ISCAS

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Tutorial Notes

Tutorial 1:
Design of Continuous-Time Filters from 0.1 Hz to 2.0 GHz

Presented by:

Edgar Sánchez-Sinencio, Texas A&M University

José Silva-Martínez, Texas A&M University

Sunday Morning, May 23, 08:30 - 11:30

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DESIGN OF CONTINUOUS-TIME

FILTERS FROM 0.1 HZ TO 2.0 GHZ

by: Edgar Sánchez-Sinencio & José Silva-Martínez,

Texas A&M University at IEEE ISCAS'04

DESIGN OF CONTINUOUS-TIME FILTERS FROM 0.1 HZ TO 2.0 GHZ

ROADMAP

- Introduction and Motivation
- Transconductance Amplifier Topologies
- Linearized OTA techniques
- Low Frequency Transconductance Amps
- LF Design Examples
- High Frequency Transconductance Amps
- HF Design Examples
- RF Filters
- Tuning
- Conclusions and Open Problems

Continuous-Time Filters from 0.1Hz to 2.0GHz

Outline

- Introduction and Motivation
- A family of Transconductance for different frequency ranges (applications).
- Common-mode feedforward and feedback strategies needed for differential output filters.
- Frequency- and Q-tuning techniques for OTA-C filters

Continuous-Time Filters from 0.1Hz to 2.0 GHz

Edgar Sánchez-Sinencio
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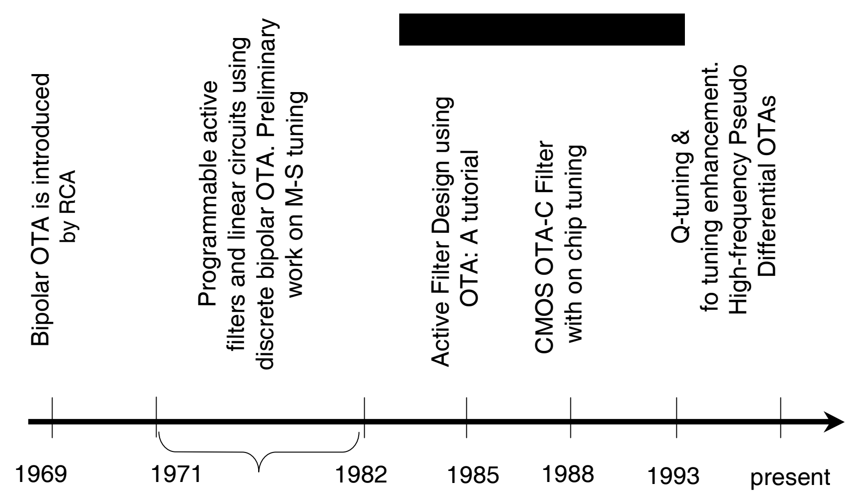
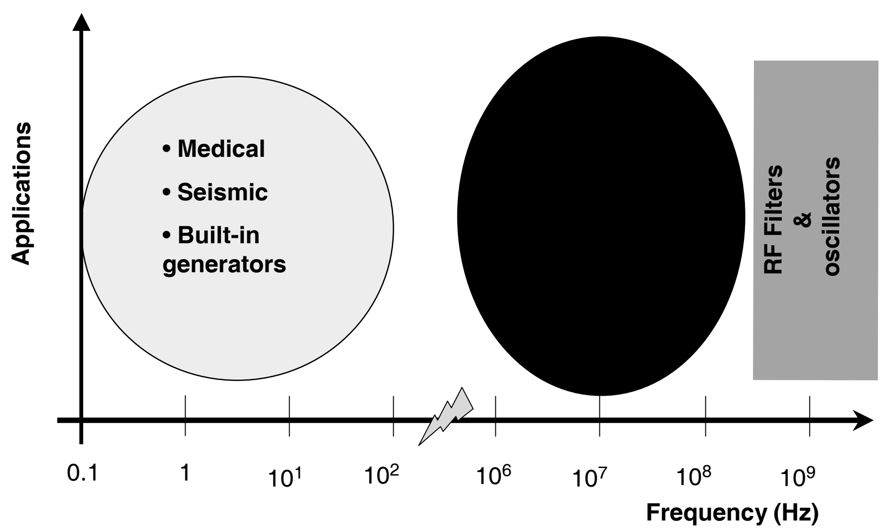
Abstract.-The bipolar transconductance amplifier (OTA) was commercially introduced in 1969 by RCA. Designers began using OTAs in the middle 80's, since then the CMOS-OTA has becoming a vital component in a number of electronic circuits, both in open loop and in closed loop applications. Here, we will focus on open loop applications. Continuous-time filters implemented with transconductance amplifiers and capacitors known as Gm-C or OTA-C are very popular for a host of applications. These applications involve frequency of operation from a few tents of a hertz up to several gigahertz. Several of those applications are in medical electronics and seismic area where the frequency range is between 0.1Hz up to 20Hz. Other applications in the audio range do not commonly use OTA-C filters because switched-capacitor techniques excel in this range. But for frequency range of a few MHz like in Intermediate Frequency (IF) filters in RF receivers OTA-C implementations are very attractive. For a few GHz range applications where the OTA becomes a simple differential pair there is number of researchers investigating LC-oscillators and filters. In this tutorial we discuss practical implementations of transconductance amplifiers oriented for wide range of applications for example in medical, IF filters, hard disk drive linear phase filters, LC-oscillators and RF filters. Furthermore the unavoidable tuning scheme to compensate the Gm/C deviations due to process technology variations is discussed. OTA single ended, fully differential and pseudo differential versions are introduced together with the common-mode feedback circuits needed for proper operation of differential architectures.

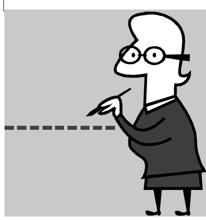


Evolution of the OTA

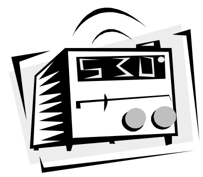
Typical applications of OTA-C filters and frequency ranges

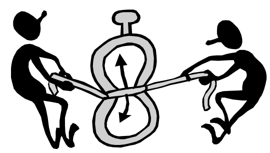


A few examples of continuous-time filters in a host of applications



ota



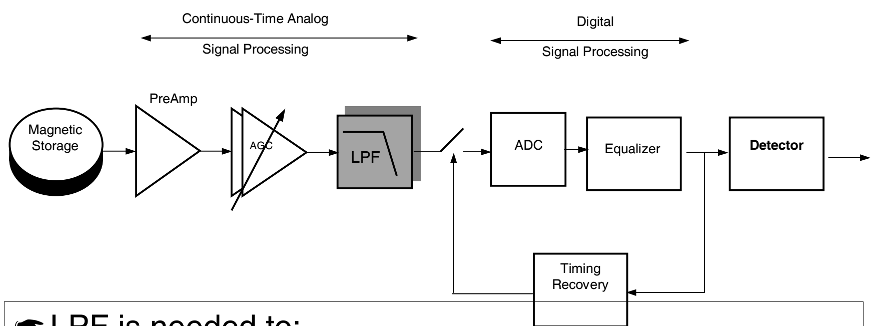


Applications for continuous time filters

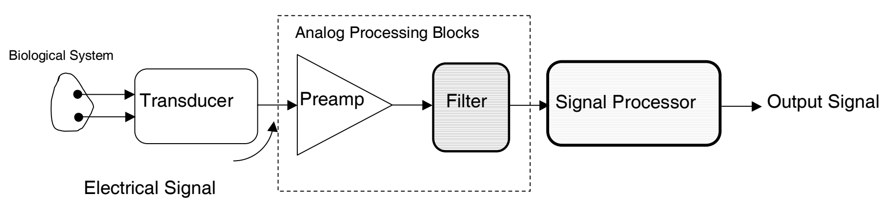
Read channel of disk drives
--
for phase equalization and smoothing the wave form

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

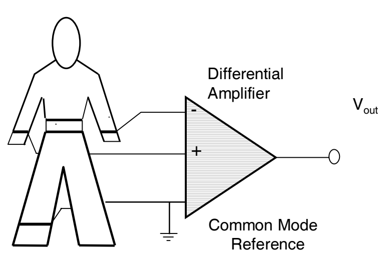
Hard Disk Driver Read Channel



- LPF is needed to:
  - Limit signal and noise bandwidth;
  - Provide anti-aliasing prior to sampling;
  - Provide significant contribution to overall equalization.



Block Diagram of a general purpose bioelectric signal acquisition system

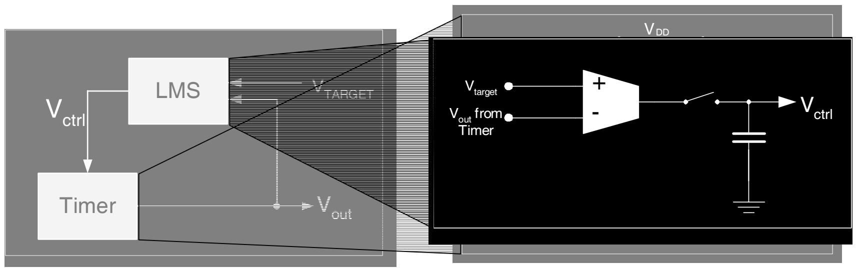


Typical configuration for the measurement of bio-potentials

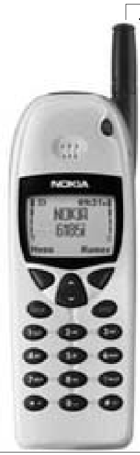
Parameter	Typical range
Gain	1-1000
Bandwidth	0.1 Hz-10KHz
Dynamic Range (DR)	60dB-100dB
CMRR	**80-140dB**
Zinput	10MΩ-1GΩ at 60Hz
Vnoise	<10nV/ $\sqrt{Hz}$
Inoise	<1µA/ $\sqrt{Hz}$

Continuous-Time Linear Ramp Implementation

LMS Integrator



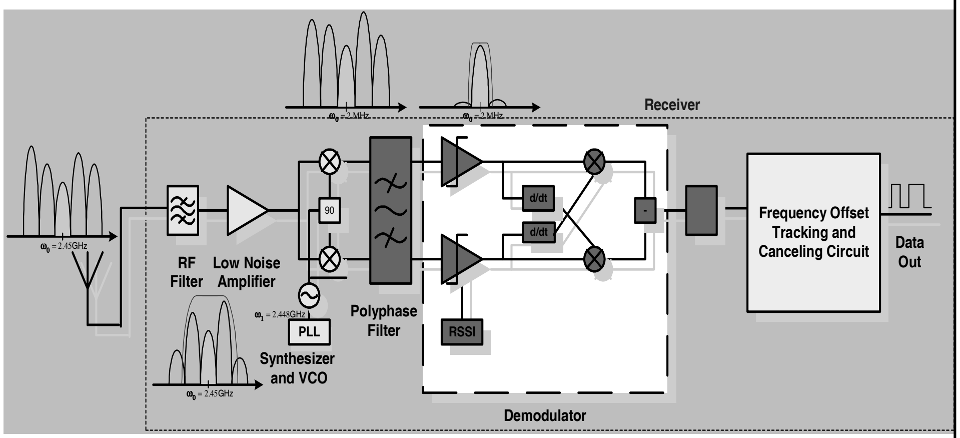
- The Timer is a cascode current source with V_{ctrl} controlling the gate voltage of the current source transistor
- The LMS block is an OTA-C integrator with a switch to control the charging of the capacitor



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

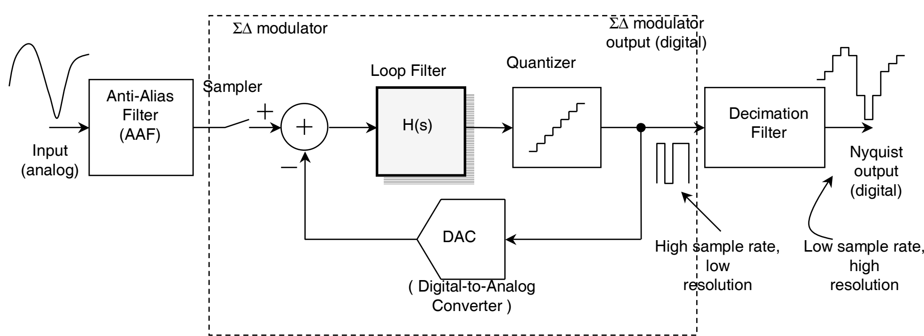
6185i digital cell phone from Nokia.

Low-IF Bluetooth Receiver



- Active polyphase filter is used to reject image and select channel.

Sigma-Delta Oversampled A/D Conversion

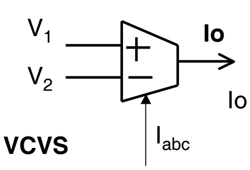


Functional level diagram of a general continuous-time sigma-delta oversampled analog-to-digital converter

A family of Transconductances for different frequency ranges applications.

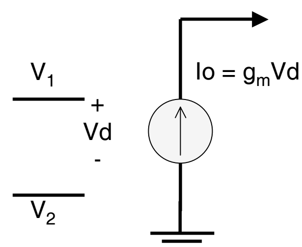
OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA)

First commercial OTA produced by RCA in 1969, i.e., CA3080



VCVS

$Io = g_m(I_{abc})(V_1 - V_2)$

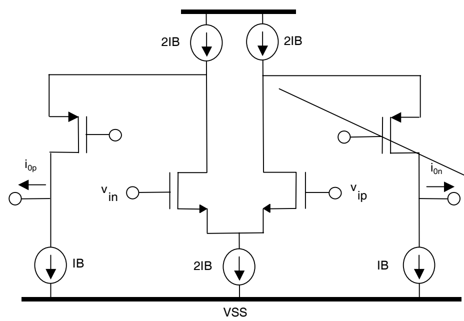


The transconductance gain “ g_m” is a function of the I_{abc} .

