System on a Chip

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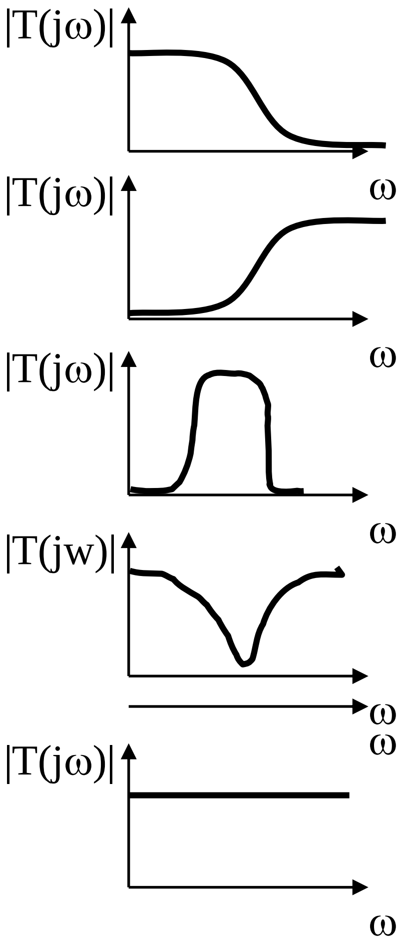
Lecture 4:
Filters

- Filters
  - General Theory
  - Continuous Time Filters

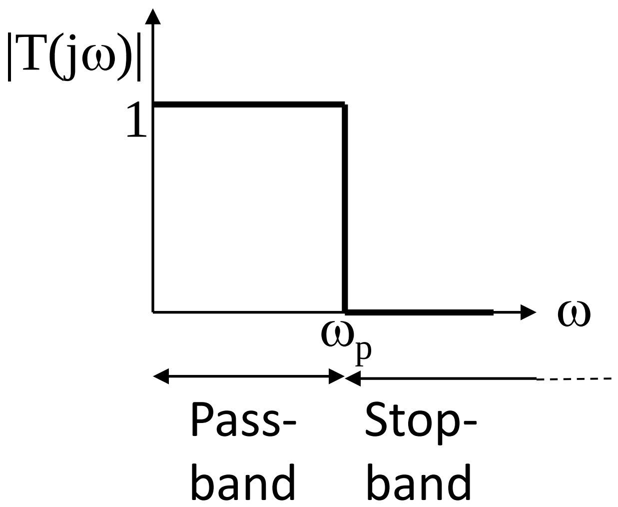
Background

- Filters are used to separate signals in the frequency domain, e.g. remove noise, tune to a radio station, etc
- 5 types of filter
  - Low pass
  - High pass
  - Band pass
  - Band stop/reject
  - All pass

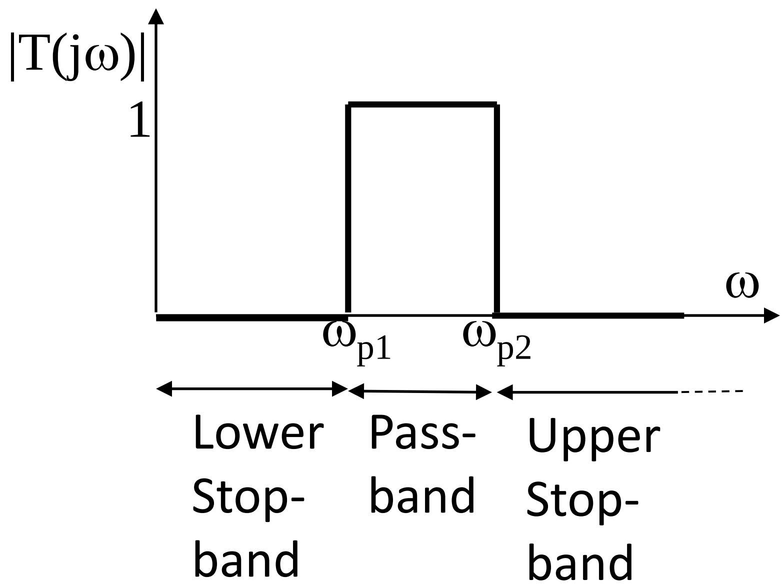


Ideal Filter

Brick-wall LP filter

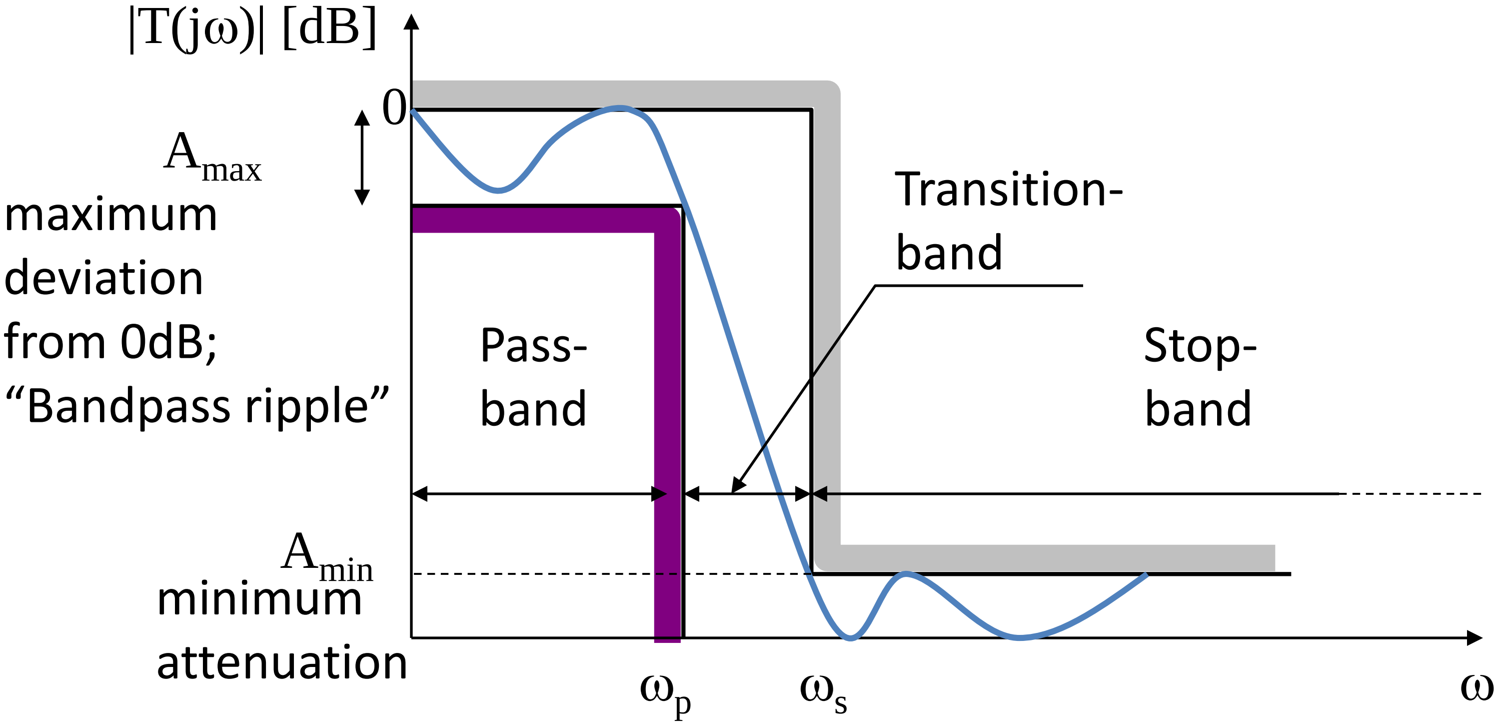


Brick-wall BP filter



- Brick-wall filters do not exist in reality
- Real filters can approximate brick-wall filters as close as required by the filter specification

“Real” LP Filter

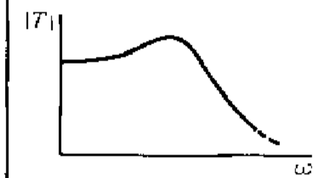
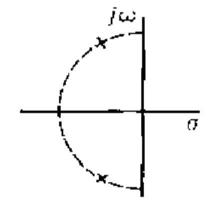
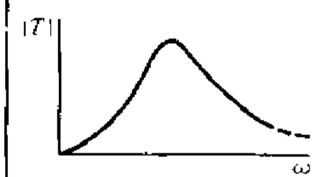
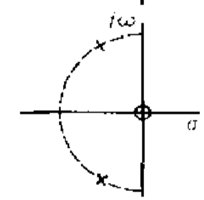
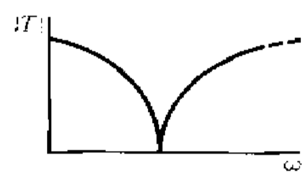
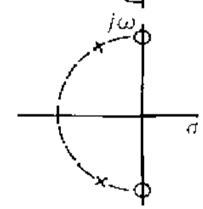
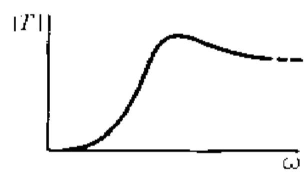
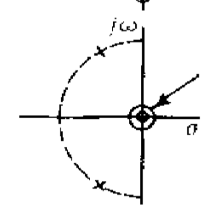
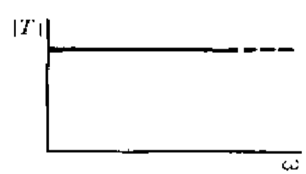
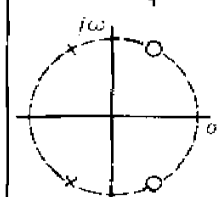


