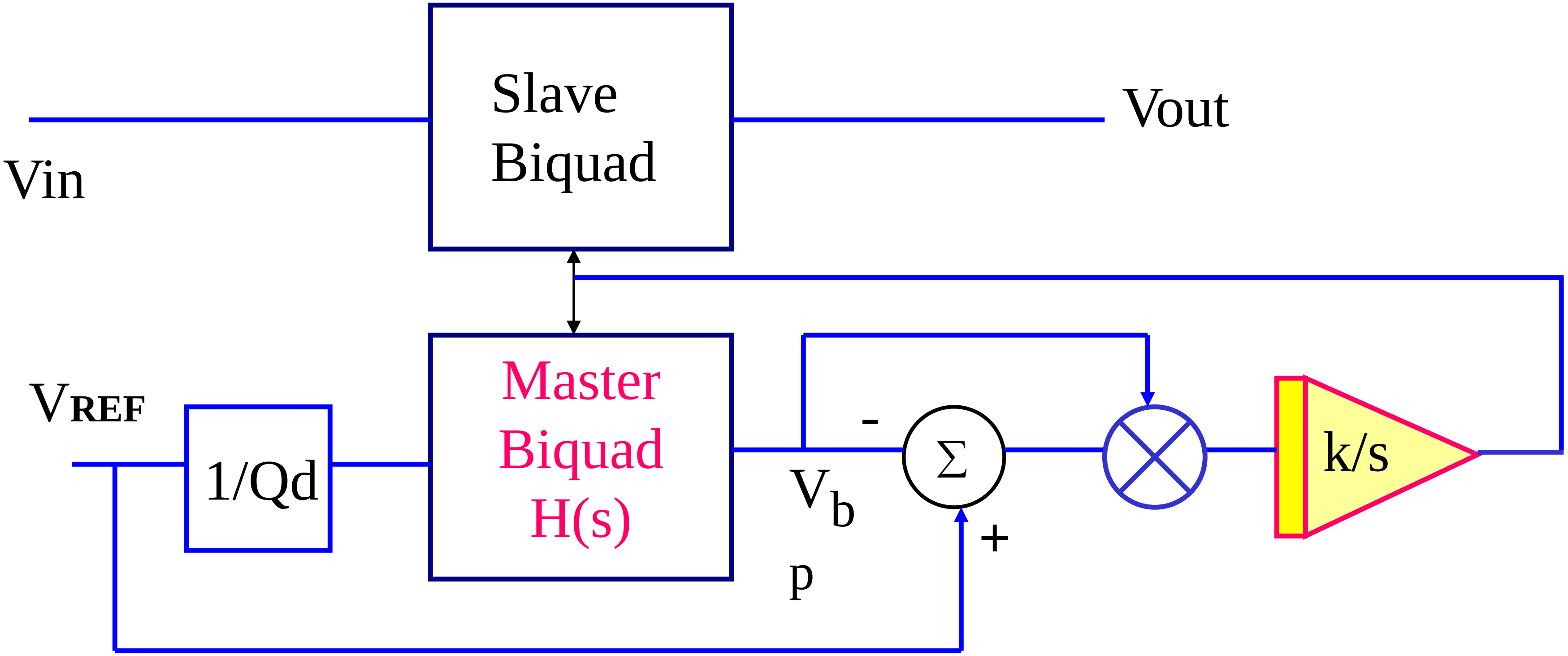
$$\frac{dW}{dt} = \mu[d(t) - y(t)] \frac{\partial \sum_{i=0}^{n} w_i x_i ,}{\partial W} = \mu[d(t) - y(t)] x_i$$

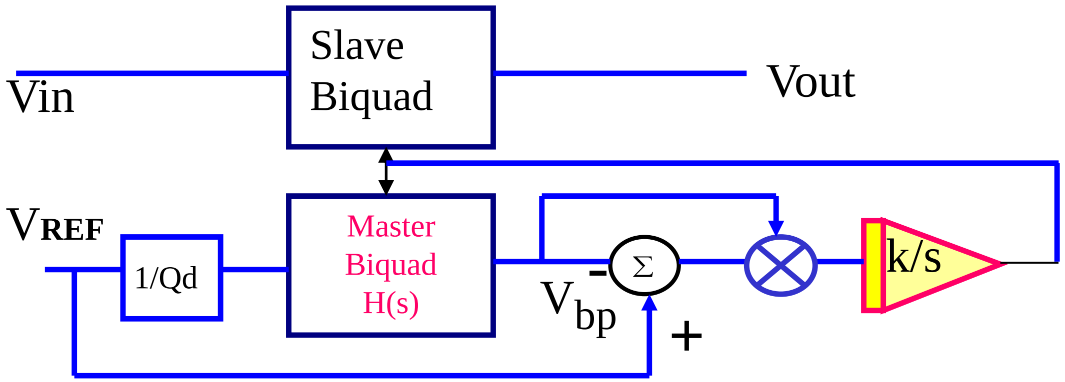
$$\dot{W} = \mu e(t) x_{xi}$$

Adaptive LMS Algorithm

$$\dot{w}_i = \mu\left[d(t) - y(t)\right]g_i(t)$$

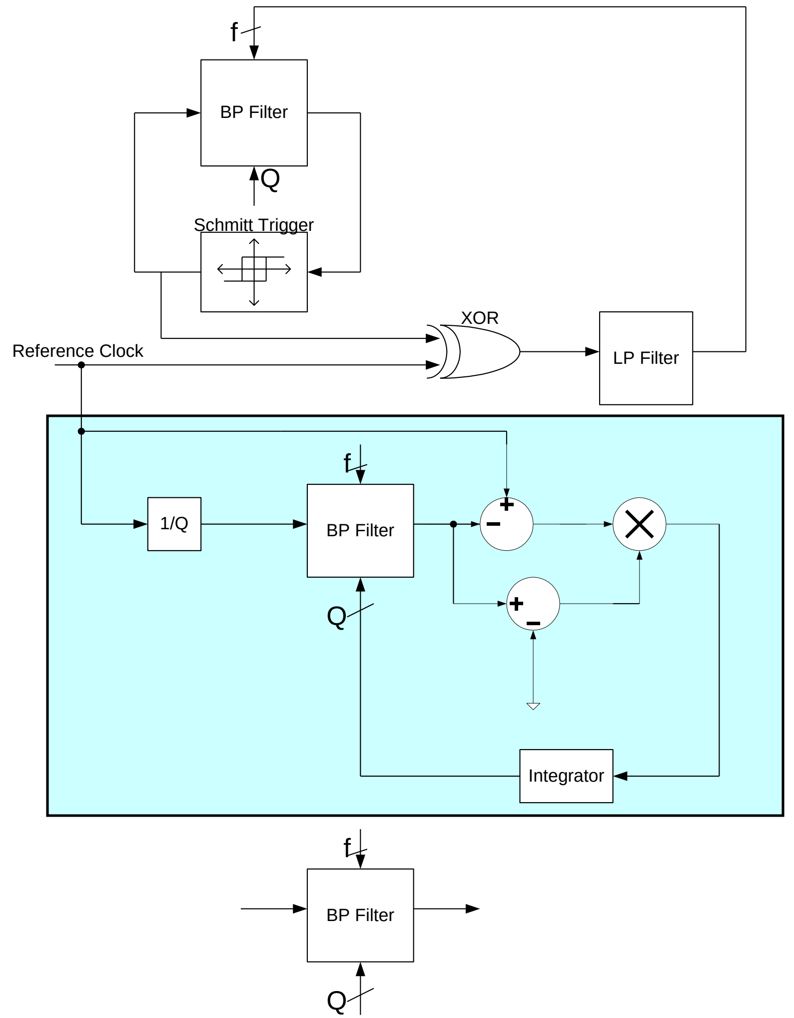
Where is the tuning signal, d(t) is the desired response, y(t) is the actual response, and g_i (t) is the gradient signal (that is the direction of tuning.





Block Diagram Solution

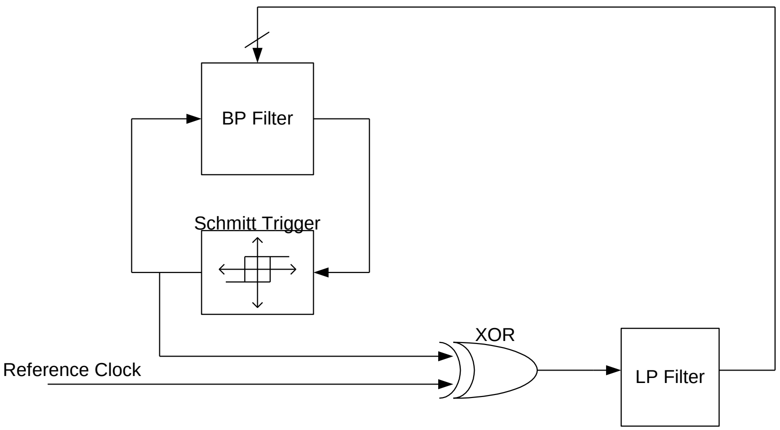
The tuning scheme implemented before

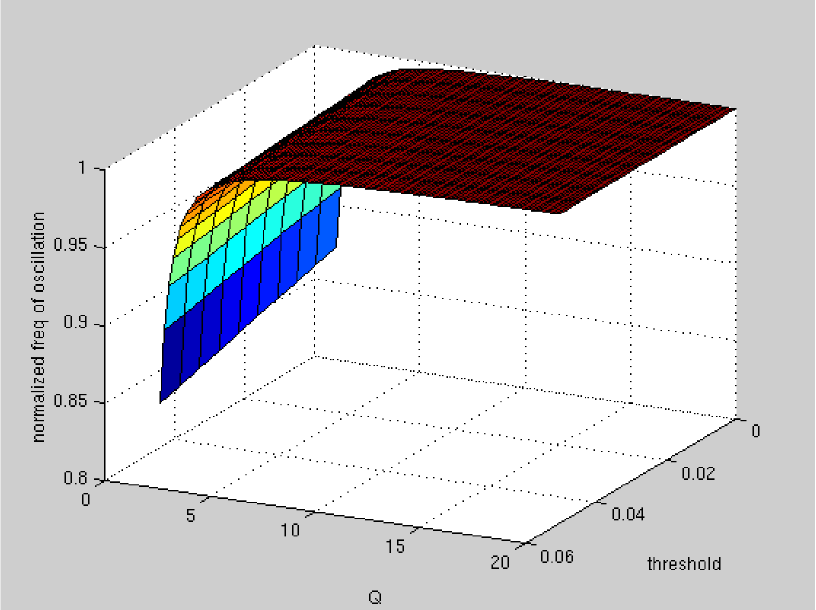


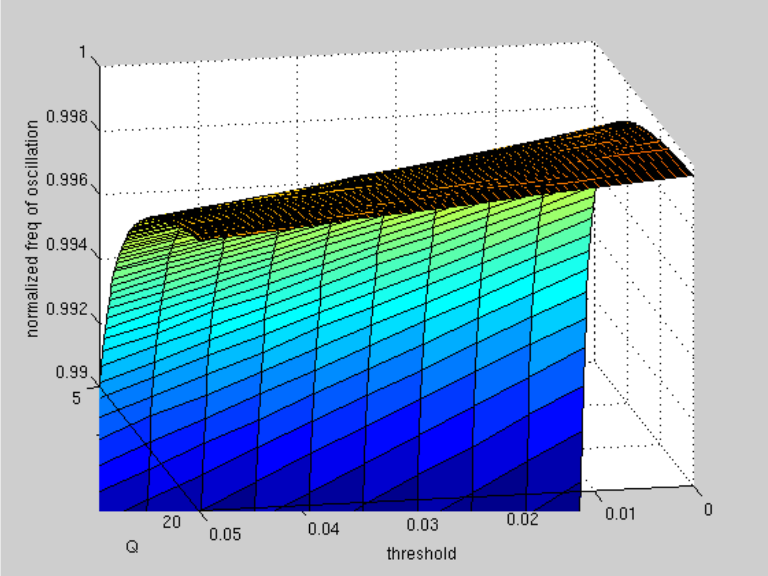
Problems in the previous scheme

- Large area overhead (may run into matching problems)
- Power hungry
- Not very robust (very low offsets required.)
- Looses accuracy at low Qs(<10) and very high Qs (~100)
- Applies only to Band-Pass filters