$\mathbf{g_m} = h_1 I_{abc}$ for bipolar and weak inversion MOSFETs

$\mathbf{g_m} = h_2 [I_{abc}]^{1/2}$ for MOSFETs in saturation

FOLDED-CASCODE OTA



POWER: $P = 4I_B(VDD + VSS)$

NOISE: $v_n^2 = \frac{16kT}{3g_m}(1 + NF)$

TRANSCONDUCTANCE:

$$g_m = \frac{g_{mo}}{1 + \frac{s}{g_{mp}/C_P}}$$

INPUT SWING: $V_{SS} + 2V_{DSAT} + V_T > V_{IN}$

OUTPUT RESISTANCE:

$$R_O \approx g_{mn}R_{ON}^2 \| g_{mp}R_{OP}^2$$

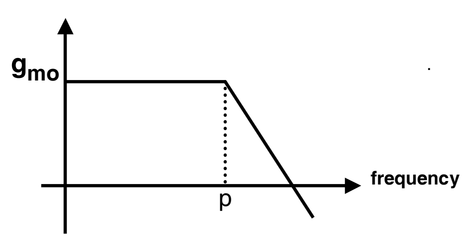
OUTPUT SWING: $V_{SS} + 2V_{DSAT} < V_{OUT} < V_{DD} - 2V_{DSATP}$

OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA) Frequency Dependence

$G_m = g_{mo} / (1 + s/p)$

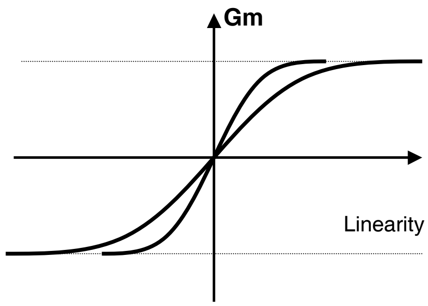
Where g_{mo} is the DC transconductance gain

p is the dominant pole which is around 10MHz to 100Mhz



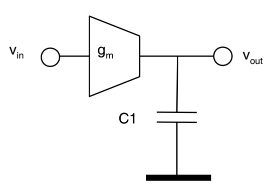
Issues about the OTA:

- Operated in open loop conditions
- High-Frequency Operation
- Poor Linearity Range



Linearity Range from 50mV to 200 mV

BASIC INTEGRATOR



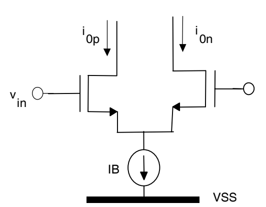
$$g_m \cong \mu_n C_{OX} \frac{W}{L}(V_{GS}-V_T) \cong \sqrt{2\mu_n C_{OX} \frac{W}{L} I_D}$$

$$HD3 \cong \frac{1}{32}\left(\frac{V_{in}}{V_{GS}-V_T}\right)^2$$

Mobility degradation:

$$\mu_n \cong \left(\frac{\mu_0}{1+\theta(V_{GS}-V_T)}\right)\left(\frac{1}{1+\frac{\theta V_{in}}{1+\theta(V_{GS}-V_T)}}\right)$$

$$HD3 \cong \frac{1}{4}\left(\frac{\theta V_{in}}{1+\theta(V_{GS}-V_T)}\right)^2$$

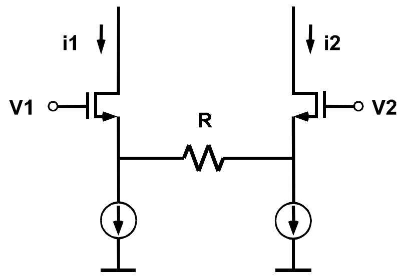


NOISE:

$$v_n{}^2 = \frac{16kT}{3g_m}(1+NF)$$

Differential Pair with Source Degeneration

Improved linearity



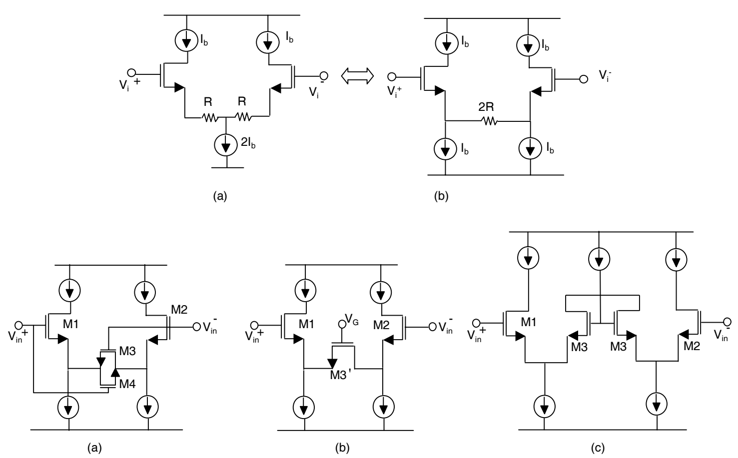
$$1/n = \frac{1}{1 + g_m R}$$

$$i_d = \left(\frac{\sqrt{2\beta_1 I_B}}{2}\right)\frac{v_d}{n}\sqrt{1-\left(\frac{v_{in}}{2n(v_{GS}-v_T)}\right)^2}$$

$HD2 = 0$ Ideally

$$HD3 = \frac{1}{32n^2}\frac{v_{in}^2}{(v_{GS}-v_T)^2}$$

g_m linearization schemes via source degeneration.

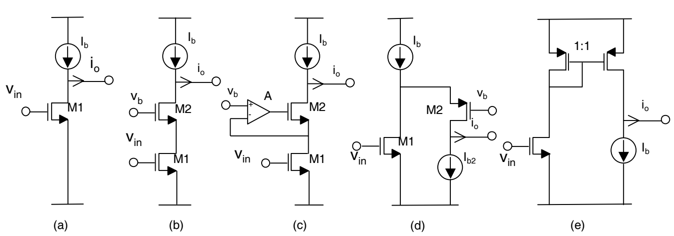


Active Source Degeneration topologies; (a) and (b) transistors biased on triode region and (c) with saturated transistors.

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)	$\dfrac{g_{m1}}{1+g_{m1}R}$ $R=1/\mu_o C_{ox}(V_{gs}-V_T)$	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.

Single-input Transconductor (ST) Implementations



Single Input (a) Negative Simple Transconductor, (b) Cascode Transconductor, (c) Enhanced Transconductor, (d) Folded-Cascode Transconductor, (e) Positive Simple Transconductor.

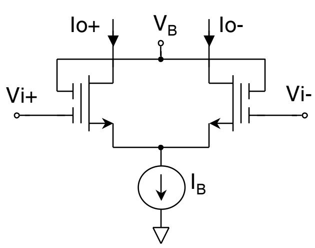
- Observe that:
  $g_m = f(I_b)$, the exact relation is a function of the transistor region of operation.
- Note that output impedance of (a) and (e) are only $1/g_{ds}$ and (b),(c) and (d) implementations have larger output impedances.

Properties of Simple (single input/ single output) Transconductors

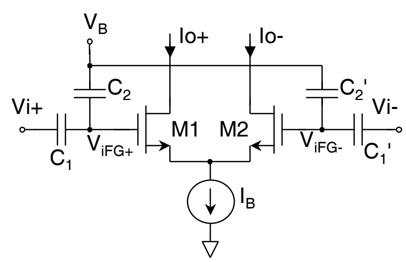
Structure/ Figure	R_{out}	Min V_{DD} *
Simple/1(a)	$\frac{1}{g_{ds1}}$	$\sqrt{\frac{2I_B}{k}} + V_{sat,I_B}$
Cascode/l(b)	$\frac{g_{m2}}{g_{ds1}\ g_{ds2}}$	$(1+m)\sqrt{\frac{2I_B}{k}} + V_{sat,I_B}$
Enhanced/1(c)	$\frac{A\ g_{m2}}{g_{ds1}\ g_{ds2}}$	$(1+m)\sqrt{\frac{2I_B}{k}} + V_{sat,I_B}$
Folded/1(d)	$\frac{g_{m2}}{g_{ds1}\ g_{ds2}}$	$\sqrt{\frac{2I_B}{k}} + V_{Tp} + V_{sat,I_B}$

* The bottom devices of the cascode pairs have an aspect ratio of $(W/L)_1/(W/L)_2=m^2$. K is a technological parameter determined by the mobility, and the gate oxide;$V_{DSAT,\ 1B}$ is the saturation voltage for the I_B current source

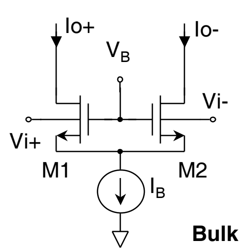
Potential Solutions for Rail-to-Rail Amplifiers:
OTAs suitable for Low Frequency Applications



FG DP Implementation

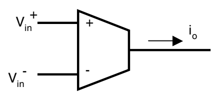


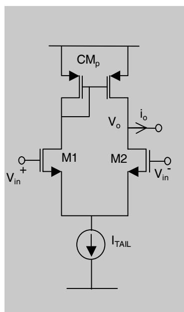
FG DP Equivalent circuit



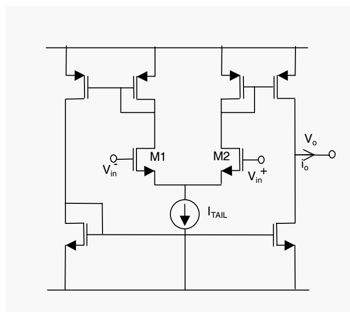
Bulk Driven DP

Three Conventional Differential Input –Single Ended OTAs

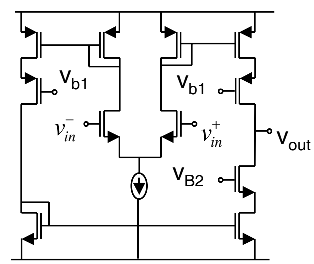




(a) Simple differential input OTA.

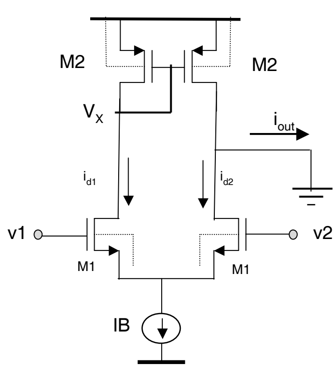


(b) Balanced OTA



(c) Cascode

Simple OTA Design Equations



$$g_m = \mu_n C_{OX} \left(\frac{W}{L}\right) V_{DSAT}$$

$$\text{Noise} \Rightarrow g_m > g_{m,min}$$

$$GBW = \frac{g_m}{C_L}$$

$$\omega_p = \frac{g_{mp}}{C_p} \cong \mu_p \frac{V_{DSAT}}{2L_p^2}$$

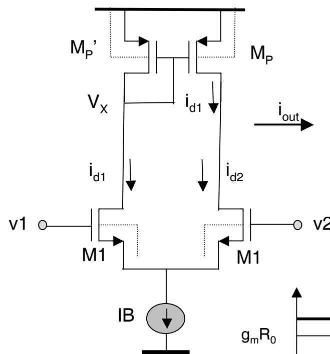
$$\text{Rout} \cong r_p \| r_n$$

V_{DSAT} is lim ited !!

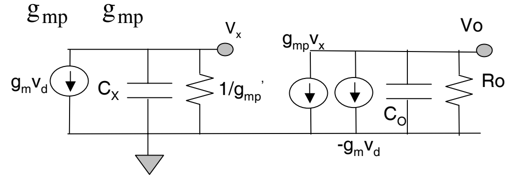
$$r_{ds} \cong \frac{V_{early} L_P}{I_D}, \qquad \omega_P \cong \frac{g_{mp}}{2C_{GSP}} \cong \frac{\mu_P V_{DSATP}}{2L_P^2}$$

$$\text{Noise}(V_{RMS}) = \sqrt{\frac{16kT}{3}} \sqrt{\frac{1}{g_{m1}}} \sqrt{1 + \frac{g_{m2}}{g_{m1}}} \left(\sqrt{BW}\right)$$

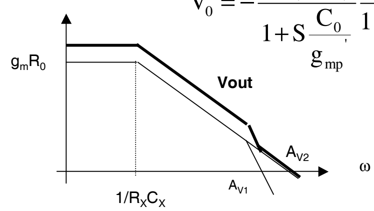
Simple OTA Frequency Response



$$\frac{1}{g_{mp}'} = \frac{1}{g_{mp}} \| R_{OL} \qquad R_O = R_{ON} \| R_{OP}$$



$$V_0 = -\frac{g_m\left(\frac{g_{mp}}{g_{mp}'}\right)}{1+S\frac{C_0}{g_{mp}'}}\frac{R_0}{1+SR_0C_0}V_d - \frac{g_mR_0}{1+SR_0C_0}V_d$$



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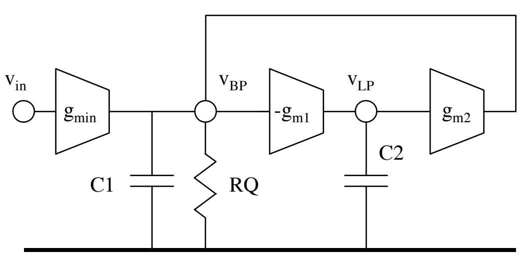
SECOND-ORDER FILTER



Bandpass output:

$$\omega_0 = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}}$$

$$Q = \sqrt{\frac{C_1}{C_2}\left(g_{m1}g_{m2}R_Q^{\;2}\right)}$$

$$A_{Vpeak} = g_{min}R_Q$$

N= number of OTAs

$$IM3 \cong \frac{3\sqrt{N}}{32}\left(\frac{V_{in}}{V_{GS}-V_T}\right)^2$$

$$g_m \cong \mu_n C_{OX}\frac{W}{L}\left(V_{GS}-V_T\right) \cong \sqrt{2\mu_n C_{OX}\frac{W}{L}I_D}$$

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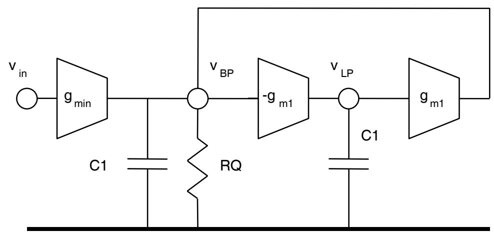
SECOND-ORDER FILTER



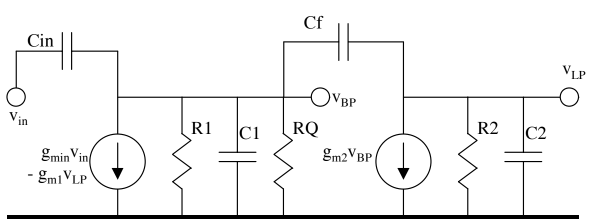
$$IM3 \cong \frac{3}{32}\left(\frac{V_{in}}{V_{GS}-V_T}\right)^2 \boxed{\left(\sqrt{3}\right)}$$

INTEGRATED NOISE:

$$v_{eq,in}(RMS) \cong \sqrt{\frac{8kT}{3\pi C1}} * \boxed{\sqrt{\frac{1}{R_Q g_{min}}}} * \sqrt{1+NF_{in}+\boxed{\frac{g_{m1}}{g_{min}}}\left(2+NF_1+NF_2\right)+\boxed{\frac{0.75}{R_Q g_{min}}}}$$

Tradeoff:
Large gain reduces the noise level but increases the harmonic distortions.

