INTEGRATED CIRCUIT CONTINUOUS TIME FILTERS

Outline - Sections
1 Introduction to Continuous Time Filters
2 Passive Filters
3 Integrators
4 Biquads
5 Filter Design
6 Filter Tuning
7 Summary
SECTION 1 - INTRODUCTION TO CONTINUOUS TIME FILTERS

Integrated Filter Categories

- Integrated Filters
  - Digital Filters
  - Analog Filters
    - Analog Continuous-Time Filters
      - Passive Filters (RC and RLC)
        - MOSFET-C Filters
      - Gm-C Filters
      - Current Mode Filters
        - Active RLC Filters
        - Log-Domain Filters
    - Analog Sampled-Data Filters
      - Switched Capacitor Filters
      - Switched Current Filters

Fig.11.1-01
Types of Filters

- A continuous time filter is a filter whose variables are continuous both in time and in amplitude.
- A discrete-time (analog sampled-data) filter is a filter whose variables are continuous in amplitude but not in time.
- A digital filter is a filter whose variables are discrete both in time and in amplitude.

Example:

A digital signal would only have the amplitude values of 0, 1, 2, through 8.
Digital Filters

Current technology constrains digital filtering to 5-10MHz or less.

Programmability is the key feature as well as unlimited resolution.

Practical problems include:
- The need for analog anti-aliasing filters
- Large chip area requirements
- Electromagnetic compatibility with low level analog signals
- Requirements for a high resolution, high speed analog to digital converter (ADC)
- Power consumption at high frequencies
**Analog Continuous-Time Filters**

Continuous-Time Filters

Advantages:
- No clock feedthrough
- No oversampling requirement
- Less dissipation than digital filters
- Higher frequency capability than switched capacitor filters
- No anti-aliasing filter required
- Uses a large body of classical filter theory and data

Disadvantages:
- Accuracy of the time constants must be achieved by tuning
- Linearity of resistor or transconductance implementations

**Discrete-Time Filters**

Advantages:
- Very accurate, tuning not required
- Switched capacitors are a well-known technique

Disadvantages:
- Clock feedthrough
- Limited in frequency by the oversampling ratio
• Requires an anti-aliasing filter
Dynamic Range of Filters

Dynamic Range:

\[
DR = \frac{P_{\text{max}}}{P_N} = \frac{V_{\text{sat}}^2}{V_N^2}
\]

where

\[P_{\text{max}} (V_{\text{sat}}) = \text{maximum signal power (voltage) at the filter output (typically for 1\% THD)}\]

\[P_N (V_N) = \text{integrated noise power (voltage) in the bandwidth of interest at filter output}\]

Noise Performance:

6th-order continuous-time filter\(^1\)

\[V_N^2 = \frac{3kTQ}{C_{\text{int}}} \quad \Rightarrow \quad DR = \frac{V_{\text{sat}}^2 C_{\text{int}}}{3kTQ}\]

Optimum DR for continuous-time filters\(^2\)

\[DR_{\text{opt}} = \frac{V_{\text{sat}}^2 C_{\text{total}}}{4kTFQ}\]

where


\( C_{total} \) = total capacitance of the filter’s integrators

\( F \) = noise factor used to account for possible excess noise contributions by nonideal devices

**Dynamic Range- Continued**

Upper bound on \( DR \) for a general high-Q bandpass filter

\[
DR_{opt} \leq \frac{V_{sat}^2 C_{total}}{2\pi k TFQ}
\]

The dependence of \( DR \) on power dissipation and bandwidth are†

\[
DR_{opt} = \frac{\eta P_{diss}}{4\pi k TFBQ^2}
\]

where

\( \eta \) = efficiency factor relating the power consumed by the filter to the maximum signal output power

\( B \) = filter bandwidth

In general, the mechanism responsible for limiting the \( DR \) in high Q filters is the regenerative gain associated with the high-Q poles.

Typical \( DR \approx 70-90 \text{dB} \) depending upon the architecture (AGC’s, etc.)

---

General Approach for Continuous and SC Filter Design

All designs start with a normalized, low pass filter with a passband of 1 radian/second and an impedance of 1Ω that will satisfy the filter specification.

1.) Cascade approach - starts with the normalized, low pass filter root locations.
2.) Ladder approach - starts with the normalized, low pass, \( RLC \) ladder realizations.
Follow-the-Leader Feedback Design (FLF)

Besides the cascade and ladder approaches, there is a third approach called follow-the-leader feedback.

The various $T_i$ are second-order bandpass transfer functions.

This structure can realize both zeros and poles.

---

**Primary Resonator Block**

Similar to FLF except that there is no feedforward paths and consequently it cannot realize complex zeros.

Easier to design and uses identical second-order blocks.
Denormalization of Filter Realizations

All filters are designed assuming a cutoff frequency of 1 radian/sec. and an impedance level of 1 ohm. In order to move the filter frequency to the desired frequency a denormalization must be performed. For active filters, the impedance denormalization is a free parameter that can be used to adjust the final values of the components.

Frequency and Impedance Denormalizations:

<table>
<thead>
<tr>
<th>Denormalization</th>
<th>Denormalized Resistance, $R$</th>
<th>Denormalized Conductance, $G_m$</th>
<th>Denormalized Capacitor, $C$</th>
<th>Denormalized Inductor, $L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency: $s = \omega p B s_n = \Omega_n s_n$</td>
<td>$R = R_n$</td>
<td>$G_m = G_{mn}$</td>
<td>$C = \frac{C_n}{\Omega_n}$</td>
<td>$L = \frac{L_n}{\Omega_n}$</td>
</tr>
<tr>
<td>Impedance: $Z = z_o Z_n$</td>
<td>$R = z_o R_n$</td>
<td>$G_m = \frac{G_{mn}}{z_o}$</td>
<td>$C = \frac{C_n}{z_o}$</td>
<td>$L = z_o L_n$</td>
</tr>
<tr>
<td>Frequency and Impedance: $Z(s) = z_o Z_n(\Omega_n s_n)$</td>
<td>$R = z_o R_n$</td>
<td>$G_m = \frac{G_{mn}}{z_o}$</td>
<td>$C = \frac{C_n}{z_o \Omega_n}$</td>
<td>$L = \frac{z_o L_n}{\Omega_n}$</td>
</tr>
</tbody>
</table>

Note that the design of switched capacitors is done in such a manner that denormalization is not necessary.
SECTION 2 - PASSIVE FILTERS

Comparison of Active and Passive Filters

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Active Filters</th>
<th>Passive Filters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dissipation</td>
<td>Requires power</td>
<td>Dissipationless</td>
</tr>
<tr>
<td>Linearity</td>
<td>Limited</td>
<td>Linear</td>
</tr>
<tr>
<td>Noise</td>
<td>Active plus thermal</td>
<td>Noiseless (except for R’s)</td>
</tr>
<tr>
<td>Size</td>
<td>Small</td>
<td>Large</td>
</tr>
<tr>
<td>Compatibility with integration</td>
<td>Good</td>
<td>Poor</td>
</tr>
<tr>
<td>Midband gain</td>
<td>Not constrained</td>
<td>Equal to less than unity</td>
</tr>
</tbody>
</table>
**Passive Filters**

Categories:
- Discrete ceramic (piezoelectric)
- Crystal
- Acoustic wave
  - Surface (SAW)
  - Bulk (BAW)
- LC

General Characteristics:
- Fractional bandwidths are small (0.1% to 3%)
- Shape factors are moderate (16 to 20dB of attenuation at 2 to 3 times the nominal bandwidth)
- Insertion loss is moderate (1.5 to 6dB)
- Cost is low ($0.3 to $3) when purchased in large quantities
- No power dissipation
- Low noise figure
### Passive Filter Performance

<table>
<thead>
<tr>
<th>Part No.</th>
<th>Type</th>
<th>Application</th>
<th>Frequency</th>
<th>BW</th>
<th>Shape Factor</th>
<th>IL</th>
<th>Price</th>
</tr>
</thead>
<tbody>
<tr>
<td>Toko HCFM8-262B</td>
<td>Ceramic</td>
<td>AM Broadcast IF</td>
<td>262 kHz</td>
<td>6kHz</td>
<td>-16dB @ ±9kHz</td>
<td>6dB</td>
<td>$1</td>
</tr>
<tr>
<td>Toko CFMR-455B</td>
<td>Ceramic</td>
<td>AM Broadcast IF</td>
<td>455kHz</td>
<td>6kHz</td>
<td>-16dB @ ±9kHz</td>
<td>6dB</td>
<td>$1</td>
</tr>
<tr>
<td>MuRata SFP450F</td>
<td>Ceramic</td>
<td>Pager IF</td>
<td>450kHz</td>
<td>6kHz</td>
<td>-40dB @ ±12kHz</td>
<td>6dB</td>
<td>-</td>
</tr>
<tr>
<td>MuRata SFE4.5MBF</td>
<td>Ceramic</td>
<td>Television Sound IF</td>
<td>4.5MHz</td>
<td>120kHz</td>
<td>-20dB @ ±270kHz</td>
<td>6dB</td>
<td>-</td>
</tr>
<tr>
<td>MuRata</td>
<td>Ceramic</td>
<td>FM Broadcast IF</td>
<td>10.7MHz</td>
<td>230kHz</td>
<td>-20dB @ ±290kHz</td>
<td>6dB</td>
<td>$0.3</td>
</tr>
<tr>
<td>ECS-10.7-15B</td>
<td>MCF</td>
<td>Cellular Phone IF</td>
<td>10.7MHz</td>
<td>25kHz</td>
<td>-40dB @ ±25kHz</td>
<td>2.5dB</td>
<td>$3</td>
</tr>
<tr>
<td>Siemens B4535</td>
<td>SAW</td>
<td>DECT IF</td>
<td>110MHz</td>
<td>1.1MHz</td>
<td>-20dB @ ±1.5MHz</td>
<td>-</td>
<td>$3</td>
</tr>
<tr>
<td>MuRata LFC30-01B0881B025</td>
<td>LC</td>
<td>Cellular RF</td>
<td>881MHz</td>
<td>25MHz</td>
<td>-20dB @ ±78MHz</td>
<td>3.5dB</td>
<td>-</td>
</tr>
<tr>
<td>Toko 6DFA-881E-11</td>
<td>Dielectric</td>
<td>Cellular RF</td>
<td>881MHz</td>
<td>25MHz</td>
<td>-20dB @ ±78MHz</td>
<td>1.8dB</td>
<td>-</td>
</tr>
<tr>
<td>Toko 6DFA-914A-14</td>
<td>Dielectric</td>
<td>Cordless Phone RF</td>
<td>914MHz</td>
<td>1MHz (0.1%)</td>
<td>-24dB @ ±45MHz</td>
<td>2.2dB</td>
<td>-</td>
</tr>
<tr>
<td>-------------------</td>
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</tr>
</tbody>
</table>
Passive RLC Filters