ω_p/ω_s: Measure for filter sharpness; filter selectivity

- Parameters required for filter synthesis: A_{max}, ω_p, A_{min}, ω_s

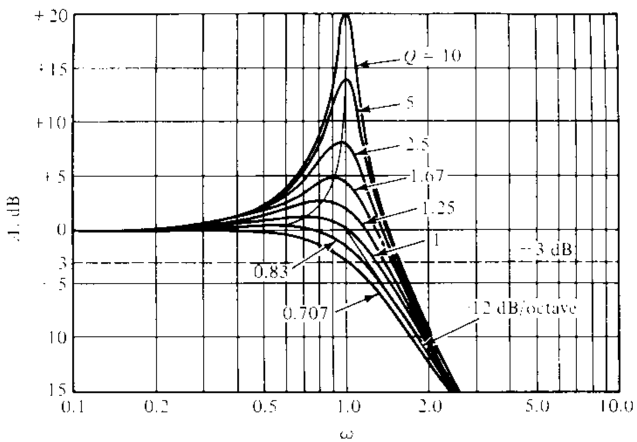
Filter Types Using Biquads

$$T(s) = K\frac{a_2 s^2 + a_1 s + a_0}{b_2 s^2 + b_1 s + b_0}$$

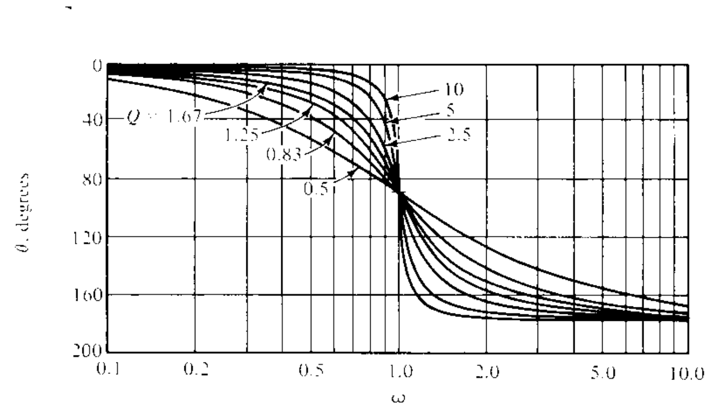
	Frequency response	*Poles/zeros*	*Name*
$T_{LP} = \dfrac{\omega_0^2}{s^2 + \dfrac{\omega_0}{Q}s + \omega_0^2}$			Lowpass
$T_{BP} = \dfrac{\dfrac{\omega_0}{Q}s}{s^2 + \dfrac{\omega_0}{Q}s + \omega_0^2}$			Bandpass
$T_{BE} = \dfrac{s^2 + \omega_0^2}{s^2 + \dfrac{\omega_0}{Q}s + \omega_0^2}$			Bandstop "notch"
$T_{HP} = \dfrac{s^2}{s^2 + \dfrac{\omega_0}{Q}s + \omega_0^2}$			Highpass
$T_{AP} = \dfrac{s^2 - \dfrac{\omega_0}{Q}s + \omega_0^2}{s^2 + \dfrac{\omega_0}{Q}s + \omega_0^2}$			Allpass

Biquadratic LP Transfer Function

Magnitude Response



Phase Response



Low-pass biquad TF

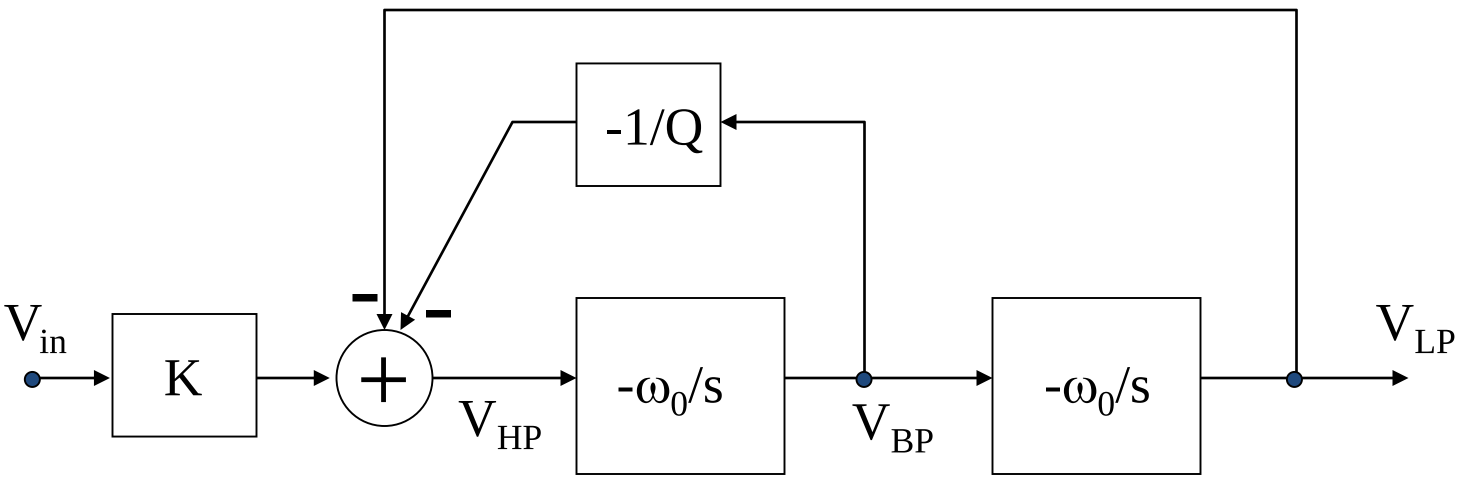
$$T(s) = K \frac{\omega_0{}^2}{s^2 + (\omega_0 / Q)s + \omega_0{}^2}$$

- Diagrams normalized to ω_0 = K = 1
- Asymptotic fall is -40 dB/dec

Biquad Block Diagram

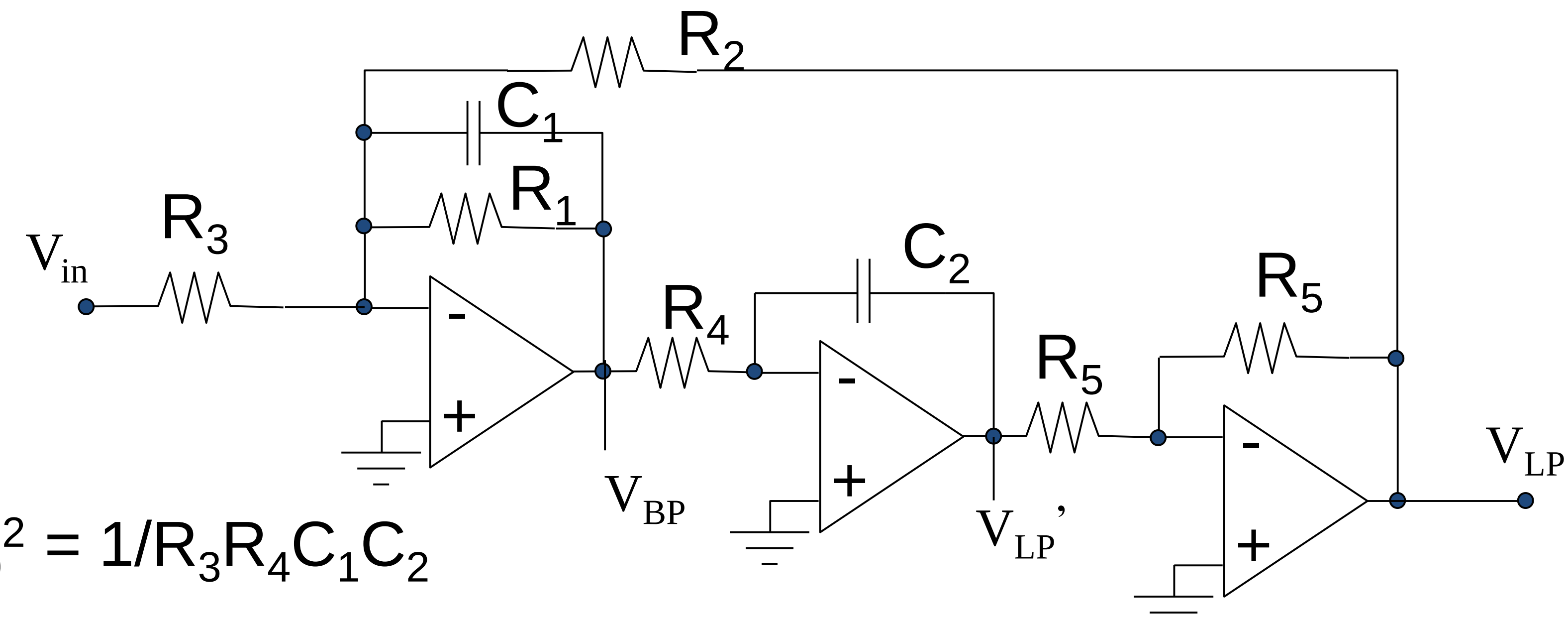
$$T(s) = \frac{V_{LP}(s)}{V_{IN}(s)} = K \frac{\omega_0^{\ 2}}{s^2 + (\omega_0 / Q)s + \omega_0^{\ 2}}$$

(K either pos. or neg.)