Output resistance effects



$$\omega_0 \cong \omega_{0ideal}\sqrt{1+\frac{1}{Q_{ideal}A_V}}$$

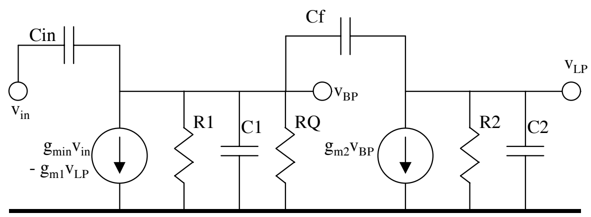
CENTER FREQUENCY IS ITTLE SENSITIVE TO A_V

BW IS QUITE SENSITIVE TO A_V

$$BW \cong BW_{ideal}\left(1+2\frac{Q}{A_V}\right)$$

$Av=g_{m1}R_1$ ($R_1=R_2$ OTA output resistance)

EFFECTS OF THE NON-DOMINANT POLE

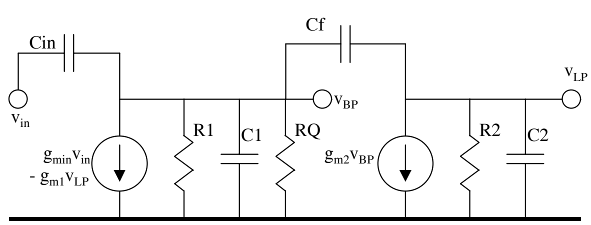


Single pole: $g_m \cong \frac{g_{m0}}{1+\frac{s}{\omega_{p1}}}$

$$\omega_0 = \omega_{0ideal}\sqrt{\frac{1}{1+\frac{2BW_{ideal}}{\omega_{P1}}}}$$ **Sensitive**

$$BW = BW_{ideal}\frac{1-2Q_{ideal}\frac{\omega_{0ideal}}{\omega_{P1}}}{1+\frac{2BW_{ideal}}{\omega_{P1}}}$$ **Quite sensitive**

PARASITIC CAPACITORS



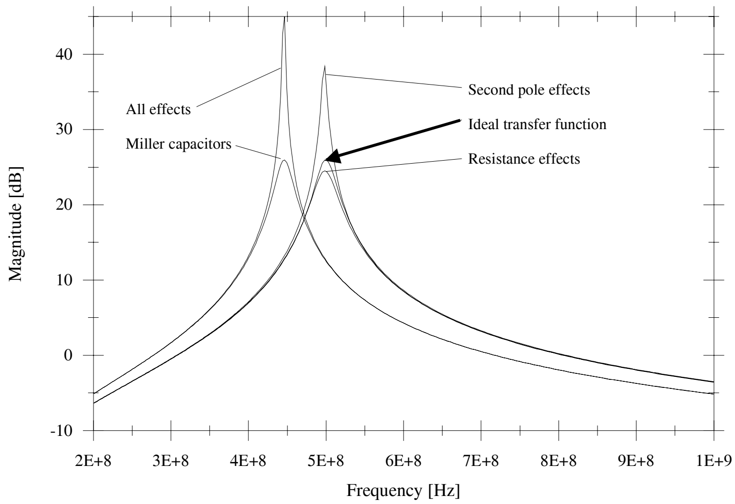
C1 and C2 are affected by the grounded parasitic capacitors (partially corrected by the automatic tuning system)

$$\omega_0 = \omega_{0ideal}\sqrt{\frac{C_1C_2}{(C_1+C_{in})(C_2+C_f)+C_2C_f}}$$

Cin introduces a high frequency zero

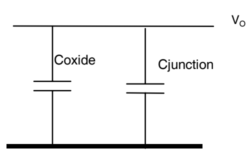
$$BW = BW_{ideal}\frac{(C_1)\left(C_2+C_f+\frac{g_{m2}-g_{m1}}{g_1}C_f\right)}{(C_1+C_{in})(C_2+C_f)+C_2C_f}$$

SMALL SIGNAL ANALYSIS



- **Little sensitve to OTA output resistances**
- **Quite sensitive to parasitic poles**
- **Parasitic capacitors should be accounted in the ATS (matching)**
- **Tune the filter itself (??)**

Parasitic capacitors



Junction capacitors are non-linear:

$$C_{total} = C_L + \frac{C_{j0}}{\sqrt{1 + \frac{v_0}{V_{DB}'}}}$$

$$HD3 \cong \frac{3}{32} \frac{C_L C_{j0}}{\left(C_L + C_{j0}\right)^2} \left(\frac{V_{in}}{V_{DB}'}\right)^2$$

If $V_{DB}' = 1$, and Vin = 200 mV

$V_{DB}' = 2\phi_f + V_{DB}$ is larger than 1 V.

$$HD3 \le \frac{1}{250} \frac{C_L C_{j0}}{\left(C_L + C_{j0}\right)^2}$$

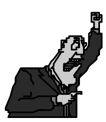
$C_L > 3\ C_{jo}$ → HD3 < 0.1 % !!

Rule of thumb: design such that $C_L > 5\ C_{jo}$

VERY LOW FREQUENCY FILTERS

The Design of Analog Circuits below 100 Hz is not trivial

- **RC > 0.001 sec**
  - if C = 10 pF then R > 100 MOHMS
  - for a 1 Hz filter (pace makers and other applications)
  - C = 10 pF, R = 628 GOHMS (Gm = 2 pA/V)
  - C = 1000 pF, R = 6.28 GOHMS (Gm = 2 nA/V)



WE NEED SMALL G_M

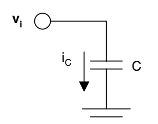
For the basic OTA-C integrator,

$$\tau = \frac{C_L}{g_m} \Leftrightarrow f = \frac{g_m}{2\pi C_L}$$

- We need very small g_m for very low frequency applications
- As an example, for τ = 1s and g_m = 16nA/V, C_L = 1.6 nF !!
- For a POLY-I POLY-II Capacitor (C/A ~ 600 aF/μm^2) in the AMI 1.2µ process, this means a Si area of 2.7 mm^2 !! => Impractical for IC's (Tiny chip area ~ 4 mm^2)

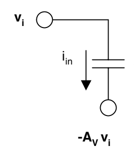
IMPEDANCE SCALERS for realization of very large time constants

Single capacitor



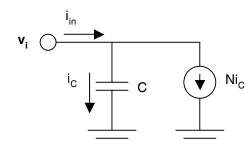
$$i_C = (sC)v_i$$

Voltage amplifier



$$i_{in} = (sC(1 + A_V))v_i$$

Current amplifier



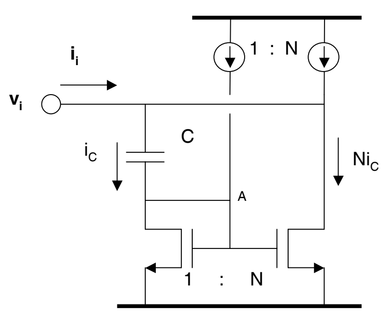
$$i_{in} = (sC(1 + N))v_i$$

Remarks:

➪**Voltage amplification is useless for low-voltage continuous-time applications.**

➪**Impedance scaler based on currrent amplification is precise for moderated N.**

CAPACITOR MULTIPLIER: CIRCUIT IMPLEMENTATION



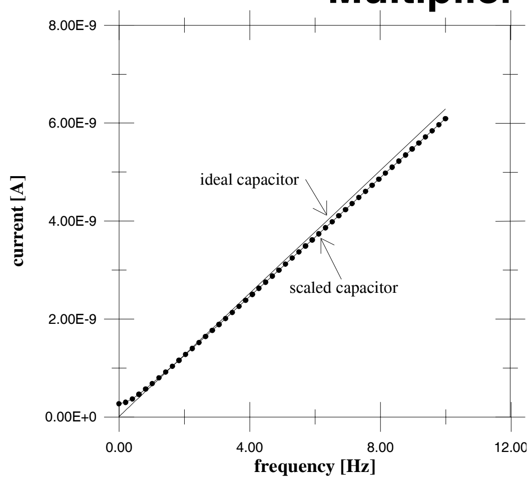
$$\frac{v_i}{i_i} = \frac{vi}{(N+1)i_C}$$

$$Z_{eq} = \frac{1}{s[(N+1)C]}$$

The following conditions must be satisfied:

➨**Low impedance at node A**
➨**Transistor output resistance might be neglected**
➨**Current gain is precise**

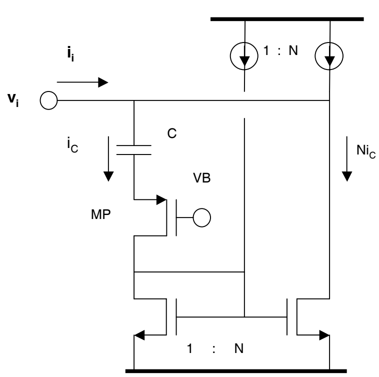
DESIGN EXAMPLE: Capacitor Multiplier



100 pF capacitor.

Scaled capacitor is 5 pF and N=19.

IMPROVING ITS FREQUENCY RESPONSE



Cascode transistor improves frequency response

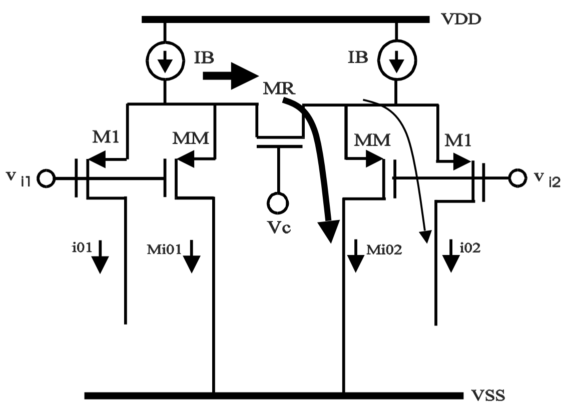
Design procedure:

MP is optimized for frequency.

N-type transistors are optimized for precision.

The loop must be stable

CURRENT DIVISION PRINCIPLE PLUS SOURCE DEGENERATION TRANSCONDUCTANCE

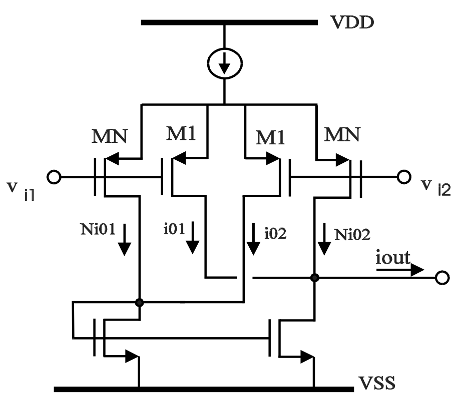


$$HDn \propto \left(\frac{v_{i1}-v_{i2}}{V_{GS}-V_T}\right)^n$$

$$\frac{i_{01}-i_{02}}{v_{i1}-v_{i2}} \cong \frac{2\mu C_{OX}}{M+1}\frac{W_R}{L_R}(V_{GS}-V_T)$$

$$M=\frac{\left(\frac{W}{L}\right)_m}{\left(\frac{W}{L}\right)_1}$$

CURRENT CANCELLATION PRINCIPLE



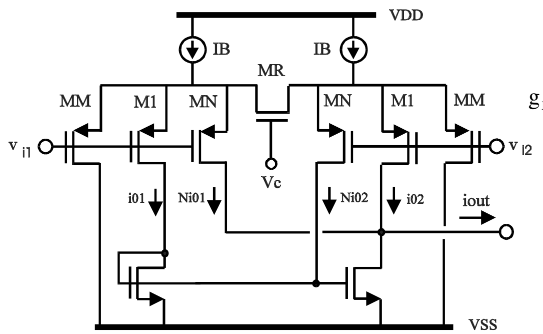
$$G_m = \frac{N-1}{N+1} g_{m1}$$

PARTIAL POSITIVE FEEDBACK !

LINEAR RANGE IS LIMITED!

GOOD NOISE PERFORMANCE

OTA FOR VERY LOW-FREQUENCY APPLICATIONS



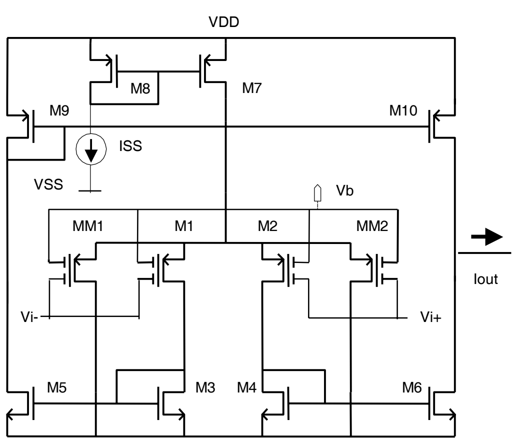
$$g_m = \frac{2(N-1)}{M+N+1}\left(\mu C_{OX}\frac{W_R}{L_R}(V_{GS}-V_T)\right)$$

$$M = \frac{\left(\frac{W}{L}\right)_{MM}}{\left(\frac{W}{L}\right)_{M1}}$$

$$N = \frac{\left(\frac{W}{L}\right)_{MN}}{\left(\frac{W}{L}\right)_{M1}}$$

3 techniques are used:
- **Source degeneration**
- **Current splitting**
- **Current cancellation**

Floating Gate plus Current Division OTA



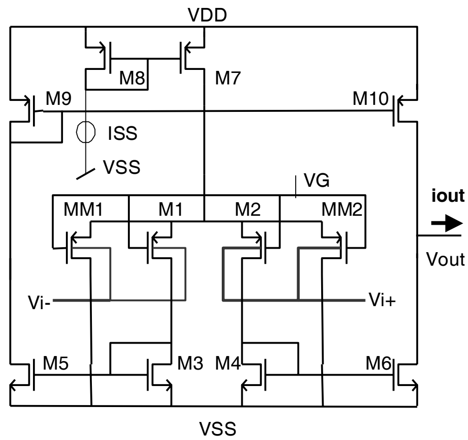
$$g_m = \frac{\alpha}{M+1}(g_m)$$

$\alpha < 1$ due to floating gates

Floating gates improve linearity but:
DC-offset might be an issue
Signal attenuation: noise might be an issue as well

VSS

Bulk Driven plus Current Deviation OTA



Bulk-Driven & Current splitting

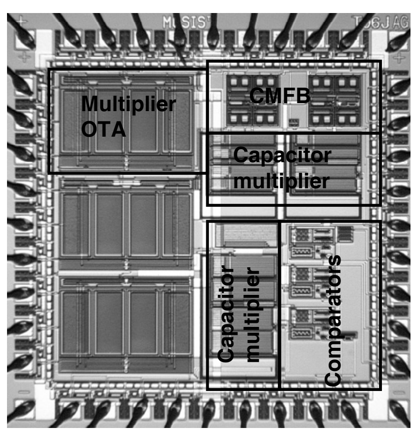
Input impedance is not very large: important issue for low frequency applications.

Gm-bulk is 4-5 times smaller than normal gm.

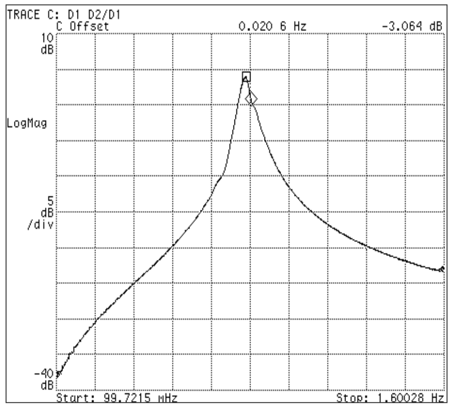
Large parasitic capacitors (well-substrate)

PSRR might be limited

Chip microphotograph of a sub-hertz filter/oscillator (JSSC-Aug-2002).

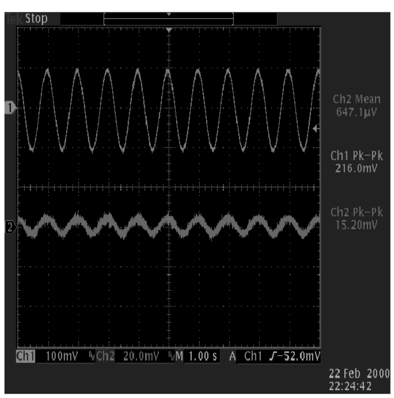


Experimental results for a 0.8 Hz BPF (0.5 μm CMOS technology).