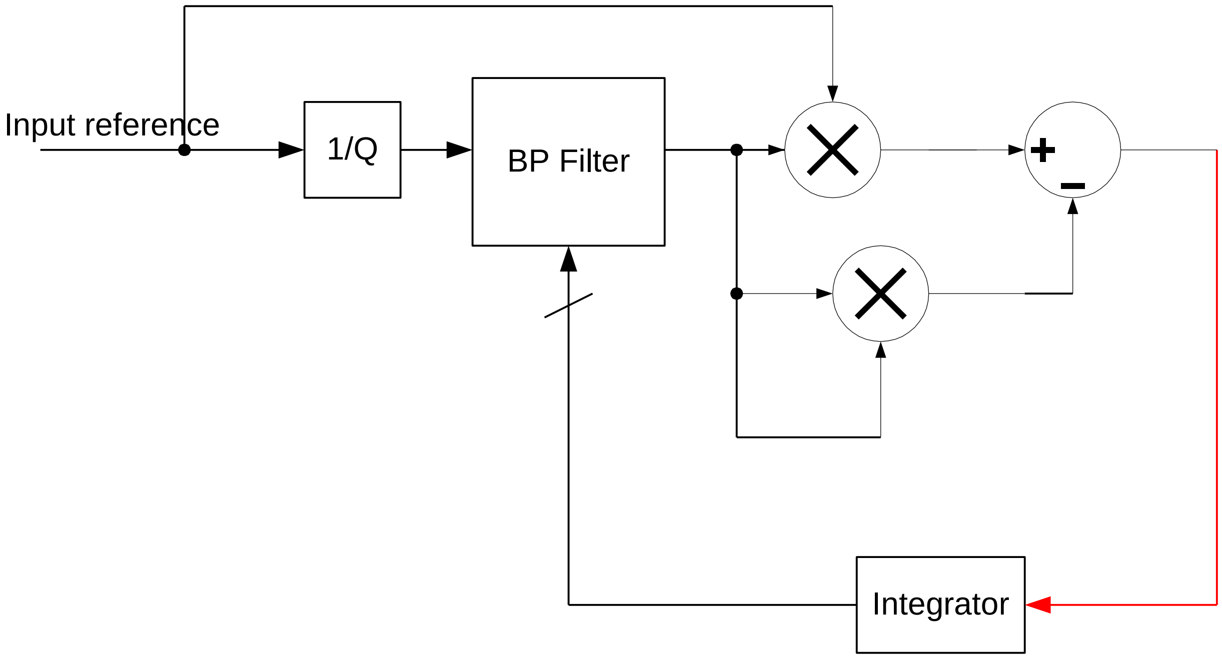
PLL


BP Filter
Schmitt Trigger
XOR
Reference Clock
LP Filter


normalized freq of oscillation
1
0.95
0.9
0.85
0.8
0
5
10
15
20
Q
0
0.02
0.04
0.06
threshold


normalized freq of oscillation
1
0.998
0.996
0.994
0.992
0.99
5
20
Q
0.05
0.04
0.03
0.02
0.01
0
threshold

Proposed Q-tuning scheme



New implementation of modified-LMS Q-tuning scheme

Tuning is independent of the shape of reference waveform

$$V_o(t) = A \times \frac{Q_a}{Q_D} \times G \times \sin\{wt + \theta\} \quad \Rightarrow \quad V_o(t) = A \times \frac{Q_a}{Q_D} \times \cos(\theta) \times \sin(wt + \theta)$$

$$G = \left| \frac{jw / w_a Q_a}{1 + jw / w_a Q_a - w^2 / w_a^2} \right| \qquad \theta = Arg\left(\frac{jw / w_a Q_a}{1 + jw / w_a Q_a - w^2 / w_a^2} \right)$$

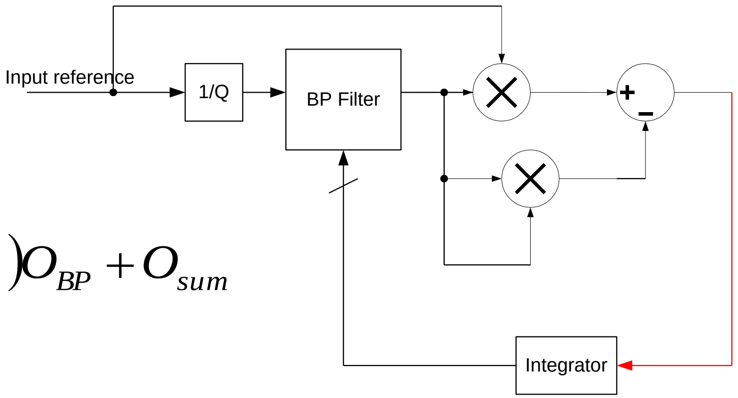
$$V_i(t) = \sum_i A_i \sin(w_i t) \quad \Rightarrow \quad V_o(t) = \sum_i A_i \frac{Q_a}{Q_D} \cos(\theta_i) \sin(w_i t + \theta_i)$$

When this input and output is processed by the tuning scheme,

$$\left\{ \frac{Q_a}{Q_D} \sum_i A_i^2 \cos^2(\theta) \; - \; \left(\frac{Q_a}{Q_D} \right)^2 \sum_i A_i^2 \cos^2(\theta) \right\} \; = \; 0$$

Improved Offset performance

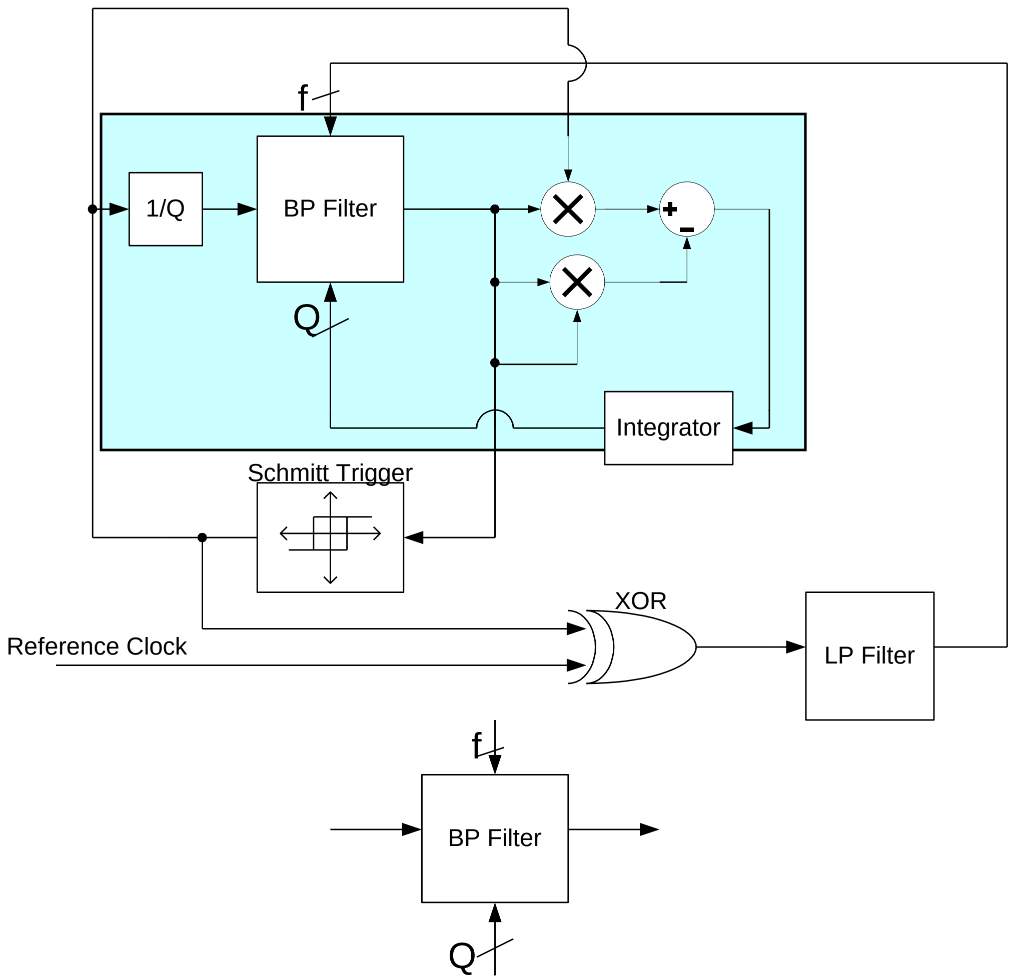
Previous offset = $G_{mul}G_{sum}^{2}(O_{in}-O_{BP})O_{BP}+G_{mul}G_{sum}O_{in}O_{sum}+G_{mul}O_{sum}^{2}+O_{mul}$



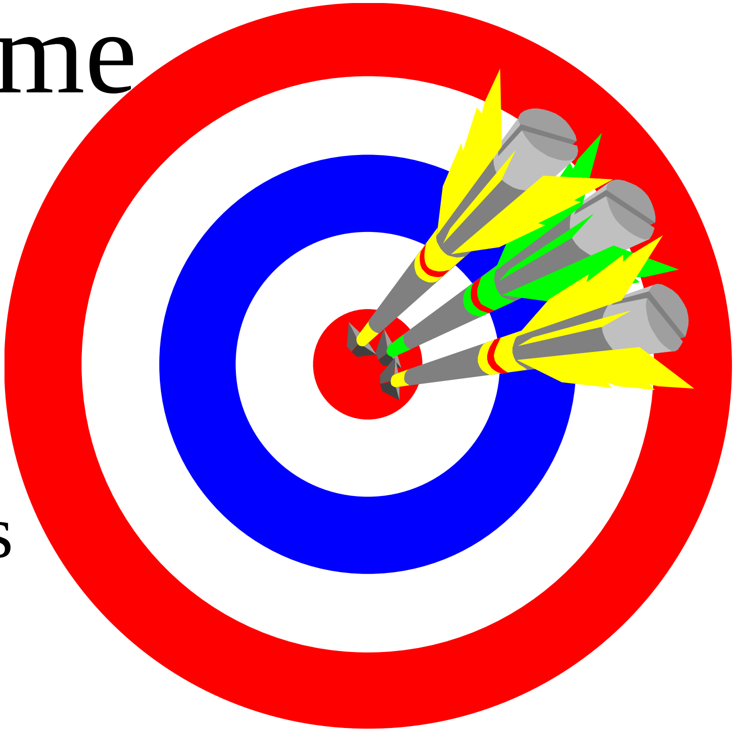
Present Offset = $G_{mul}G_{sum}(O_{in}-O_{BP})O_{BP}+O_{sum}$

- Reduced offset => improved accuracy

The new tuning scheme



Improvements over the previous tuning scheme



- Area overhead decreased

  (Previous scheme => 2 extra filters

  New scheme => 1 extra filter)

- Eases the matching restrictions

  (Previous tuning scheme => match 3 filters

  New tuning scheme => match 2 filters)