Passive filters consist of:

- Inductors
- Capacitors

Realizations:

1.) Geometrically centered
2.) Arbitrary (difficult)

Advantages of passive filters:

- Low noise (no noise)
- High frequency - up to the GHz range

Disadvantage of passive filters:

- Poor IC compatibility
- Difficult to tune

Lowpass Filter with zeros at infinity

Lowpass Filter with \( j \omega \) zeros

Fig.12.6-4
RC Filters (Polyphase Filters)

A polyphase filter is a fully symmetric RC network with multiple inputs.

- Depending on the phase and amplitude relation of the inputs, it rejects some inputs and passes others.
- Every input vector set can be decomposed into the unit vector sets.

\[
\text{Input frequency} = \frac{1}{2\pi RC}
\]

Lag:
\[
T(j\omega) = \frac{1}{j\omega RC + 1}
\]
\[
\text{Arg}[j\omega] = -\tan^{-1}(\omega RC)
\]

Lead:
\[
T(j\omega) = \frac{j\omega RC}{j\omega RC + 1}
\]
\[
\text{Arg}[j\omega] = 90^\circ - \tan^{-1}(\omega RC)
\]
**RC Polyphase Filters - Continued**

Polyphase filters will add the counter-clockwise phase sequences and reject the clockwise phase sequences.

![Diagram](image-url)

**Counter-clockwise quadrature sequence**

- \( V/0^\circ \)
- \( V/90^\circ \)
- \( V/180^\circ \)
- \( V/270^\circ \)

Input frequency = \( \frac{1}{2\pi RC} \)

**Clockwise quadrature sequence**

- \( V'/0^\circ + 45^\circ \)
- \( V'/90^\circ - 45^\circ \)
- \( V'/270^\circ - 45^\circ \)
- \( V'/90^\circ \)

Input frequency = \( \frac{1}{2\pi RC} \)

Fig. 10-PF2
Applications of Polyphase Filters

1.) Quadrature Signal Generation:

Quadrature signals are generated from a differential signal by means of the polyphase filter.

2.) Unwanted Signal Rejection:

If the desired signal and the unwanted signal have opposite sequences, the unwanted signal can be rejected by the polyphase filter.
**Wideband Polyphase Filters**

Staggered polyphase filters:

- Wideband image rejection can be obtained with staggering several polyphase stages
- Wider polyphase filter band $\Rightarrow$ Large number of stages $\Rightarrow$ More loss

Four stages of polyphase filters in cascade with different center frequencies.

Fig10-PF4
Practical Design Issues of Polyphase Filters

1.) Component Matching:

From Monte Carlo simulations it has been found that:

Ultimate Image Rejection > -20·log_{10}(\sigma_{Component})

(For 60dB image rejection a 0.1% matching between R’s and C’s is required)

The matching requirements will set the minimum area of the R’s and C’s

Comments:

- In general, \( \sigma^2_{Component} \propto \frac{1}{\text{Area}} \)
- Measurements on poly resistors of a 1µm CMOS process gave \( \sigma < 0.1\% \) for a resistor area of 2800µm²
- Approximate matching of capacitors is \( \sigma \approx 0.1\% \) for a capacitor area of 220µm²

2.) Cutoff Frequency of the Resistors:

At high frequencies, the polysilicon acts as a transmission line and does not provide the desired phase shift.

\[ f_{\text{Resistor}} = \frac{1}{2\pi L^2 R_{\text{s}} C_{\text{ox}}} \]

where \( L \) is the length of the resistor and \( R_{\text{s}} \) is the sheet resistance of the polysilicon and \( C_{\text{ox}} \) is the oxide capacitance. This property of the resistors sets the maximum length of the resistors.
Practical Design Issues of Polyphase Filters - Continued

3.) Capacitive Loading Mismatch:

- Symmetric loading has no effect on the image rejection property of the polyphase filter.
- Low impedance loads ($Z_{\text{Load}}$) increases the polyphase filter loss.
- Matching between the output loads is required

Normally, the load is either another polyphase filter which is sufficiently matched by the gate capacitance of the amplifier following the polyphase filter.

Rule of thumb from worst case simulations:

If $C_{\text{Load}} < C_{\text{Poly}}$, the ultimate image rejection $> -20\log_{10}\left(\frac{\sigma_{\text{Load}} C_{\text{Load}}}{C_{\text{Poly}}}\right)$

If $C_{\text{Load}} > C_{\text{Poly}}$, the ultimate image rejection $> -20\log_{10}(\sigma_{\text{Load}})$
Practical Design Issues of Polyphase Filters - Continued

4.) Differential Output instead of Quadrature Differential Output:

- Last polyphase stage can provide up to 3dB of gain with small loading capacitors
- Dummy capacitors try to match the output loading on all branches
- Because of the matching issue, this is useful in high image rejection systems only if $C_{Load} \ll C_{Poly}$

- Outputs load each other, therefore the maximum gain is 0 dB
- This is an order of magnitude less sensitive to load mismatches if $C_{Load} \ll C_{Poly}$ and almost zero sensitivity to load mismatch if $C_{Load} > C_{Poly}$
Practical Design Issues of Polyphase Filters - Continued

5.) Polyphase Input and Output Impedance at Pole Frequency:

At the polyphase pole frequency, the input and output impedance of the polyphase filter is $R || (1/sC)$ which is independent of the load and source impedance.

$$Z_{in} = R || (1/sC)$$

at $f = 1/(2\pi RC)$

6.) Input Impedance at DC:

If the load is a capacitor, then the input impedance at DC is high.

If the polyphase filter is driven by a current source a peak is experienced at low frequencies. (Should be driven by a low impedance source)
Practical Design Issues of Polyphase Filters - Continued

7.) Polyphase Noise:
   • The resistors in the polyphase filters generate noise at the output ($v_n^2 = 4kTR_{poly}$)
   • If several polyphase filters are in cascade, the noise of the previous stage is attenuated by the following. Therefore, the last stage noise becomes dominant.
   • There is a tradeoff between the loading of the polyphase filter on the preceding stage (which sets the loss) and the noise of the polyphase filter which establishes the resistor.

8.) Polyphase Voltage Gain and Loss:

Therefore,

$$|V_{out}| = \sqrt{2} \left| \frac{Z_{Load}}{Z_{Load} + Z_{Poly}} \right| |V_{in}|$$

a.) Output open circuit: $Z_{Load} >> Z_{Poly}$ ⇒ 3dB gain
b.) $Z_{Load} = Z_{Poly}$ ⇒ 3dB loss
c.) At high frequencies, capacitive loads cause severe loss
**Layout Issues for Polyphase Filters**

- 60dB image rejection requires 0.1% matching among $R$’s and $C$’s of the polyphase filter.

- Therefore, careful layout is vital to preserve the matching property.

Comments:

1.) Common centroid layout is used to cancel the process gradients
2.) The dummy elements eliminate systematic mismatch
3.) Minimized total width of the layout further decreases the gradient.
**Layout of Polyphase Filters - Continued**

- Each interconnect line has the same length and the same number of corners using the following serpentine structure:

![Serpentine Structure Diagram](image1)

- The capacitance on interconnect lines is balanced.

![Capacitance Balance Diagram](image2)
SECTION 3 - INTEGRATORS

The Role of the Integrator in Active Filters

In most active filters, the summing integrator is the key building block (the primitive).

\[ V_o(s) = \pm \frac{k_1}{s}V_1(s) \pm \frac{k_2}{s}V_2(s) \cdots \pm \frac{k_n}{s}V_n(s) \]

Classical realization:

\[ H(j\omega) = \frac{V_{out}(j\omega)}{V_{in}(j\omega)} = \frac{-1}{j\omega RC} \]
**Integrator Nonidealities**

The op amp is approximated as:

\[ A(s) = \frac{A_o}{1 + \frac{s}{\omega_a}} = \frac{GB}{s + \omega_a} \]

The integrator behavior is degraded by:
- Finite \( A_o \) in the low frequency region
- Finite \( GB \) in the high frequency region
**Integrators - Continued**

Biquad uses ±summing integrator with damping for second-order stages.

![Biquad integrator diagram]

Leapfrog uses direct integrator implementation.

![Leapfrog integrator diagram]
OTA-C ($G_M$-C) Integrators

Integrator:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{GM}{sC}$$

What is a $G_M$?

A transconductance amplifier with a linear relationship between $I_{out}$ and $V_{in}$.

Advantages:
- High frequency
- Simple

Disadvantages:
- $G_M$ must be linear and tunable
- Sensitive to parasitic capacitances at the output nodes
Differential output requires common mode feedback
**Nonideal Performance of the OTA-C Integrator**

We will consider the single-ended version for purposes of simplicity.