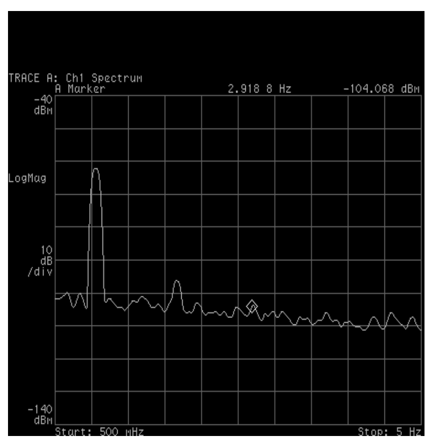


BULK DRIVEN OTA

EXPERIMENTAL RESULTS (0.5 μm)



Input Ch1 214mVpp @ 1 Hz
Output Ch2 15.2mVpp



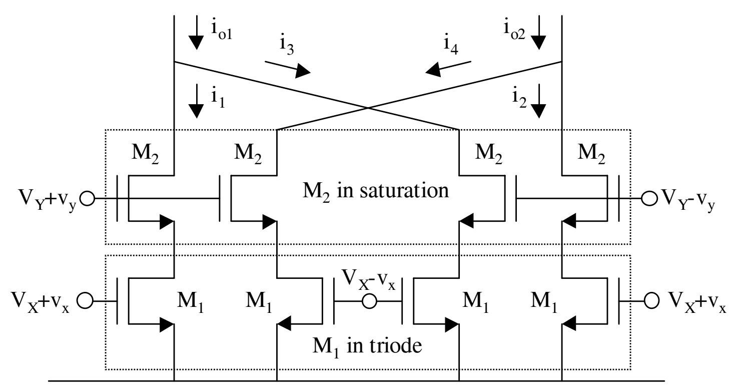
THD ~ -39dBm ~ 1.1% @214mVpp, 1Hz

EXPERIMENTAL RESULTS FOR THE DIFFERENT OTA DESIGNS

PARAMETER	REFERENCE	SD+CD	FG+CD	BD+CD
G_M (nA/V)	9.4	9.3	9.2	9.4
HD_3(%)	0.9@162mV$_{pp}$	1.0@242mV$_{pp}$	1.1@330mV$_{pp}$	0.9@900mV$_{pp}$
Input noise (μVrms)	18.1	26.1	39.1	104.7
SNR@1%HD_3(dB)	69.9	70.3	69.5	69.6
I_{BIAS}(nA)	2.6	120	232	560

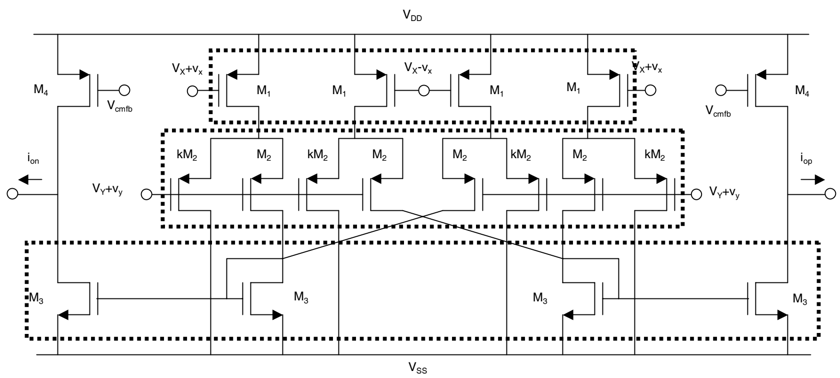
Key:
SD source degeneration
CD current division
FG floating gate
BD bulk driven

OTA for Low-Frequency Applications with Current cancellation techniques



Basic circuit is an analog multiplier

OTA with current splitting and cancellation techniques



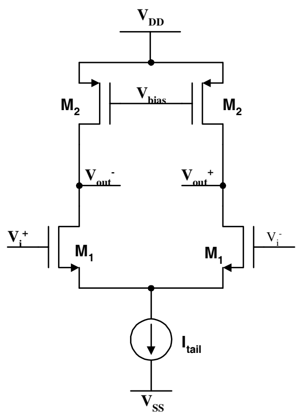
Basic circuit is an analog multiplier

Fully Differential and Pseudo Differential OTAs

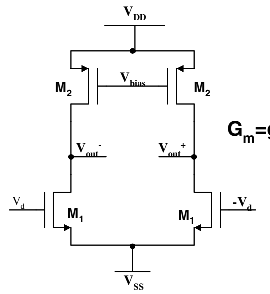
Common - Mode Feedforward and
Feedback strategies needed for
differential output filters

Fully Differential OTA Characteristics

Simple Differential OTA
With tail Current Source

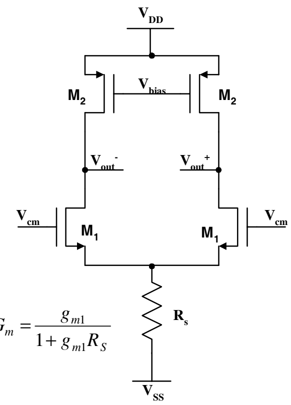


Differential-Mode



$G_m = g_{m1}$

Common-Mode

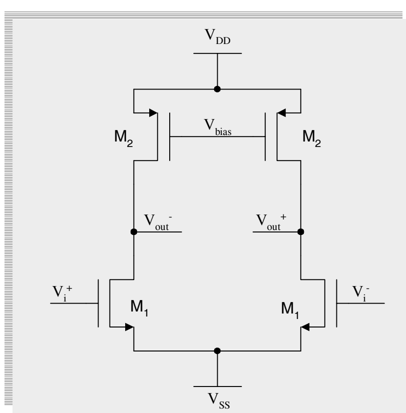


$$G_m = \frac{g_{m1}}{1 + g_{m1} R_S}$$

✘ Limited linear input range
✘ Limited tuning range

✔ Reasonable Common-mode gain
✔ Reasonable PSRR

Pseudo Differential Transconductance

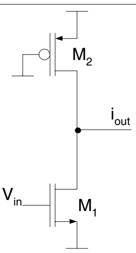


Simple Pseudo Differential OTA

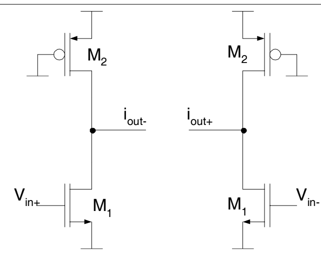
❑ Advantages
✔ Suitability for low voltage
✔ Wider common-mode input range

❑ Disadvantages
✘ Poor common-mode gain
$A_{CM}=A_{DM}>>1$
✘ Poor PSRR
✘ Low output impedance
✘ Need for fast and strong Extra CMFB Circuit to
(1) Fix output common-mode voltage
(2) Suppress common-mode signals

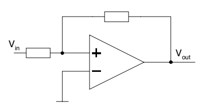
Derivation of CMFF Pseudo-differential **OTA**

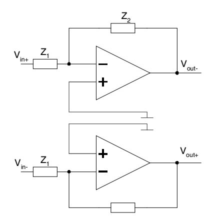


Single ended OTA circuit

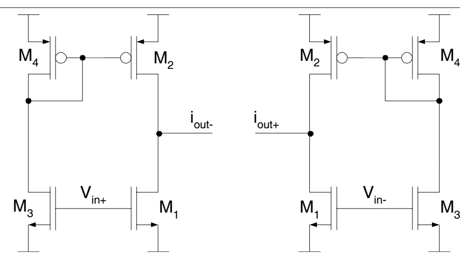


Circuit of OTA for differential input

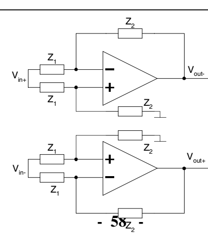




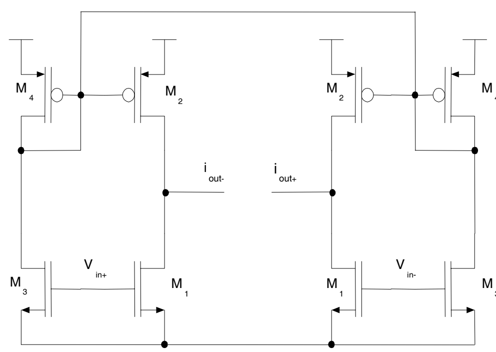
CMFF Techniques for Pseudo-differential **OTAS**

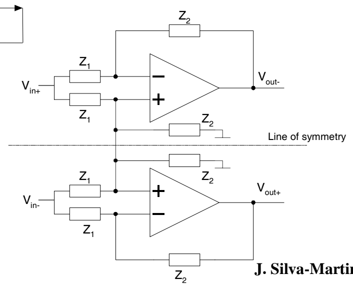


Circuit of OTA for common mode signals

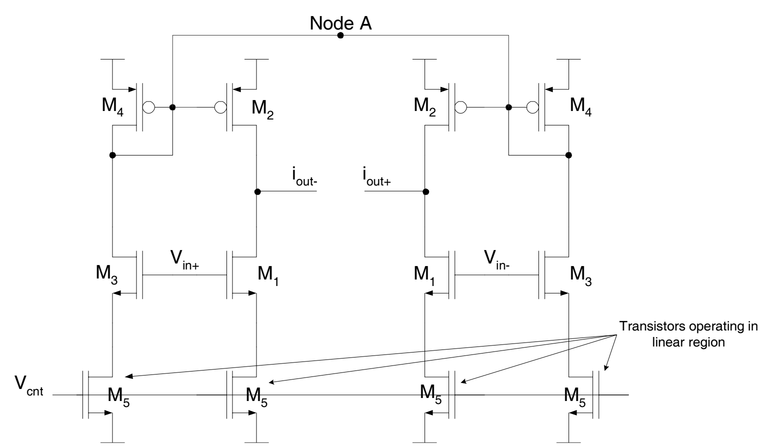


Fully-balanced, fully-symmetric CMFF OTA



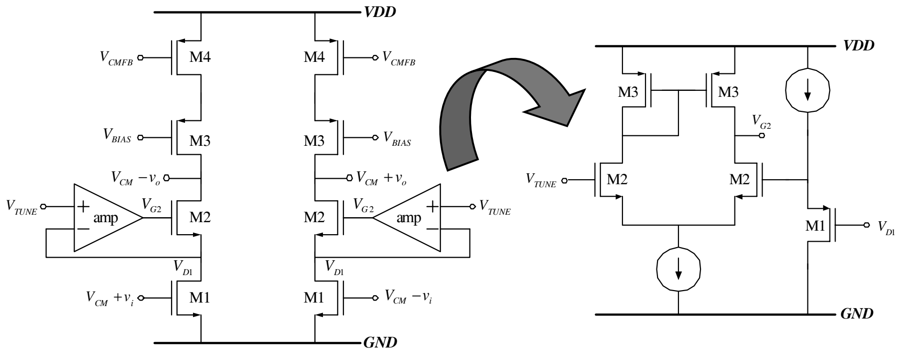


OTA with improved linearity



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

Pseudo-Differential OTA with large output impedance

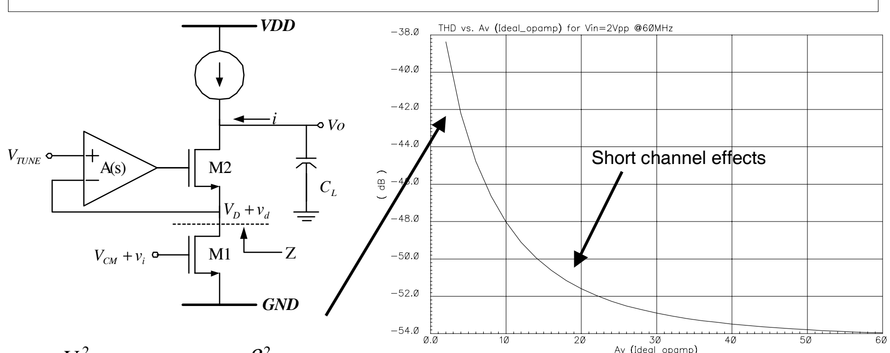


Pseudo-differential OTA

RGC amplifier "amp"

- Transistors M1 operate in Linear region: Wide linear range; Large tuning range;
- RGC loop: Fix V_{DS1} ➨ provides better linearity.

OTA Design Issues: RGC Loop

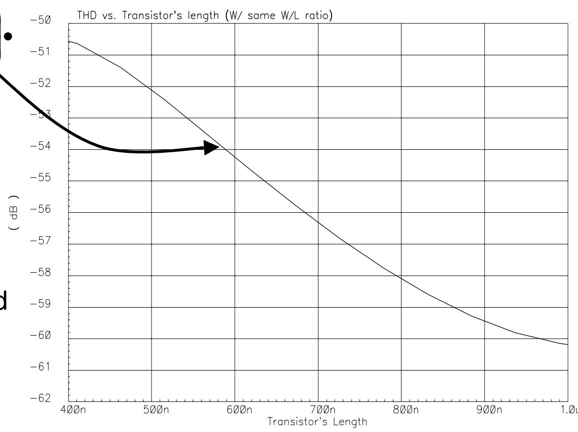


$$HD3 \approx \frac{V_i^2}{4} \cdot \frac{\beta^2}{[A(s)g_{m2} + \beta(V_{CM} - V_{TN} - V_D)]^2}$$

- OTA Gm's Linearity: Limited by how well the drain voltage of the input transistor is fixed.

Design Issues: Short-Channel Effects

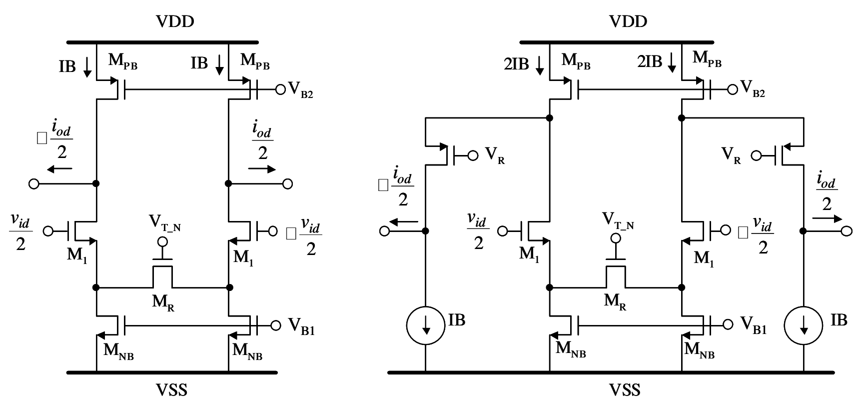
$$\mu_{eff} = \mu_0\left(\frac{1}{1+\theta(V_{GS}+v_{gs}-V_T)}\right)\bullet\left(\frac{1}{1+\frac{1}{L\varepsilon_c}\cdot(V_{DS}+v_{ds})}\right)$$



- ☛ Transversal electric field's effect is reduced by RGC loop;
- ☛ Trade-off:

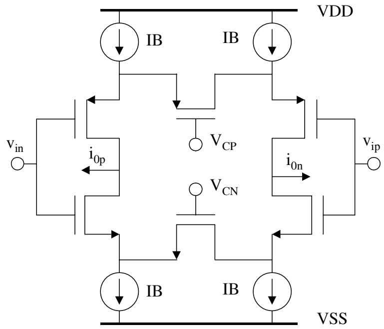
THD vs. Frequency response

TYPICAL OTA ARCHITECTURES WITH SOURCE DEGENERATION



$$G_m = \frac{g_{m1}}{1+N_n}\sqrt{1-\left(\frac{v_{id}}{2(1+N_n)V_{DSAT1}}\right)^2} = \frac{\mu_n C_{OX} V_{DSAT1}\left(\frac{W}{L}\right)_1}{1+N_n}\sqrt{1-\left(\frac{v_{id}}{2(1+N_n)V_{DSAT1}}\right)^2}$$

OTA based on Complementary Differential Pairs



TRANSCONDUCTANCE:

$$G_m = \frac{g_{mn}}{1+N_N} + \frac{g_{mp}}{1+N_P}$$

DISTORTION:

$$HD3 \cong \frac{1}{32}\left(\frac{V_{in}}{(V_{GS}-V_T)(N+1)}\right)^2$$

OUTPUT RESISTANCE:

$$R_{out} \approx \frac{g_{mn} r_{on} R_N^2}{2} \Bigg\| \frac{g_{mp} r_{op} R_P^2}{2}$$