- Improves accuracy of tuning

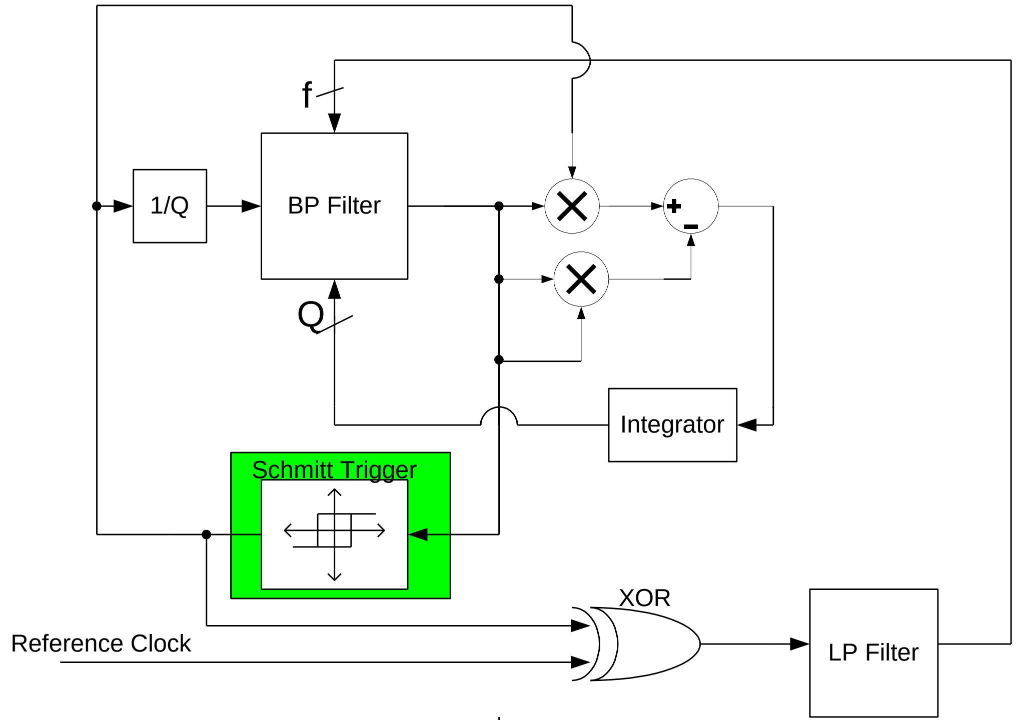
  (New tuning scheme is more tolerant to offsets than the previous one)

Circuits to be designed

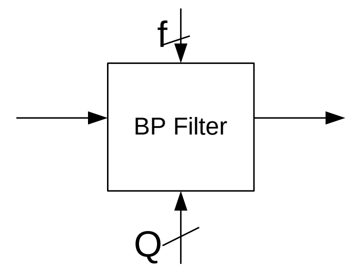
- Comparator
- Attenuator
- Multiplier
- LPF → outside the IC using Opamp
- Differential difference adder
- Integrator → outside the IC using Opamp

(Both macro model & transistor level are used in simulations for the OpAmp)

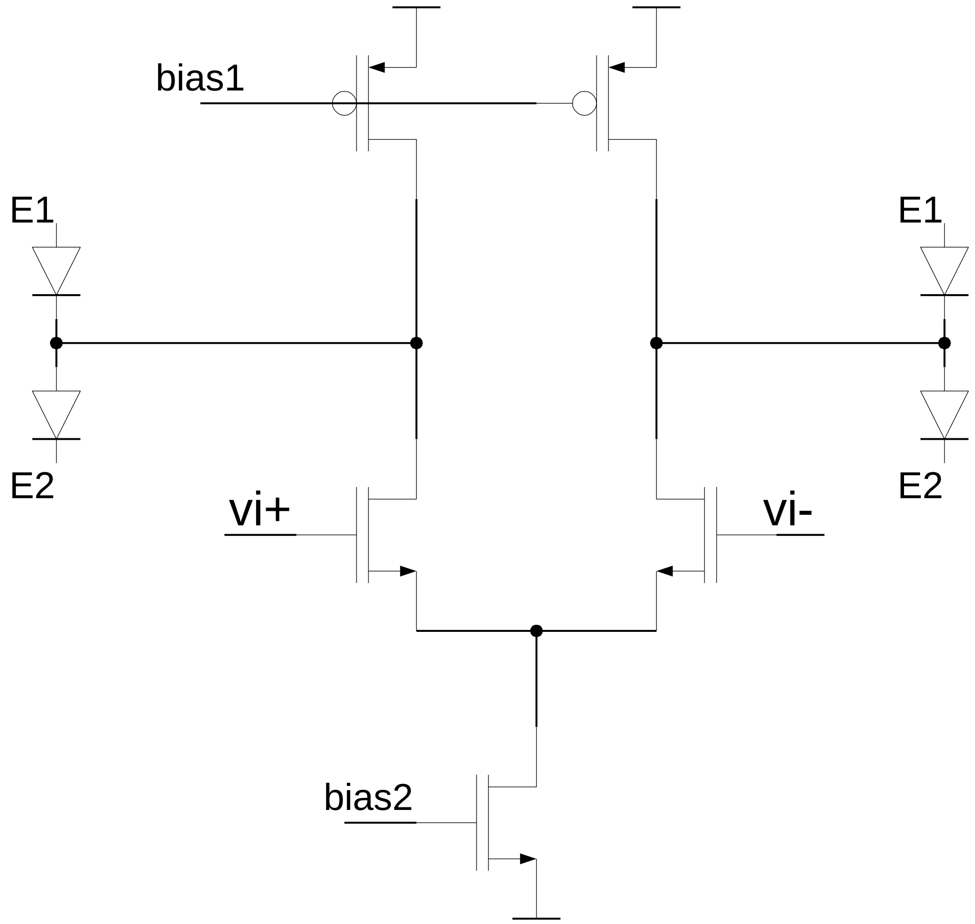
Comparator



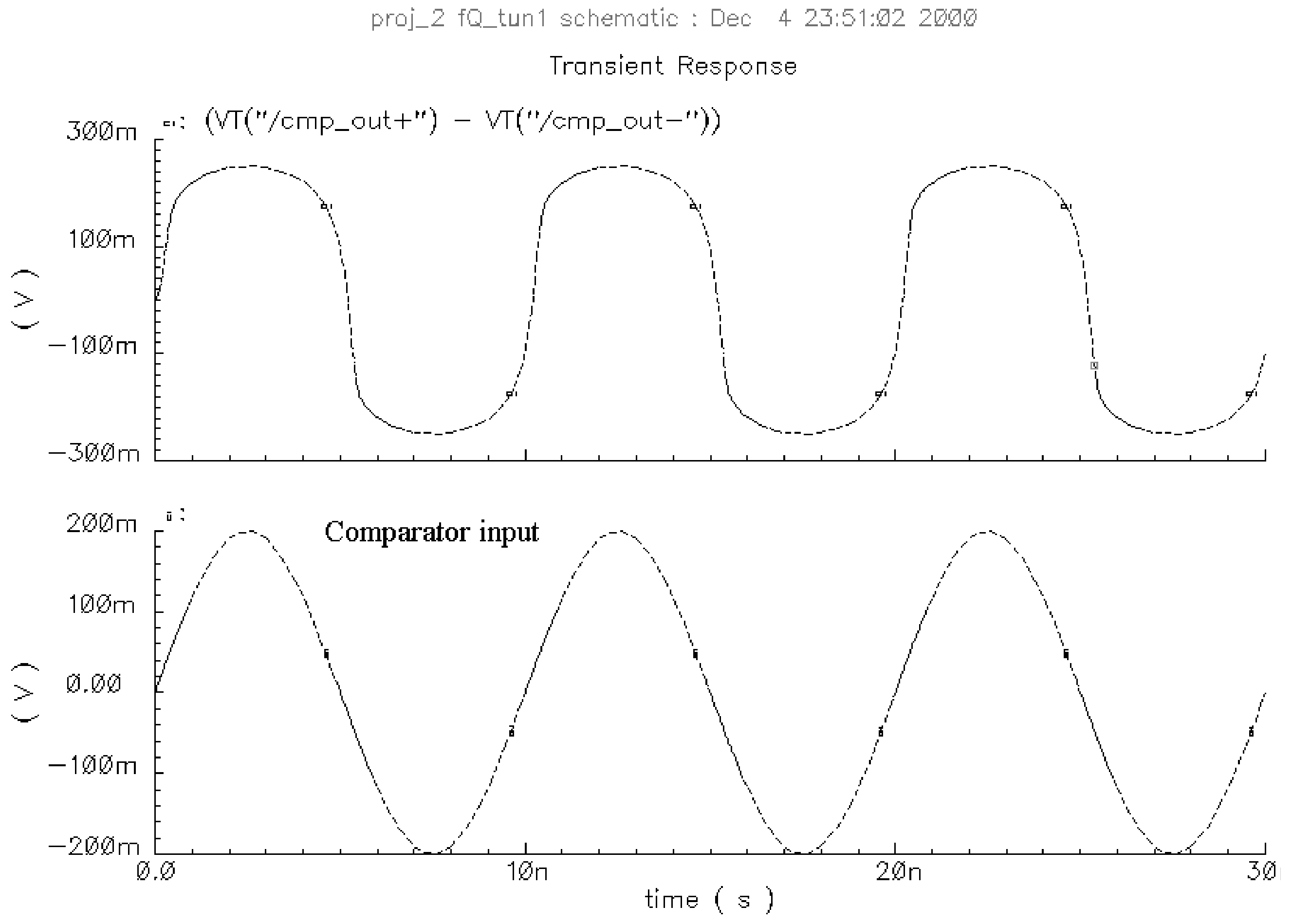
- **Non-linear amplifier**
  - **Gain should be as close to unity to improve THD**
  - **If less than unity, no oscillations**
- **Rate of change of gain wrt input should be high (should be very non-linear)**
  - **cannot use complex circuits**
  - **DIODE**



