EE247
Lecture 5

• Summary last lecture
• Today
  – Effect of integrator non-idealities on continuous-time filter behavior
  – Various integrator topologies utilized in monolithic filters
    • Resistor + C based filters
    • Transconductance (gm) + C based filters
    • Switched-capacitor filters
  – Continuous-time filters
    • Facts about monolithic Rs & Cs and its effect on integrated filter characteristics
    • Opamp MOSFET-C filters
    • Opamp MOSFET-RC filters
    • Gm-C filters

Lecture 5
Summary Last Lecture

• Ladder Type Filters
  – All pole ladder type filters
    • Convert to integrator based form
    • Example: 5th order Butterworth filter
  – High order ladder type filters incorporating zeros
    • 7th order elliptic filter in the form of ladder RLC with zeros
      – Sensitivity to component variations
      – Compare with cascade of biquads
  \( \rightarrow \text{Doubly terminated LC ladder filters} \rightarrow \text{Lowest sensitivity to component variations} \)
    • Convert to integrator based form utilizing SFG techniques
    • Example: Single-ended & differential high order filter implementation
Effect of Integrator Non-Idealities on Filter Performance

- Ideal integrator characteristics
- Real integrator characteristics:
  - Effect of finite DC gain
  - Effect of integrator non-dominant poles

**Ideal Integrator**

\[
\text{Ideal Intg.} \quad \text{opamp DC gain} = \infty
\]

\[
H(s) = \frac{-\omega_0}{s}
\]

\[
\omega_0 = \frac{1}{RC}
\]
Ideal Integrator Quality Factor

\[ H(s) = \frac{-\omega_0}{s} = -\frac{\omega_0}{j\omega} \]

Since \( Q \) is defined as:

\[ H(j\omega) = \frac{1}{R(\omega) + jX(\omega)} \]

\[ Q = \frac{X(\omega)}{R(\omega)} \]

Then:

\[ Q_{\text{ideal}} = \infty \]

Real Integrator Non-Idealities

\[ H(s) = \frac{-\omega_0}{s}, \quad -90^\circ \]

\[ H(s) = \frac{-a}{(1 + \frac{s}{\omega_1})(1 + \frac{s}{p^2})(1 + \frac{s}{p^3}) \ldots}, \quad -90^\circ \]
Effect of Integrator Finite DC Gain on Q

Example: $P1/\omega_0 = 1/100 \rightarrow$ phase error $\cong +0.5$ degree

Effect of Integrator Finite DC Gain on Q

- Phase lead @ $\omega_0$

$\rightarrow$ Droop in the passband
Effect of Integrator Non-Dominant Poles

Example: $\omega_0 / P_2 = 1/100 \rightarrow$ phase error $\approx -0.5$ degree

Effect of Integrator Non-Dominant Poles

- Phase lag @ $\omega_0$

$\rightarrow$ Peaking in the passband
In extreme cases could result in oscillation!
Effect of Integrator Non-Dominant Poles & Finite DC Gain on $Q$

\[ \angle - \pi/2 + \arctan \frac{P1}{\omega_0} - \arctan \frac{\omega_0}{p_i} = -90^\circ \]

Note that the two terms have different signs → Can cancel each other’s effect!

Integrator Quality Factor

**Real Intg.**

\[ H(s) = \frac{-a}{(1 + \frac{s}{\omega_0})(1 + \frac{s}{p_2})(1 + \frac{s}{p_3})} \]

Based on the definition of $Q$ and assuming that:

\[ \frac{\omega_0}{p_2,3,\ldots} << 1 \quad \text{&} \quad a >> 1 \]

It can be shown that in the vicinity of unity-gain-frequency:

\[ Q_{real} = \frac{1}{a - \omega_0 \sum_{i=2}^{\infty} \frac{1}{p_i}} \]
Example:
Effect of Integrator Finite Q on Bandpass Filter Behavior

\[ 0.5 \psi_{\text{lead}} @ \omega_0 \quad \text{Integrator DC gain} = 100 \]

\[ 0.5 \psi_{\text{excess}} @ \omega_\text{intg} \]

Example:
Effect of Integrator Finite Q on Filter Behavior

\[ \psi_{\text{error}} @ \omega_\text{intg} = 0 \]

\[ (0.5 \psi_{\text{lead}} - 0.5 \psi_{\text{excess}}) @ \omega_\text{intg} \]