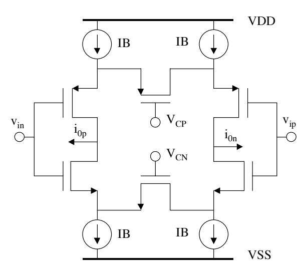
OUTPUT SWING IS LIMITED:

$$V_{IN} - V_{TN} < V_{OUT} < V_{IN} + V_{TP}$$

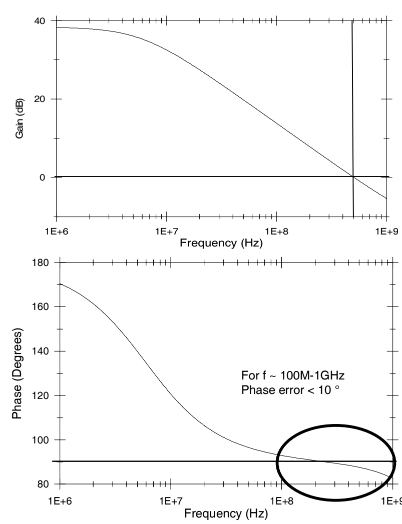
DC GAIN:

$$A_V \approx A_{VN}$$

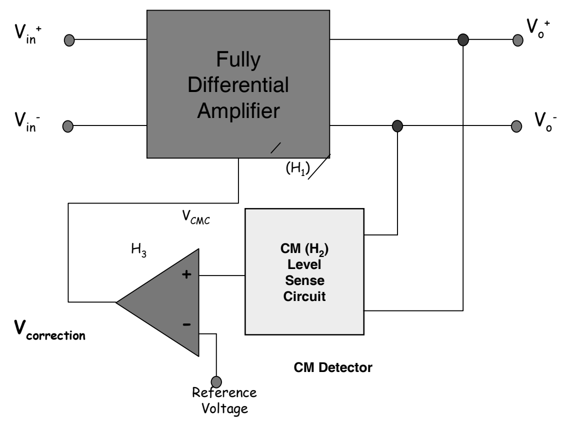
OTA Results



Transistor	W, L	Bias current
MP	200, 0.6 µm	200 µA
MCP	15, 0.6 µm	0 µA
MN	200, 0.6 µm	200 µA
MCN	30, 0.6 µm	0 µA

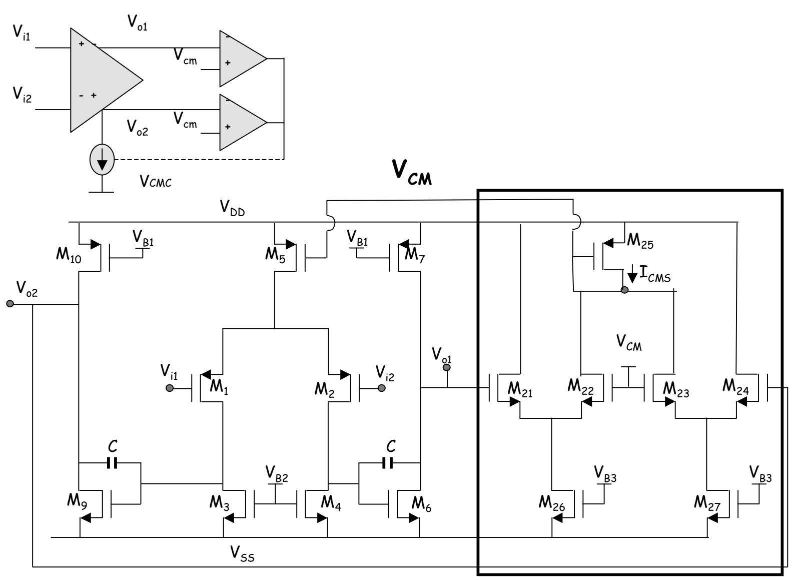


Common-Mode Feedback Circuits

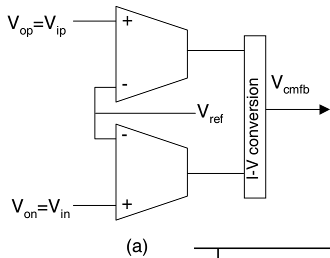


Conceptual Architecture of Common-Mode Feedback Loop

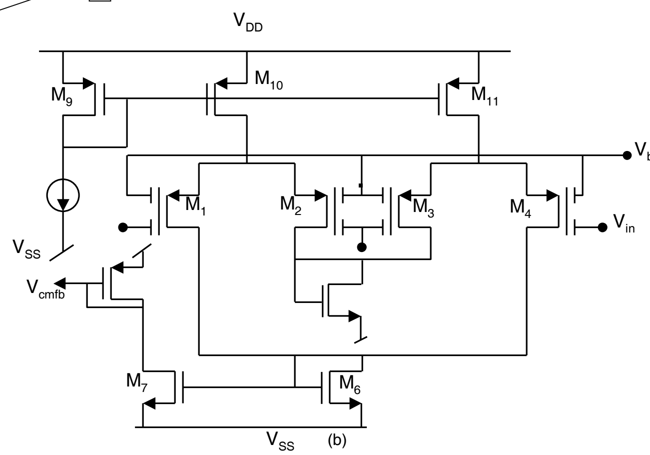
Example of a compensated Op Amp and a CM sense circuit



Op Amp **CM Detector**



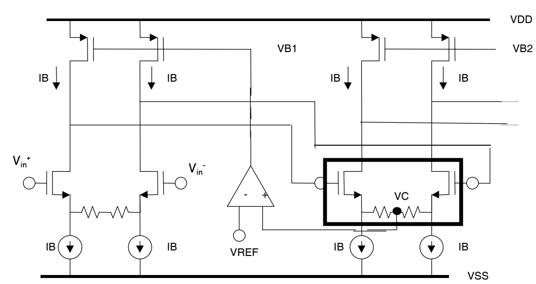
Common-mode feedback circuit.
(a) Block diagram. (b) Circuit using floating gates.

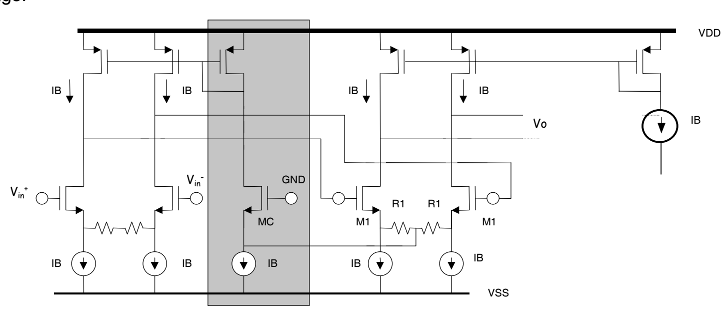


CMFB FOR OTAS: Solution 1

Typical OTA connection in pseudo- differential OTA-C based circuits.

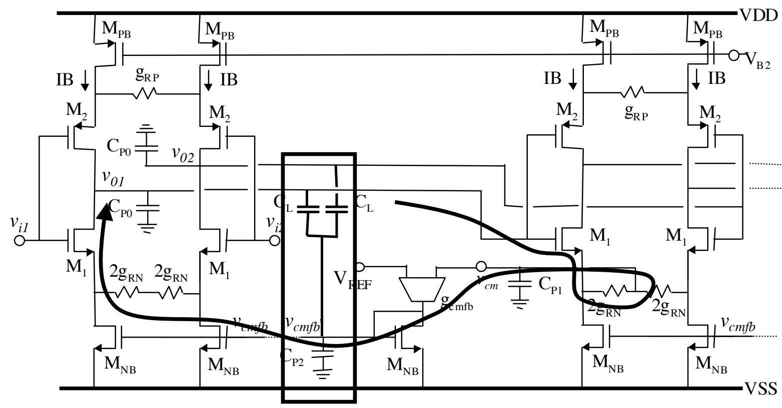
The common-mode voltage is obtained from the input of the following stage.



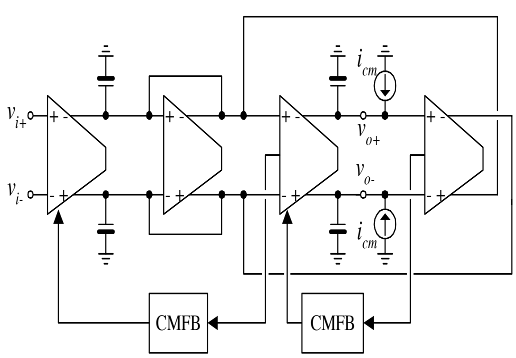


Pseudo-differential OTAs including the CMFB for the first one with good PSRR

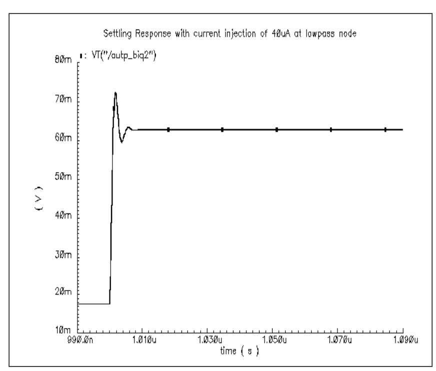
COMMON-MODE FEEDBACK USING BYPASSING THE LOW-FREQUENCY SECTION



Testing the COMMON-MODE FEEDBACK in a biquadratic filter



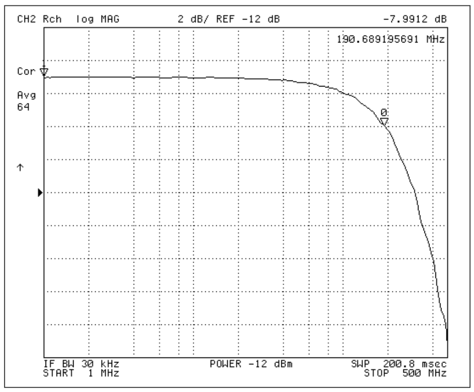
Biquadratic section with a common-mode input current: pulse of 40 mA.

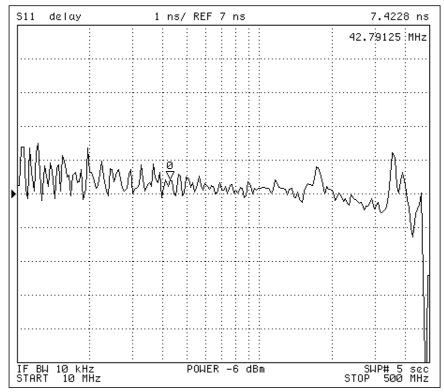


Pulse response of the CMFB embedded in a biquadratic section

Frequency response

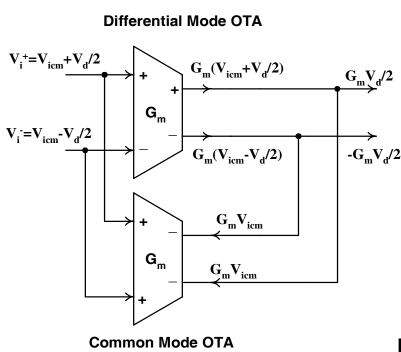
Magnitude and Group Delay

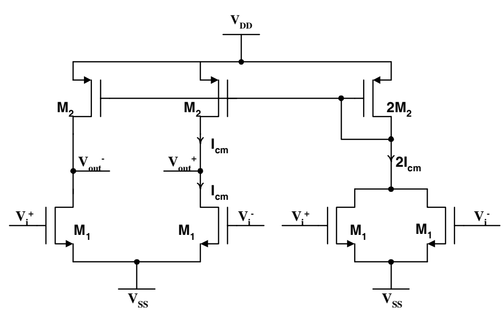




200 MHz linear phase filter: Group delay ripple is < +/- 1 nsec up to 400 MHz

A PD Solution 2

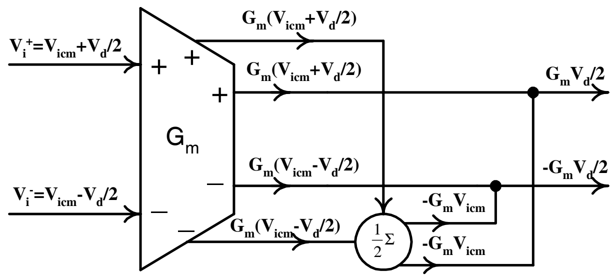




Pseudo differential OTA With CMFF

- ✔ CMFF is applied to cancel the common mode input signal
- ✘ Add load to the driving stage, input capacitance doubles
- ✘ CMFB is still needed

Proposed OTA Block Diagram (solution 3)



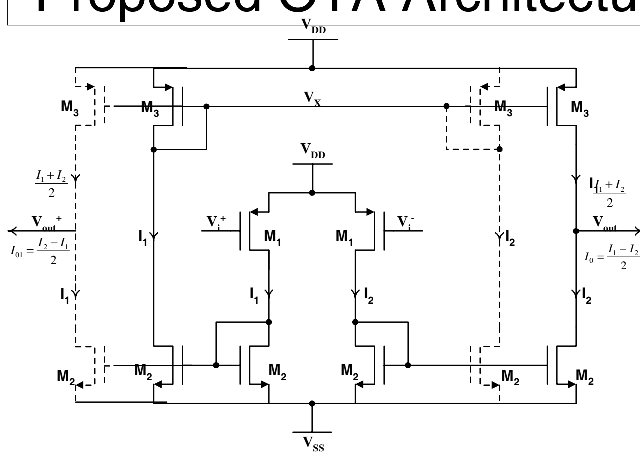
❑ Common-mode detection using the same differential transconductance by making copies of the current

✔ Input capacitance is not increased

✔ CMFF is inherently achieved

✔ CMFB can be easily arranged

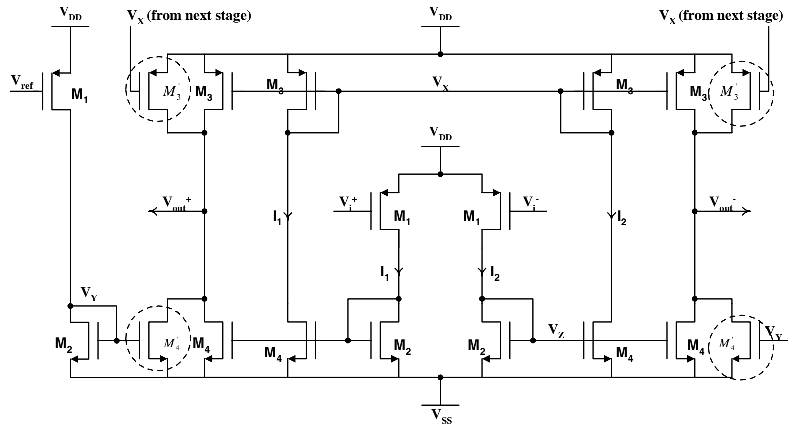
Proposed OTA Architecture



❑ Inherent common-mode detection

❑ Inherent common-mode Feedforward

[redacted]



- ❑ CMFB is arranged exploiting the direct connection of the OTAs
- ❑ Avoid using a separate common-mode detector
- ❑ Differential-mode signals and common-mode signals share basically the same loop

Small Signal Analysis

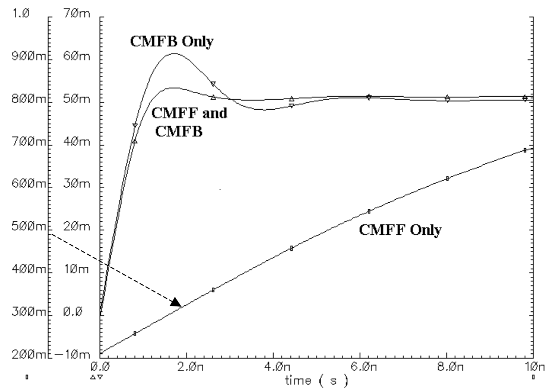
- ❑ The path from the differential signal to the ouput encounters one pole
- ❑ The other path is a common-mode path