Circuit of differential comparator

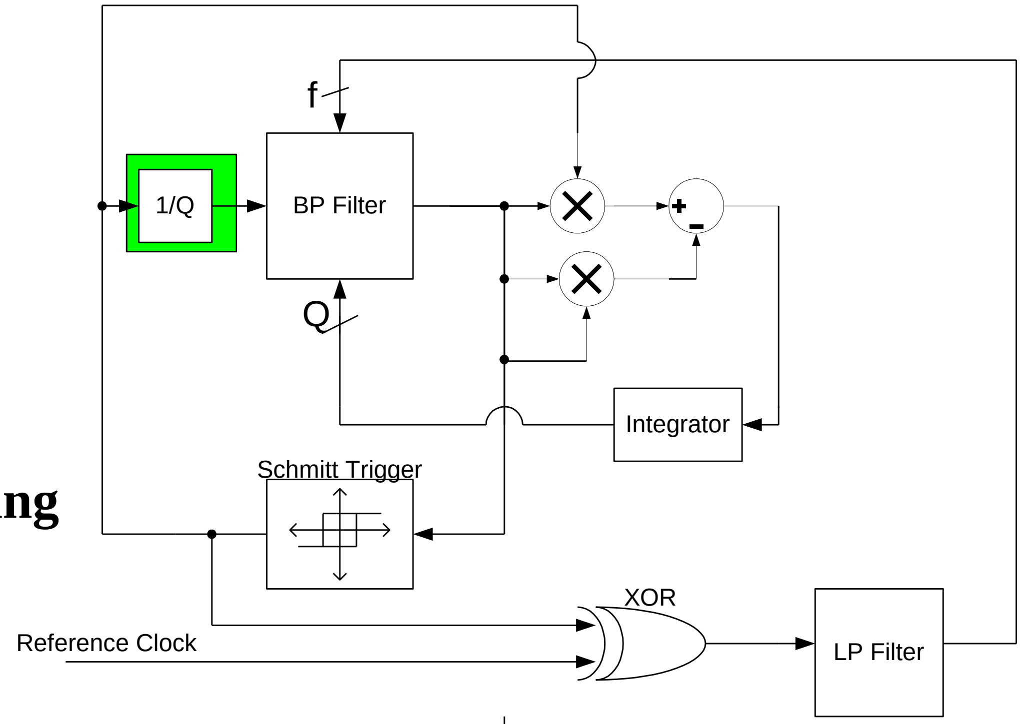


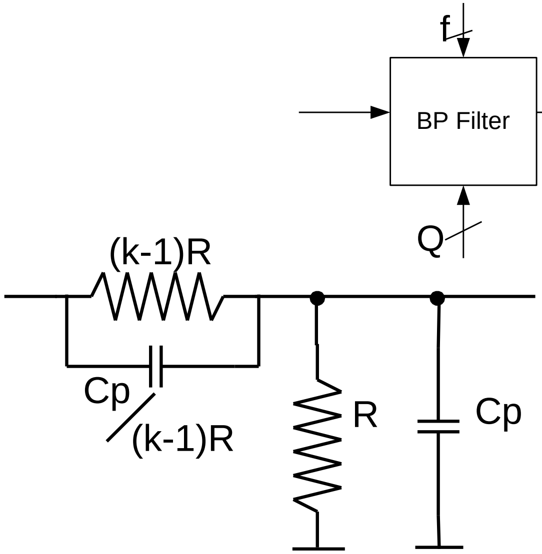
Comparator characteristics



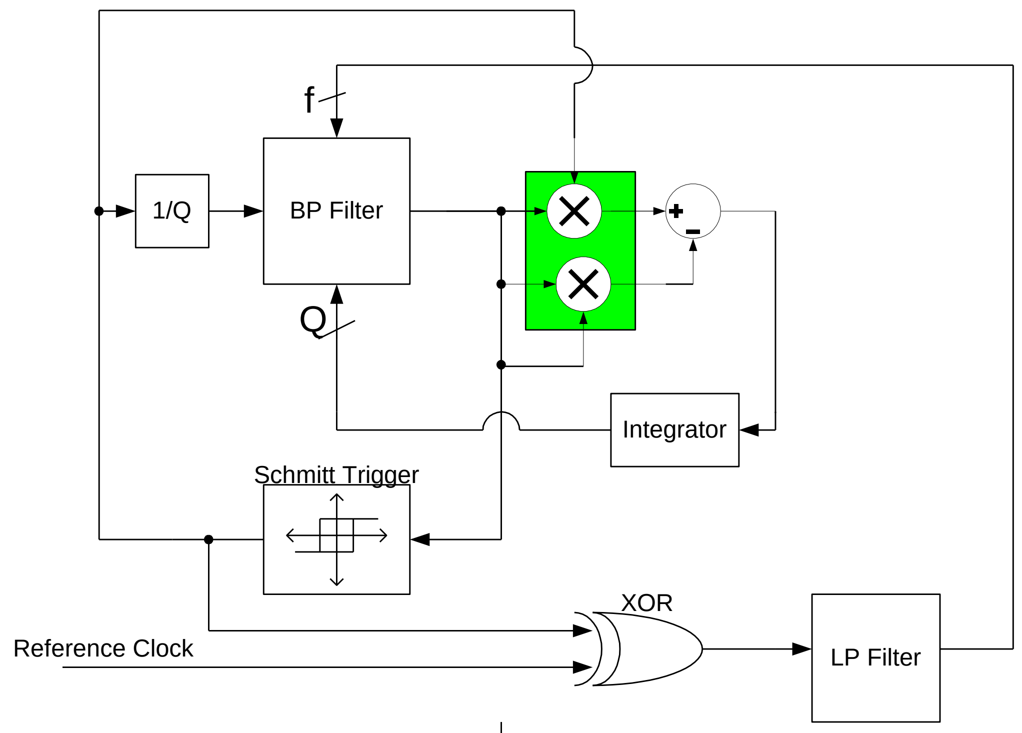
Attenuator

- **Capacitor**
  - **Large capacitors for matching**
  - **Large capacitors → Large loading**
- **Resistor**
  - **Larger resistors for matching**
  - **Large resistors → Small loading**
  - **Should take parasitic capacitor into consideration**

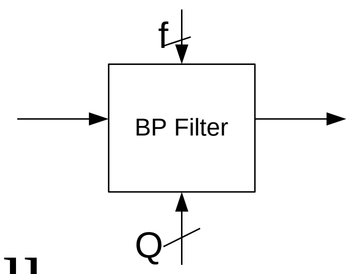




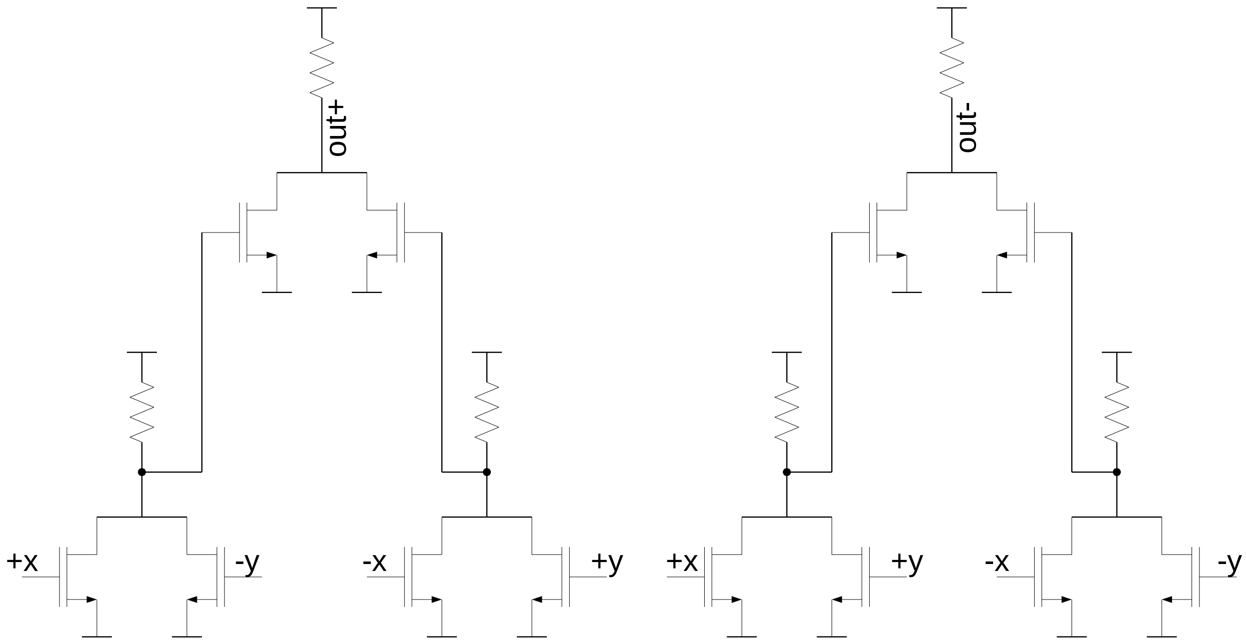
Multiplier



- Constraints
  - Symmetric
  - Good frequency response
  - Good CMRR
  - Gain should not be very small

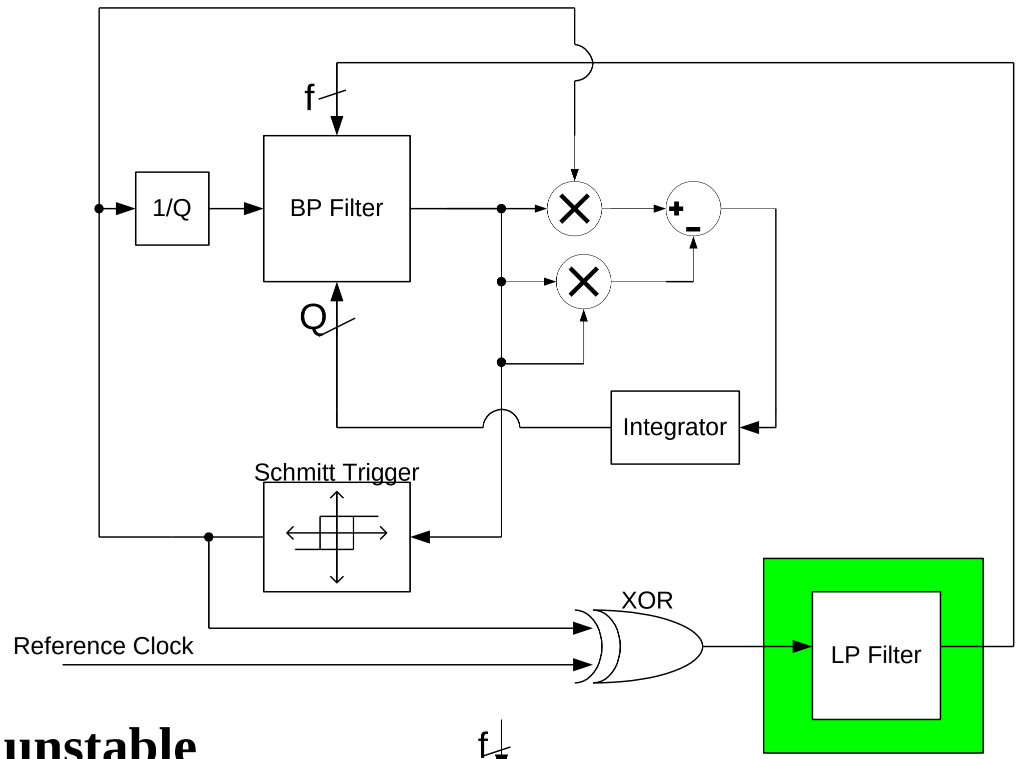


Multiplier

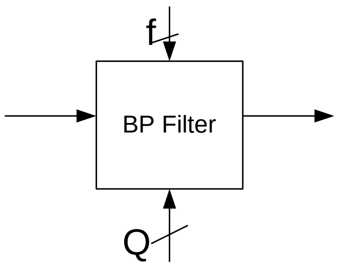


$$Multiplier\ \ gain = \frac{-32\,V_{DD}^{3}}{\left(V_1 - V_T\right)^3} \qquad CM\ \ gain = \frac{4\,V_{DD}^{2}}{\left(V_1 - V_T\right)^2}$$

LPF

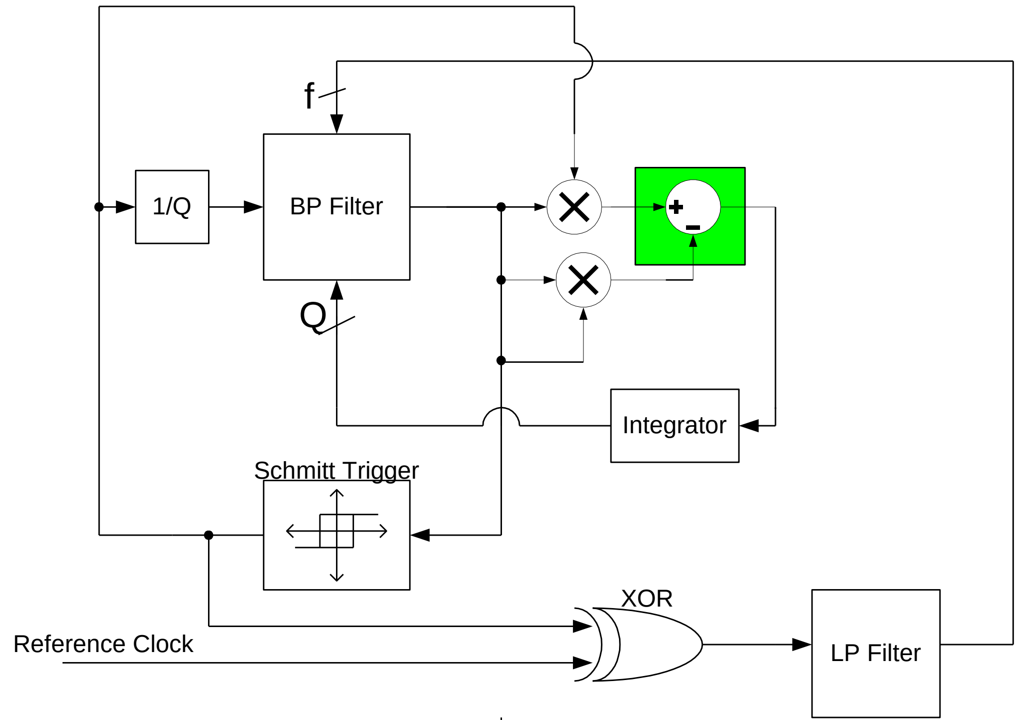


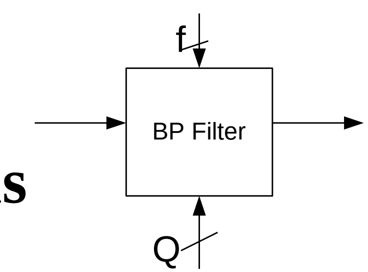
- **Constraints**
  - **High gain ⇒PLL might be unstable**
  - **Low gain ⇒ small pull-in range**
  - **low cut-off freq ⇒ small pull-in range**
  - **High cut-off freq ⇒ Jitter noise**
  - **Single ended output**