- Universal Active Filter: realizes LP, HP, and BP

Tow-Thomas Biquad Realization



$\omega_0^2 = 1/R_3R_4C_1C_2$

$Q = \mathrm{Sqrt}(R_1^2C_1/R_2R_4C_2)$

$K = -R_2/R_3$

(R_5 arbitrarily chosen)

V_{LP}: Inverting LP Filter

V_{LP}': Non-inverting LP Filter

N-th Order Filter

$$T(s) = K\frac{(s-z_1)(s-z_2)\cdots(s-z_M)}{(s-p_1)(s-p_2)\cdots(s-p_N)}$$

Number of poles determines order

- zeros are obviously placed in stopband
- for stability: M≤N; N-M zeros at $\omega = \infty$
- for stability: Re{pi} < 0
- no general optimisation algorithms known

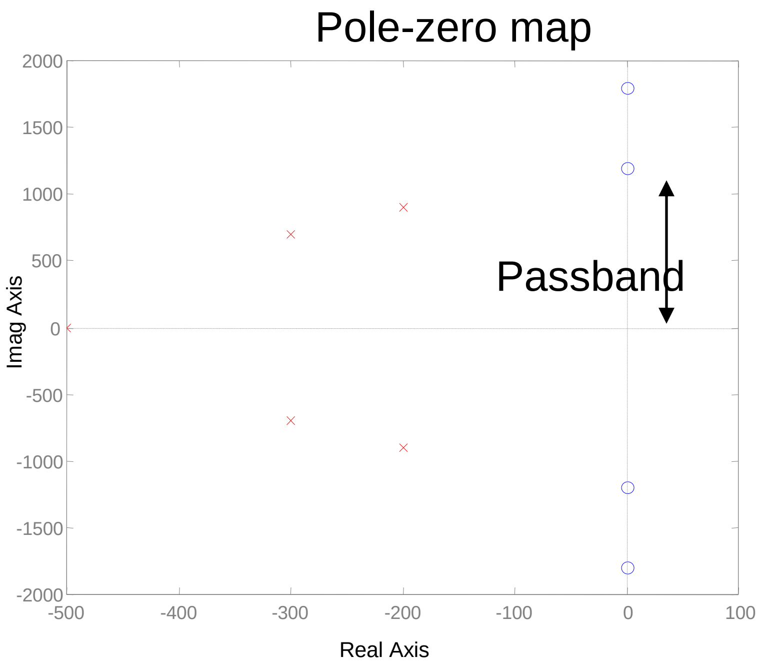
- Special Case: all zeros at $\omega = \infty$; all pole filter

$$T(s) = K\frac{1}{(s-p_1)(s-p_2)\cdots(s-p_N)}$$

Example: 5-th Order Filter

Poles: P[1..5] = 1.0e+002 *[-3.0000 + 7.0000i
-3.0000 - 7.0000i
-2.0000 + 9.0000i
-2.0000 - 9.0000i
-5.0000]

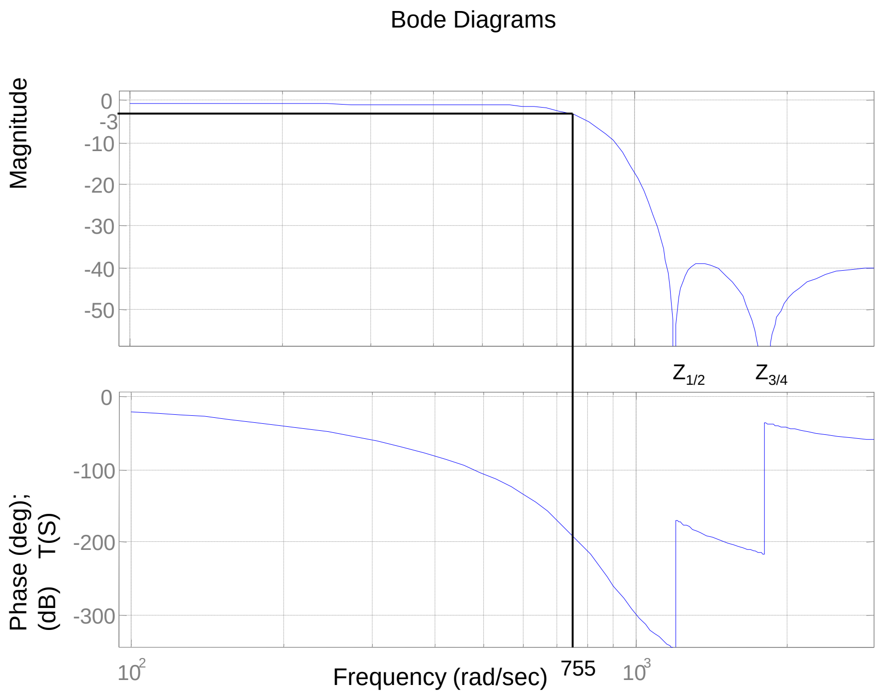
Zeros: Z[1..4] = 1.0e+003 *[0 + 1.2000i
0 - 1.2000i
0 + 1.8000i
0 - 1.8000i]

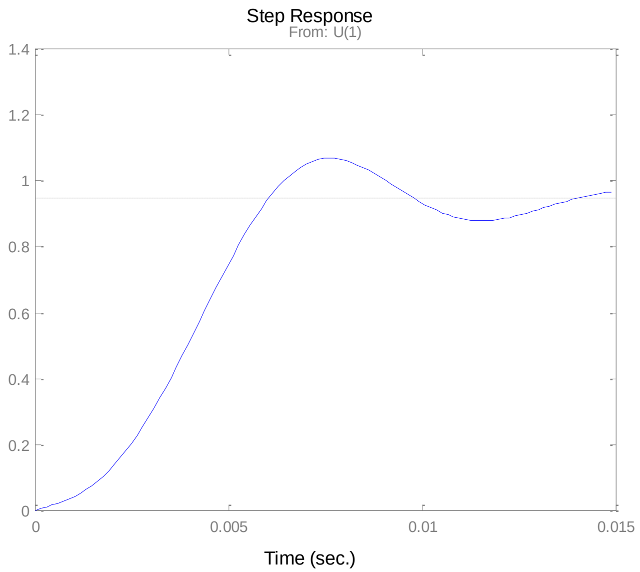


$$T(s) = \frac{50s^4 + 1.44 \cdot 10^8 s^2 + 2.3 \cdot 10^{14}}{s^5 + 1500s^4 + 2.17 \cdot 10^6 s^3 + 1.58 \cdot 10^9 s^2 + 8.64 \cdot 10^{11} s + 2.47 \cdot 10^{14}}$$

Example: 5-th Order Filter

$$T(s) = \frac{50s^4 + 1.44\cdot 10^8 s^2 + 2.3\cdot 10^{14}}{s^5 + 1500s^4 + 2.17\cdot 10^6 s^3 + 1.58\cdot 10^9 s^2 + 8.64\cdot 10^{11} s + 2.47\cdot 10^{14}}$$





Butterworth LP Filter

- Make $T(j\omega)$ so that:

$$|T(j\omega)|^2 = \frac{1}{1+\left(\frac{\omega}{\omega_0}\right)^{2N}}$$

- N: Filter order
- All pole filter

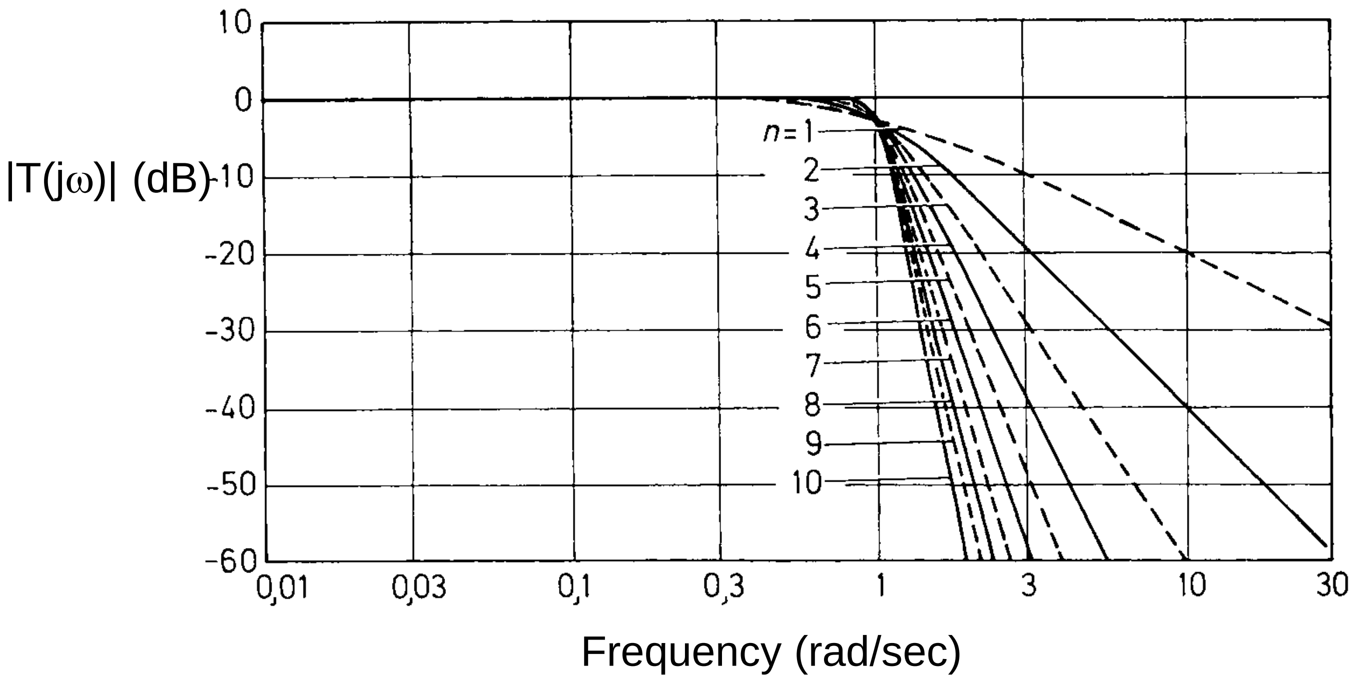
ω_0: $|T(j\omega)|$ has dropped by 3 dB

Normalized Butterworth Polynomials:

For ω_0=1:

N	**Denominator of T(s)**
1	$(s+1)$
2	$(s^2+1.1414s+1)$
3	$(s+1)(s^2+s+1)$
4	$(s^2+0.765s+1)(s^2+1.848s+1)$
5	$(s+1)(s^2+0.618s+1)(s^2+1.618s+1)$