Integrator DC gain = 100 & P2 @ 100, \omega_0
Summary
Effect of Integrator Non-Idealities on Q

\[ Q_{\text{ideal}}^{\text{intg.}} = \infty \]

\[ Q_{\text{real}}^{\text{intg.}} = \frac{1}{\Delta - \alpha_0 \sum_{n=2}^{\infty} \frac{1}{n}} \]

- Amplifier DC gain reduces the overall Q in the same manner as series/parallel resistance associated with passive elements
- Amplifier poles located above integrator unity-gain frequency enhance the Q
  - If non-dominant poles close to unity-gain freq. \( \rightarrow \) Oscillation
- Depending on the location of unity-gain-frequency, the two terms can cancel each other out!

Various Integrator Based Filters

- Continuous Time
  - Resistive element based
    - Opamp-RC
    - Opamp-MOSFET-C
    - Opamp-MOSFET-RC
  - Transconductance (Gm) based
    - Gm-C
    - Opamp-Gm-C

- Sampled Data
  - Switched-capacitor Integrator
Integrator Implementation
Opamp-RC & Opamp-MOSFET-C & Opamp-MOSFET-RC

\[ V_o = -\omega_o \frac{s}{s + 1} \]
where \( \omega_o = \frac{1}{R C} \)

---

Integrator Implementation
Gm-C & Opamp-Gm-C

\[ V_o = \frac{-\omega_o}{s} \]
where \( \omega_o = \frac{G_m}{C} \)
Integrator Implementation
Switched-Capacitor

\[
\int_{f_1}^{f_2} T = 1/f_{\text{clk}} \quad \text{for} \quad f_{\text{signal}} \ll f_{\text{clk}}
\]

\[
V_0 = \frac{f_{\text{clk}} \cdot C_s}{C_f} \int v_{\text{in}} \, dt
\]

\[
\omega_0 = f_{\text{clk}} \times \frac{C_s}{C_f}
\]

Main advantage: Critical frequency function of ratio of Caps & clock freq.

\[\rightarrow\] Critical filter frequencies (e.g. LPF -3dB freq.) very accurate

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Few Facts About Monolithic \(Rs \& Cs \& Gms\)

- Monolithic continuous-time filter critical frequency set by \(RxC\) or \(GmxC\)
- Absolute value of integrated \(Rs \& Cs \& Gms\) are quite variable
  - \(Rs\) vary due to doping and etching non-uniformities
    - Could vary by as much as \(\sim \pm 20 \text{ to } 40\%\) due to process & temperature variations
  - \(Cs\) vary because of oxide thickness variations and etching inaccuracies
    - Could vary \(\sim \pm 10 \text{ to } 15\%\)
  - \(Gms\) typically function of mobility, oxide thickness, current, device geometry …
    - Could vary \(\sim \pm 40\%\) or more with process & temp. & supply voltage

\[\rightarrow\] Continuous-time filter critical frequency could vary by over \(+50\%\)
Few Facts About Monolithic Rs & Cs

- While absolute value of monolithic Rs & Cs and $gms$ are quite variable, with special attention paid to layout, $C$ & $R$ & $gms$ quite well-matched
  - *Ratios* very accurate and stable over time and temperature
- With special attention to layout (e.g. interleaving, use of dummy devices, common-centroid geometries...):
  - Capacitor matching $\ll 0.1\%$
  - Resistor matching $< 0.1\%$
  - Gm matching $< 0.5\%$

Impact of Process Variations on Filter Characteristics

Facts about RLC filters

- $\omega_{-3dB}$ determined by absolute value of $Ls$ & $Cs$
  
  \[ C_{RLC}^C = C_1 \times C_{Norm}^C = \frac{C_{Norm}^C}{R \times \omega_{-3dB}} \]

- Shape of filter depends on *ratios* of normalized $Ls$ & $Cs$
  
  \[ L_{RLC}^R = L_2 \times L_{Norm}^R = \frac{L_{Norm}^R \times R^*}{\omega_{-3dB}} \]

[Diagram of RLC Filters]
Effect of Monolithic R & C Variations on Filter Characteristics

- Filter shape (whether Elliptic with 0.1dB Rpass or Butterworth..etc) is a function of ratio of normalized Ls & Cs in RLC filters
- Critical frequency (e.g. $\omega_{3dB}$) function of absolute value of Ls & Cs
- Absolute value of integrated Rs & Cs & Gms are quite variable
- Ratios very accurate and stable over time and temperature

→ What is the effect of on-chip component variations on monolithic filter frequency characteristics?