- **Built using external components for good control**

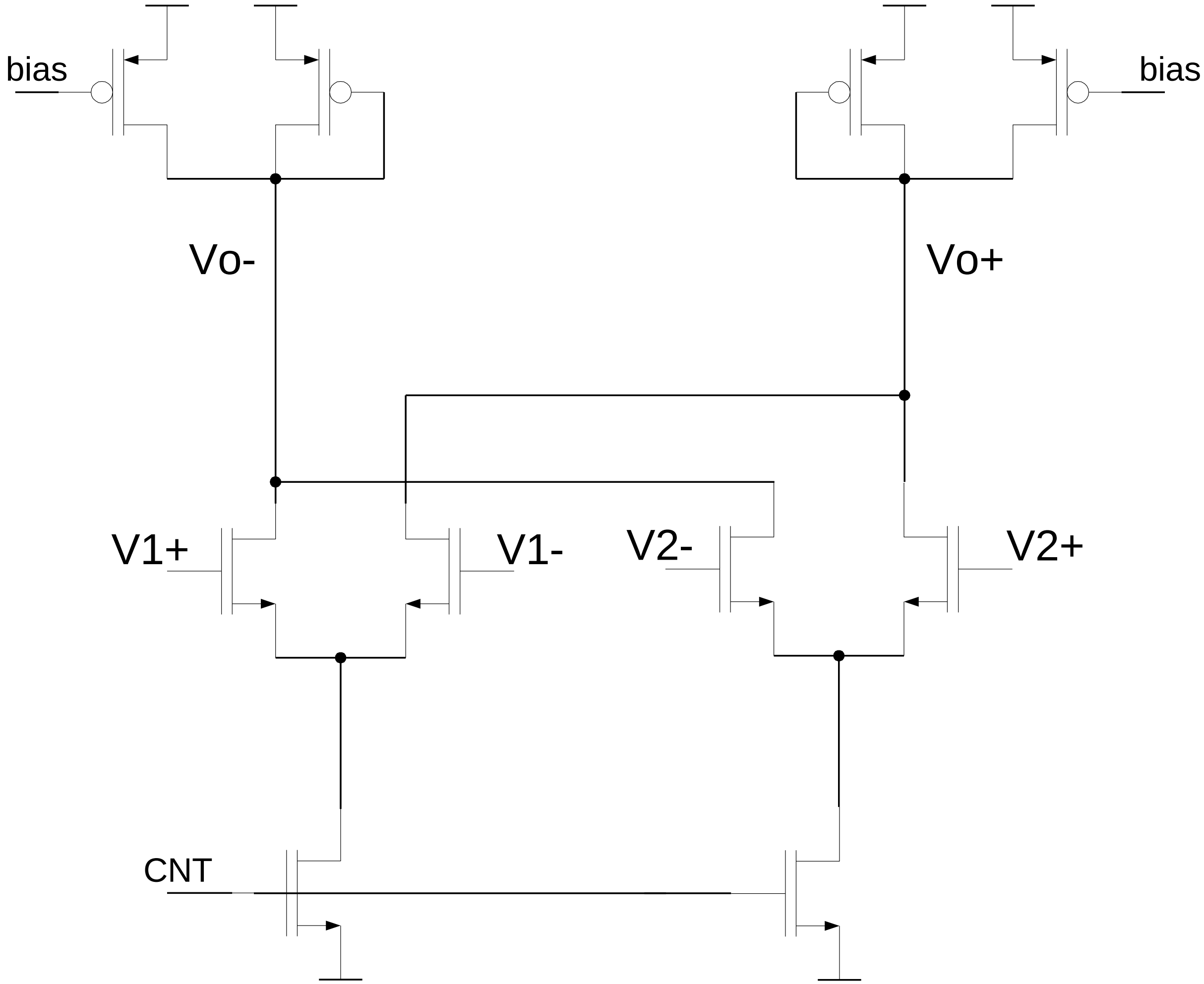
Differential difference adder





- **Add/Subtract two differential signals**
  - **High gain ⇒Q tuning loop unstable**
  - **Low gain ⇒ Lesser accuracy**
  - **Need not have a good frequency response**

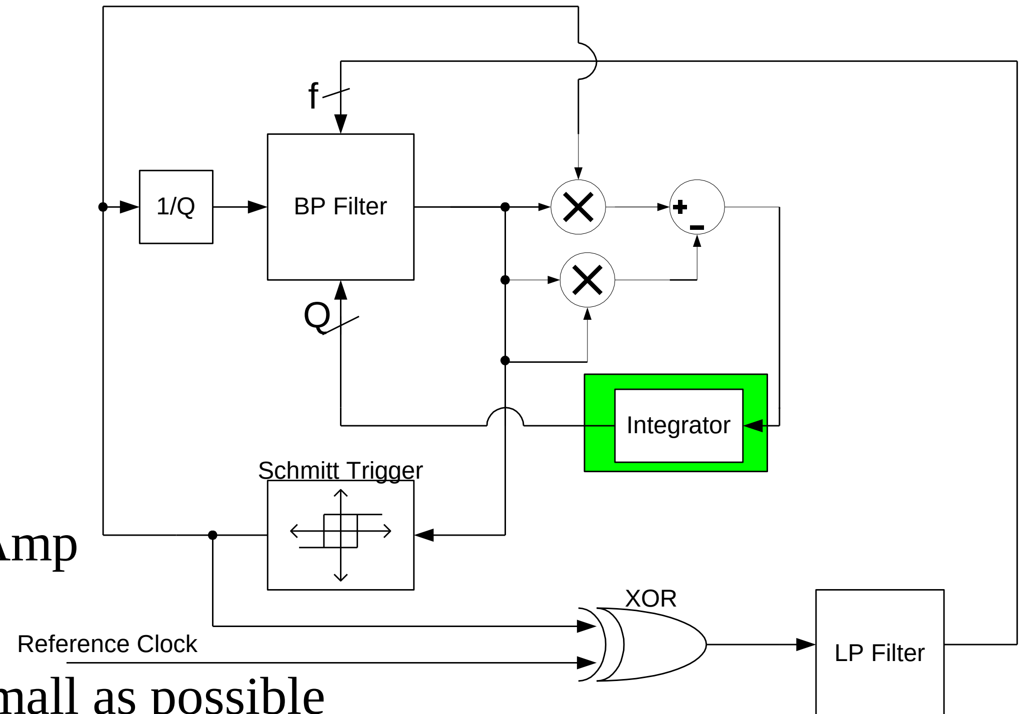
DDA circuit

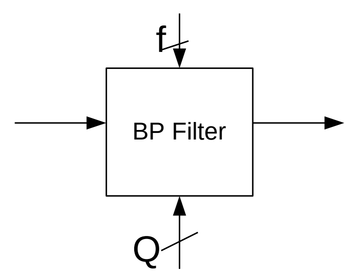


Integrator

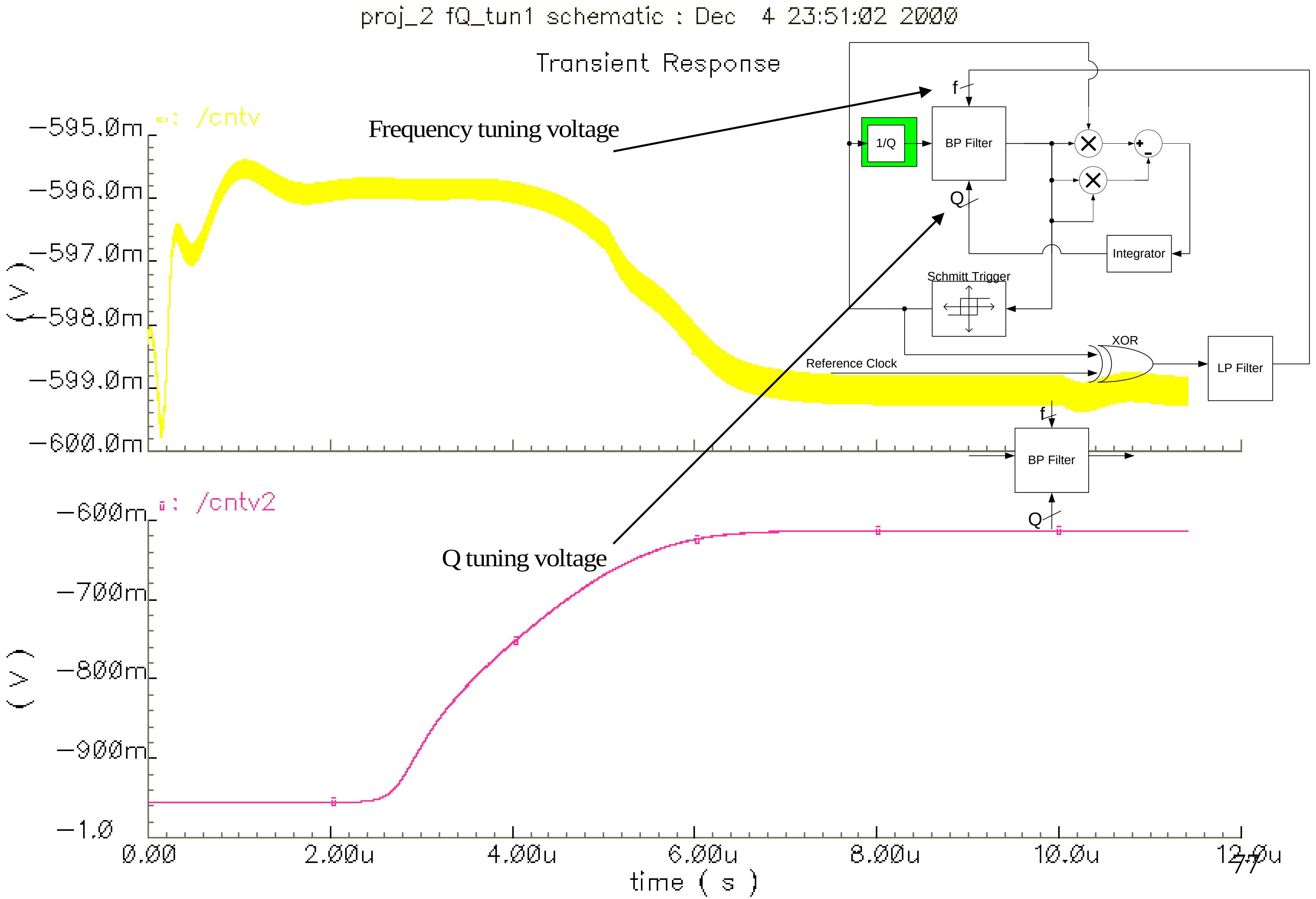
– Very high gain required to minimize Q tuning errors
– Frequency compensated Op-Amp in open loop can be used
– 3dB frequency should be as small as possible
– Phase margin as large as possible