BW-LP Frequency Response

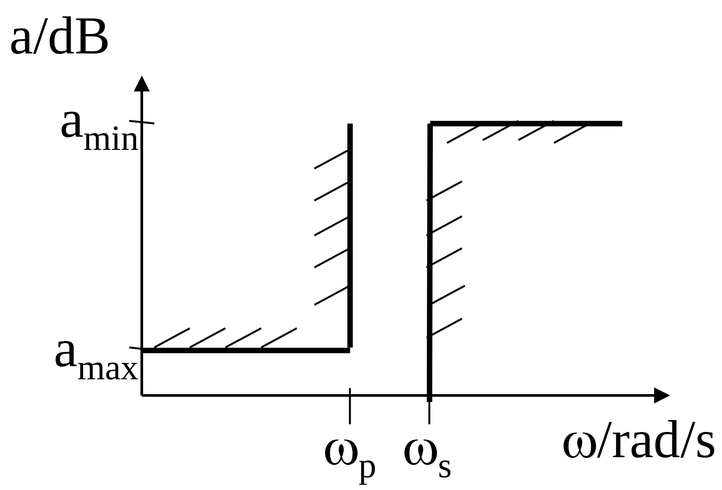


- maximally flat in passband i.e. the first 2N-1 derivatives of $|T(j\omega)|$ are 0 at $\omega=0$
- $|T(j\omega)|$ monotonically falling
- not steepest roll-off

BW-LP Design

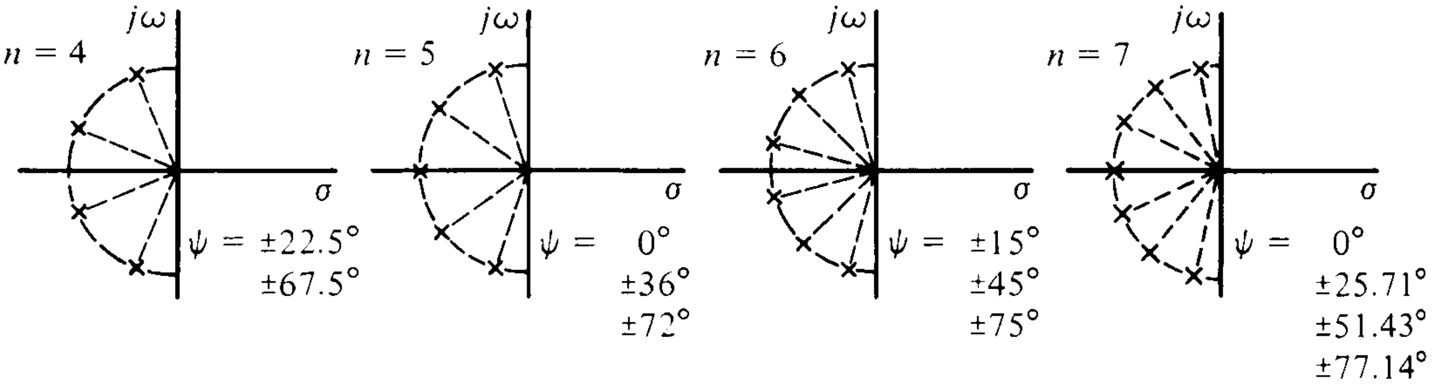
$$\omega_0 = \frac{\omega_p}{\left(10^{a_{\max}/10} - 1\right)^{\frac{1}{2N}}}$$

$$N = \frac{\log\left(\dfrac{10^{a_{\min}/10} - 1}{10^{a_{\max}/10} - 1}\right)}{2\log\left(\dfrac{\omega_p}{\omega_s}\right)}$$



- Design a filter so that in the passband $|T(j\omega)|$ has fallen not more than by a_{max} and in the stopband the minimum attenuation is a_{min}
- Find ω_0 and N

BW Pole Locations



- Poles located on a circle around the origin
- $\psi_k = 90° (2k + N - 1)/N \quad k = 1,2,\ldots,2N$
- If N is odd, then there is a pole at $\psi = 0$, if N is even there are poles at $\psi = \pm 90°/N$
- Poles are separated by $\psi = 180°/N$

Chebychev LP Filter

- Make T(jω) so that:

$$|T(j\omega)|^2 = \frac{1}{1+\varepsilon^2 C_N{}^2(\omega)}$$

$$C_N(\omega) = \cos(N\cos^{-1}(\omega)) \quad for\ \omega < 1$$
$$C_N(\omega) = \cosh(N\cosh^{-1}(\omega)) \quad for\ \omega > 1$$

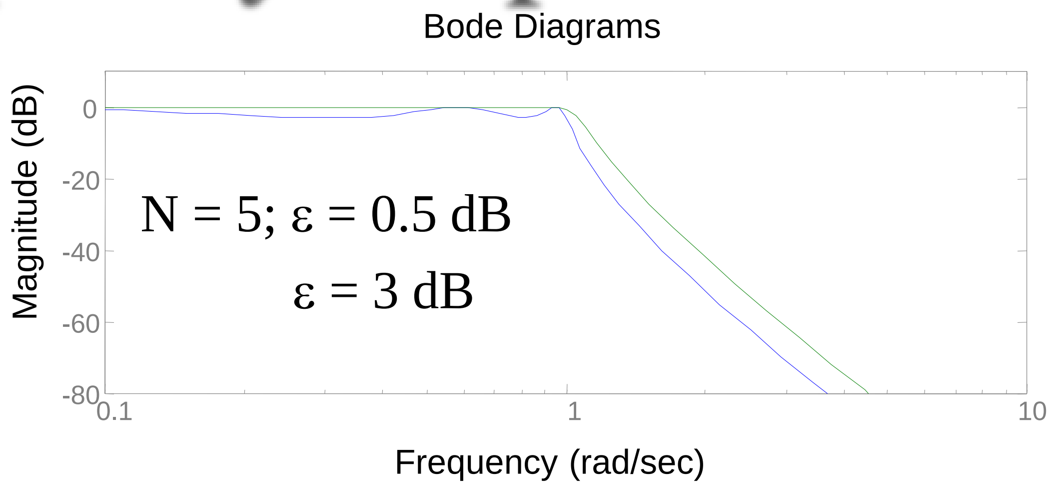
- N: Filter order
- All pole filter
- Normalized for $\omega_0 = 1$
- ε: design parameter; determines ripple

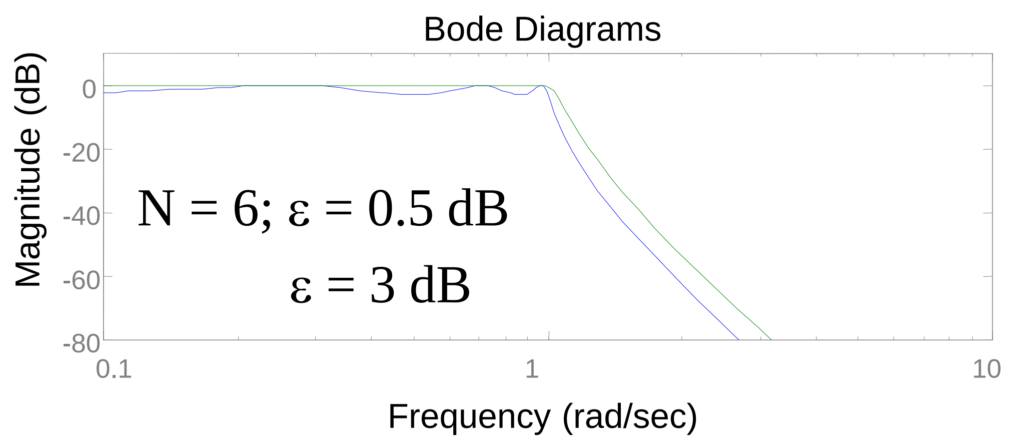
Chebychev Polynomials; Denominator of T(s):

N	ε=0.3493; (0.5 dB ripple)	ε=0.5089; (1 dB ripple)
1	(s+2.863)	(s+1.965)
2	(s^2+1.425s+1.516)	(s^2+1.098s+1.103)
3	(s+0.626)(s^2+0.626s+1.142)	(s+0.494)(s^2+0.494s+0.994)
4	(s^2+0.351s+1.064)(s^2+0.845s+0.356)	(s^2+0.297s+0.987)(s^2+0.674s+0.279)
5	(s+0.362)(s^2+0.224s+1.036)(s^2+0.586s+0.477)	(s+0.289)(s^2+0.179s+0.988)(s^2+0.468s+0.429)

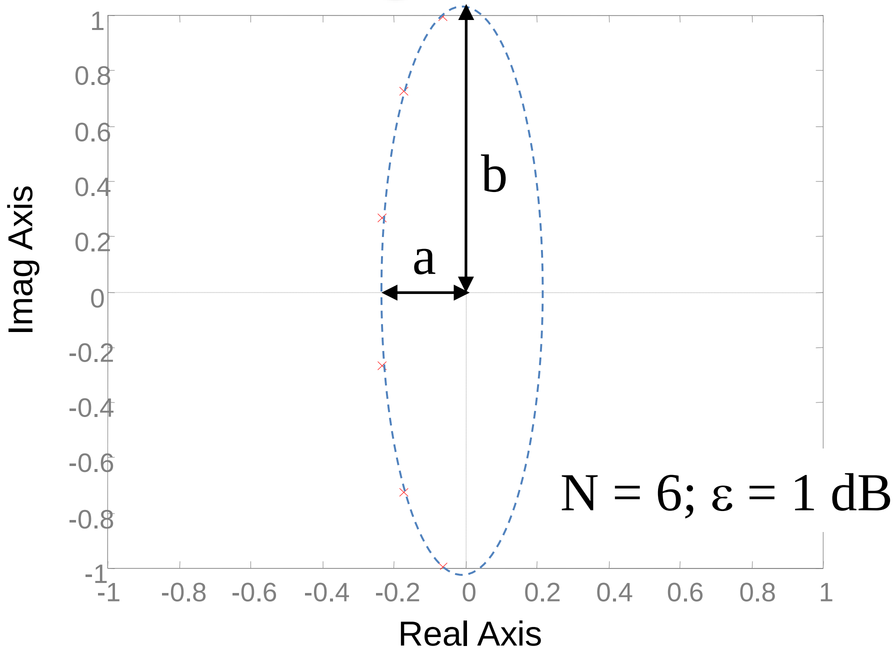
CC-LP Frequency Response

- **Properties:**
- Ripples in Bandpass between
- $\omega = 0$ and $\omega = 1/(1+e^2)^{0.5}$
- $|H(j1)| = 1/(1+\varepsilon^2)^{0.5}$ for all N
- $|H(0)| = 1$ for N odd
- $= 1/(1+\varepsilon^2)^{0.5}$ for N even
- steeper roll-off than Butterworth
- Implementation: see Butterworth example





