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

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

6185i digital cell phone from Nokia.
All multi media applications -- Anti aliasing before ADC and smoothing after DAC

CMP-35 portable MP3 player
LOSSY OTA-C INTEGRATORS
OTA-C Two Integrator Loop Filters

Two Integrator Loop Filters

Vin1 \rightarrow g_{m3} \rightarrow Vout

Vin2 \rightarrow g_{m4} \rightarrow Vo1

Vo1 \rightarrow g_{m2} \rightarrow Vo2

Vin1 \rightarrow C_1 \rightarrow Vin2

Vo3 \rightarrow C_2
How to generate the zeros of the filter?

Canonic OTA-C Biquad

\[ V_{01} = \frac{s^2 C_1 C_2 V_C + s C_1 g_{m2} c V_B + g_{m1} g_{m2} V_A}{s^2 C_1 C_2 + s C_1 g_{m2} + g_{m1} g_{m2}} \]
General Second-Order OTA-C Filter

For \( V_1 = V_2 = V_3 \)

\[
H(s) = \frac{C_2}{C_2 + C'_2} s^2 + s \left( \frac{g_{b1}}{C_2} + \frac{g_{b0}}{C_1} \right) + \frac{g_{m2}}{C_2} \frac{g_{m3}}{C_1 (C_2 + C'_2)} + \frac{g_{m1} g_{m2}}{C_1 (C_2 + C'_2)}
\]

\[
\omega_0^2 = \frac{g_{m1} g_{m2}}{C_1 (C_2 + C'_2)} \quad ; \quad \omega \frac{Z}{Z} = \frac{g_{b0}}{C_1} \frac{g_{m2}}{C_2}
\]

\[
\frac{\omega_0}{Q} = \frac{g_{m3}}{(C_2 + C'_2)} \quad ; \quad \frac{\omega}{Q} = \frac{g_{b1}}{C_2}
\]

\[
|H(0)| = \frac{g_{b0}}{g_{m1}} \quad ; \quad |H(\infty)| = \frac{C_2}{C_2 + C'_2}
\]
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 $C_1$ becomes $C_1/a$.

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:

\[ V_1 = \frac{1}{sC_1} g_{m1} (V_i - V_0) \]  \hspace{1cm} (1)

\[ V_0 = (g_{m2} V_1 - g_{m3} V_0) \frac{1}{sC_2} \]  \hspace{1cm} (2)

(1) into (2)

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

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

\[ \omega_0^2 = \frac{g_{m1}g_{m2}}{C_1C_2}, \quad BW = \frac{\omega_0}{Q} = \frac{g_{m3}}{C_2} \]

\[ Q = \frac{1}{g_{m3}} \sqrt{\frac{g_{m1}g_{m2}C_2}{C_1}} = \frac{C_2\omega_0}{g_{m3}} \]
Now let's assume the transconductance is characterized by:

\[ g_m = g_{mo} e^{-s/\omega_p} \approx g_{mo} \left( 1 - s / \omega_p \right) \text{ for } \omega_p \ll \omega_0. \]

Under this condition the excess phase can be expressed as \( \phi_E \approx \omega_0 / \omega_p \).

Note that ideally \( \phi_E = 0^\circ \).

Then,

\[
H_{LP}(s) = \frac{g_{mol} g_{mo2} \left( 1 - s / \omega_{p1} \right) \left( 1 - s / \omega_{p2} \right)}{s^2 C_1 C_2 + s C_1 g_{mo3} \left( 1 - s / \omega_{p3} \right) + g_{mol} g_{mo2} \left( 1 - s / \omega_{p1} \right) \left( 1 - s / \omega_{p2} \right)}
\]

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

Then the actual \( \omega_{oa} \) and \( BW_a \) become

\[
\omega_{oa}^2 = \frac{g_{mol} g_{mo2}}{C_1 C_2 + \frac{g_{mol} g_{mo2}}{\omega_{p1} \omega_{p2}} - \frac{C_1 g_{mo3}}{\omega_{p3}}}
\]

\[
BW_a = \frac{\omega_{oa}}{Q_a} = \frac{C_1 g_{mo3} - \left( \frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}} \right) g_{mol} g_{mo2}}{C_1 C_2 - \frac{C_1 g_{mo3}}{\omega_{p3}} + \frac{1}{\omega_{p1} \omega_{p2}}}
\]
Let us also assume that \( \omega_{p1} = \omega_{p2} = \omega_p \), then \( \omega_{oa} \approx \omega_o \), thus,