Built using external components

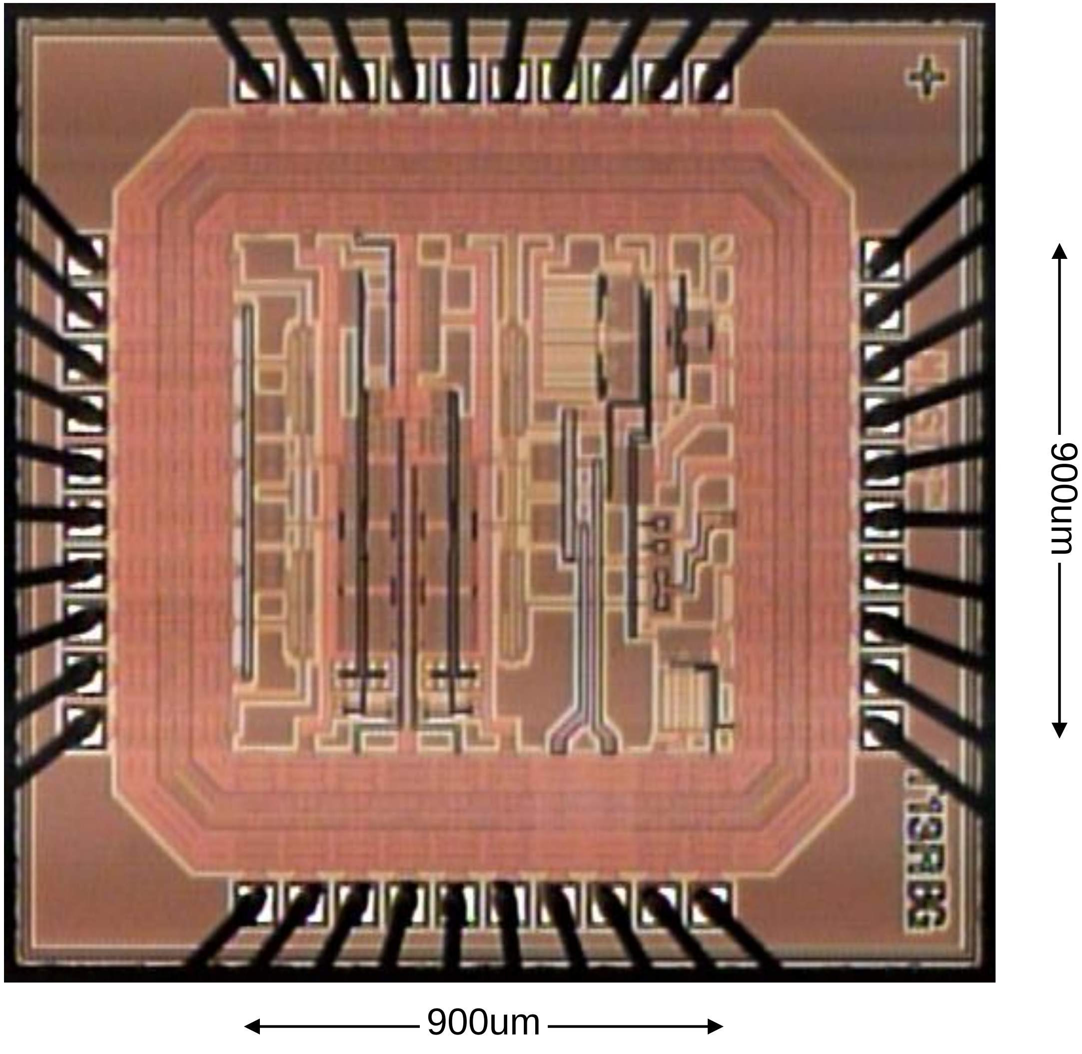




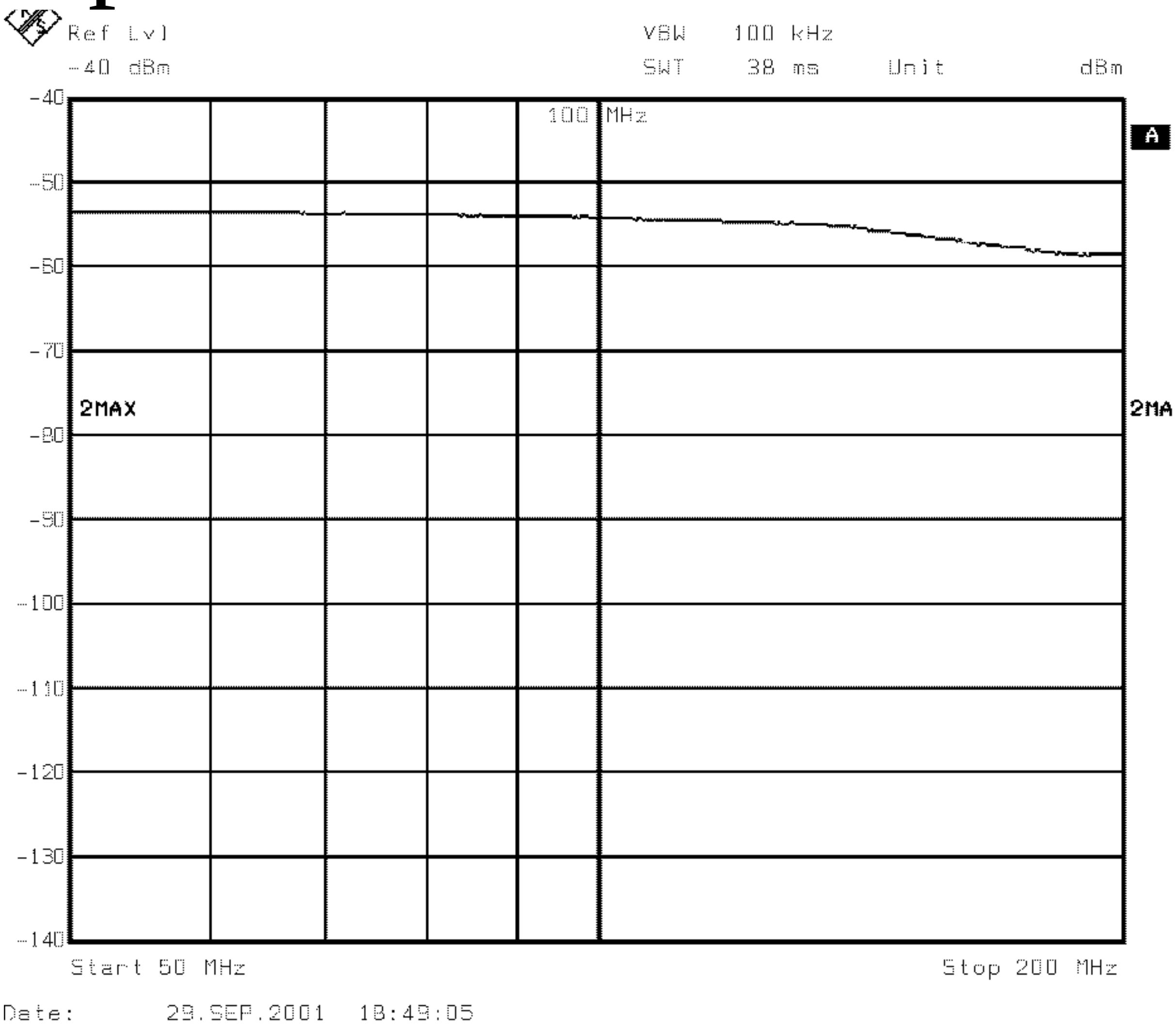
Simulated results for tuning scheme



Die Photograph

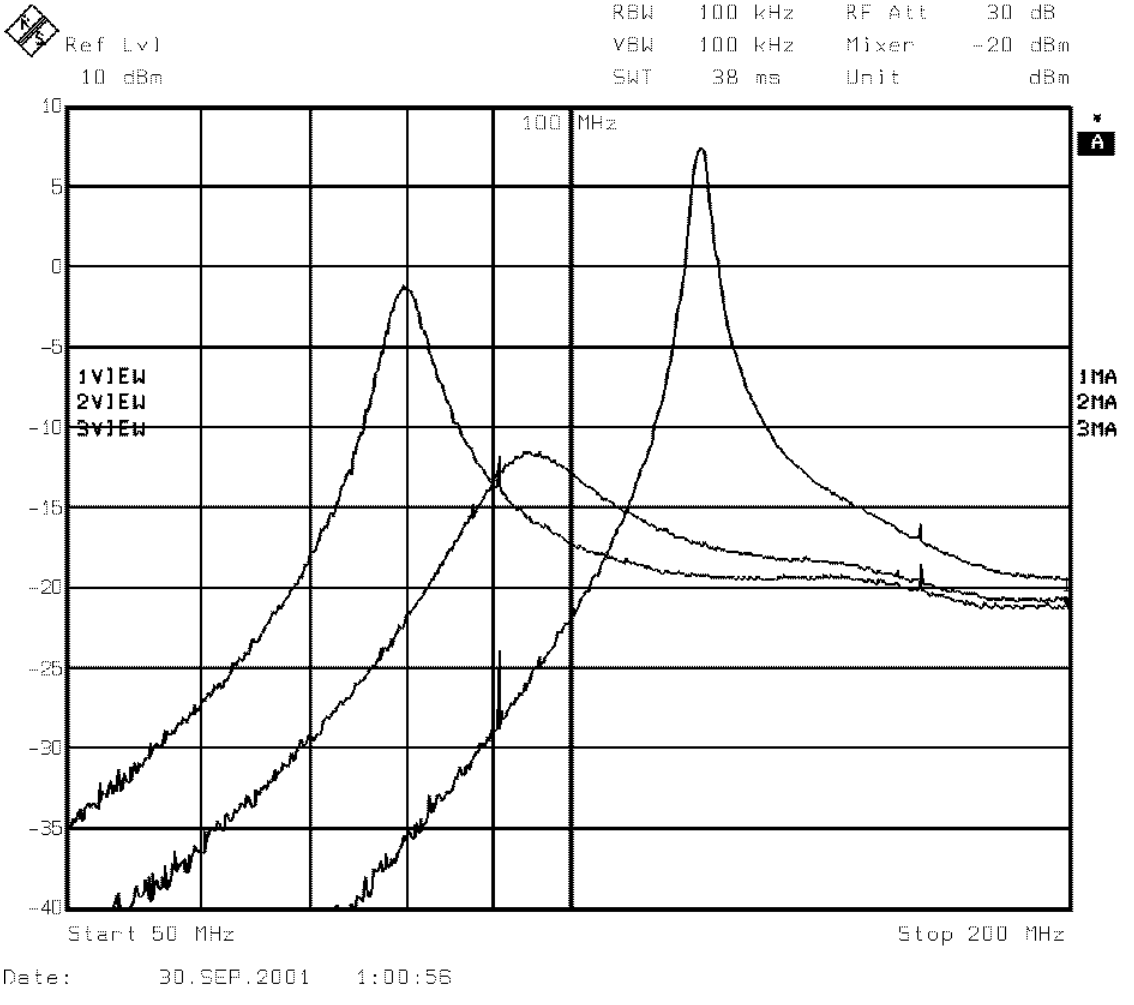


Experimental results .on



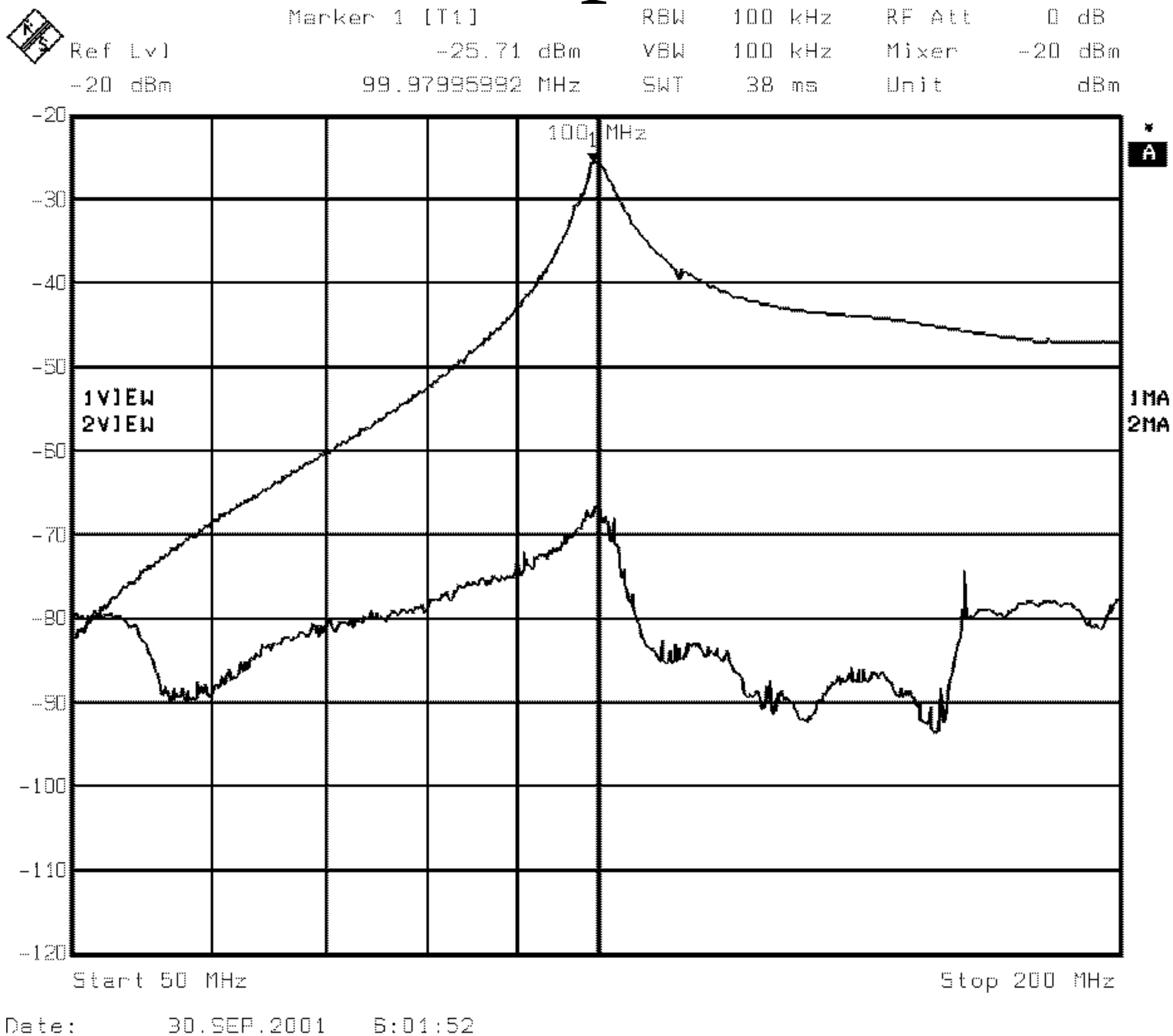
This response should be subtracted from other plots to get actual response