Chebychev Pole Locations



- Minor axis:

$a = \sinh(1/N \sinh^{-1}(1/\varepsilon))$

- Major axis:

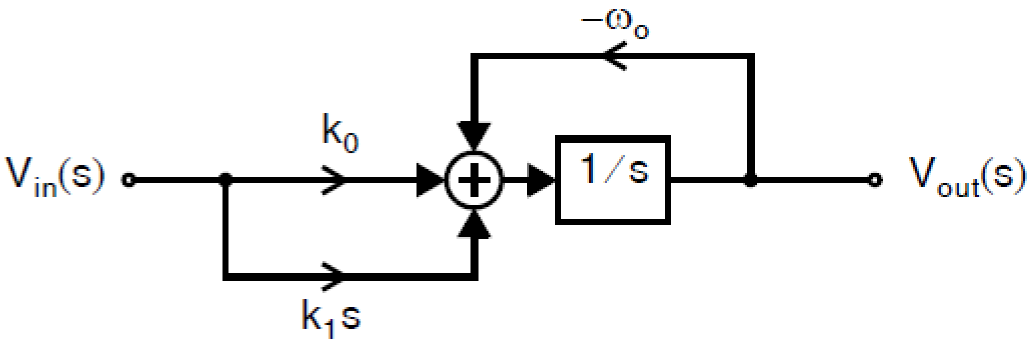
$b = \cosh(1/N \cosh^{-1}(1/\varepsilon))$

- Poles located on an ellipse around the origin; narrow ellipse means poles closer to imag. axis $\Rightarrow$ larger ripples. Wider ellipse $\Rightarrow$ small ripples; approaches Butterworth filter
- $\sigma_k = -\sinh(1/N \sinh^{-1}(1/ \varepsilon)) \sin((2k-1)\pi/2N)$
- $\omega_k = -\cosh(1/N \sinh^{-1}(1/ \varepsilon)) \cos((2k-1)\pi/2N)$

Motivation

- **Switched Capacitor Filters**
  - Pro: Accurate transfer-functions
  - Pro: High linearity, good noise performance
  - Con: Limited in speed
    - Clock rate must be greater than twice the signal frequency
  - Con: Requires anti-aliasing filters
- **Continuous-time filters**
  - Con: Moderate transfer-function accuracy (requires tuning circuitry)
  - Con: Moderate linearity
  - Pro: High-speed
  - Pro: Good noise performance
- Required building blocks:
  - Integrators, summers and gain stages
    - Allow to realise any rational function, hence any integrated continuous-time filter
    - Any rational transfer function with real-valued coefficients may be factored into first- and second-order terms

First Order Filter

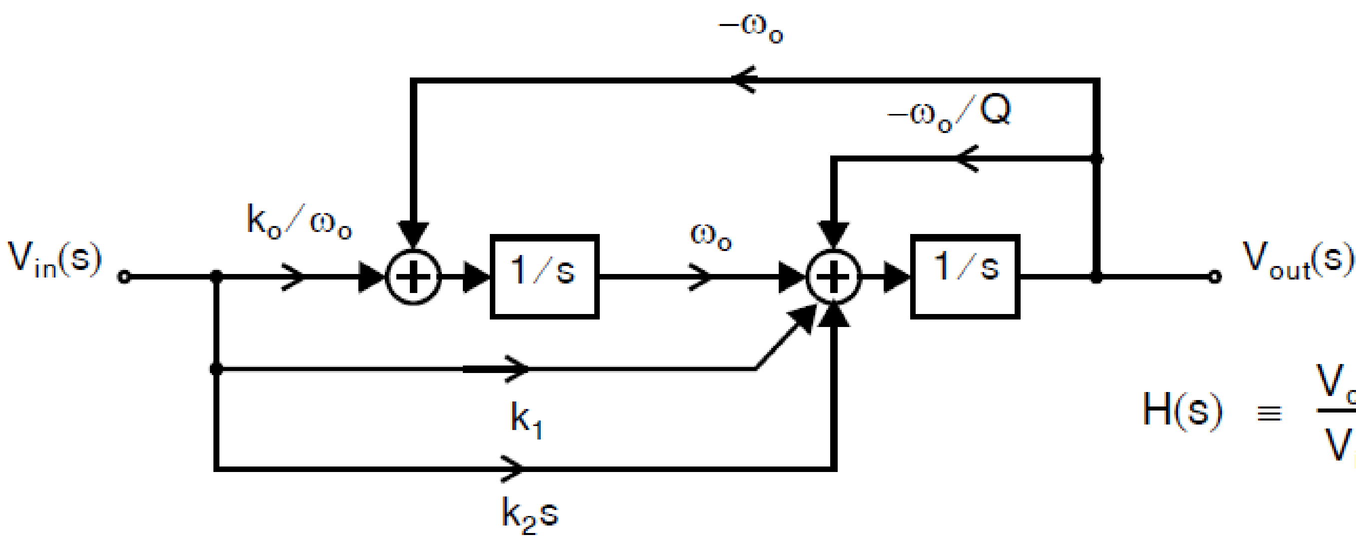


$$H(s) \equiv \frac{V_{out}(s)}{V_{in}(s)} = \frac{k_1 s + k_0}{s + \omega_0}$$

block diagram of a first-order continuous-time filter

- first-order continuous-time filter requires one integrator, one summer, and up to three gain elements
- In general: *One integrator is required for each pole in an analog filter*

Second Order Filter

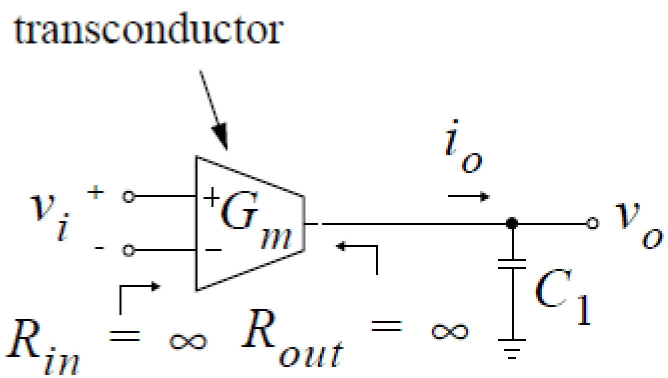


$$H(s) \equiv \frac{V_{out}(s)}{V_{in}(s)} = \frac{k_2 s^2 + k_1 s + k_o}{s^2 + \left(\frac{\omega_o}{Q}\right) s + \omega_o^2}$$

block diagram of a second-order continuous-time filter

- Two integrators are required to realise the two poles
- For stability: ω_0/Q must be positive
  - One integrator must have feedback around it, hence the integrator is "lossy"
  - A large feedback coefficient ω_0/Q results in a very lossy integrator, hence the Q is low
    - $Q<1/2$: both poles are real; $Q>1/2$: poles are complex-conjugate pairs

G_m-C Integrators



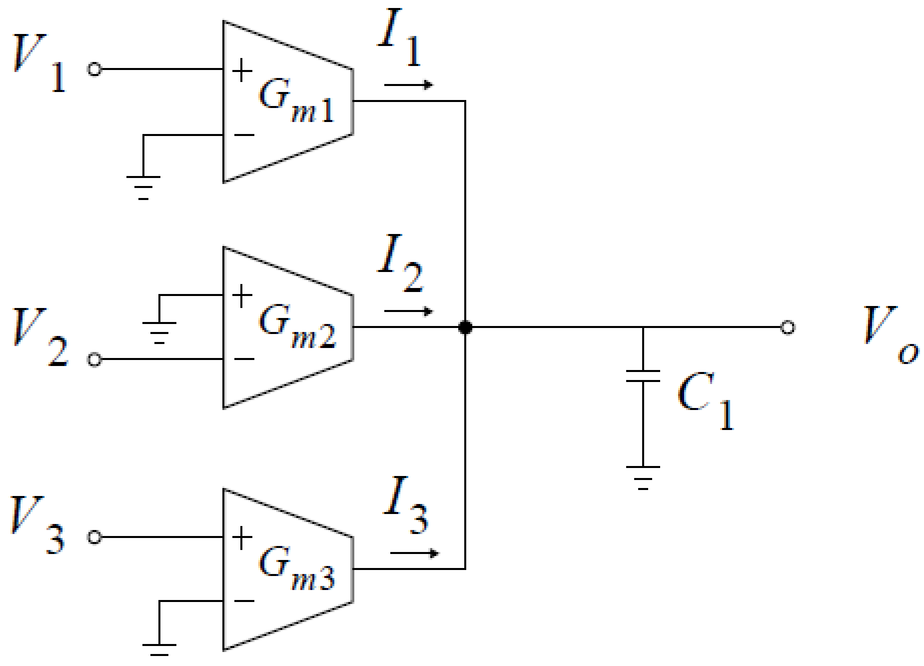
$$i_o = G_m v_i$$

$$V_o = \frac{I_o}{sC_1} = \frac{G_m V_i}{sC_1} \equiv \left(\frac{\omega_{ti}}{s}\right) V_i$$

$$\omega_{ti} = \frac{G_m}{C_1}$$

- Use a **transconductor** (or OPA) to build an integrator: $i_o = G_m v_i$
- Output current is linearly related to input voltage
- Output impedance is ideally infinite
- OTA (**operational transconductance amplifier**) has a high G_m value but is not usually linear

Multiple Input G_m-C Integrators



$$V_o = \frac{1}{sC_1}(G_{m1}V_1 - G_{m2}V_2 + G_{m3}V_3)$$

Example

- What Gm is needed for an integrator having a unity gain frequency of ω_{ti}= 20 MHz when C=2 pF?