\[
\begin{align*}
\text{BW}_a &= \frac{\omega_{oa}}{Q_a} = \frac{C_1g_{m3} - \frac{2}{\omega_{p1}}g_{mo}g_{m2}}{C_1C_2 - \frac{C_1g_{m3}}{\omega_p} + \frac{1}{\omega_p^2}} \approx \frac{C_1g_{m3} - \frac{2}{\omega_p}g_{mo}g_{m2}}{C_1C_2} \\
\text{BW}_a &\approx \frac{g_{m3} - g_{mo}g_{m2}}{C_2} \cdot 2 = \text{BW} - \omega_{oa}^2 \cdot 2 = \text{BW} - \frac{2\omega_{oa}^2}{\omega_{p1}} \\
Q_a &= \frac{\omega_{oa}}{\text{BW}_a} \approx \frac{1 \cdot \omega_{oa}}{g_{m3} - \omega_{oa}^2} \frac{2}{\omega_{p1}} \\
Q_a &= \frac{C_2\omega_{oa}}{g_{m3}} - \frac{2\omega_{oa}}{\omega_{p1}} = \frac{Q}{1 - \frac{Q_2\omega_{oa}}{\omega_{p1}}} = \frac{Q}{1 - \frac{2\omega_{oa}}{\omega_{p1}}Q}
\end{align*}
\]

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

\[
Q_a \approx \frac{Q}{1 - 2\phi_E Q} \approx Q(1 + 2\phi_EQ)
\]

\( \text{BW}_a = \text{BW} - 2\omega_{oa}\phi_E \)
Furthermore, if $A_{vo} = g_m R_o$ is taken into account, then

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

If $A_{vo} = 500$

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

Note that:

- $Q_a \downarrow$ when $A_{vo} \downarrow$
- $BW_a \downarrow$ $Q_a \uparrow$ when $\phi_E \uparrow$

$$Q_a \approx \frac{Q}{1 - 2\phi E Q} \approx Q(1 + 2\phi_E Q)$$

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

Table:

<table>
<thead>
<tr>
<th>$Q$</th>
<th>$Q_a$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.996</td>
</tr>
<tr>
<td>5</td>
<td>4.902</td>
</tr>
<tr>
<td>10</td>
<td>9.6</td>
</tr>
<tr>
<td>50</td>
<td>41.667</td>
</tr>
</tbody>
</table>
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

<table>
<thead>
<tr>
<th>Poles frequency*</th>
<th>[ \frac{\sqrt{g_{m1}g_{m2}}}{C_1C_2}\sqrt{1 + \left( \frac{g_{o1}}{g_{m1}} \right) \left( \frac{g_{m3} + g_{o2} + g_{o3}}{g_{m2}} \right) - \left( \frac{g_{m1}}{\omega_{P1}} \right) \left( \frac{g_{m2}}{\omega_{P2}} \right)} ]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth*</td>
<td>[ BW_{ideal}(1 - error) \approx \left( \frac{g_{m3}}{C_1} \right) \left( 1 - \frac{g_{o1} + g_{o2} + g_{o3}}{g_{m3}} \right) - 2Q \frac{\omega_0}{\omega_{P1,2}} \right) ]</td>
</tr>
</tbody>
</table>

* \( \omega_{P1,2} \) and \( g_{o1,2} \) are the non-dominant pole and output conductance, respectively.
Lossy Integrator With Positive Feedback

\[ \frac{V_o}{V_{in}} = \frac{-g_{m1} Z}{1 - g_{m1} Z} = - \frac{g_{m1}}{s C + (g_{m2} - g_{m1})} \]
Low-Frequency, High-Q OTA-C Biquad

\[
\begin{align*}
V_{02} &= \frac{g_{m1}g_{m2}/C_1C_2}{s^2 + s(g_{m1} - g_{m2})/C_1 + g_{m1}g_{m2}/C_1C_2} = \frac{g_{m1}g_{m2}/C_1C_2}{D(s)} \\
V_{01} &= \frac{g_{m1}/C_2(s + (g_{m2} - g_{mQ})/C_1)}{D(s)}
\end{align*}
\]

Resonator

\[D(s) = s^2 + s(g_{m1} - g_{m2})/C_1 + g_{m1}g_{m2}/C_1C_2\]
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 $G_m(s)$. 
Active Frequency Compensation Transconductor


\[
I_0 = g_m(V_1 - V_2)
\]

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

\[
\omega \text{ depends on } I_{SS}
\]

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

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

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

It is possible to make

\[
\omega_{eff} >> \omega_p, \omega_N
\]
Phase compensation techniques

In a Biquad:

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

\[ i_o = (g_{m1} - g_{m2})(V_1 - V_2) \]

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

Recall that

\[ g_m = \frac{g_{mo}}{1 + s/\omega_p} \approx g_{mo} \left(1 - s/\omega_p\right) \approx g_{mo} e^{-s/\omega_p} \]
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 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 $C_{\text{int}}$
- Let the capacitance used in two integrator loop be $C_{\text{ext}}$
- Effective transconductance:
  \[ g_{m1}^{\text{eff}} = \frac{g_{m1}}{1 + \frac{g_{m1}}{g_{ds5}}} \]
- CMFB loop gain =
  \[ \frac{g_{m1}^{\text{eff}} \cdot g_{m6} \cdot g_{m2}}{g_{m4} \cdot g_{m4} \cdot g_{ds2}} \cdot \frac{1}{\left(1 + \frac{sC_{\text{int}}}{g_{m4}}\right)^2 \left(1 + \frac{sC_{\text{ext}}}{g_{ds2}}\right)} \]
• Gain \((I_o/V_i)\) = \(g_{ds5} \left\{ 2 - \left( 1 + \frac{4 \beta_1}{g_{ds5}} (V_G - V_T) \right)^{-\frac{1}{2}} \right\}\)