Sources of nonideal behavior are:

1.) Finite output resistance

2.) Frequency dependence of $G_m$

3.) RHP zero due to feedforward (found in Miller compensation)
Finite Output Resistance of the OTA-C Integrator

Model:

\[
\frac{V_{\text{out}}}{V_{\text{in}}} = -\frac{G_m R_o s C}{R_o + \frac{1}{s C} + 1}
\]

Frequency Response:

Solution: Make \( R_o \) large by using cascode configuration
**Finite Bandwidth of the OTA-C Integrator**

Model:

\[
\frac{V_{out}}{V_{in}} = \frac{-G_m(0)}{sC(s\tau_a+1)}
\]

Frequency Response:

![Frequency Response Graph](image1)

Solution:

Use higher \(f_T\) MOSFETs (smaller channel lengths)
**RHP of the OTA-C Integrator**

Model:

\[ V_{out} = -\frac{G_m s C_z}{s (C + C_z)} \approx -\frac{(G_m s C_z)}{s C} \]

Frequency Response:

- **Magnitude Response:**
  - \( V_{out} \) vs. \( V_{in} \)
  - Frequency \( \omega \) vs. dB

- **Phase Shift Response:**
  - Phase shift vs. Frequency \( \omega \)
  - Critical frequency \( \omega_c \)
  - Phase shift from \( 0^\circ \) to \( 180^\circ \)

![Diagram of OTA-C Integrator](Fig.11.4-06)

![Frequency Response Diagram](Fig.11.4-07)
Solution:
Neutralization techniques
OTA-C Integrators

Simple OTA:
Suffers from poor linearity, signal-to-noise limitations, and low output resistance

Folded-Cascode OTA (M3 in active region):

\[ G_m = \frac{g_{m1}}{1 + g_{m1}R_{M3}} \approx \frac{1}{R_{M3}} = K_n \frac{W_3}{L_3} (V_C - V_{S3} - V_{Tn}) \]
Suffers from linearity and excessive power dissipation†

OTA-C Integrators - Continued

Linearization Schemes:

(a.) M3-M4 work in a saturation-active mode for positive $V_{in}$ and in an active-saturation mode for negative $V_{in}$. Can result in a linear operation. $I$ varies $G_M$. 

(b.) M1 and M2 are in active region. Current is proportional to $V_{DS1}$ ($V_{DS2}$). $I_D$ varies $G_M$. 
OTA-C Integrators - Continued

Tradeoff between power and linearity:

Low-power consumption OTA

Linearized OTA

Fig.11.4-11
OTA-C Integrators - Continued

Practical implementation of an OTA with 90µA of power supply current.
OTA-C Integrators - Continued

Biasing Circuit for the previous OTA:
Linearized OTA-C Integrator

![ OTA-C Integrator Circuit ]

This circuit provides a linear transconductance relationship as follows:

Let $K_N'(W/L)_N = A K_P'(W/L)_P$

\[ V_{GSP} - |V_{TP}| = \sqrt{A} \ (V_{GSN} - V_{TN}) \]

and

\[ V_{GSN} + V_{GSP} = V_{in} + V \]

where $V_1 = V_2 = V$

Combining these two equations gives

\[ V_{GSN} - V_{TN} = \frac{V_{in} + V - V_{TN} - |V_{TP}|}{\sqrt{A} + 1} \]

\[ \Rightarrow \]

\[ i_D = \frac{1}{2} K_N \frac{W}{L} \frac{(V_{in} + V - V_{TN} - |V_{TP}|)^2}{A + 1 + 2\sqrt{A}} \]

\[ \Rightarrow \]

\[ G_m = \frac{i_D}{V_{in}} = 2 \cdot \frac{K_N \frac{W}{L}}{A + 1 + 2\sqrt{A}} \]

\( \text{Tuning?} \)
\textbf{GM-C-OTA Integrator}

Integrator:

\[
\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \frac{I_{\text{out}}}{V_{\text{in}}} \cdot \frac{V_{\text{out}}}{I_{\text{out}}} = GM \frac{1}{sC} = \frac{GM}{sC}
\]

Advantages:
- Avoids influence of parasitic capacitances.
- High frequency (20MHz, 6th-order filter for disk-drive read channels)
- Tuning can be done by current multiplication of $I_{\text{out}}$

Disadvantages:
- More noise - two active elements ($GM$ and OTA)
MOSFET-C-Op Amp Integrator

Integrator:

\[
\begin{align*}
R & \quad 2C \\
\text{v}_1 & \quad \text{v}_2 \\
\text{i}_1 & \quad \text{i}_2
\end{align*}
\]

Differential input resistance is (see Sec. 4.2):

\[
R_{in} = 2R = \frac{v_1-v_2}{i_1-i_2} = \frac{v_1-v_2}{\beta (V_{C1}-V_{C2})(v_1-v_2)} = \frac{1}{\beta (V_{C1}-V_{C2})}, \quad v_1, v_2 \leq \min\{V_{C1}-V_T, V_{C2}-V_T\}
\]

Advantages:
- Straight-forward implementation of RC active filters
- Insensitive to parasitics

Disadvantages:
- Requires input currents
Amplifiers must be able to drive the resistors
**MOSFET-C-OTA Filters**

Integrator:

Replace the op amp of MOSFET-C-OP AMP filters with a high-gain OTA. At high frequencies, if the $G_M$ is large, the circuit behaves as an op amp.

![Diagram of MOSFET-C-OTA Integrator](image)

- Typically requires bipolar devices to get high-enough $G_M$.
- Has been used to build filters at video frequencies.
MOSFET-C-Op Amp Integrator Performance

Ideal Performance:

Uses FETs in the triode region to achieve a linear four-quadrant multiplier.

Ideal Operation (Op Amp Gain = $\infty$):

\[
I_1 = K' \left( \left( V_{GS} + \frac{V_y}{2} - V_T \right) \frac{V_x}{2} - \frac{1}{2} \left( \frac{V_x}{2} \right)^2 \right)
\]

\[
I_2 = K' \left( \left( V_{GS} - \frac{V_y}{2} - V_T \right) \frac{V_x}{2} - \frac{1}{2} \left( \frac{V_x}{2} \right)^2 \right)
\]

\[
I_3 = K' \left( \left( V_{GS} - \frac{V_y}{2} - V_T \right) \frac{V_x}{2} - \frac{1}{2} \left( \frac{V_x}{2} \right)^2 \right)
\]

\[
I_4 = K' \left( \left( V_{GS} + \frac{V_y}{2} - V_T \right) \frac{-V_x}{2} - \frac{1}{2} \left( \frac{-V_x}{2} \right)^2 \right)
\]

\[
V_{out} = \frac{K'}{sC} (V_{o^+} - V_{o^-}) = \frac{K'}{sC} (I_4 - I_3 - I_1 + I_2) = \frac{K'}{sC} \left( \frac{V_x V_y}{2} + \frac{V_x V_y}{2} \right) = \frac{K'}{sC} V_x V_y
\]

\[
\therefore V_{out} = \frac{K'}{sC} V_{in} V_C
\]

As long as the following conditions are satisfied, the integrator will be linear.

\[
V_{inp}, V_{inn} \leq \min\{(V_{y^+} - V_T), (V_{y^-} - V_T)\}
\]
MOSFET-C-Op Amp Integrator Performance

Previous circuit with a finite op amp gain (A):

\[ V_{out} = V_o^{+} - V_o^{-} = A \left( V_i^{+} - V_i^{-} \right) = AV_{in} \quad \Rightarrow \quad V_i^{+} = \frac{V_o}{2A} \quad \text{and} \quad V_i^{-} = -\frac{V_o}{2A} \]

\[ I_1 = K' \left[ \left( V_{GS} + \frac{V_y}{2} - \frac{V_{out}}{2A} - V_T \left( \frac{V_x}{2} - \frac{V_{out}}{2A} \right) \right) - \frac{1}{2} \left( \frac{V_x}{2} - \frac{V_{out}}{2A} \right)^2 \right] \]

\[ I_2 = K' \left[ \left( V_{GS} - \frac{V_y}{2} - \frac{V_{out}}{2A} - V_T \left( \frac{V_x}{2} - \frac{V_{out}}{2A} \right) \right) - \frac{1}{2} \left( \frac{-V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right] \]

\[ I_3 = K' \left[ \left( V_{GS} + \frac{V_y}{2} + \frac{V_{out}}{2A} - V_T \left( \frac{V_x}{2} + \frac{V_{out}}{2A} \right) \right) - \frac{1}{2} \left( \frac{-V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right] \]

\[ I_4 = K' \left[ \left( V_{GS} + \frac{V_y}{2} + \frac{V_{out}}{2A} - V_T \left( \frac{-V_x}{2} + \frac{V_{out}}{2A} \right) \right) - \frac{1}{2} \left( \frac{-V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right] \]

\[ V_{out} = \frac{K'/sC}{1 + \frac{1}{A}} \left( +I_4 - I_3 - I_1 + I_2 \right) \]

\[ = \frac{K'/sC}{1 + \frac{1}{A}} \left[ \frac{V_xV_y}{2} + \frac{V_TV_{out}}{A} - \frac{V_{GS}V_{out}}{A} + \frac{V_xV_y}{2} - \frac{V_{GS}V_{out}}{A} + \frac{V_TV_{out}}{A} \right] \]

\[ V_{out} \left[ \left( 1 + \frac{1}{A} \right) \frac{2K'/sC}{A} \left( V_{GS} - V_T \right) \right] = \frac{K'}{sC} V_xV_y \]
\[ V_{out} = \frac{K'/sC}{1 + \frac{1}{A} + \frac{2K'/sC}{A} (V_{GS} - V_T)} \cdot V_{in} \cdot V_C \]
MOSFET-C-Op Amp Integrator Performance