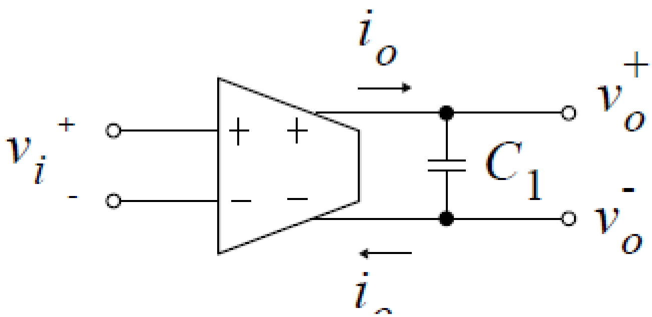
$$\begin{aligned} G_m &= 2\pi \times 20MHz \times 2pF \\ &= 0.251\ mA/V \end{aligned}$$

- Or equivalently: G_m=1/3.98kΩ

- This is related to the unity gain frequency by:

$$2\pi \times 20MHz = \frac{1}{3.98k\Omega \times 2pF}$$

Fully Differential Integrators



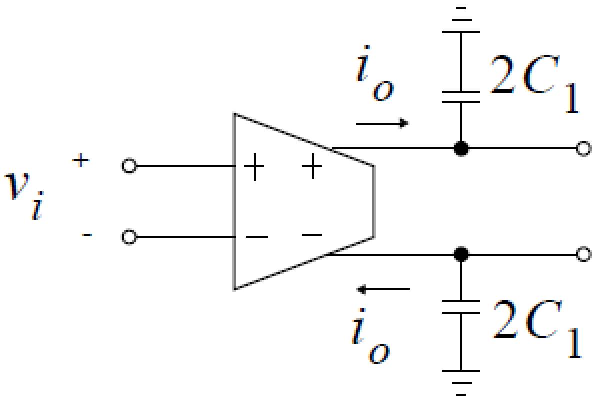
$$V_o = \frac{I_o}{sC_1} = \frac{G_m V_i}{sC_1} \qquad \omega_{ti} = \frac{G_m}{C_1}$$

$$v_o = v_o^+ - v_o^-$$

$$i_o = G_m v_i$$

- Use a single capacitor between differential outputs
- Requires some sort of common-mode feedback to set output common-mode voltage
- Needs some extra caps for compensating common mode feedback loop

Fully Differential Integrators



$$V_o^+ = I_o / s(2C_1)$$

$$V_o^- = -I_o / s(2C_1)$$

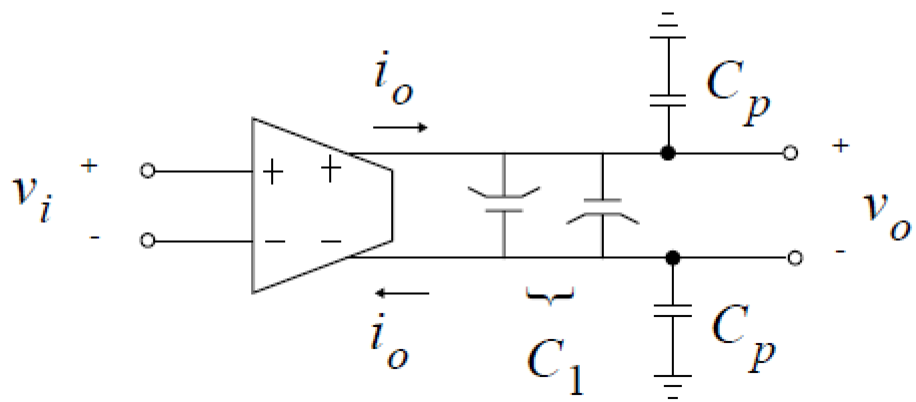
$$v_o = v_o^+ - v_o^-$$
$$i_o = G_m v_i$$

$$\omega_{ti} = \frac{G_m}{C_1}$$

$$V_o = \frac{2I_o}{s(2C_1)} = \frac{G_m V_i}{sC_1}$$

- Use two grounded capacitors
- Still requires common-mode feedback but compensation caps for common-mode feedback can be the same grounded capacitors

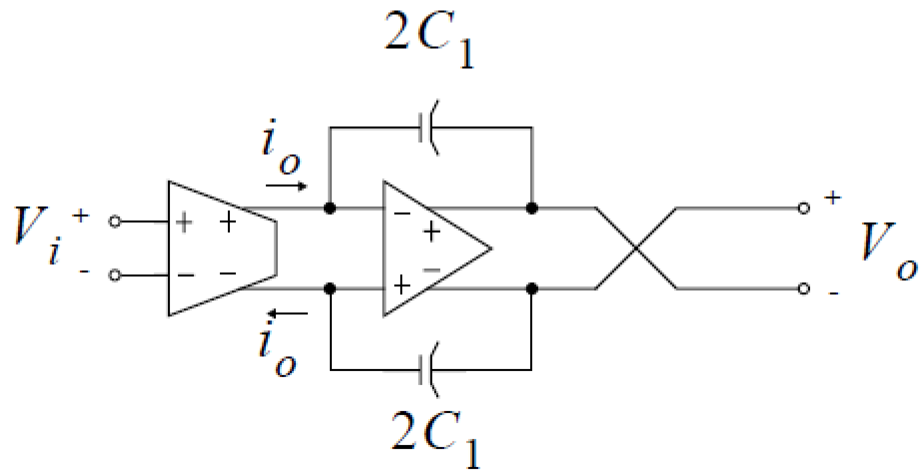
Fully Differential Integrators



$$V_o = \frac{G_m V_i}{s(C_1 + C_p/2)} \qquad \omega_{ti} = \frac{G_m}{(C_1 + C_p/2)}$$

- Integrated capacitors have top and bottom plate parasitic capacitances
- To maintain symmetry, usually 2 parallel caps used as shown above
- Note that parasitic capacitance affects time-constant and cause non-linearity

G_m-C Opamp Integrator



$$V_o = \frac{2I_o}{s(2C_1)} = \frac{G_m V_i}{sC_1} \qquad \omega_{ti} = \frac{G_m}{C_1}$$

- Use an extra Opamp to improve linearity and noise performance
- Also known as a "Miller Integrator"
- The gain of extra Opamp reduces the effect of parasitic capacitances
- Cross coupling of output wires to maintain positive integration coefficient

G_m-C Opamp Integrator

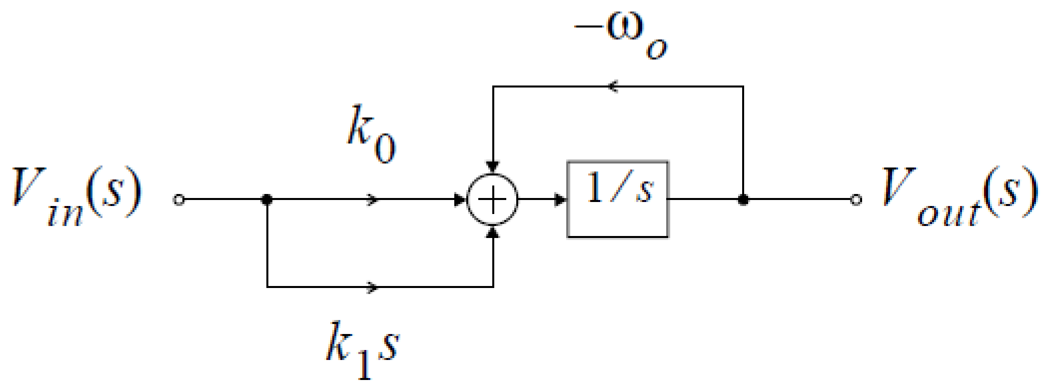
Advantages

- Effect of parasitic caps reduced by opamp gain — more accurate time-constant and better linearity
- Less sensitive to noise since transconductor output is low impedance (due to opamp feedback)
- cell drives virtual Gnd — output-impedance of G_m cell can be lower and smaller voltage swing needed

Disadvantages

- Lower operating speed because it now relies on feedback
- Larger power dissipation
- Larger silicon area

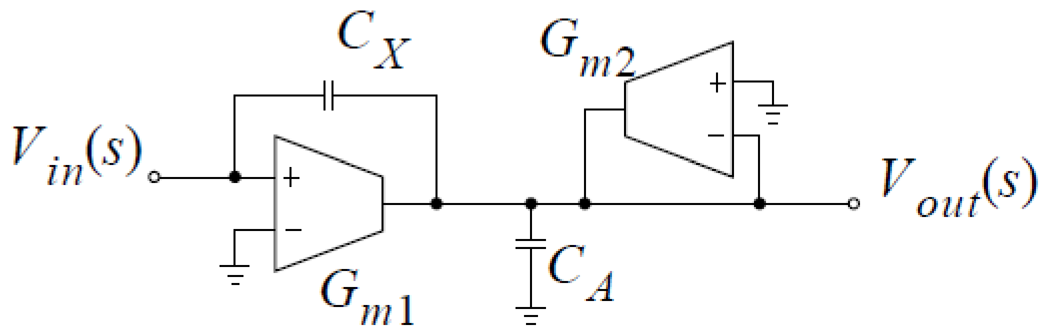
First Order Filter



General first-order transfer-function: $\dfrac{V_{out}(s)}{V_{in}(s)} = \dfrac{k_1 s + k_0}{s + \omega_0}$