Impact of Process Variations on Filter Characteristics

$\tau_1 = C_{RLC} R^* = \frac{C_{Norm}}{\omega_{3dB}}$

$\tau_2 = \frac{L_{RLC}}{R^*} = \frac{I_{Norm}}{\omega_{3dB}}$

$\tau_1 = C_{Norm} \quad \tau_2 = I_{Norm}$
Impact of Process Variations on Filter Characteristics

\[
\begin{align*}
\gamma_1 &= C_{i1} \frac{R_1}{\omega_{3dB}} \\
\gamma_2 &= C_{i2} \frac{R_2}{\omega_{3dB}} \\
\gamma_1 \gamma_2 &= C_{i1} R_1 \frac{C_{Norm}}{L_{Norm}} \\
\gamma_1 \gamma_2 &= C_{i2} R_2 \frac{C_{Norm}}{L_{Norm}}
\end{align*}
\]

Variation in absolute value of integrated \(Rs\) & \(Cs\) \(\Rightarrow\) change in critical freq. \((\omega_{3dB})\)

Since \textit{Ratios} of \(Rs\) & \(Cs\) very accurate

\(\Rightarrow\) Continuous time monolithic filters fully retain their shape even with absolute component value change

Example: LPF Worst Case Corner Frequency Variations

- While absolute value of on-chip RC (gm-C) time-constants vary by as much as 100% (process & temp.)
- With proper precautions, excellent matching can be achieved:
  - Well-preserved relative amplitude & phase vs freq. characteristics
  - Need to adjust (tune) continuous-time filter critical frequencies only

Nominal Bandwidth

Detailed passband (note shape is well-retained)
Tunable Opamp-RC Filters

Example:
- 1st order Opamp-RC filter is designed to have a corner frequency of 1.6MHz
- Assuming process variations of:
  - C varies by +10%
  - R varies by -25%
- Build the filter in such a way that the corner frequency can be adjusted post-manufacturing.

Filter Corner Frequency Variations

- Assuming expected process variations of:
  - Maximum C variations by +10% → C_{min}=9pF, C_{max}=11pF
  - Maximum R variations by -25% → R_{min}=7.5K, R_{max}=12.5K
- Corner frequency ranges from 23.57MHz to 11.57MHz

→ Corner frequency varies by +48% & -27%
Variable Resistor or Capacitor

• Make provisions for either R or C to be adjustable (this example adjustable R)
• Monolithic Rs can only be made adjustable in discrete steps (not continuous)

\[
\frac{R_{\text{max}}}{R_{\text{nom}}} = \frac{f_{\text{max}}}{f_{\text{nom}}} = 1.48
\]
\[
\rightarrow R_{\text{max}}^\text{nom} = 14.8k\Omega
\]
\[
\frac{R_{\text{min}}}{R_{\text{nom}}} = \frac{f_{\text{min}}}{f_{\text{nom}}} = 0.72
\]
\[
\rightarrow R_{\text{min}}^\text{nom} = 7.2k\Omega
\]

Tunable Resistor

• Maximum C variations by \(+/-10\%\) \(C_{\text{min}}=9pF, C_{\text{max}}=11pF\)
• Maximum R variations by \(+/-25\%\) \(R_{\text{min}}=7.5K, R_{\text{max}}=12.5K\)
  \(\Rightarrow\) Corner frequency varies by \(+48\% & -27\%\)
• Assuming \(n=3\) bit (0 or 1) control signal for adjustment

\[
R_1 = R_{\text{min}}^\text{nom} = 7.2k\Omega
\]
\[
R_2 = \left(\frac{R_{\text{max}}^\text{nom} - R_{\text{min}}^\text{nom}}{2n-1}\right) \times \frac{2^n-1}{2^n} = (14.8k - 7.2k)\frac{4}{7} = 4.34k\Omega
\]
\[
R_3 = \left(\frac{R_{\text{max}}^\text{nom} - R_{\text{min}}^\text{nom}}{2n-2}\right) \times \frac{2^n-2}{2^n} = (14.8k - 7.2k)\frac{2}{7} = 2.17k\Omega
\]
\[
R_4 = \left(\frac{R_{\text{max}}^\text{nom} - R_{\text{min}}^\text{nom}}{2n-3}\right) \times \frac{2^n-3}{2^n} = (14.8k - 7.2k)\frac{1}{7} = 1.08k\Omega
\]