Filter response



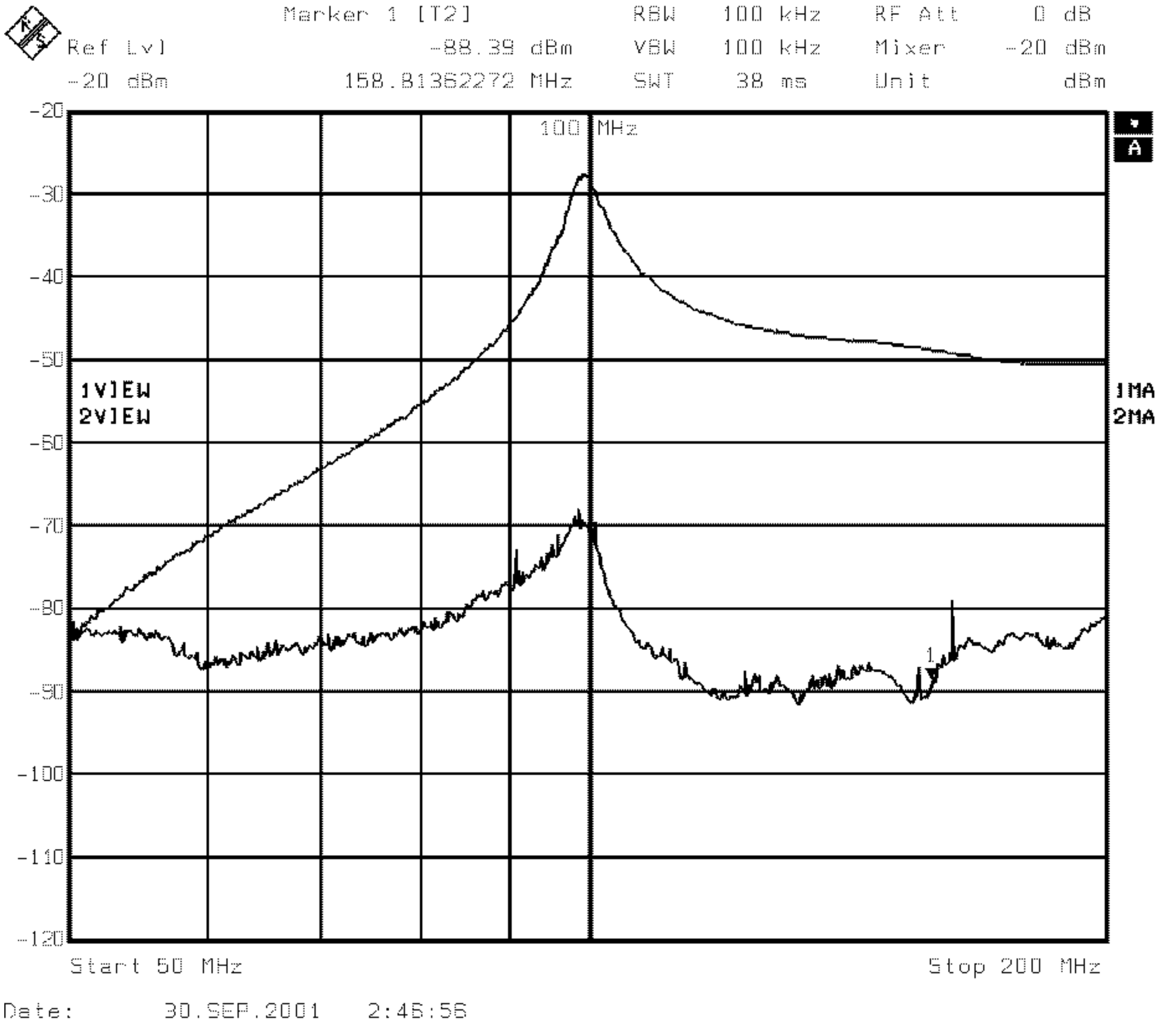
- Qs of 16, 5 and 40 at 80,95 and 110 MHz

DM-CM response of the filter



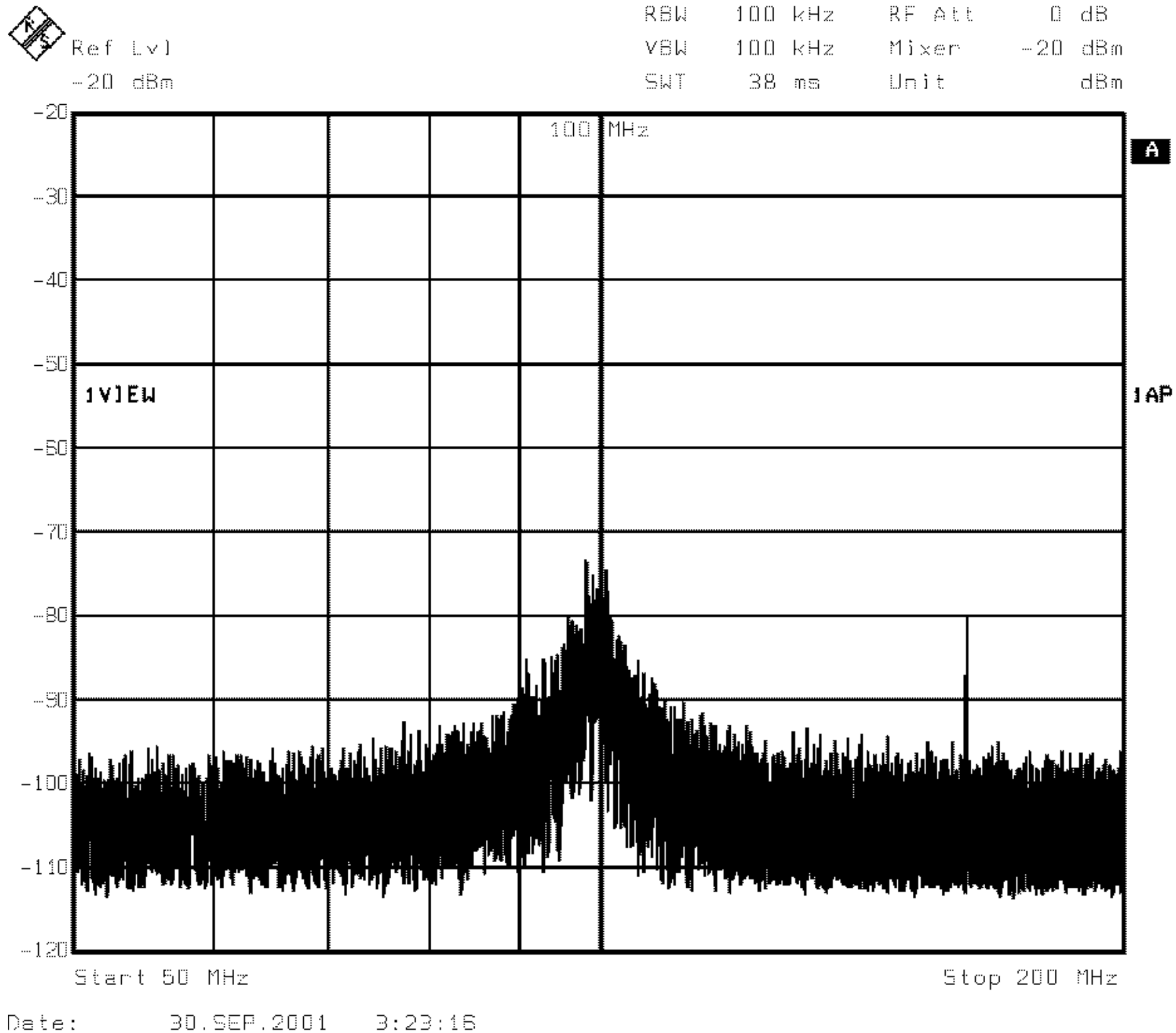
- CMRR is more than 40dB in the band of interest

Supply response of the filter



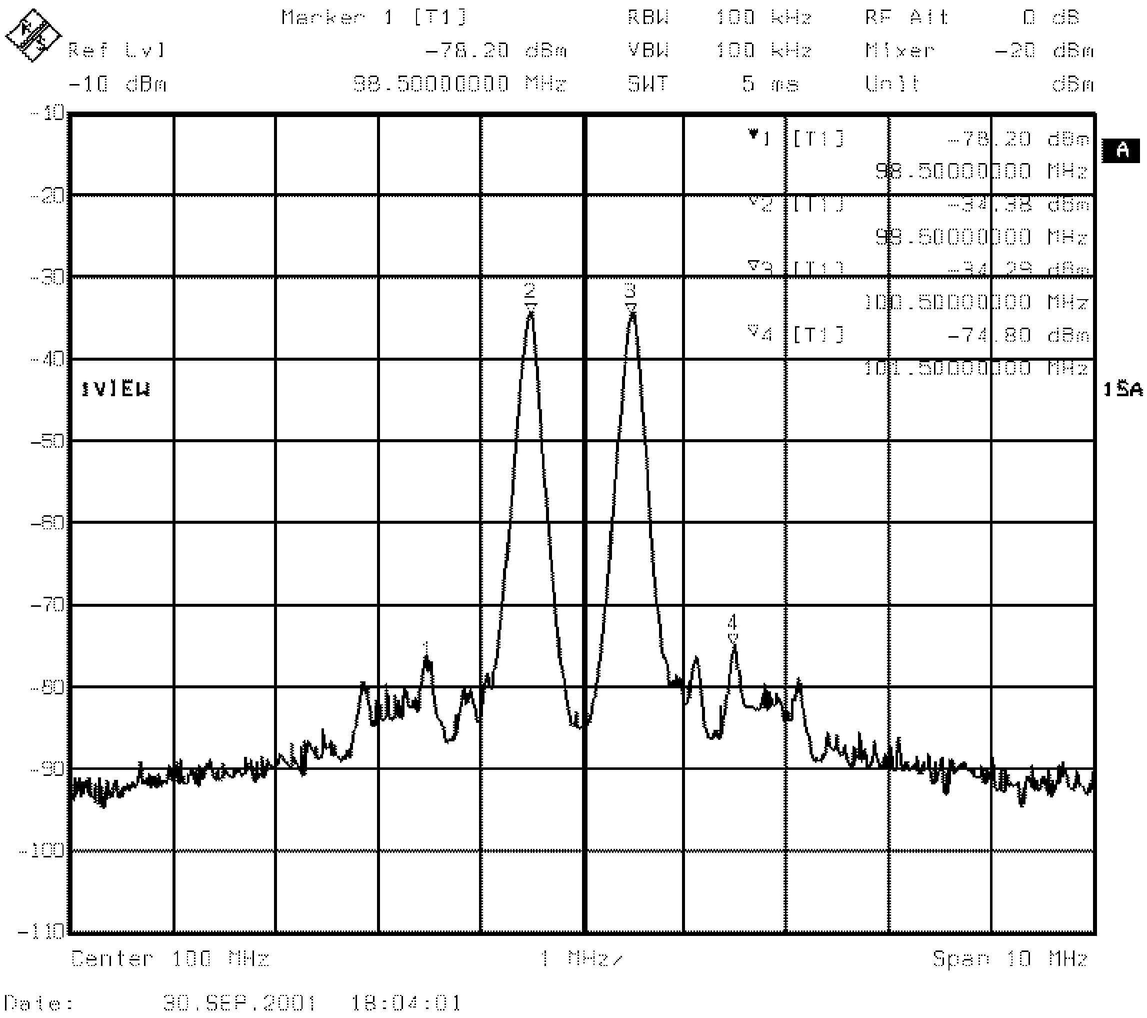
- PSRR$^-$ is more than 40dB in the band of interest