- Built with a single integrator and two feed-ins branches
- ω_0 sets the pole frequency

First Order Filter



- Can show that the transfer function is given by (using a current equation at the output node):

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{sC_X + G_{m1}}{s(C_A + C_X) + G_{m2}} = \frac{s\left(\frac{C_X}{C_A + C_X}\right) + \left(\frac{G_{m1}}{C_A + C_X}\right)}{s + \left(\frac{G_{m2}}{C_A + C_X}\right)}$$

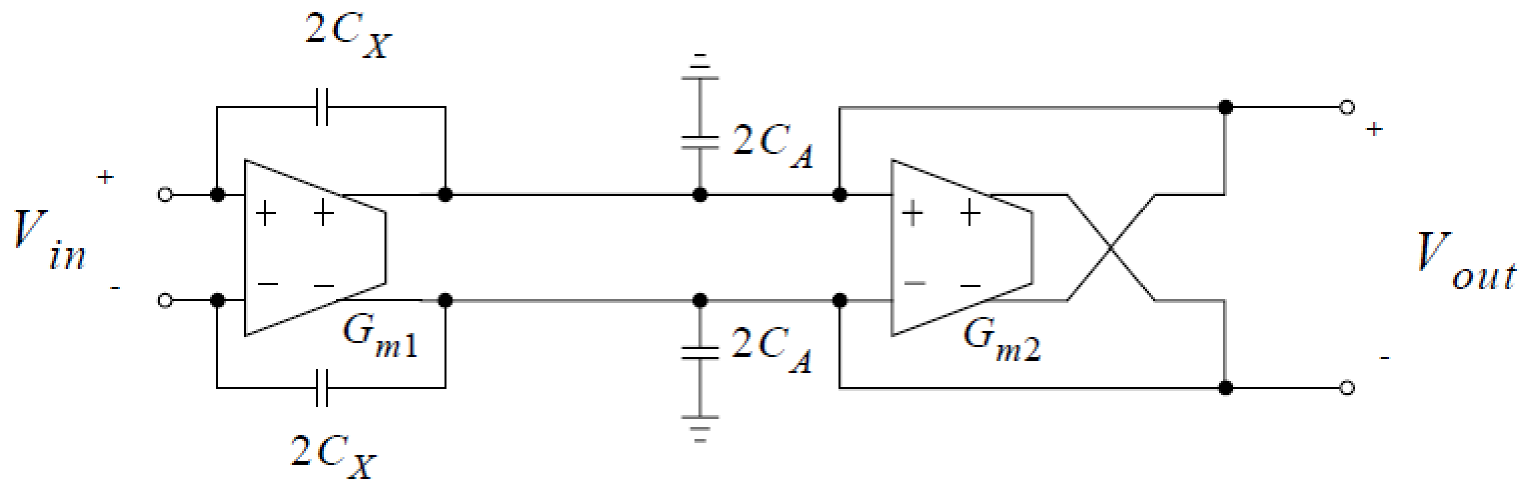
- Equating with the block diagram transfer function:

$$C_X = \left(\frac{k_1}{1 - k_1}\right) C_A \text{where } \ (0 \le k_1 < 1)$$

$$G_{m1} = k_0(C_A + C_X)$$

$$G_{m2} = \omega_o(C_A + C_X)$$

Fully-Differential First-Order Filter



- Same equations as single-ended case but cap sizes doubled

$$C_X = \left(\frac{k_1}{1-k_1}\right) C_A \text{where } \ (0 \le k_1 < 1)$$

$$G_{m1} = k_0(C_A + C_X)$$

$$G_{m2} = \omega_o(C_A + C_X)$$

- Can realize $k_1<0$ by cross-coupling wires at C_x

Example

- Find fully-diff values when dc gain = 0.5, a pole at 20 MHz and a zero at 40MHz. Assume C_A=2pF

$$H(s) = \frac{0.25(s + 2\pi \times 40MHz)}{(s + 2\pi \times 20MHz)} = \frac{0.25s + 2\pi \times 10MHz}{s + 2\pi \times 20MHz}$$

K_1=0.25, $k_0=2\pi \times 10^7$, $\omega_0=4\pi \times 10^7$

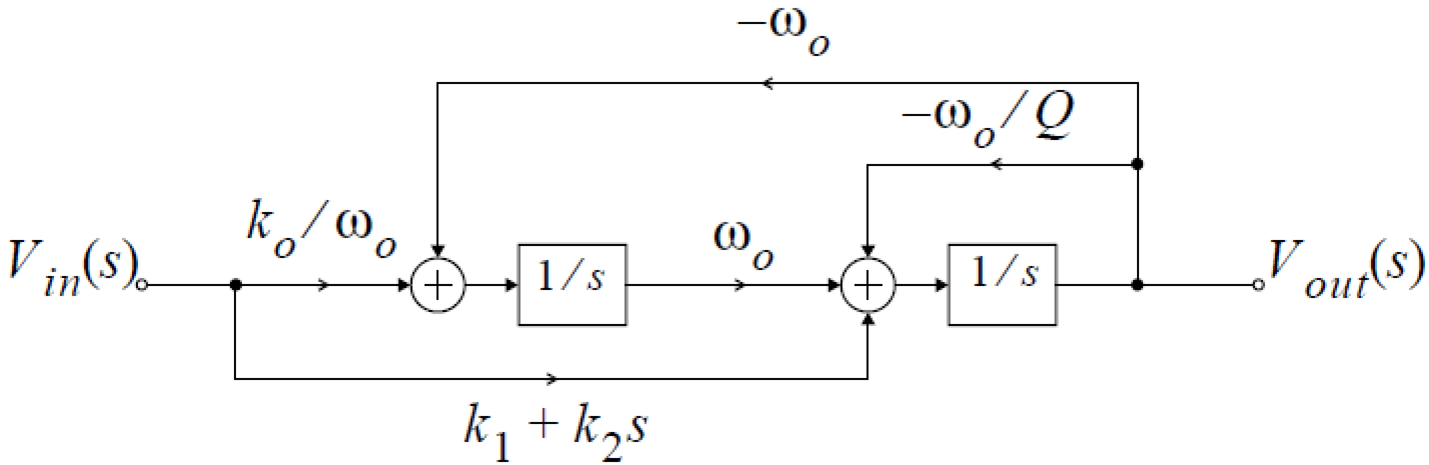
- So:

$$C_X = 2pF \times \frac{0.25}{1 - 0.25} = 0.667pF$$

$$G_{m1} = 2\pi \times 10^7 \times 2.667pF = 0.168\ mA/V$$

$$G_{m2} = 4\pi \times 10^7 \times 2.667pF = 0.335\ mA/V$$

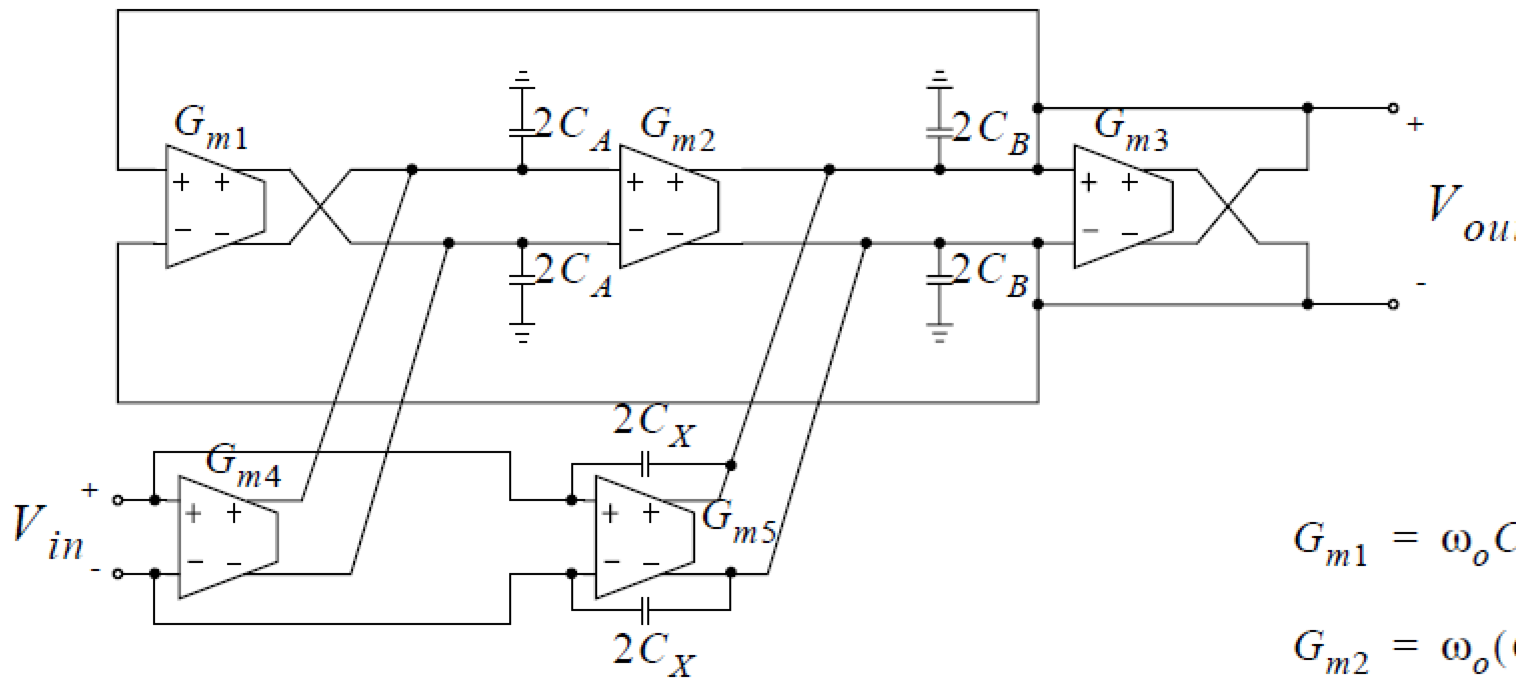
Second Order Filter



$$H(s) \equiv \frac{V_{out}(s)}{V_{in}(s)} = \frac{k_2 s^2 + k_1 s + k_o}{s^2 + \left(\frac{\omega_o}{Q}\right)s + \omega_o^2}$$