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

• To improve linearity, use larger resistor for source degeneration.
• Differential gain = \( \frac{g_{m1,eff}}{g_{ds2}} \)

• Common mode gain = \( \left[ 1 - \frac{1}{1 + \frac{sC_{int}}{g_{m4}}} \right] \times \frac{g_{m1,eff}}{g_{ds2}} + \frac{g_{m1,eff}}{g_{m4}} \)

• CMRR\textsubscript{(DC)} = \( \frac{g_{m4}}{g_{ds2}} \)
• Gain from +ve supply = 1

• Gain from -ve supply =

\[
1 - \frac{1}{1 + \frac{sC_{int}}{g_{m4}}} \times \frac{g_{m1,eff}}{g_{ds2}} + \frac{g_{m1,eff}}{g_{m4}}
\]

• Gain from VSS is less than gain from VDD. So, output should be measured wrt VDD

• PSRR is same as CMRR
Output noise current =

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

\[ + i_{n2}^2 + i_{n4}^2 \times \left( \frac{g_{m2}}{g_{m4}} \right)^2 + i_{n6}^2 \times \left( \frac{g_{m2}}{g_{m4}} \left( 1 + \frac{sC_{int}}{g_{m4}} \right) \right)^2 \]
Simplified noise expression

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

<table>
<thead>
<tr>
<th></th>
<th>NMOS</th>
<th>PMOS</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Speed</strong></td>
<td>Faster</td>
<td>Slower</td>
</tr>
<tr>
<td><strong>Device noise</strong></td>
<td>Low thermal noise</td>
<td>Low flicker noise</td>
</tr>
<tr>
<td><strong>Linearity</strong></td>
<td>Bulk effect degrades linearity</td>
<td>No bulk effect</td>
</tr>
<tr>
<td><strong>Substrate noise</strong></td>
<td>Higher due to common substrate</td>
<td>Can be better shielded</td>
</tr>
</tbody>
</table>
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
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

MLL, Impulse, Freq syn ....
The most accurate scheme so far


- 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:

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

\[ \dot{W} = \mu e(t) x_i x_i \]
Adaptive LMS Algorithm

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

Where \( \mu \) 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.)
- Loses accuracy at low Qs(<10) and very high Qs (~100)
- Applies only to Band-Pass filters
PLL

Reference Clock → XOR → LP Filter

BP Filter → Schmitt Trigger
Proposed Q-tuning scheme

New implementation of modified-LMS Q-tuning scheme
Tuning is independent of the shape of reference waveform

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

\[
G = \left| \frac{jw}{w_a Q_a} \right|_{1+\frac{jw}{w_a Q_a - w^2/w_a^2}} \quad \theta = \text{Arg} \left( \frac{jw}{w_a Q_a} \right) \\
\frac{G}{1+\frac{jw}{w_a Q_a - w^2/w_a^2}}
\]

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

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 = \( G_{mul} \cdot G_{sum}^2 \cdot (O_{in} - O_{BP}) \cdot O_{BP} + G_{mul} \cdot G_{sum} \cdot O_{in} \cdot O_{sum} + G_{mul} \cdot O_{sum}^2 + O_{mul} \)

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

- Reduced offset \Rightarrow 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

\[ (\text{VT}('/\text{cmp\_out\_+}') - \text{VT}('/\text{cmp\_out\_}')) \]

**Transients Response**

**Comparator input**
Attenuator

- **Capacitor**
  - Large capacitors for matching
  - Large capacitors \(\rightarrow\) Large loading

- **Resistor**
  - Larger resistors for matching
  - Large resistors \(\rightarrow\) Small loading
  - Should take parasitic capacitor into consideration
Multiplier

• Constraints
  – Symmetric
  – Good frequency response
  – Good CMRR
  – Gain should not be very small
Multiplier gain = \(-\frac{32 V_{DD}^3}{(V_1 - V_T)^3}\)

CM gain = \(\frac{4 V_{DD}^2}{(V_1 - V_T)^2}\)
LPF

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

- Built using external components for good control
Differential difference adder

- Add/Subtract two differential signals
  - High gain $\Rightarrow$ Q tuning loop unstable
  - Low gain $\Rightarrow$ 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

proj_2 fQ_tun1 schematic: Dec 4 23:51:02 2000

Transient Response

Frequency tuning voltage

Q tuning voltage

BP Filter
LP Filter
Schmitt Trigger
XOR
Reference Clock

1/Q
Integrator

BP Filter

LP Filter
Die Photograph
Experimental results

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
• 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

- IM₃ of 45dB when the input signal is 44.6mV
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.
References

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