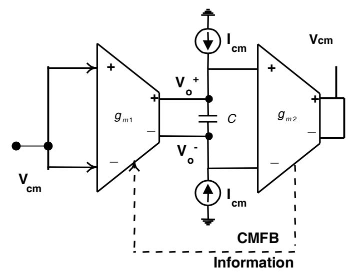
$$g_m(s) = \frac{i_{od}}{v_d} \cong g_{m1}\frac{g_{m2}}{g_{m2} + sC_Z} = \frac{g_{m1}}{1 + s/\omega_{nd}} \qquad \omega_{nd} = \frac{g_{m2}}{C_Z}$$

$$\Delta\phi \cong -\tan^{-1}(\omega/\omega_{nd1})$$

$$\min V_{DD} = \max\{(V_{TN} + V_{ov1} + V_{ov2} + V_{peak}), (|V_{TP}| + V_{ov3} + V_{ov4})\}$$

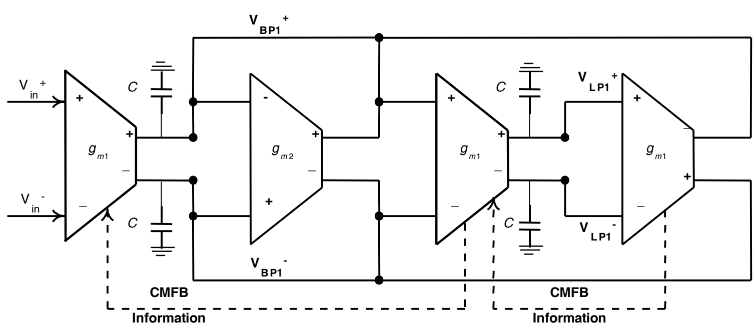
Simulation Results



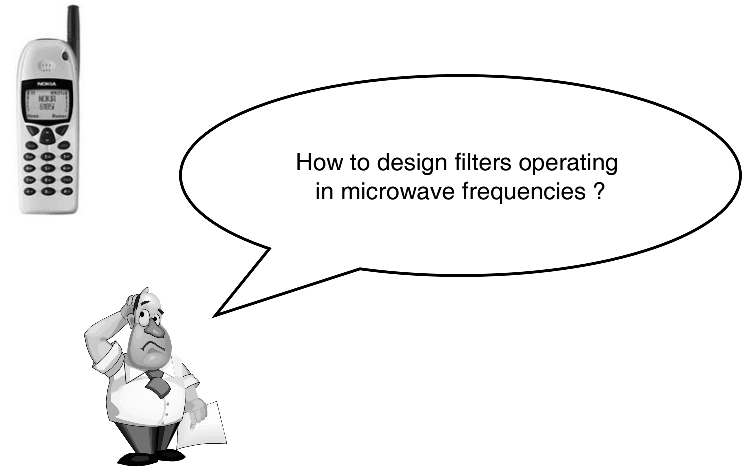


Output voltage applying common-mode current step (I_{cm})

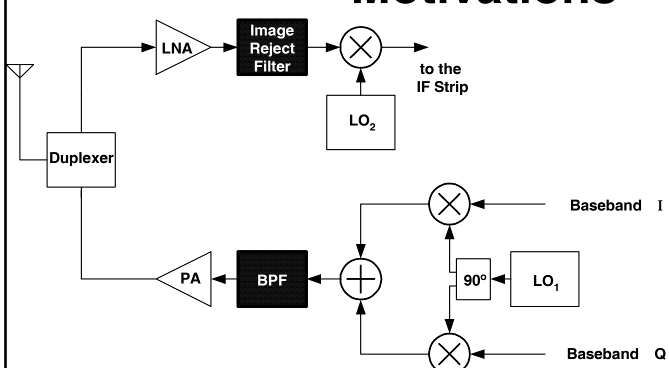
How to use OTAs as CM Detector?



- A 2nd Order Filter is used as an example
- Exploit direct connection of the cascaded OTAs in the filter
- Differential OTA used as CM detector also

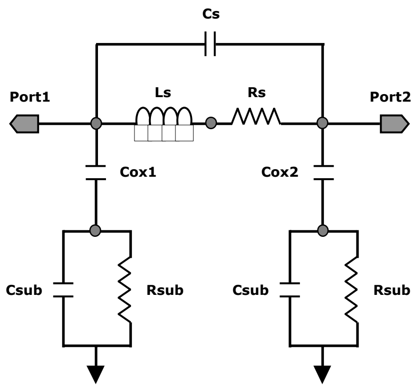


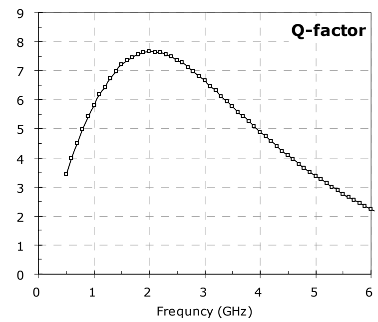
Problem Definition and Motivations



- Transceiver front-ends are the sections where the signal processing is performed at the frequency of the transmitted and received signals.
- Frequency range of interest is from 900 MHz to 2.4 GHz.
- The BPF is off-chip for high performance increasing the cost and the form factor.

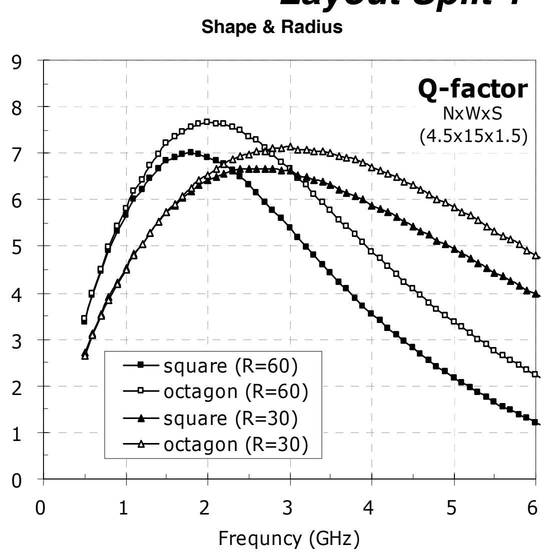
Equivalent Circuit & Calculation

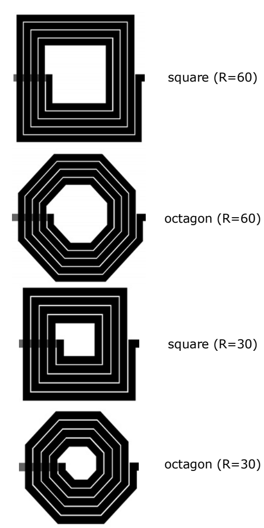




Equivalent Circuit **Parameter Calculation**

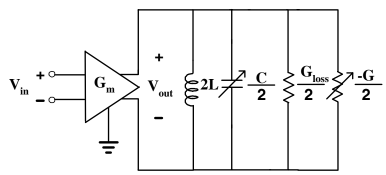
Layout Split 1



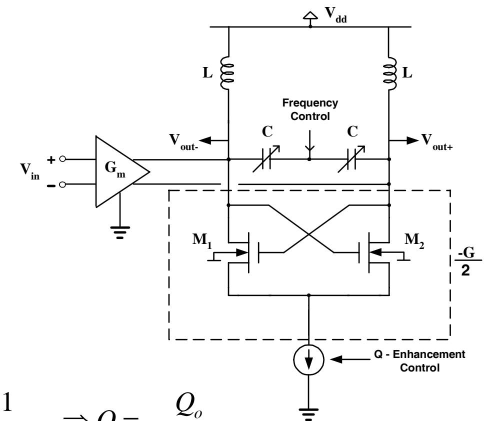


- **Shape : Octagon > Square**
- **Radius : 60 > 30**

Q-Enhancement Bandpass Filter



(a)



(b)

$$Q_o \equiv \frac{1}{\omega_o L G_{loss}} \Rightarrow Q = \frac{Q_o}{1 - \dfrac{G}{G_{loss}}}$$

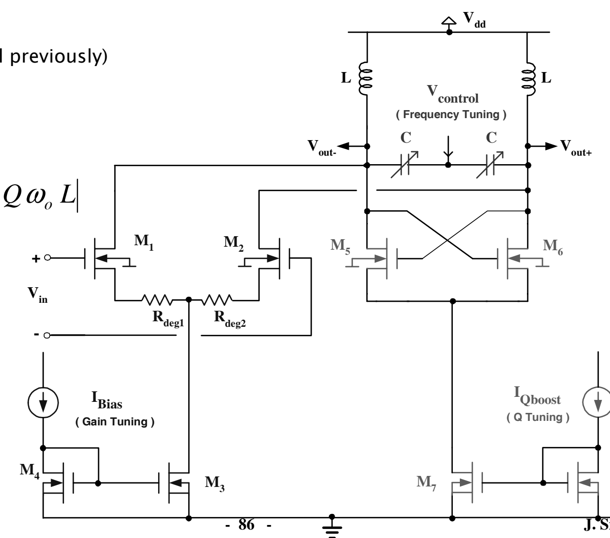
$$(G_{loss} > G \quad for\ stability)$$

A CMOS Programmable RF Bandpass Filter

Programmable in:
- Peak Gain (not exploited previously)
- Filter Q
- Center Frequency

$$|H(j\omega_o)| \cong |G_m(j\omega_o) Q \omega_o L|$$

$$\omega_o \cong \frac{1}{\sqrt{LC}}$$



A CMOS Programmable Bandpass Filter

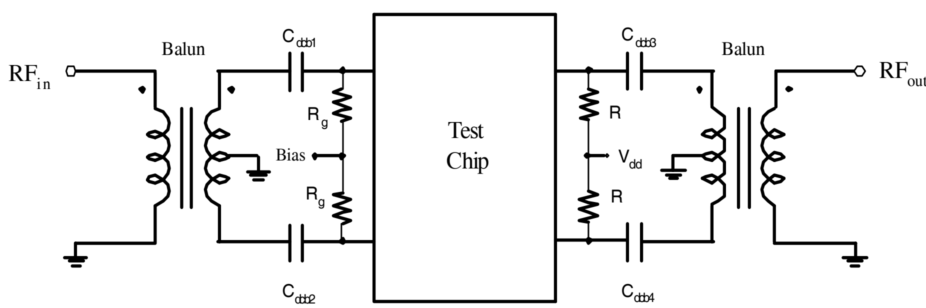
- The peak gain programmability through the input G_m stage.

$$\left|H(j\omega_o)\right| \cong \left|\frac{G_m(j\omega_o)}{G_{loss}-G}\right| = \left|G_m(j\omega_o)\,Q\,\omega_o\,L\right|$$

- Increasing Q also increases the peak gain.

- If ω_o and Q are fixed, the peak gain can be modified through G_m.

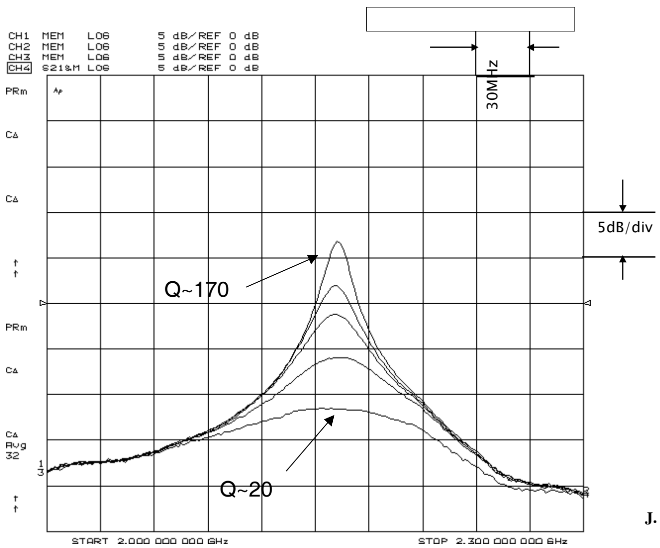
The Test Setup



- The measurements were performed on chips enclosed in TQFP44.
- All the measurement results to be presented except the power consumption are of the entire combination of the filter, output buffer and external passive circuitry providing DC bias and impedance matching.

Measured Q-Tuning

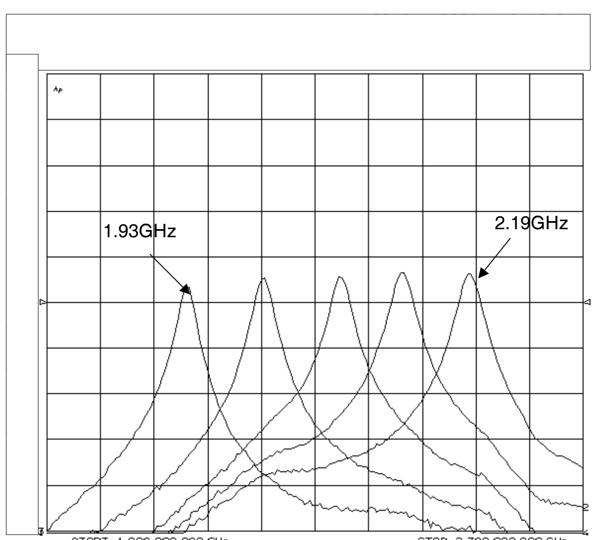
More than 3 octaves at f_o=2.16GHz



E. Sanchez-Sinencio J. Silva-Martinez

Measured Frequency Tuning

13% around 2.1GHz with Q~100



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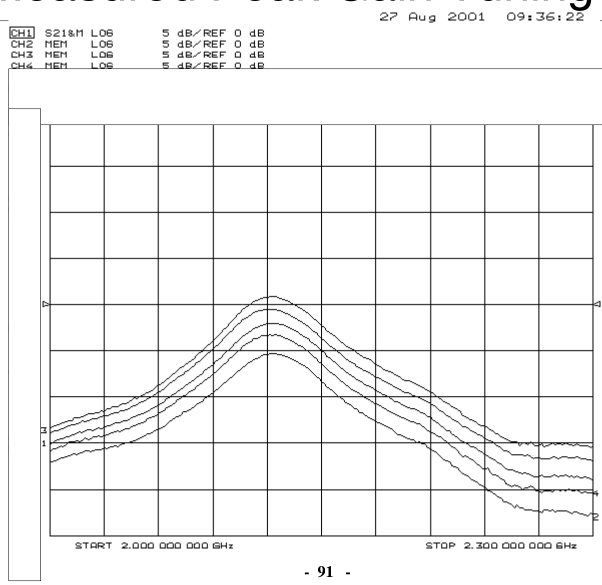
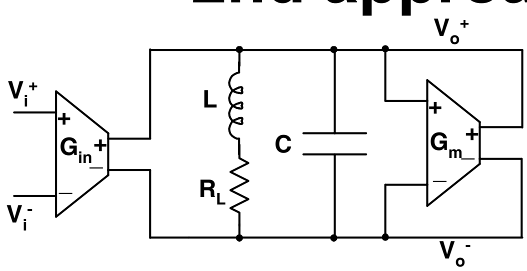
Measured Peak Gain Tuning



Around 2 octaves with f_o=2.12GHz and Q=40

Providing gain at the ω_o of an image-reject filter is useful in a receiver front-end after the LNA, to relax the NF spec of the mixer.