Tuning resolution \(= 1.08k/10k = 10\%\)
Tunable Opamp-RC Filter

```
<table>
<thead>
<tr>
<th>D2</th>
<th>D1</th>
<th>D0</th>
<th>Rnom</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>1</td>
<td>7.2K</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
<td>8.28K</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
<td>9.37K</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>14.8K</td>
</tr>
</tbody>
</table>
```

Post manufacturing:
- Set all Dx
- Measure -3dB frequency
  - If frequency too high decrement D to D-1
  - If frequency too low increment D to D+1
  - If frequency within 10% of the desired corner frequency → stop

For higher order filters, all filter integrators tuned simultaneously

Tunable Opamp-RC Filters

Summary

- Program Cs and/or Rs to freq. tune the filter
- All filter integrators tuned simultaneously
- Tuning in discrete steps & not continuous
- Tuning resolution limited
- Switch parasitic C & series R can affect the freq. response of the filter
Opamp RC Filters

- Advantages
  - Since resistors are quite linear, linearity only a function of opamp linearity
    - good linearity

- Disadvantages
  - Opamps have to drive resistive load, low output impedance is required
    - High power consumption
  - Continuous tuning not possible
  - Tuning requires programmable Rs and/or Cs
Integrator Implementation
Opamp-RC & Opamp-MOSFET-C & Opamp-MOSFET-RC

\[
\int_{t_0}^{t} \begin{cases} \dot{V}_C + \frac{1}{R_C} V_C = \frac{V_{in}}{R} \end{cases}
\]

where

\[
\frac{V_o}{V_{in}} = \frac{-\omega_0}{s} \quad \text{where} \quad \omega_0 = \frac{1}{RC}
\]

Use of MOSFETs as Resistors

\[
\text{R replaced by MOSFET} \rightarrow \text{Continuously variable resistor:}
\]

MOSFET IV characteristic:
Use of MOSFETs as Resistors

Single-Ended Integrator

\[ I_D = \mu C_{ox} \frac{W}{L} \left( (V_{gs} - V_{th}) V_{ds} - \frac{V_{ds}^2}{2} \right) \]

\[ I_D = \mu C_{ox} \frac{W}{L} \left( (V_{gs} - V_{th}) V_i - \frac{V_i^2}{2} \right) \]

\[ G = \frac{\partial I_D}{\partial V_i} = \mu C_{ox} \frac{W}{L} (V_{gs} - V_{th} - V_i) \]

\[ \Rightarrow \text{Tunable by varying } V_G: \]

Problem: Single-ended MOSFET-C Integrator → Effective R non-linear
Note that the non-linearity is mainly 2nd order type

Use of MOSFETs as Resistors

Differential Integrator

Opamp-MOSFET-C

[Diagram]

\[ I_D = \mu C_{ox} \frac{W}{L} \left( (V_{gs} - V_{th} - \frac{V_{th}}{2}) V_{ds} \right) \]

\[ I_{D1} = \mu C_{ox} \frac{W}{L} \left( (V_{gs} - V_{th}) V_i - \frac{V_i}{2} \right) \]

\[ I_{D2} = -\mu C_{ox} \frac{W}{L} \left( (V_{gs} - V_{th} + \frac{V_{th}}{2}) V_i \right) \]

\[ I_{D1} - I_{D2} = \mu C_{ox} \frac{W}{L} (V_{gs} - V_{th}) V_i \]

\[ G = \frac{\partial (I_{D1} - I_{D2})}{\partial V_i} = \mu C_{ox} \frac{W}{L} (V_{gs} - V_{th}) \]

• Non-linear term cancelled!
• Admittance independent of \( V_i \)

Problem: Threshold voltage dependence
MOSFET-C Integrator

• For the Opamp-RC integrator, opamp input stays at $0V$ (virtual gnd.)