Previous circuit with a finite op amp gain \((A)\) and a threshold variation \((\Delta V_T^+\) and \(\Delta V_T^-)\):

\[
I_1 = K' \left[ \left( V_{GS} + \frac{V_y}{2} - \frac{V_{out}}{2A} - V_T - \Delta V_T^+ \right) \left( \frac{V_x}{2} - \frac{V_{out}}{2A} \right) - \frac{1}{2} \left( \frac{V_x}{2} - \frac{V_{out}}{2A} \right)^2 \right]
\]

\[
I_2 = K' \left[ \left( V_{GS} - \frac{V_y}{2} - \frac{V_{out}}{2A} - V_T - \Delta V_T^+ \right) \left( -\frac{V_x}{2} - \frac{V_{out}}{2A} \right) - \frac{1}{2} \left( \frac{V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right]
\]

\[
I_3 = K' \left[ \left( V_{GS} - \frac{V_y}{2} + \frac{V_{out}}{2A} - V_T - \Delta V_T^- \right) \left( \frac{V_x}{2} + \frac{V_{out}}{2A} \right) - \frac{1}{2} \left( \frac{V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right]
\]

\[
I_4 = K' \left[ \left( V_{GS} + \frac{V_y}{2} + \frac{V_{out}}{2A} - V_T - \Delta V_T^- \right) \left( -\frac{V_x}{2} + \frac{V_{out}}{2A} \right) - \frac{1}{2} \left( \frac{V_x}{2} + \frac{V_{out}}{2A} \right)^2 \right]
\]

\[
V_{out} = \frac{K'/sC}{1 + \frac{1}{A}} (I_4 - I_3 - I_1 + I_2)
\]

\[
= \frac{K'/sC}{1 + \frac{1}{A}} \left[ \frac{V_x V_y}{2} + \frac{\Delta V_T^+ V_o}{A} + \frac{V_T V_o}{2A^2} - \frac{V_{GS} V_o}{A} + \frac{V_x V_y}{2} + \frac{\Delta V_T^- V_o}{A} + \frac{V_T V_o}{2A^2} - \frac{V_{GS} V_o}{A} \right]
\]

\[
V_o = V_{out} = \frac{K'/sC}{1 + \frac{1}{A}} \left[ \frac{V_x V_y}{2} - \frac{2V_{out}}{A} (V_{GS} - V_T) + \frac{V_{out}}{A} (\Delta V_T^+ + \Delta V_T^-) \right]
\]
\[
V_{out} = \frac{K'/sC}{\left(1 + \frac{1}{A}\right) \left[1 + \frac{2K'/sC}{1 + A} (V_{GS} - V_T) - \frac{K'/sC}{1 + A} (\Delta V_T^+ + \Delta V_T^-) \right]} \cdot V_{in} \cdot V_C
\]
True Active RC Filters

Integrator:

Advantages:
- Good linearity
- Wide dynamic range

Disadvantages:
- Op amp must be able to drive resistances
- Requires large area
**Log Domain Integrator**

Log domain filters are essentially $g_m$-$C$ filters with an exponential relationship for $g_m$.

A log domain lowpass filter:

**Analysis:**

Note that $V_{BE1} + V_{BE2} = V_{BE3} + V_{BE4}$

$\Rightarrow \quad i_1i_2 = i_3i_4 \quad \Rightarrow \quad i_{in}I_b = i_3i_{out}$

But, $i_3 = i_C + I_{damp} = C \frac{dv_{BE4}}{dt} + I_{damp}$

and

$$C \frac{dv_{BE4}}{dt} = CV_T \frac{d}{dt} \ln \left( \frac{i_{out}}{I_s} \right) = \frac{CV_T}{i_{out}} \frac{di_{out}}{dt} = \frac{CV_T}{i_{out}} s \cdot i_{out}$$

$\therefore \quad i_{in}I_b = (sCV_T + I_{damp})i_{out}$

$$\Rightarrow \quad \frac{i_{out}}{i_{in}} = \frac{I_b}{sCV_T + I_{damp}} = \frac{I_b}{CV_T} \frac{1}{s + \frac{I_{damp}}{CV_T}}$$

(set $I_{damp} = 0$ for true integrator)

**Comments:**

- The corner frequency of the filter depends on temperature (tuning is required)
• Lack of pure logarithmic VI characteristics and finite $\beta$ of the BJTs lead to linearity degradation
• Good for high frequency (100MHz)
Log Domain Filters - Continued

Differential log-domain integrator:

![Differential log-domain integrator diagram]

Fig. 10-LD1
**Current Mode Integrator**

Circuit:

\[
\begin{align*}
\text{Summation of currents at node } a \text{ gives:} \\
i_n - i_p - i_f &= sCv_a + g_mv_a \\
\text{Also,} \\
-i_f &= g_mv_a \\
\text{Combining,} \\
i_n - i_p - i_f &= \frac{-sC_i_f}{g_m} - i_f \rightarrow i_f = \frac{g_m}{sC} (i_p - i_n) \\
\text{Thus,} \\
i_{\text{out}} &= K_i_f = (i_p - i_n)
\end{align*}
\]

Use the dc current sources (I and KI) to tune the transconductances.
Current Mode Integrators - Continued

Current-Mode Integrator Nonidealities:

Small signal model-

Analysis gives,

\[ i_f = \frac{\frac{g_m}{4g_{ds}}}{\left(1 + \frac{sC_2}{4g_{ds}}\right)\left(1 + \frac{sC_1}{g_m}\right)} \left[\left(1 + \frac{sg_{ds}C_2}{g_m^2}\right)p - \left(1 - \frac{sC_1}{g_m}\right)n\right] \]

- Dominant pole moves from the origin to \(-4g_{ds}/C_2\)
- Unity gain frequency is \(g_m/C_2\)
- The undesired capacitance, \(C_1\), inversely affects the frequency of the non-dominant pole \((g_m/C_1)\) causing an undesirable excess phase shift at the integrator unity gain frequency
- If \(C_2 \gg C_1\), and one uses cascode techniques to reduce the output conductances, Q factors exceeding 20 can be achieved at high frequencies
Current Mode Filters - Continued

Fully-Balanced Current Mode Integrator:

Ideally,

\[ i_{op} - i_{on} = \frac{g_m}{sC} (i_p - i_n) \]

In reality,

\[ i_{op} - i_{on} = A \left( \frac{1 - \frac{s}{z_1}}{1 + \frac{s}{p_1}} \right) (i_p - i_n) \]

where

\[ A = \frac{g_m - g_{ds}}{g_{ds}} \]
\[ z_1 = \frac{g_m - g_{ds}}{2C_{gd}} \]

and

\[ p_1 = \frac{g_{ds}}{C + 4C_{gd}} \]
To achieve high Q factor, the RHP zero should be as large as possible.
SECTION 4 - BIQUADS

**The Biquad**

A biquad has two poles and two zeros.

- Poles are complex and always in the LHP.
- The zeros may or may not be complex and may be in the LHP or the RHP.

Transfer function:

\[
H_a(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{-\left(K_2s^2 + K_1s + K_0\right)}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2} = K \left(\frac{(s-z_1)(s-z_2)}{(s-p_1)(s-p_2)}\right)
\]

- Low pass: zeros at \(\infty\)
- Bandstop: zeros at \(\pm j\omega_o\)
- High pass: zeros at 0
- Allpass: Poles and zeros are complex
- Bandpass: One zero at 0 and the other at \(\infty\)
- conjugates
Development of a Biquad Realization

Rewrite $H_a(s)$ as:

$$s^2V_{out}(s) + \frac{\omega_o s}{Q}V_{out}(s) + \omega_o^2 V_{out}(s) = -(K_2 s^2 + K_1 s + K_0)V_{in}(s)$$

Dividing through by $s^2$ and solving for $V_{out}(s)$, gives

$$V_{out}(s) = \frac{-1}{s} \left[ (K_1 + K_2)s V_{in}(s) + \frac{\omega_o}{Q} V_{out}(s) + \frac{1}{s} (K_0 V_{in}(s) + \omega_o^2 V_{out}(s)) \right]$$

If we define the voltage $V_1(s)$ as

$$V_1(s) = \frac{-1}{s} \left[ \frac{K_0}{\omega_o} V_{in}(s) + \omega_o V_{out}(s) \right]$$

then $V_{out}(s)$ can be expressed as

$$V_{out}(s) = \frac{-1}{s} \left[ (K_1 + K_2)s V_{in}(s) + \frac{\omega_o}{Q} V_{out}(s) - \omega_o V_1(s) \right]$$

Synthesizing the voltages $V_1(s)$ and $V_{out}(s)$, gives (assuming inverting integrators)

![Diagram](image_url)
Example of a Biquad Realization

Connecting the two previous circuits gives:

For negative integrators-

\[ V_{in}(s) \xrightarrow{K_0/\omega_o} -\frac{1}{s} V_1(s) \xrightarrow{-\omega_o} -\frac{1}{s} V_{out}(s) \]

For positive integrators-

\[ V_{in}(s) \xrightarrow{K_0/\omega_o} \frac{1}{s} V_1(s) \xrightarrow{\omega_o} \frac{1}{s} V_{out}(s) \]
More Biquads

KHN:

![KHN Biquad Diagram](image)

Tow-Thomas Biquad:

![Tow-Thomas Biquad Diagram](image)

Needs some additions to make it a universal biquad.
**OTA-C BIQUADS**

**Building Blocks for OTA-C Biquads**

Integrator

\[ V_o = \frac{V_{i1}-V_{i2}}{sC} = \frac{G_{m1}V_{in}}{sC} \]