2nd approach



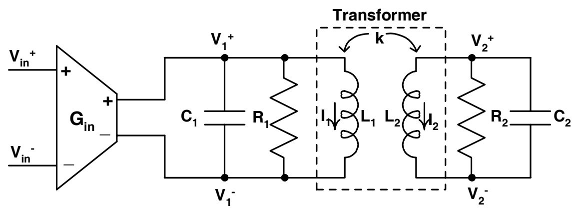
$$\omega_0 \cong \frac{1}{\sqrt{LC}}\sqrt{1-\frac{1}{Q_0^2}}$$

$$Q \cong \frac{Q_0}{1-(G_m/G_{loss})}\sqrt{1-\frac{1}{Q_0^2}}$$

$$G_{loss} \cong R_L(\omega_o C)^2$$

- ☺ Replace off-chip filters
- ☺ Eliminate need of impedance matching
- ☺ Reduce power, area, and cost
- ☹ Integrated spiral inductors are lossy
- ☹ Positive feedback is needed for Q-enhancement

Dual Resonator Bandpass Filter

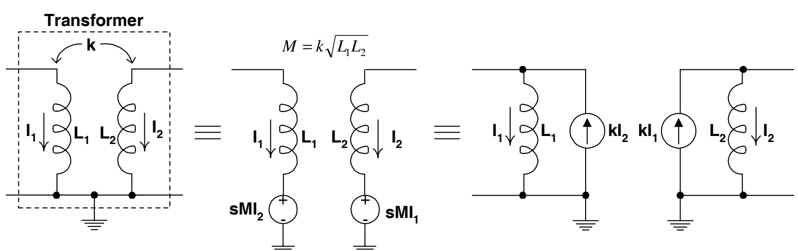


$$R = R_1 = R_2 = K_2 Q\sqrt{L/C}$$

- ❑ Two magnetically coupled resonators
- ❑ 4th order filter
- ❑ ω_0 is fixed by the LC product
- ❑ Coupling coefficient k is used to tune Q

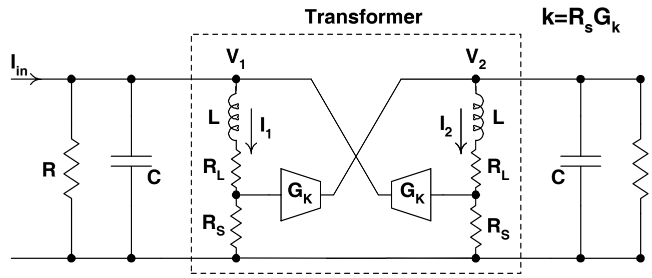
$$LC = (1/\omega_0^2)$$
$$k = K_1/Q$$

Transformer Models



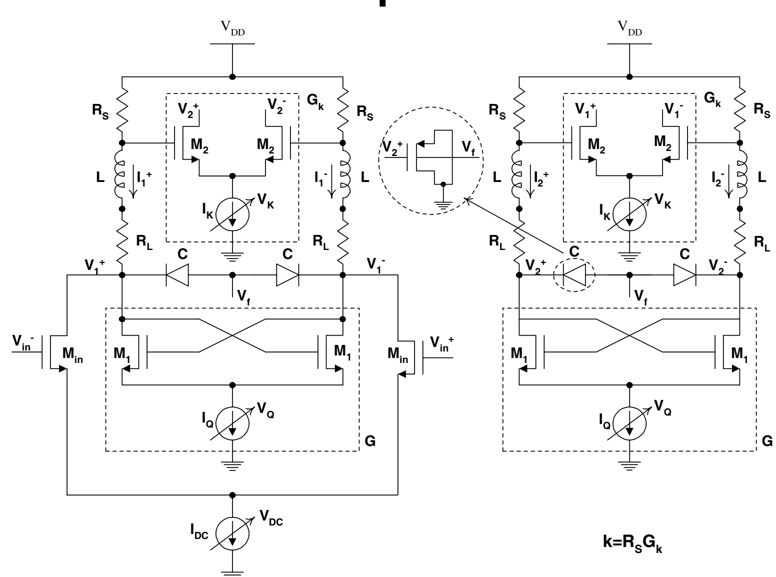
- ❑ Transformer can be replaced by induced currents
- ❑ Due to losses, induced currents are not in phase
- ❑ Severe passband ripples
- ❑ Coupling neutralization to maintain a flatband response
- ❑ Emulate the transformer action by electric coupling

Emulation of Magnetic Coupling

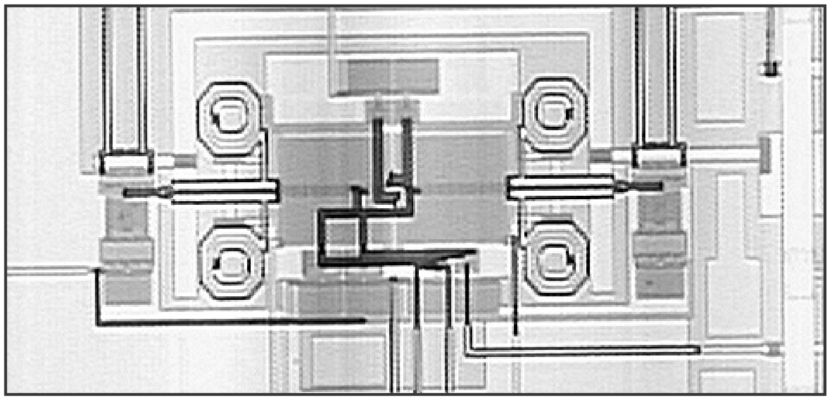


- ❑ $k=G_kR_s$(≅0.01)
- ❑ Provide possibility of BW tuning while maintain flatband
- ❑ Placing inductors apart to diminish magnetic coupling

Circuit Implementation

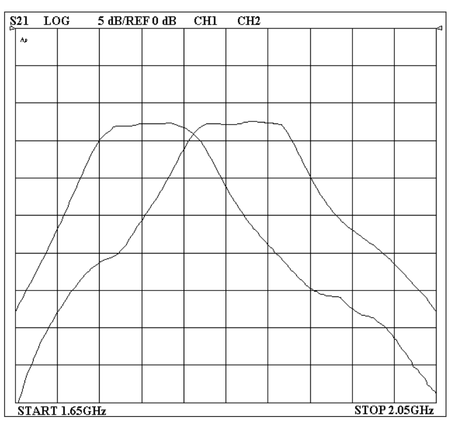


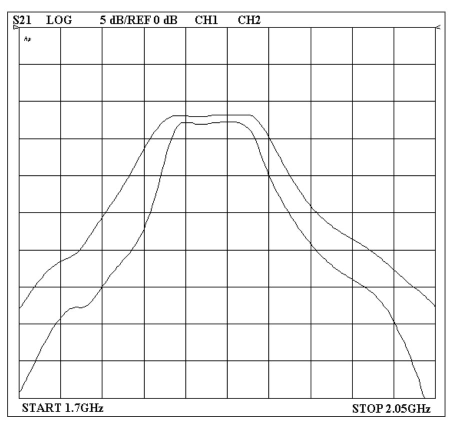
Chip Micrograph



X=500μm, Y= 300μm

Measurement Results

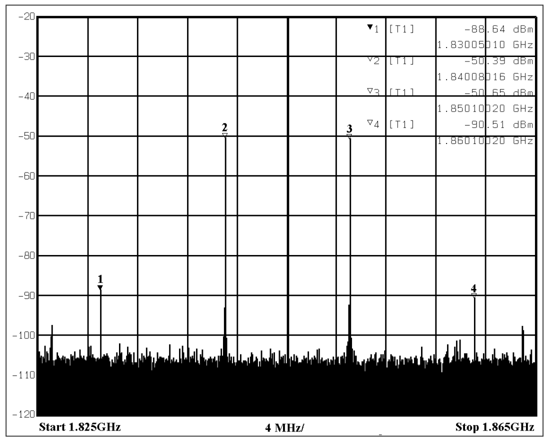




Frequency tuning
f_0={1.77GHz,1.86GHz}

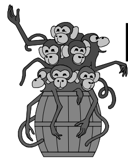
Bandwidth tuning
BW={70MHz,100MHz}

Measurement Results



Two Tone intermodulation distortion -34dBm each

Frequency- and Q-tuning techniques for OTA-C filters



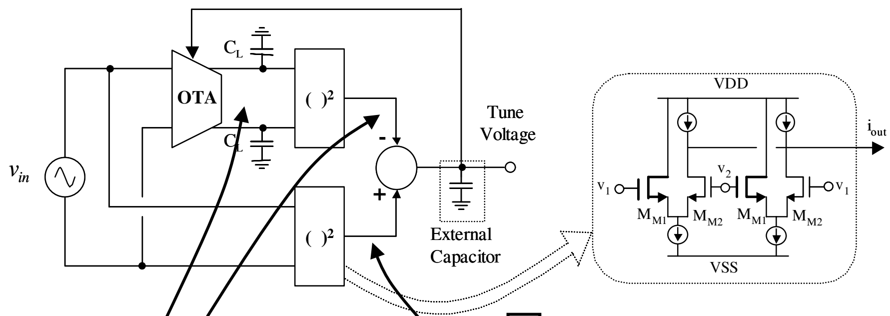
Need for Automatic Tuning

- Process variations can change f_o and Q by at least 20%
- Parameters also change with temperature and time(aging)
- Automatic tunining is a critical issue for the optimal performance of continuous-time circuits.

Methods of tuning

- Master-Slave
- Based on trigonometric properties
- Based on filter phase information
- Pre-tuning
- Burst tuning
- Switching techniques

Automatic Tuning Based on Power Comparison



$$v_o = \frac{G_m}{C_L}\int v_{in}\,dt = \left(\frac{f_u}{f_r}\right)V\cos(2\pi f_r t)$$

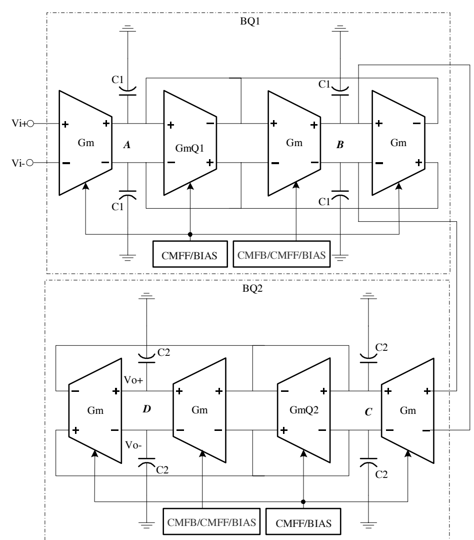
$$v_0^2 = \boxed{\frac{V^2}{2}\left(\frac{f_u}{f_r}\right)^2} - \frac{1}{2}\left(\frac{f_u}{f_r}\right)^2 V^2\cos(4\pi f_r t)$$

$$v_0^2 = \boxed{\frac{V^2}{2}} - \frac{1}{2}V^2\cos(4\pi f_r t)$$

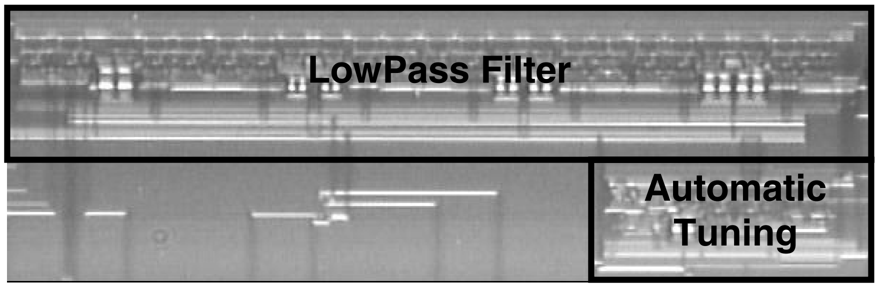
DC offsets are quite critical!!

4th-Order Equiripple Linear Phase Filter

- **Two biquads Gm-C filter;**
- **C1 & C2 are total capacitance at node A,B, and C, D;**
- **One CM control per node;**
- **Node A&C--Low impedance node: CMFF/Bias;**
- **Node B&D--CMFB/CMFF/Bias;**

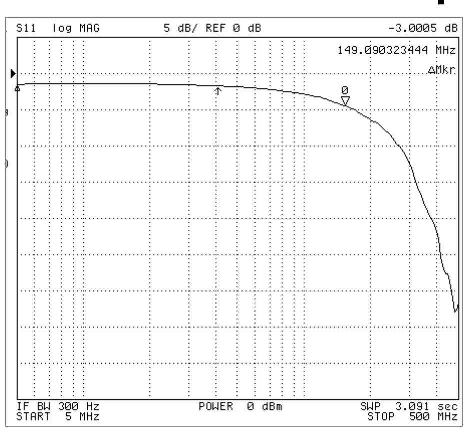


Microphotograph of the chip



CMOS 0.35 μm technology

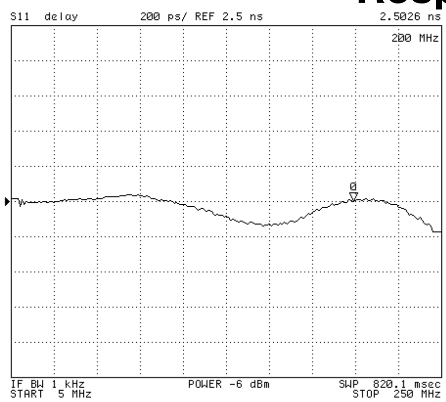
Experimental Results: Frequency Response



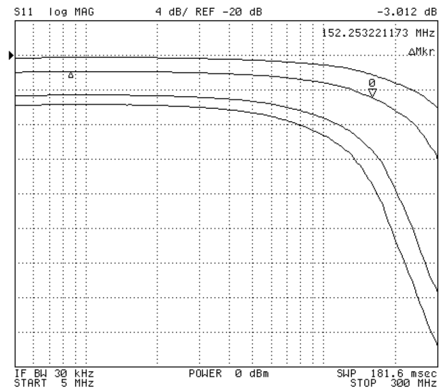
Filter's Magnitude Response (BW=150MHz)

Filter's Phase Response (BW=150MHz)

Experimental Results: Frequency Response



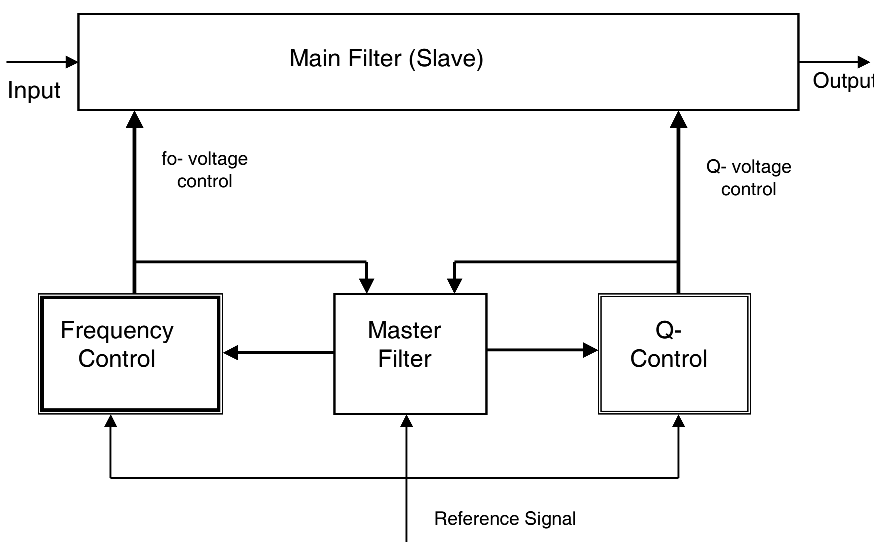
Filter's Group delay Response (BW=150MHz)



Filter's Magnitude Response
with automatic frequency tuning

Master-Slave Tuning Concept

Master-Slave Tuning Concept



Frequency Tuning

Phased-Locked Loop (PLL)

- Most widely used scheme
- Accurate (less than 1% error is reported)
- **Square wave input reference**
- Only Phase - Frequency Detector, and LPF are the additional components
- It may take a large area overhead

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

Q Tuning Schemes

Based on an envelope detector and a switched- capacitor integrator. It yields an accuracy of about 30%

Modified LMS

- Q-accurate of about 1%
- It does not use envelope detector
- **Square wave input, any periodic function is sufficient**
- Independent of frequency tuning