• For the MOSFET-C integrator, opamp input stays at the voltage $V_x$ which is a function of 2nd order MOSFET non-linearities

$\Rightarrow$ Common-mode voltage sensitivity

Use of MOSFET as Resistor Issues

• Distributed nature of gate capacitance & channel resistance results in infinite no. of high-frequency poles $\Rightarrow$ excess phase
• Filter performance mandates well-matched MOSFETs $\Rightarrow$ long channel devices
• Excess phase increases with $L^2$
  $\Rightarrow$ Tradeoff between matching and integrator $Q$
  $\Rightarrow$ This type of filter limited to low frequencies
Example:
Opamp MOSFET-C Filter

• Suitable for low frequency applications
• Issues with linearity
• Linearity achieved ~40-50dB
• Needs tuning

5th Order Elliptic MOSFET-C LPF with 4kHz Bandwidth


Improved MOSFET-C Integrator

\[ I_D = \mu C_{in} \frac{W}{T} \left( V_{GL} - V_{TH} \right) \]
\[ I_{D1} = \mu C_{in} \frac{W}{T} \left( V_{GL1} - V_{TH} \right) \]
\[ I_{D3} = \mu C_{in} \frac{W}{T} \left( V_{GL2} - V_{TH} \right) \]
\[ I_{X1} - I_{D1} = I_{D3} - I_{X2} \]
\[ I_{X2} = \mu C_{on} \frac{W}{T} \left( V_{GL2} - V_{TH} \right) \]

No threshold dependence
First order Common-mode non-linearity cancelled
Linearity achieved in the order of 60-70dB

R-MOSFET-C Integrator

- Improvement over MOSFET-C by adding resistor in series with MOSFET
- Voltage drop primarily across fixed resistor → small MOSFET Vds → improved linearity
- Linearity in the order of 90dB possible
- Generally low frequency applications


R-MOSFET-C Lossy Integrator

Negative feedback around the non-linear MOSFETs improves linearity
Compromises frequency response accuracy

Example:
Opamp MOSFET-RC Filter

5th Order Bessel MOSFET-RC LPF -22kHz bandwidth
THD -> -90dB for 4Vp-p 2kHz input signal

• Suitable for low frequency applications
• Significant improvement in linearity compared to MOSFET-C
• Needs tuning


Operational Amplifiers (Opamps) versus Operational Transconductance Amplifiers (OTA)

Opamp
Voltage controlled voltage source

• Low output impedance
• Output in the form of voltage
• Can drive R-loads
• Good for RC filters, OK for SC filters
• Extra buffer adds complexity, power dissipation

OTA
Voltage controlled current source

• High output impedance
• In the context of filter design called gm-cells
• Output in the form of current
• Cannot drive R-loads
• Good for SC & gm-C filters
• Typically, less complex compared to opamp higher freq. potential
• Typically lower power
Integrator Implementation

**Gm-C & Opamp-Gm-C**

\[
\frac{V_O}{V_{in}} = -\frac{\omega_o}{s} \quad \text{where} \quad \omega_o = \frac{G_m}{C}
\]

---

**Gm-C Filters**

**Simplest Form of CMOS Gm-C Integrator**

- MOSFET in saturation region:
  \[
  I_d = \frac{\mu C_{ox} W}{2} (V_{gs} - V_{th})^2
  \]
- \( Gm \) is given by:
  \[
  s_m = \frac{\partial I_d}{\partial V_{gs}} = \mu C_{ox} \frac{W}{2L} (V_{gs} - V_{th})
  \]
  \[
  = \frac{1}{2} \frac{I_d}{V_{gs} - V_{th}}
  \]
  \[
  = \frac{1}{2} \left( \frac{1}{2} \mu C_{ox} \frac{W}{L} \right)^{1/2}
  \]

*Id varied via Vcontrol → gm tunable via Vcontrol*

Gm-C Filters
Simplest Form of CMOS Gm-C Integrator

- MOSFET in saturation region $G_m$ is given by:
  $s_m = \frac{\partial I_d}{\partial V_{gs}} = 2 \left( \frac{1}{2} \mu C_{ox} \frac{W}{L} I_d \right)^{1/2}$
  $I_d$ varied via $V_{control}$
  $g_m$ tunable via $V_{control}$

- Critical frequency continuously tunable via $V_{control}$
  \[ V_{os} = -\frac{\omega_o}{s} \]
  where \[ \omega_o = \frac{s_m^{M1,2}}{2 \times C_{int} g} \]

Second Order Gm-C Filter

- Simple design
- Tunable
- $Q$ function of device ratios:
  \[ Q = \frac{s_m^{M1,2}}{g_m^{M1,4}} \]
Filter Frequency Tuning Techniques