Amplifier

\[ V_o = \frac{G_{m1}(V_{i1}-V_{i2})}{G_{m2}} = \frac{G_{m1}}{G_{m2}}V_{in} \]

Summing Amplifier

\[ V_o = \frac{G_{m1}V_{i1}}{G_{m2}} + \frac{G_{m2}V_{i2}}{G_{m4}} + \frac{G_{m3}V_{i3}}{G_{m4}} \]

Summing Integrator

\[ V_o = \frac{G_{m1}V_{i1}}{sC} + \frac{G_{m2}V_{i2}}{sC} + \frac{G_{m3}V_{i3}}{sC} \]

Fig. 11.5-06
A Single-Ended Bandpass OTA-C Biquad

![A Single-Ended Bandpass OTA-C Biquad](image)

Transfer function:

\[
\frac{V_{out}(s)}{V_{in}(s)} = \frac{\frac{G_M}{C} s}{s^2 + \frac{G_M}{QC} s + \left(\frac{G_M}{C}\right)^2}
\]
A Differential Bandpass OTA-C Biquad

Transfer function:

\[
\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \frac{\frac{G_M}{C} s}{s^2 + \frac{G_M}{QC} s + \left(\frac{G_M}{C}\right)^2}
\]
Single-Ended OTA-C Biquad

Analysis:

\[ \sum_{i=0}^{\infty} V_a \Rightarrow sC_a V_a - G_m V_{in} + G_m V_o = 0 \]

Solving for \( V_a \):

\[ V_a = -\frac{-G_m V_{in} + G_m V_o}{sC_a} \]

\[ \sum_{i=0}^{\infty} V_o \Rightarrow sC_o V_o + sC_z (V_o - V_{in}) - G_m V_o = 0 \]

Combining equations gives,

\[ sC_o V_o + sC_z (V_o - V_{in}) - G_m \frac{-G_m V_{in} + G_m V_o}{sC_a} + G_m V_o = 0 \]

Which gives,

\[ \left( sC_o + sC_z + \frac{G_m G_m}{sC_a} + G_m \right) V_o = \left( sC_z + \frac{G_m G_m}{sC_a} \right) V_{in} \]

\[ V_o = \frac{s^2 C_z C_a + G_m G_m}{s^2 C_a (C_z + C_b) + sG_m C_a + G_m G_m} \]

\[ \therefore \quad \frac{V_o}{V_{in}} = \frac{s^2 C_z C_a + G_m G_m}{s^2 + \frac{sG_m C_a}{C_z + C_b} + \frac{G_m G_m}{C_a (C_z + C_b)}} \]
A biquad with complex zeros on the $j\omega$ axis.
**Differential OTA-C Biquad**

Differential version of the previous biquad:

![Differential OTA-C Biquad Diagram](Fig.11.5-075)
Example 4-1

Use the biquad of the previous page to design the following transfer function where $K_2 = 4.49434 \times 10^{-1}$, $K_0 = 4.580164 \times 10^{12}$, $\omega_o = 2.826590 \times 10^6$ and $Q = 6.93765$.

$$H_a(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{K_2 s^2 + K_0}{s^2 + \frac{\omega_o}{Q} s + \omega_o^2}$$

Find the values of all transconductances and capacitors and let $C_a = 2 \text{ pF}$, $C_b = 35 \text{ pF}$ and $G_{m1} = G_{m2}$.

Solution

Equating $H_a(s)$ to the previous transfer function gives

$$K_2 = \frac{C_z}{C_b + C_z}, \quad K_0 = \frac{G_{m2} G_{m4}}{C_a (C_b + C_z)}, \quad \frac{\omega_o}{Q} = \frac{G_{m3}}{C_b + C_z}, \quad \text{and} \quad \omega_o^2 = \frac{G_{m1} G_{m2}}{C_a (C_b + C_z)}$$

Thus, $C_z = \frac{K_2 C_b}{1 - K_2} = \frac{0.44943 \cdot 35 \text{ pF}}{1 - 0.44943} = 28.2264 \text{ pF}$

$$G_{m1} = G_{m2} = \frac{\omega_o \sqrt{C_a (C_b + C_z)}}{2.826590 \times 10^6 \sqrt{2(35+28.2264)}} = 31.785 \mu \text{S}$$

$$G_{m3} = \frac{\omega_o}{Q} \frac{C_a (C_b + C_z)}{(35+28.2264) \times 10^{-12}} = 25.760 \mu \text{S}$$

Note that $\frac{K_0}{\omega_o^2} = \frac{G_{m4}}{G_{m1}}$ which gives $G_{m4} = \frac{K_0}{\omega_o^2} G_{m1} = \frac{4.580164 \times 10^{12}}{(2.826590 \times 10^6)^2} 31.785 \mu \text{S} = 18.221 \mu \text{S}$
A Fully Differential, OTA-C General Biquad

Transfer function:

\[
\frac{V_{out}}{V_{in}} = \frac{\left(\frac{C_x}{C_x+C_B}\right)s^2 + \left(\frac{G_{m5}}{C_x+C_B}\right)s + \left(\frac{G_{m2}G_{m4}}{C_A(C_x+C_B)}\right)}{s^2 + \left(\frac{G_{m3}}{C_x+C_B}\right)s + \left(\frac{G_{m1}G_{m2}}{C_A(C_x+C_B)}\right)}
\]
**Two-Thomas Lowpass Biquad**

We shall use this circuit later in a filter example.

![Biquad Circuit Diagram](attachment:fig11514.png)

**Analysis:**

\[ \sum i = 0 \text{ at } V_a \Rightarrow sC_1 V_a - G_{m1}(V_{in} - V_{out}) = 0 \Rightarrow sC_1 V_a = G_{m1}(V_{in} - V_{out}) \]

\[ \sum i = 0 \text{ at } V_{out} \Rightarrow sC_2 V_{o2} + G_{m3} V_{out} - G_{m2} V_a = 0 \]

Combining equations gives,

\[ sC_2 V_{o2} + G_{m3} V_{out} - G_{m2} \left( \frac{G_{m1}(V_{in} - V_{out})}{sC_1} \right) = 0 \Rightarrow (s^2 C_1 C_2 + s C_1 G_{m3} + G_{m1} G_{m2}) V_{out} = G_{m1} G_{m2} V_{in} \]

\[ \therefore \frac{V_{out}}{V_{in}} = \frac{G_{m1} G_{m2}}{s^2 C_1 C_2 + s C_1 G_{m3} + G_{m1} G_{m2}} = \frac{G_{m1} G_{m2}}{C_1 C_2} \]
Nonideal Effects in OTA Biquads

1.) Finite OTA output resistance.

Use structures that can absorb the output resistance.

The admittances, \( Y_1 \) and \( Y_2 \), consist of a capacitor in parallel with a resistor. The resistor can “absorb” the finite output resistances of the OTAs.

2.) Finite OTA bandwidth.

\[
\frac{V_{out}}{V_{in}} = G_m(0) \frac{sRC + 1}{sC(s\tau_a + 1)}
\]

If \( RC = \tau_a \), then the dominant OTA pole is canceled

An alternate approach:

\[
G_{meq} = (G_{m1}(0) - G_{m2}(0)) \left[ 1 - s \frac{G_{m1}(0) \tau_{a1} - G_{m2}(0) \tau_{a2}}{G_{m1}(0) - G_{m2}(0)} \right]
\]

\[
G_{meq} = G_{m1}(0) - G_{m2}(0) \quad \text{if} \quad G_{m1}(0) \tau_{a1} = G_{m2}(0) \tau_{a2}
\]
MOSFET-C BIQUADS

General Second-Order Biquad

Transfer function:

\[
\frac{V_{out}}{V_{in}} = -\frac{\left(\frac{C_1}{C_B}\right)s^2 + \left(\frac{G_2}{C_B}\right)s + \left(\frac{G_1G_3}{C_AC_B}\right)}{s^2 + \left(\frac{G_5}{C_B}\right)s + \left(\frac{G_3G_4}{C_AC_B}\right)}
\]
SECTION 5 - IC FILTER DESIGN

A Design Approach for Lowpass Cascaded Filters

1.) From $T_{PB}$, $T_{SB}$, and $\Omega_n$ (or $A_{PB}$, $A_{SB}$, and $\Omega_n$) determine the required order of the filter approximation, $N$.

2.) From the appropriate tables find the normalized poles of the approximation.

3.) Group the complex-conjugate poles into second-order realizations. For odd-order realizations there will be one first-order term.

4.) Realize each of the terms using first- and second-order blocks.

5.) Cascade the realizations in the order from input to output of the lowest-Q stage first (first-order stages generally should be first).

6.) Denormalize the filter if necessary.

More information can be found elsewhere\textsuperscript{1\textsuperscript{2}\textsuperscript{3}\textsuperscript{4}}.

---


Example 5-1 - Fifth-order, Low Pass, Continuous Time Filter using the Cascade Approach

Design a cascade, switched capacitor realization for a Chebyshev filter approximation to the filter specifications of $T_{PB} = -1$dB, $T_{SB} = -25$dB, $f_{PB} = 100$kHz and $f_{SB} = 150$kHz. Give a schematic and component value for the realization. Also simulate the realization and compare to an ideal realization.

**Solution**

First we see that $\Omega_n = 1.5$. Next, recall that when $T_{PB} = -1$dB that this corresponds to $\varepsilon = 0.5088$. We find that $N = 5$ satisfies the specifications ($T_{SB} = -29.9$dB).