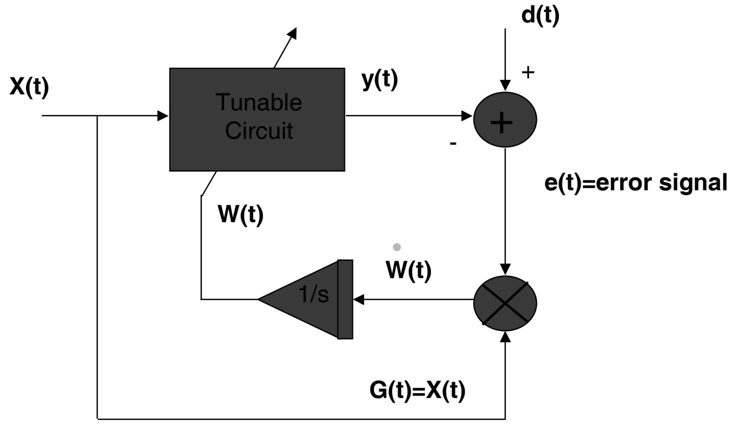
Adaptive LMS Algorithm: Introduction

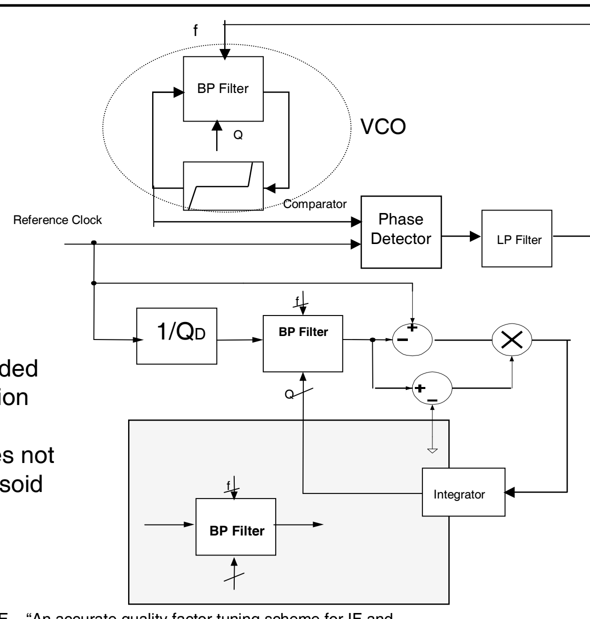
- Called Adaptive "Least-Mean-Squares" Algorithm because it learns by minimizing the *mean-square error (MSE)* between a desired response and the actual response of a system
- Minimizes error by updating system coefficients through a feedback loop

Adaptive LMS Algorithm: Theory

- Using the steepest descent algorithm to minimize the MSE we obtain:
- $\dot{W}(t) = k[d(t) - y(t)]G(t) = k[e(t)]G(t)$
  - W(t) = tuning signal
  - d(t) = desired system output
  - y(t) = actual system output
  - G(t) = tuning gradient (partial derivative of y(t) with respect to W(t))
  - k = adaptation constant

LMS Algorithm: Block Diagram (Linear System)

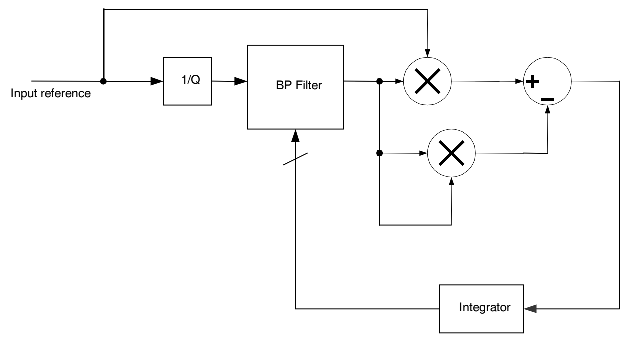




- Three Filters are needed
- An accurate attenuation 1/QD block is required
- Reference signal does not need to be a pure sinusoid

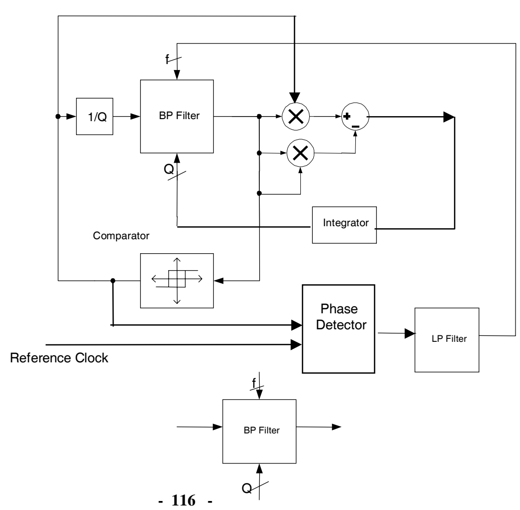
Stevenson, J.M.; Sanchez-Sinencio, E "An accurate quality factor tuning scheme for IF and high-Q continuous-time filters". IEEE Journal of Solid-State Circuits, Volume: 33 No.12 , Dec. 1998 , Page(s): 1970 -1978

An enhanced Q-tuning scheme

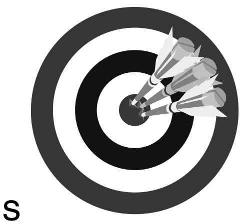


New implementation of modified-LMS Q-tuning scheme. Note that the LMS has been implemented in a different way yielding a structure with less offset voltages. See reference for more details.

The enhanced tuning scheme



Improvements over the previous Tuning scheme comparison



- Area overhead decreased

  (Previous scheme => 2 extra filters

  New scheme => 1 extra filter)

- Eases the matching restrictions

  (Previous tuning scheme => match 3 filters

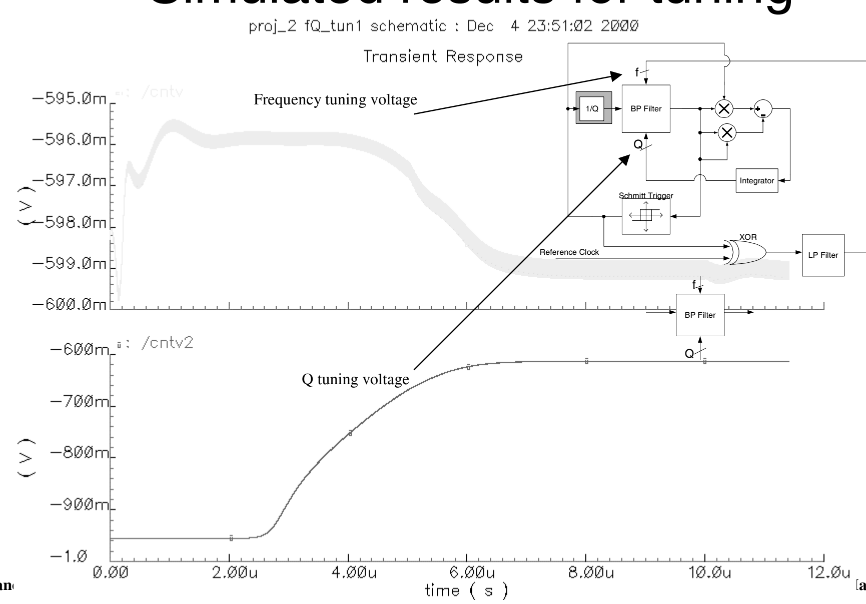
  New tuning scheme => match 2 filters)

- Improves accuracy of tuning

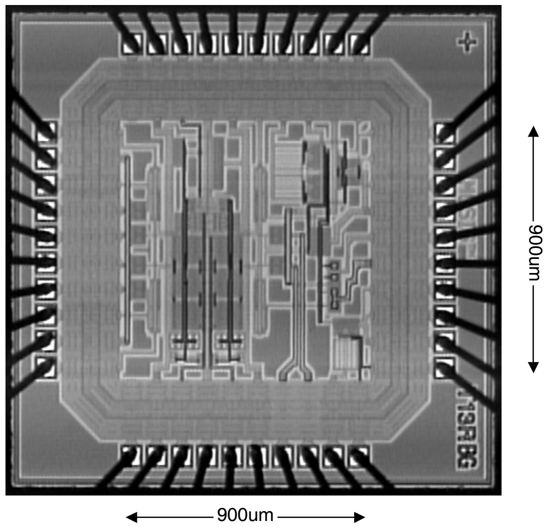
  (New tuning scheme is more tolerant to offsets than the previous one)

  The Q-tuning loop speed can be equal to the f_0-tuning loop for optimal performance

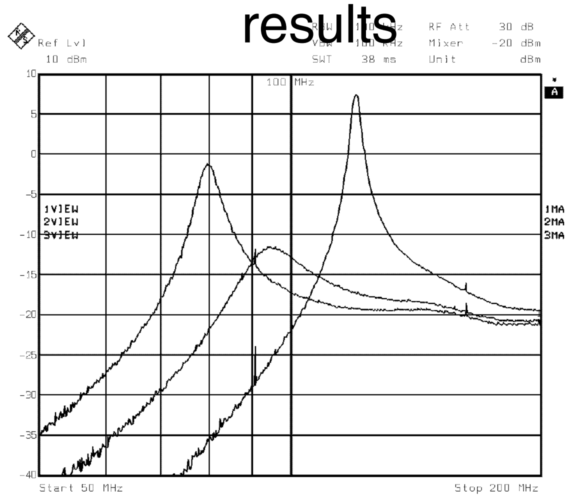
Simulated results for tuning



Die Photograph

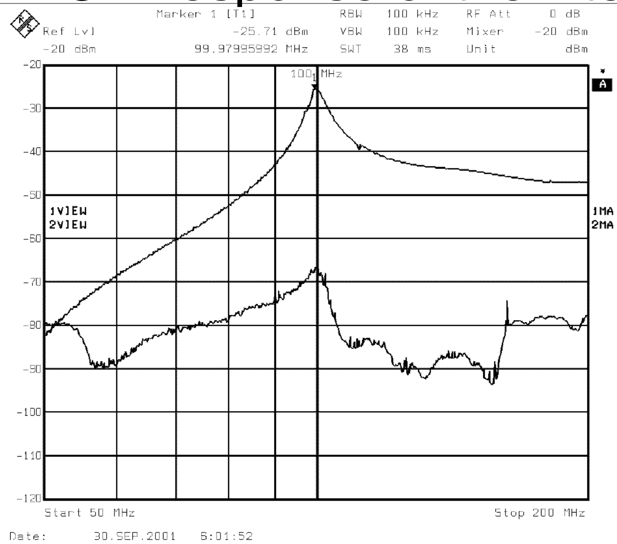


Experimental Q tuning range results



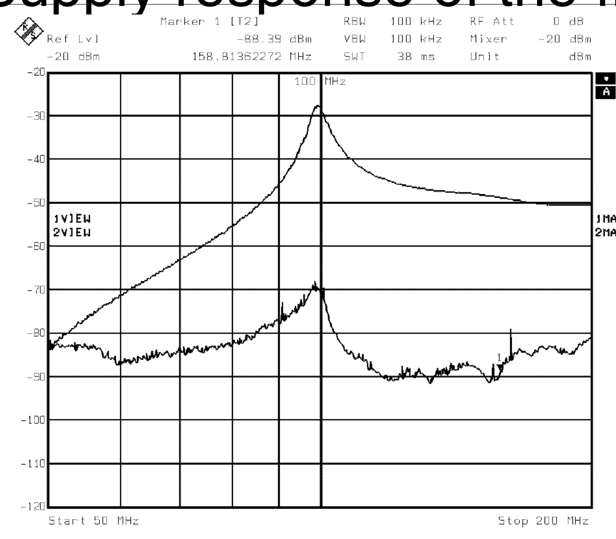
- Qs of 16, 5 and 40 at 80,95 and 110 MHz

DM-CM response of the filter



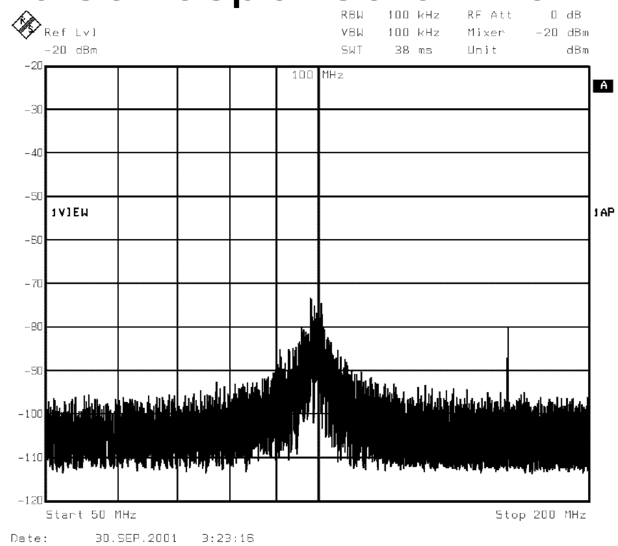
- CMRR is more than 40dB in the band of interest

Supply response of the filter



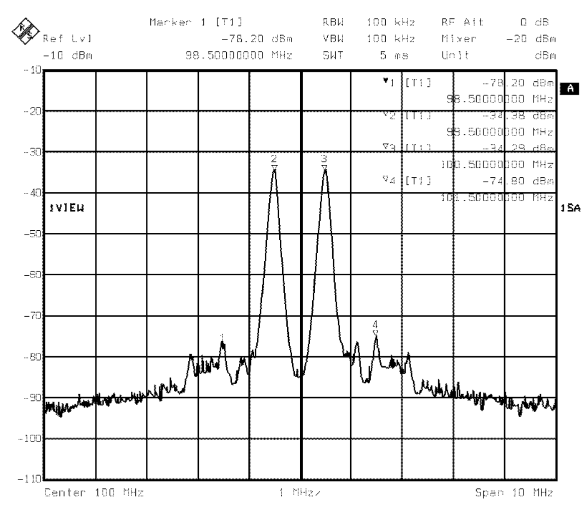
- $PSRR^-$ is more than 40dB in the band of interest

Noise response of the filter



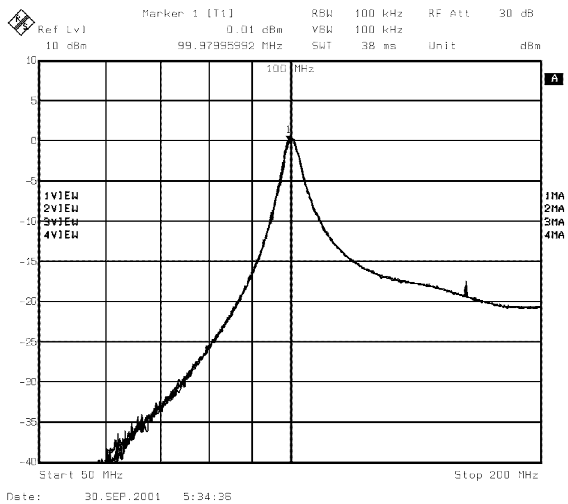
- Total integrated noise power at the output= -60dBm

Two-tone inter-modulation test



IM_3 of < -40dB when the input signal is 44.6mV

Filter response for four different ICs



- The tuning works!

Open problems

- Efficient solutions with SNDR> 70 dB for 1Vpk-pk are needed for wireline applications
  - frequency range 10 MHz - 50 MHz
  - Noise density < -148 dBm
  - IM3 ~ -65 dB A 1 Vpk-pk
  - Accuracy ~ 1-2 %
- Solutions for next generation of read channels
  - Frequency range 200 MHz - 800 MHz
  - Gain boosting ~ 20 dB
  - Accuracy ~ 5 %
  - Very low power and small area
- RF Filters
  - Small noise figure
  - High-Q solutions
  - Precision better than 1 % (efficient automatic tuning circuitry)

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