Noise response of the filter



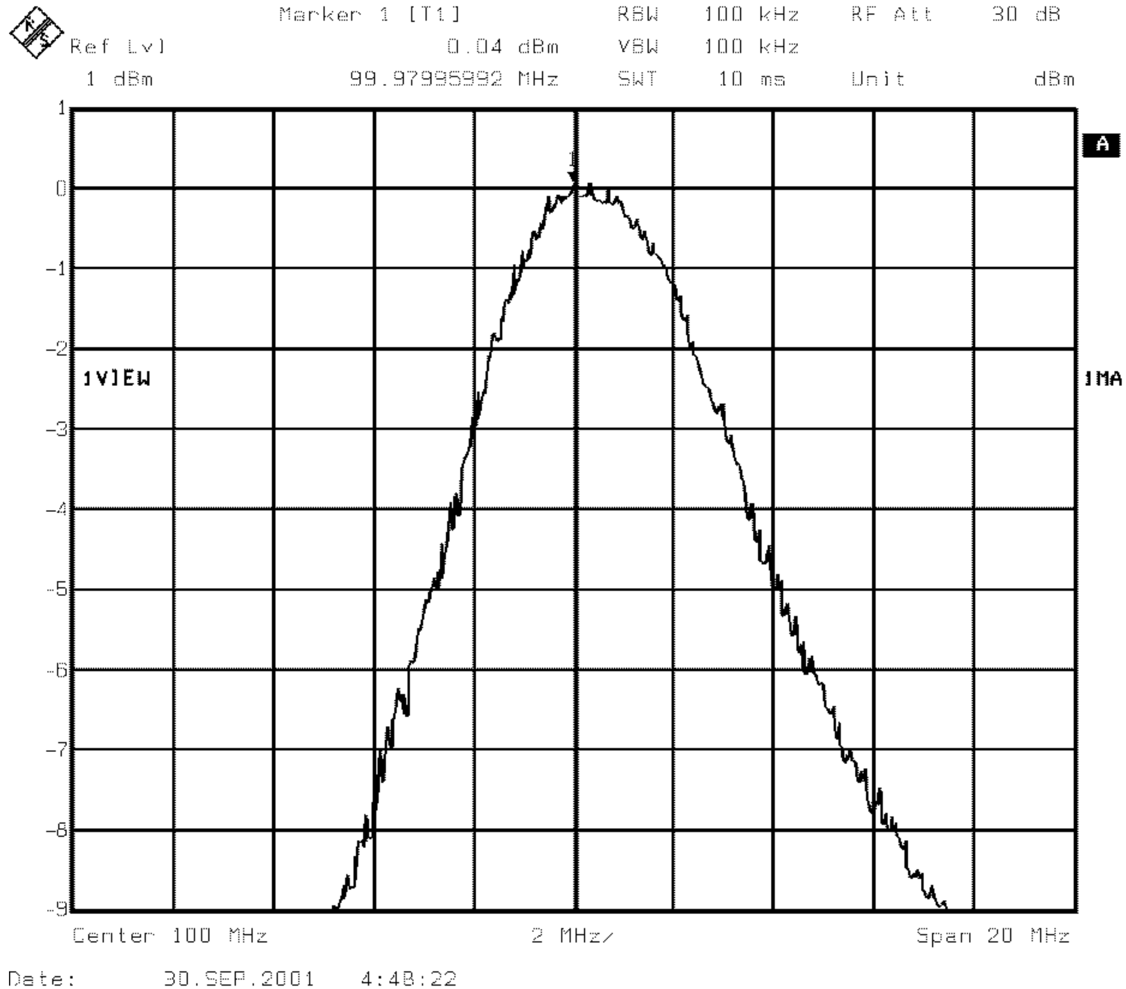
- Total integrated noise power at the output= -60dBm

Two-tone inter-modulation test



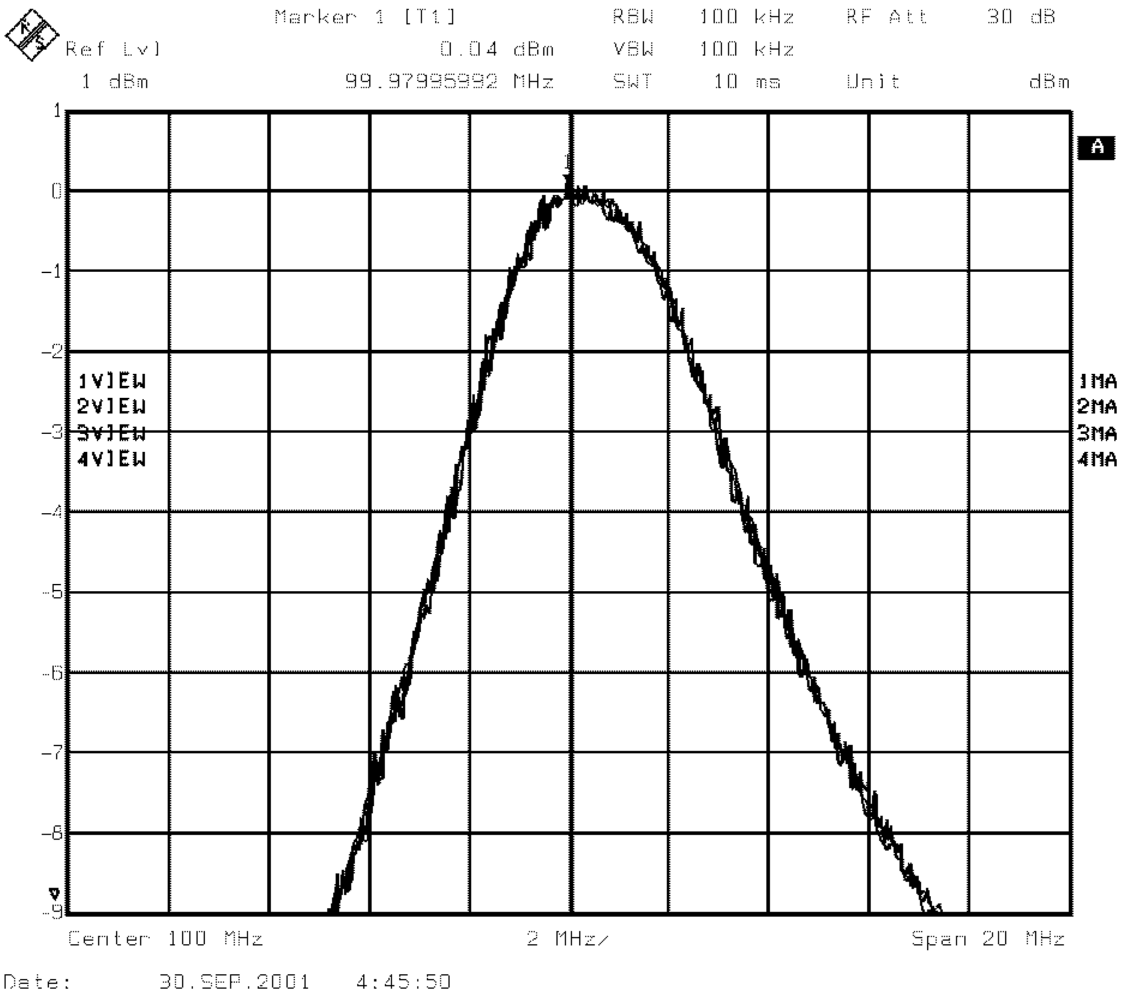
- IM$_3$ of 45dB when the input signal is 44.6mV

Filter response when tuned to Q=20



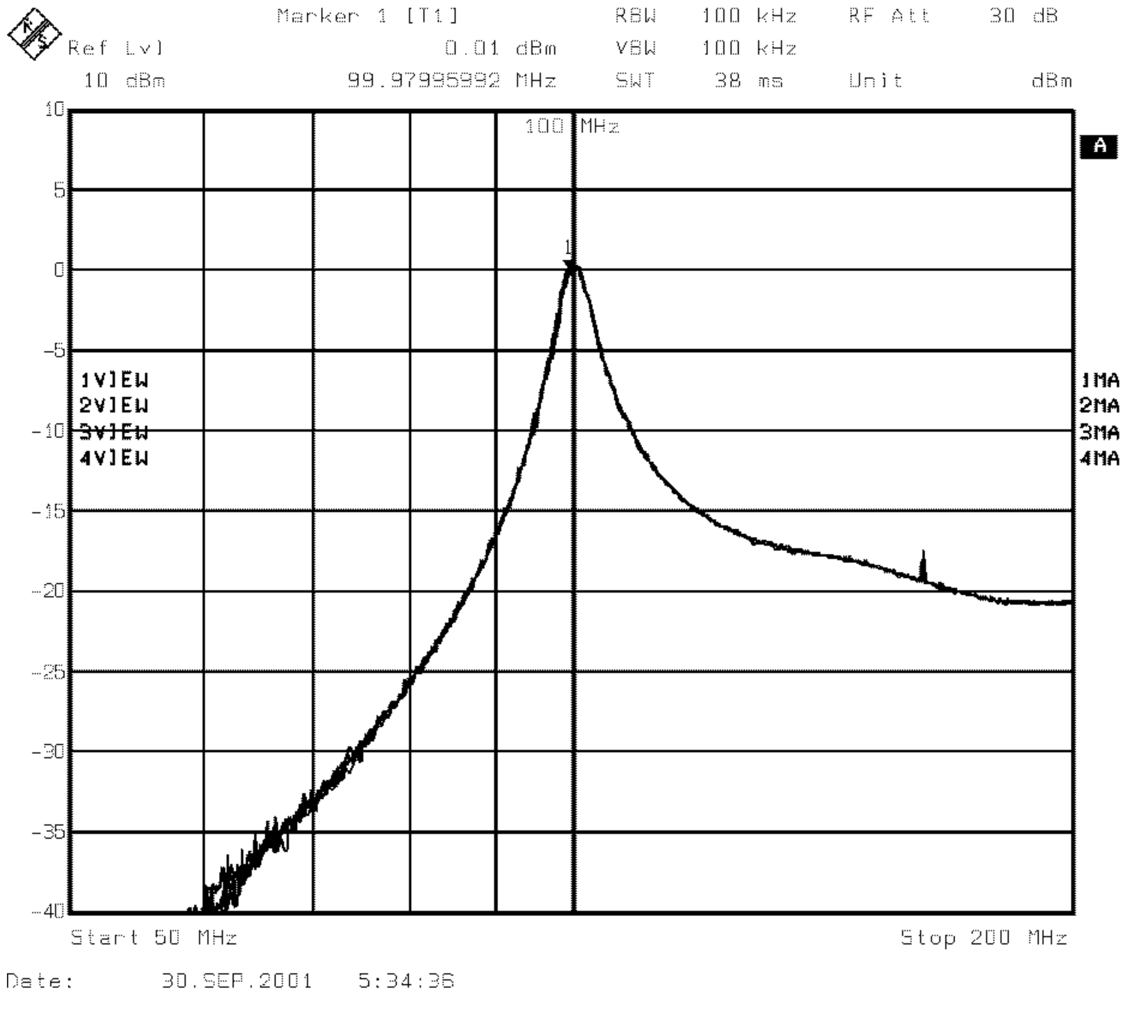
Both bandwidth and gain corroborate that accuracy of tuning is around 1%

Filter response for four different ICs



- Tuning accuracy is around 1%

Filter response for four different ICs



- The tuning works!

Conclusions

- A new high-frequency fully-differential OTA is designed.
- A band pass filter with f=100MHz and Q=20 is designed using the new OTA in AMI0.5um
- A new tuning scheme for BP filters that overcomes many of the problems faced by previous scheme is implemented.

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