- Component trimming

- Automatic on-chip filter tuning
  - Continuous tuning
    - Master-slave tuning
  - Periodic off-line tuning
    - Systems where filter is followed by ADC & DSP, existing hardware can be used to periodically update filter freq. response

Example: Tunable Opamp-RC Filter

Post manufacturing:
- Usually at wafer-sort tuning performed
- Measure -3dB frequency
  - If frequency too high decrement D to D-1
  - If frequency too low increment D to D+1
  - If frequency within 10% of the desired corner freq. stop

Not practical to require end-user to tune the filter
→ Need to fix the adjustment at the factory
Trimming

- Component trimming
  - Build fuses on-chip,
    - Based on measurements @ wafer-sort blow fuses selectively by applying high current to the fuse
      - Expensive
      - Fuse regrowth problems!
      - Does not account for temp. variations & aging
  - Laser trimming
    - Trim components or cut fuses by laser
      - Even more expensive
      - Does not account for temp. variations & aging

Example: Tunable/Trimmable Opamp-RC Filter

\[
\begin{array}{c|c|c|c|c}
D2 & D1 & D0 & R_{\text{nom}} \\
\hline
1 & 1 & 1 & 7.2K \\
1 & 1 & 0 & 8.28K \\
1 & 0 & 1 & 9.37K \\
\vdots & \vdots & \vdots & \vdots \\
0 & 0 & 0 & 14.8K \\
\end{array}
\]
Automatic Frequency Tuning

• By adding additional circuitry to the main filter circuit
  – Have the filter critical frequency automatically tuned
    → Expensive trimming avoided
    → Accounts for critical frequency variations due to temperature and effect of aging

Master-Slave Automatic Frequency Tuning

• Following facts used in this scheme:
  – Use a replica (master) of the main filter (called the slave) in the tuning circuitry
  – Place the replica in close proximity of the main filter
  – Use the tuning signal generated to tune the replica, to also tune the main filter
  – In the literature, this scheme is called master-slave tuning!
Master-Slave Frequency Tuning
Reference Filter (VCF)

- Use a biquad for master filter (VCF)
- Utilize the fact that @ the frequency $f_0$ the lowpass (or highpass) outputs are 90 degree out of phase wrt to input

$$\frac{V_{LP}}{V_{in}} = \frac{l}{s^2 + \frac{s}{\omega_0} + 1}$$

@ $\omega = \omega_0$, $\phi = -90^\circ$

- Apply a sinusoid at the desired $f_0$
- Compare the LP output phase to the input
- Based on the phase difference
  - Increase or decrease filter critical freq.

Master-Slave Frequency Tuning
Reference Filter (VCF)

$$V_{tune} = -K \times V_{ref}^{rms} \times V_{LP}^{rms} \times \cos \phi$$

Input Signal Frequency
Master-Slave Frequency Tuning
Reference Filter (VCF)

- By closing the loop, feedback tends to drive the error voltage to zero.
  \[ \text{Locks } f_0, \text{ the critical frequency of the filter to the accurate reference frequency} \]
- Typically the reference frequency is provided by a crystal oscillator with accuracies in the order of few ppm

Master-Slave Frequency Tuning
Reference Filter (VCF)

• Issues to be aware of:
  – Input reference tuning signal needs to be sinusoid → Disadvantage since clocks are usually available as square waveform
  – Reference signal feed-through to the output of the filter can limit filter dynamic range (reported levels or about 100μVrms)
  – Ref. signal feed-through is a function of:
    • Reference signal frequency wrt filter passband
    • Filter topology
    • Care in the layout
    • Fully differential topologies beneficial

Master-Slave Frequency Tuning
Reference Voltage-Controlled-Oscillator (VCO)

• Instead of VCF a voltage-controlled-oscillator (VCO) is used
• VCO made of replica integrator used in main filter
• Tuning circuit operates exactly as a conventional phase-locked loop (PLL)
• Tuning signal used to tune main filter

Master-Slave Frequency Tuning
Reference Voltage-Controlled-Oscillator (VCO)

• Issues to be aware of:
  – Design of stable & repeatable oscillator challenging
  – VCO operation should be limited to the linear region of the amp or else the operation loses accuracy
  – Limiting the VCO signal range to the linear region not a trivial design issue
  – In the case of VCF based tuning ckt there was only ref. signal feedthrough. In this case, there is also the feedthrough of the VCO signal!!
  – Advantage over VCF based tuning → Reference input signal square wave (not sin.)