Find the roots for the Chebyshev approximation with $\varepsilon = 0.5088$ for $N = 5$. Therefore we can express the normalized lowpass transfer function as,

$$T_{LPn}(s_n) = T_1(s_n)T_2(s_n)T_3(s_n) = \left[ \frac{0.2895}{s_n + 0.2895} \right] \frac{0.9883}{s_n^2 + 0.1789s_n + 0.9883} \frac{0.4293}{s_n^2 + 0.4684s_n + 0.4293}.$$

Next, we design each of the three stages individually.
Example 5-1 - Continued

Stage 1 - First-order Stage

Let us select OTA-C shown to realize the first-order stage. It is easy to show that the transfer function is given as

\[ T_1(s_n) = \frac{G_{m11n}}{sC_{11n} + G_{m21n}} \]

Equating to \( T_1(s_n) \) gives \( G_{m11n} = G_{m21n} = 0.2895 \text{S} \) and \( C_{11n} = 1\text{F} \).

Next, we unnormalize these values using a value of \( \Omega_n = 10^5 \cdot 2\pi \) and an arbitrary impedance scaling of \( z_o = 10^5 \). Thus, we get the following denormalizes values of

\[ G_{m11} = G_{m21} = \frac{0.2895 \text{S}}{10^5} = 2.895 \mu\text{S} \quad \text{and} \quad C_{11} = \frac{1\text{F}}{10^5 \cdot 2 \times 10^5 \pi} = 15.9 \text{pF} \]

Stage 2 - Second-order, High-Q Stage

The next product of \( T_{Lp_n}(s_n) \) is

\[ T_2(s_n) = \frac{0.9883}{s_n^2 + 0.1789s_n + 0.9883} = \frac{G_{m12}G_{m22}}{s^2 + \frac{G_{m32}C_{22}}{G_{m32}C_{22}} + \frac{G_{m12}G_{m22}}{C_{12}C_{22}}} \]

Assume that \( G_{m12n} = G_{m22n} \) and \( C_{12n} = C_{22n} = 1\text{F} \). Equating coefficients gives, \( G_{m12n} = G_{m22n} = 0.99413 \text{S} \) and \( G_{m32n} = 0.1789 \text{S} \).

Denormalizing with \( \Omega_n = 10^5 \cdot 2\pi \) and \( z_o = 10^5 \) gives \( G_{m12} = G_{m22} = 9.94133 \mu\text{S} \), \( G_{m32} = 1.789 \mu\text{S} \).
and $C_{12} = C_{22} = 15.9 \text{pF}$. 
**Example 5-1 - Continued**

Stage 3 - Second-order, Low-$Q$ Stage

The next product of $T_{LPn}(s_n)$ is

$$T_3(s_n) = \frac{0.4293}{s^2+0.4684s+0.4293} = \frac{G_{m13}G_{m23}}{C_{13}C_{23}} + \frac{G_{m33}G_{m13}G_{m23}}{sC_{23}^2 + C_{13}C_{23}}$$

Assume that $G_{m13} = G_{m23}$ and $C_{13} = C_{23} = 1\text{F}$. Equating coefficients gives $G_{m13} = G_{m23} = 0.6552\text{S}$ and $G_{m32} = 0.4684\text{S}$.

Denormalizing with $\Omega_n = 10^5\cdot 2\pi$ and $z_o = 10^5$ gives $G_{m13} = G_{m23} = 6.552\mu\text{S}$, $G_{m33} = 4.684\mu\text{S}$, and $C_{13} = C_{23} = 15.9\text{pF}$.

Overall realization:

All capacitors are 15.9pF and $G_{m11} = G_{m21} = 2.895\mu\text{S}$, $G_{m12} = G_{m22} = 9.94133\mu\text{S}$, $G_{m32} = 1.789\mu\text{S}$, $G_{m13} = G_{m23} = 6.552\mu\text{S}$, and $G_{m33} = 4.684\mu\text{S}$.
Example 5-1 - OTA-C Cascade Filter

Simulation results:

```plaintext
EXAMPLE 5-1 - OTA-C Cascade Filter
.OPTION LIMPTS=1000
VIN 1 0 DC 0 AC 1.0
R11 1 0 10000MEG
G11 0 2 1 0 2.895U
C11 2 0 15.9P
R2 2 0 10000MEG
G21 2 0 2 0 2.895U
G12 0 3 2 4 9.94133U
C12 3 0 15.9P
R3 3 0 10000MEG
R4 4 0 10000MEG
G32 4 0 4 0 1.789U
G13 0 5 4 6 6.552U
C13 5 0 15.9P
R5 5 0 10000MEG
G23 0 6 5 0 6.552U
C23 6 0 15.9P
G33 6 0 6 0 4.684U
.AC DEC 10 1 10MEG
.PRINT AC VDB(6) VP(6) VDB(4) VP(4) VDB(2) VP(2)
```
G22 0 4 3 0 9.94133U
C22 4 0 15.9P

.PROBE
.END
Low-Power, High Accuracy Continuous Time 450 KHz Bandpass Filter

450 KHz Bandpass filter specifications:

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lower Stopband</td>
<td>400 KHz</td>
</tr>
<tr>
<td>Lower Passband</td>
<td>439 KHz</td>
</tr>
<tr>
<td>Upper Passband</td>
<td>461 KHz</td>
</tr>
<tr>
<td>Upper Stopband</td>
<td>506 KHz</td>
</tr>
<tr>
<td>Passband Ripple</td>
<td>&lt;0.5dB</td>
</tr>
<tr>
<td>Stopband Attenuation</td>
<td>&gt;55dB</td>
</tr>
<tr>
<td>Tuning Resolution</td>
<td>&lt;1%</td>
</tr>
<tr>
<td>Power Consumption</td>
<td>≈2mA</td>
</tr>
<tr>
<td>Total Inband Noise</td>
<td>&lt;340μVrms</td>
</tr>
</tbody>
</table>

Cascaded-biquad BFP structure:

To compromise between the group delay and complexity, a Chebyshev approximation of 12th order is used.

![Diagram](image.png)
**455 KHz Filter - Continued**

Table 3: Values of the parameters of a biquad

<table>
<thead>
<tr>
<th>Biquad</th>
<th>$K_2$</th>
<th>$K_0$</th>
<th>$\omega_0$</th>
<th>$Q$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>4.494340x10^{-1}</td>
<td>4.580164x10^{12}</td>
<td>2.826590x10^{6}</td>
<td>6.93765</td>
</tr>
<tr>
<td>2</td>
<td>7.782568x10^{-1}</td>
<td>4.875459x10^{12}</td>
<td>2.826590x10^{6}</td>
<td>6.937844</td>
</tr>
<tr>
<td>3</td>
<td>3.926305x10^{-1}</td>
<td>2.251889x10^{12}</td>
<td>2.669664x10^{6}</td>
<td>17.80233</td>
</tr>
<tr>
<td>4</td>
<td>3.723485x10^{-1}</td>
<td>4.144164x10^{12}</td>
<td>2.992737x10^{6}</td>
<td>17.80241</td>
</tr>
<tr>
<td>5</td>
<td>6.946693x10^{-1}</td>
<td>1.340316x10^{12}</td>
<td>2.992739x10^{6}</td>
<td>17.80286</td>
</tr>
<tr>
<td>6</td>
<td>2.790028x10^{-1}</td>
<td>9.230595x10^{12}</td>
<td>2.669660x10^{6}</td>
<td>17.80292</td>
</tr>
</tbody>
</table>
455 KHz Filter - Continued

Biquad Design:

![Biquad Diagram](image)

<table>
<thead>
<tr>
<th>Biquad</th>
<th>$G_{m1}(S)$</th>
<th>$G_{m2}(S)$</th>
<th>$G_{m3}(S)$</th>
<th>$G_{m4}(S)$</th>
<th>$C_a$(pF)</th>
<th>$C_a$(pF)</th>
<th>$C_z$(pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3.06493x10^{-5}</td>
<td>3.06493x10^{-5}</td>
<td>1.384522x10^{-5}</td>
<td>1.58119x10^{-5}</td>
<td>4.0</td>
<td>70</td>
<td>30</td>
</tr>
<tr>
<td>2</td>
<td>2.82619x10^{-5}</td>
<td>2.82619x10^{-5}</td>
<td>1.179439x10^{-5}</td>
<td>1.553489x10^{-5}</td>
<td>4.8</td>
<td>30</td>
<td>70</td>
</tr>
<tr>
<td>3</td>
<td>3.636792x10^{-5}</td>
<td>3.636792x10^{-5}</td>
<td>1.376188x10^{-5}</td>
<td>1.034177x10^{-5}</td>
<td>3.8</td>
<td>70</td>
<td>50</td>
</tr>
<tr>
<td>4</td>
<td>3.481898x10^{-5}</td>
<td>3.481898x10^{-5}</td>
<td>1.125281x10^{-5}</td>
<td>1.449968x10^{-5}</td>
<td>7.8</td>
<td>60</td>
<td>20</td>
</tr>
<tr>
<td>5</td>
<td>3.909740x10^{-5}</td>
<td>3.909740x10^{-5}</td>
<td>1.064108x10^{-5}</td>
<td>5.265746x10^{-5}</td>
<td>10.4</td>
<td>60</td>
<td>8</td>
</tr>
<tr>
<td></td>
<td>3.902157x10^{-5}</td>
<td>3.902157x10^{-5}</td>
<td>1.188262x10^{-5}</td>
<td>4.548467x10^{-5}</td>
<td>6.8</td>
<td>60</td>
<td>40</td>
</tr>
</tbody>
</table>
**455 KHz Filter - Continued**

Noise minimization of the OTA:

The equivalent input-referred noise of the OTA is,

\[
v_n^2 = \sum_{i=1,4,10,12} 2v_{ni}^2 = 2 \left( \frac{g_{mi}}{g_{m1}} \right)^2 e_{ni}^2 \sum_{i=1,4,10,12}
\]

where

\[
v_n^2 = \text{Noise power of transistors M1, M4, M10, and M12}
\]

Using the definition of flicker-noise spectral density we get

\[
e_{ni}^2 = S(f) = \frac{K_F}{C_{ox}} \Delta f
\]

Using the definition of flicker-noise spectral density we get

\[
e_{ni}^2 = S(f) = \frac{K_F}{C_{ox}} \Delta f
\]

Substituting into the above equation gives

\[
v_n^2 = \sum_{i=1,4,10,12} 2 \left( \frac{g_{mi}}{g_{m1}} \right)^2 \left[ \sum_{i=1,4,10,12} \frac{K_F}{C_{ox}} \Delta f + 4kT \left( \frac{2}{3} \frac{\Delta f}{g_{mi}} \right) \right]
\]

It can be shown that the equivalent input-referred noise is minimized in the bandwidth of interest if

---

Fig.11.6-08
\[ L_1 \approx \sqrt{\left[ \frac{1}{L_4^2} + \frac{K_{Fp} K_P' \cdot 2}{K_{Fn} K_N' \cdot L_{10}^2} \right]^{-1}} \approx \sqrt{\left[ 1 + \frac{1.3 \times 10^{-24} \cdot 50 \mu}{3.6 \times 10^{-24} \cdot 140 \mu^2} \right]^{-1}} \approx 1 \mu m \]
455KHz Filter - Continued

Nominal Frequency Response:

Fig. 8: Frequency-responses of the 450-KHz BPF - $V_c = 2.1$ V
455KHz Filter - Continued
Temperature dependence of the filter:

![Diagram of 455KHz Bandpass Filter Simulation]

Fig. 10: Frequency-response of the 450-KHz BPF - $V_c = 2.1\text{ V} - T \in [-40^\circ\text{C}, 27^\circ\text{C}, 85^\circ\text{C}]$ - Typical

We will consider the tuning of this filter in the next section.
RF Image Reject Filter

In many RF and IF applications, bandpass filters are very difficult because large Q exacerbates the difficulties in achieving the desired performance.

A possible solution is to use notch filtering in place of bandpass filtering. The following is an example of a image reject filter suitable for GSM applications.

Typical heterodyne receiver:
RF Image Reject Filter

Principle:

Does not require high-Q inductors
No stability problems
No excessive noise and linearity degradation
RF Image Reject Filter

Filter Architecture:

Comments:
- Uses low-Q, on-chip spiral inductors
- Linearity and noise is determined by the transconductors and the impedance level

Transfer function:

\[
G_m(s) = \frac{I_{out}}{V_{in}}(s) = g_m \frac{s^2 + s(\frac{r_s}{L_s} - \frac{g_{mz}}{C}) + \left(1 - \frac{g_{mz} r_s}{L_s C}\right)}{s^2 + \frac{r_s}{L_s} + \frac{1}{L_s C}} = g_m \frac{s^2 + \frac{\omega_z}{Q_z} s + \omega_z^2}{\omega_p^2 s + \omega_p^2}
\]

where
\[
\omega_z = \sqrt{\frac{1 - g_{mz} r_s}{L_s C}} \quad Q_z = \frac{\omega_z}{\frac{r_s}{L_s} - \frac{g_{mz}}{C}}
\]
**RF Image Reject Filter**

Transconductors:

\[ I_o = g_m V_{in} \]

Comments:
- Simple source-coupled differential pair is used for less noise
- Minimum size transistors is selected to minimize parasitic capacitance
- Linearity is a function of the gate-source overdrive voltage and hence the tail current
RF Image Reject Filter

Frequency Response:

Tuning to be considered later.
**Q-Enhanced LC Filters**

Second-order, Bandpass Filter:

1GHz capability in bipolar.
Q-Enhanced LC Filters - Continued

CMOS Q-enhanced Filter (Kuhn†):

Second-order circuit-

Differential architecture allows:

- Negative resistances implemented by positive feedback (M2A & M2B)
- Reduces power supply noise
- Second-order nonlinearities are cancelled
- Increases the signal swing

Results:

Q's of up to 10,000 at 100MHz

Ladder Filter Design (Low Pass)

1.) From $T_{BP}$, $T_{SB}$, and $\Omega_n$ (or $A_{PB}$, $A_{SB}$, and $\Omega_n$) determine the required order of the filter approximation.

2.) From tables similar to Table 9.7-3 and 9.7-2 find the RLC prototype filter approximation.

3.) Write the state equations and rearrange them so each state variable is equal to the integrator of various inputs.

4.) Realize each of rearranged state equations by continuous time integrators or switched capacitor integrators.

5.) Denormalize the filter if necessary.
**Example 5-2 - Fifth-order, Low Pass, OTA-C Filter using the Ladder Approach**

Design a ladder, OTA-C realization for a Chebyshev filter approximation to the filter specifications of $T_{BP} = -1dB$, $T_{SB} = -25dB$, $f_{PB} = 100kHz$ and $f_{SB} = 150 kHz$. Give a schematic and component value for the realization. Also simulate the realization and compare to an ideal realization. Adjust your design so that it does not suffer the -6$dB$ loss in the pass band. (Note that this example should be identical with Ex. 5-1.)

**Solution**

From Ex. 5-1, we know that a 5th-order, Chebyshev approximation will satisfy the specification. The corresponding low pass, $RLC$ prototype filter is

\[
\begin{align*}
V_{in}(s_n) &\rightarrow L_5n=2.1349 \, \text{H} & L_3n=3.0009 \, \text{H} & L_{1n}=2.1349 \, \text{H} \\
+ &\rightarrow C_{4n}=1.0911 \, \text{F} & C_{2n}=1.0911 \, \text{F} & + \\
\rightarrow &\rightarrow 1 \, \text{Z} & 1 \, \text{Z} & V_{out}(s_n) \\
&\rightarrow \rightarrow \rightarrow \rightarrow \rightarrow \rightarrow \\
&\rightarrow \rightarrow \rightarrow \rightarrow \rightarrow \rightarrow \\
\end{align*}
\]

Next, we must find the state equations and express them in the form of an integrator. Fortunately, the above results can be directly used in this example.
Example 5-2 - Continued

\[ L_{1n}: \quad V'_1(s) = \frac{R'}{s_n L_{1n}} \left[ V_{in}(s) - V_2(s) - \left( \frac{R_{0n}}{R'} \right) V'_1(s) \right] \]  

(1)

This equation can be realized by the OTA-C integrator shown which has one noninverting input and two inverting inputs. The transfer function for this integrator is

\[ V_1(s) = \frac{1}{C_{1n}s_n} \left[ G_{m11n} V_{in}(s) - G_{m21n} V_2(s) - G_{m31n} V'_1(s) \right] \]  

(2)

Choosing \( L_{1n} = C_{1n} = 2.1349F \) gives \( G_{m11n} = G_{m21n} = G_{m31n} = 1S \) assuming that \( R_{0n} = R' = 1\Omega \). Also, double the value of \( G_{m11n} (G_{m11n} = 2S) \) in order to gain 6dB and remove the -6dB of the RLC prototype.

\[ C_{2n}: \quad V_2(s) = \frac{1}{s_n R'C_{2n}} \left[ V'_1(s) - V'_3(s) \right] \]  

(3)

This equation can be realized by the OTA-C integrator shown which has one noninverting input and one inverting input. As before we write that

\[ V_2(s) = \frac{1}{s_n C_{2n}} \left[ G_{m12n} V'_1(s) - G_{m22n} V'_3(s) \right] \]  

(4)

Choosing \( C_{2n} = 1.0911F \) gives \( G_{m12n} = G_{m22n} = 1S \)
Example 5-2 - Continued

\[ L_{3n}: \quad k_3(n) = \frac{R'}{sL_{3n}} [V_2(s_n) - V_4(s_n)] \]  \hspace{1cm} (5)

Eq. (5) can be realized by the OTA-C integrator shown which has one noninverting input and one inverting input. For this circuit we get

\[ V_3'(s_n) = \frac{1}{s_nC_{3n}} \left[ G_{m13n}V_2(s_n) - G_{m23n}V_4(s_n) \right] \]  \hspace{1cm} (6)

Choosing \( L_{3n} = C_{3n} = 3.0009F \) gives \( G_{m13n} = G_{m23n} = 1S \)

\[ C_{4n}: \quad V_4(s_n) = \frac{1}{sR'C_{4n}} \left[ V_2'(s_n) - \left( \frac{R'}{R_{6n}} \right) V_{out}(s_n) \right] \]  \hspace{1cm} (7)

Eq. (7) can be realized by the OTA-C integrator shown with one noninverting and one inverting input. As before we write that

\[ V_4(s_n) = \frac{1}{s_nC_{4n}} \left[ G_{m14n}V_3'(s_n) - G_{m24n}V_{out}(s_n) \right] \]  \hspace{1cm} (8)

Choosing \( C_{4n} = 1.0911F \) gives \( G_{m14n} = G_{m24n} = 1S \)

\[ L_{5n}: \quad V_{out}(s_n) = \frac{R_{6n}}{sL_{5n}} [V_4(s_n) - V_{out}(s_n)] \]  \hspace{1cm} (9)

The last state equation, Eq. (9), can be realized by the OTA-C integrator shown which has one noninverting input and one inverting input. For this circuit we get

\[ V_{out}(s_n) \approx \frac{1}{s_nC_{5n}} \left[ G_{m15n}V_4(s_n) - G_{m25n}V_{out}(s_n) \right]. \]  \hspace{1cm} (10)
Choosing $L_{5n} = C_{5n} = 2.1439\,\text{F}$ gives $G_{m_{14n}} = G_{m_{24n}} = 1\,\text{S}$
Example 5-2 - Continued

Realization:

To denormalize, \( \Omega_n = 200,000\pi \) and pick \( z_o = 10^5 \).

\[
\therefore C_1 = 33.9780\text{pF}, \quad C_2 = 17.3654\text{pF}, \quad C_3 = 47.7608\text{pF}, \quad C_4 = 17.3654\text{pF}, \quad \text{and} \quad C_5 = 33.9780\text{pF}
\]

All transconductances are \( G_{mi} = 10\mu\text{S} \).
Example 5-2 - Continued

Simulation of Example 5-2:

![Graph showing frequency response with dB scale and frequency axis from 10 Hz to 1 M Hz]
10th Order Bandpass Ladder OTA-C Filter

An example of the leapfrog BPF structure - Order N = 10
Log Domain Filters

Fourth-order log-domain bandpass filter:

Fig. 10-LD2
Log Domain Filters - Continued

Measured results of the fourth-order, bandpass filter of previous page:

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency tuning range</td>
<td>50MHz - 130MHz</td>
</tr>
<tr>
<td>Integrator bias current</td>
<td>220µA @ f_o = 130MHz</td>
</tr>
<tr>
<td>Integrating capacitors</td>
<td>8pF and 32pF</td>
</tr>
<tr>
<td>Power consumption with V_CC = 5V</td>
<td>233mW @ f_o = 130MHz</td>
</tr>
<tr>
<td>Quality factor</td>
<td>~</td>
</tr>
<tr>
<td>3rd-order intermodulation distortion</td>
<td>-45.6dB @ f_o = 83MHz</td>
</tr>
<tr>
<td>Output current 3rd-order intercept point</td>
<td>-14.5dBm @ f_o = 83MHz</td>
</tr>
<tr>
<td>Output noise power density @ f_o = 83MHz</td>
<td>-152.4 dBm/Hz</td>
</tr>
</tbody>
</table>
SECTION 6 - FILTER TUNING

**Tuning Methods For Continuous Time Filters**

In all tuning methods, an on-chip reference circuit is monitored and tuned. The main filter becomes tuned by virtue of matching with the on-chip reference circuit.

Common Techniques for Automatic Filter Tuning:

- **Resistive Tuning:**
  - Main Filter
  - $V_c$
  - $G$ or $G_M$
  - Resistance Comparator
  - External Resistor
  - Comparator

- **Phase Tuning:**
  - Main Filter
  - $V_c$
  - VCO
  - Phase Comparator
  - Clock

- **Frequency Tuning:**
  - Main Filter
  - $V_c$
  - VCF
  - Phase Comparator
  - External Signal
**Tuning Methods**

Indirect method:
- The filter is tuned in place (“in situ”). A master filter which is not in the system is tuned and the tuning signals are applied to the slave filter which is in the system.
- This can be done at a high rate so the filter is constantly being tuned or infrequently such as at power up or during some predetermined calibration period.

Direct method:
- The filter is taken out of the circuit and tuned. If another filter has been tuned it can be inserted in the circuit while the other filter is being tuned.
Master-Slave (Indirect) Tuning Scheme

Comments:
• Filter (slave) does not need to be disconnected from the system
• Two filters are required
**Direct Tuning**

Example:

Tuning procedure:
1.) The filter is take apart into several first- or second-order sections.
2.) Each section is tuned to the center frequency.

Comments:
- Necessary switches may influence the filter performance
• Need to remember the tuning voltages (memory)
Tuning a High-Q Bandpass Filter using Direct Tuning

Block diagram for the direct tuning of the previous 455KHz bandpass filter†.

Tuning procedure:
1.) The filter is removed from the circuit and a step voltage applied.
2.) The number of cycles in the ringing waveform and their period is used to tune the filter.
3.) Most tuning algorithms work better with a linear frequency tuning voltage relationship.

A Linear, Differential-Compensated Tuning Scheme

Avoids the problem of nonlinearity in the tuning scheme.

Oscillations between two states of tuning iteration can occur. This becomes a problem as the frequency closely approaches the desired frequency.
A Tuning Scheme that Avoids the Oscillation Problem

Operation:
1.) A coarse tuning cycle using the LDC algorithm is used to find the actual frequency.
2.) The difference between the actual frequency and desired frequency is used to begin a successive approximation cycle to fine-tune the filter.
Example of the High-Q Filter Step Response

Generation of a square wave from the step response:

\[
\text{Amplitude} = g_m R_o \approx \frac{R_{upper} |R_{M3}(g_m r_{ds})^2}{R_{M3}} = (g_m r_{ds})^2 \text{ if } R_{upper} \gg R_{M3}(g_m r_{ds})^2
\]

Tuning time is approximately 800μs to an accuracy within ±1%.
**Constant-Q Tuning**

By taking the advantage of the OTA topology and proper sizing, the Q-factor of the filter will remain constant during tuning as illustrated below.
### Simulation Results of the 450KHz Filter

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Simulated Results</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center frequency</td>
<td>450±1.8KHz</td>
</tr>
<tr>
<td>Bandwidth (-3dB)</td>
<td>22 KHz</td>
</tr>
<tr>
<td>Gain at 450 KHz (Max.)</td>
<td>≈ 8dB</td>
</tr>
<tr>
<td>Tunable frequency</td>
<td>225KHz to 675KHz</td>
</tr>
<tr>
<td>Tuning time</td>
<td>800µs</td>
</tr>
<tr>
<td>Total in-band noise</td>
<td>314µV&lt;sub&gt;rms&lt;/sub&gt;</td>
</tr>
<tr>
<td>Maximum single-level</td>
<td>&lt;100dBµV</td>
</tr>
<tr>
<td>Power dissipation</td>
<td>2.2mA from 3V</td>
</tr>
<tr>
<td>Power Supply</td>
<td>3V</td>
</tr>
</tbody>
</table>
**Tuning of the RF Image Reject Filter**

Filter:

![Filter Diagram]

Direct tuning method:

![Direct Tuning Diagram]
Filter becomes a voltage controlled oscillator (VCO).
**Tuning Algorithm**

Since the poles of the VCO and zeros of the filter are identical, tuning the poles tunes the zeros.

Need to tune both the frequency and the notch depth.

Root locations:
Notch Filter and Tuning Circuits
Notch Filter and Tuning Circuits - Continued

Entire circuit:
Phase Detector and Lowpass Filter

[Diagram of a phase detector and lowpass filter circuit]

[Graph showing the output voltage (V) vs. phase (rad)]
Amplitude-Locked Loop

Input rectified at the emitters of Q1 and Q2.
Tuning Response
Frequency tuning voltage

Q tuning voltage →

VCO output

Frequency response →
Tuning From One Frequency to Another

Frequency control voltages for 1.175 GHz and 1.085 GHz

Q control voltage for 1.175 GHz and 1.085 GHz

Frequency response for 1.175 GHz and 1.085 GHz
RF Image Reject Filter Layout

Simulation results: Notch attenuation over 200KHz ≈ 20-25dB, NF ≈ 3-4dB
### SECTION 7 - SUMMARY

**Comparison of Filter Types**

<table>
<thead>
<tr>
<th>Filter Type</th>
<th>Advantages</th>
<th>Disadvantages</th>
<th>Frequency</th>
<th>Applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Digital</td>
<td>Precision, dynamic range, programmability</td>
<td>Power consumption, chip area, aliasing, ADC requirements, external clock requirement</td>
<td>&lt;10MHz</td>
<td>Low IF filtering, baseband filtering and signal processing</td>
</tr>
<tr>
<td>Passive LC</td>
<td>Dynamic range, stability</td>
<td>Quality factor, chip area</td>
<td>&gt;100MHz</td>
<td>Power amplifier harmonic suppression, low Q RF preselection</td>
</tr>
<tr>
<td>Electro Acoustic</td>
<td>Dynamic range, stability</td>
<td>Process modifications, chip area</td>
<td>&gt;100MHz</td>
<td>RF preselection, IF filtering</td>
</tr>
<tr>
<td>Switched Capacitor</td>
<td>Precision</td>
<td>Dynamic range at high Q, aliasing, external clock requirement</td>
<td>&lt;10MHz</td>
<td>Low frequency, moderate Q IF filtering, baseband filtering</td>
</tr>
<tr>
<td>Gm-C</td>
<td>Frequency of operation</td>
<td>Dynamic range at high Q, tuning requirement</td>
<td>&lt;100MHz</td>
<td>Moderate Q IF filtering, baseband filtering</td>
</tr>
<tr>
<td>Q-enhancement LC</td>
<td>Dynamic range, stability</td>
<td>Chip area, tuning requirement</td>
<td>&gt;100MHz</td>
<td>RF preselection, IF filtering</td>
</tr>
<tr>
<td>Current Mode Filters</td>
<td>Frequency of operation</td>
<td>Dynamic range at high Q, tuning requirement</td>
<td>&lt;150MHz</td>
<td>Moderate Q IF filtering, baseband filtering</td>
</tr>
</tbody>
</table>
**What is the Future of IC Filters?**

- To be attractive, integrated circuit filters must be:
  - Low power
  - Accurate
  - Small area
  - Low noise
  - Large dynamic range (linear)
- Active filters will be limited to around 100MHz and will be implemented by OTA-C, log-domain, current mode techniques. The key is to reduce power, area and noise.
- The higher the Q in bandpass filters, the more difficult the filter is to implement.
- RF filtering can be done using notch filters or filters with \( j\omega \) axis zero.
- Submicron CMOS (<0.25\( \mu \)m) will allow filters to work at higher frequencies but the success of these filters depends on clever circuit techniques.
- SiGe BiCMOS probably will allow integrated circuit filters up to 10GHz using clever circuit techniques.
- Switched capacitor filters continue to be the technique of choice up to several MHz.