- Block diagram: see lecture on switched capacitor circuits
  - Modified to have positive integrators

Differential Second Order Filter (Biquad)



$$C_X = C_B\left(\frac{k_2}{1-k_2}\right) \text{where } (0 \le k_2 < 1)$$

$$G_{m1} = \omega_o C_A$$

$$G_{m2} = \omega_o(C_B + C_X)$$

$$G_{m3} = \frac{\omega_o(C_B + C_X)}{Q}$$

$$G_{m4} = (k_o C_A)/\omega_o$$

$$G_{m5} = k_1(C_B + C_X)$$

Differential Second Order Filter (Biquad)

- Transfer function:

$$H(s) \equiv \frac{V_{out}(s)}{V_{in}(s)} = \frac{s^2\left(\frac{C_X}{C_X + C_B}\right) + s\left(\frac{G_{m5}}{C_X + C_B}\right) + \left(\frac{G_{m2}G_{m4}}{C_A(C_X + C_B)}\right)}{s^2 + s\left(\frac{G_{m3}}{C_X + C_b}\right) + \left(\frac{G_{m1}G_{m2}}{C_A(C_X + C_B)}\right)}$$

- Note that there is a restriction on the high-frequency gain coefficient k_2 as in the first-order case
- Note that G_{m3} sets the damping of this biquad
- G_{m1} and G_{m2} form two integrators with unity-gain frequencies of ω_0/s

Example

- Find values for a bandpass filter with a centre frequency of 20 MHz, a Q value of 5, and a centre frequency gain of 1
- Assume $C_A = C_B = 2$ pF

$$H(s) \equiv \frac{V_{out}(s)}{V_{in}(s)} = \frac{Gs\frac{\omega_o}{Q}}{s^2 + s\frac{\omega_o}{Q} + \omega_o^2}$$

- where G =1 Is the gain at the center frequency

Example

- Since $\omega_0 = 2\pi \times 20\text{MHz}$ and $Q = 5$, we find:

$$k_1 = G\frac{\omega_o}{Q} = 2.513\times10^7 \text{rad/s}$$

- Since k_0 and k_2 are zero, we have $C_x = C_{mA} = 0$
- The transconductance values are:

$$G_{m1} = \omega_o C_A = 0.2513\ \text{mA/V}$$

$$G_{m2} = \omega_o(C_B + C_X) = 0.2513\ mA/V$$

$$G_{m3} = G_{m5} = k_1 C_B = 50.27\mu\text{A/V}$$

CMOS Tranconductors

- A large variety of methods
- Best approach depends on application
- Two main classifications: triode or active transistor based

Triode vs. Active

- Triode based tends to have better linearity
- Active tend to have faster speed for the same operating current

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Triode Tranconductors

- Recall n-channel triode equation

$$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left((V_{GS} - V_{tn})V_{DS} - \frac{V_{DS}^2}{2}\right)$$

- Conditions to remain in triode

$$V_{DS} < V_{eff} \qquad \text{where} \quad V_{eff} = V_{GS} - V_{tn}$$

- or equivalently: $V_{GS} > V_{DS} + V_{tn}$

- Above models are only reasonably accurate
  - Higher order terms are not modelled
- Not nearly as accurate as exponential model in BJTs
- Use fully-differential architectures to reduce even order distortion terms — also improves common mode noise rejection
  - The third order term dominates

Fixed Bias Triode Tranconductors

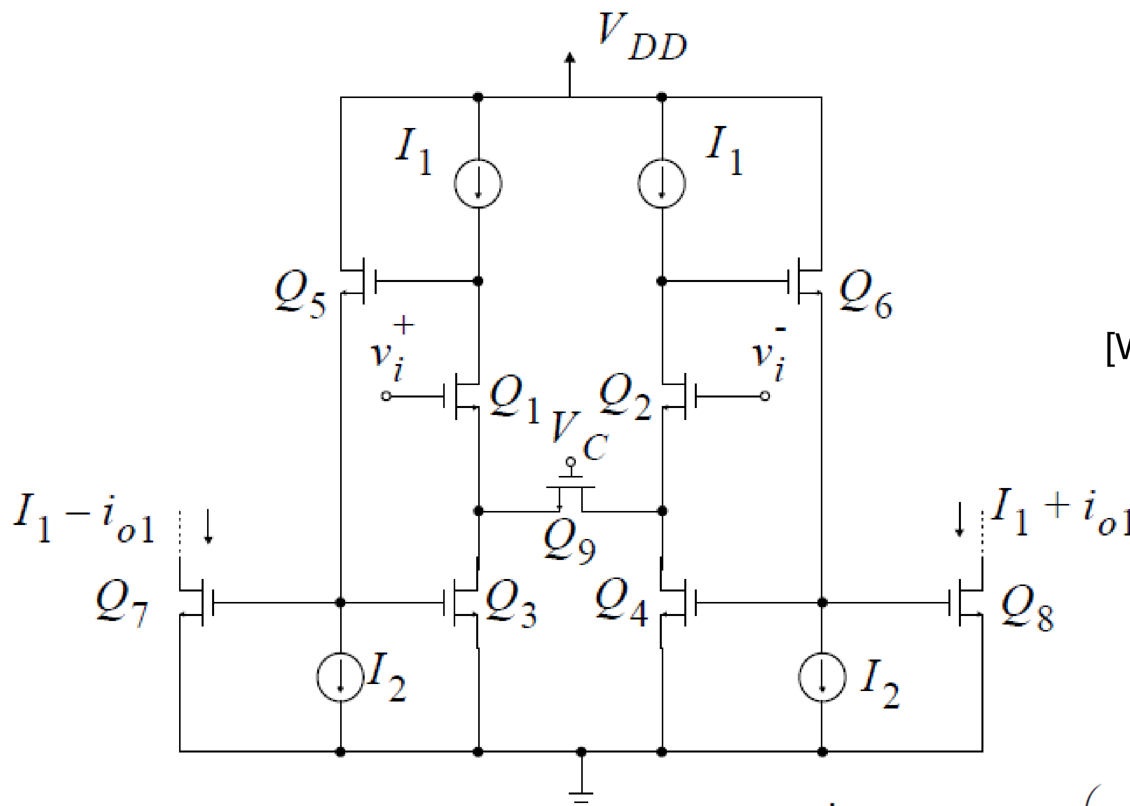
- Use a small v_{DS} voltage so v^2_{DS} term goes to zero
  - Drain current is approximately linear with applied v_{DS}. Transistor in triode becomes a linear resistor

$$r_{DS} \equiv \left(\frac{\partial i_D}{\partial v_{DS}}\right)^{-1}\Bigg|_{v_{DS}=0}$$

- Resulting in: $r_{DS} = \left(\mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS} - V_{tn})\right)^{-1}$

- Can use a triode transistor where a resistor would normally be used — resistance value is tunable

Fixed Bias Triode Tranconductors

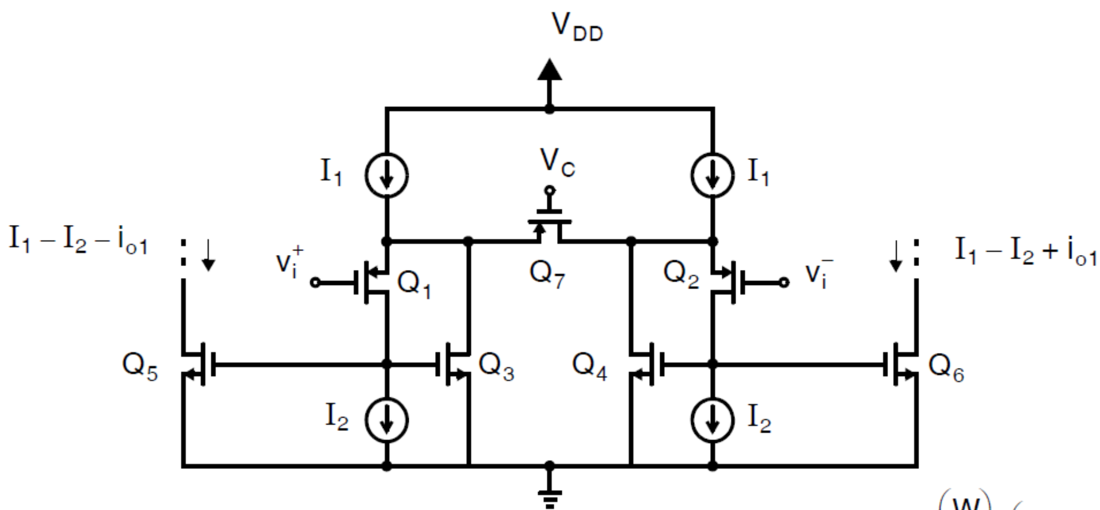


[Welland, 1994]

$$G_m = \frac{i_{o1}}{v_i^+ - v_i^-} = \mu_n C_{ox} \left(\frac{W}{L}\right)_9 (V_{gs9} - V_{tn})$$

- Q9 is in the triode region
  - transconductor has a variable transconductance value that can be adjusted by changing the value of V_{gs9}
- Moderate linearity

Fixed Bias Triode Tranconductors

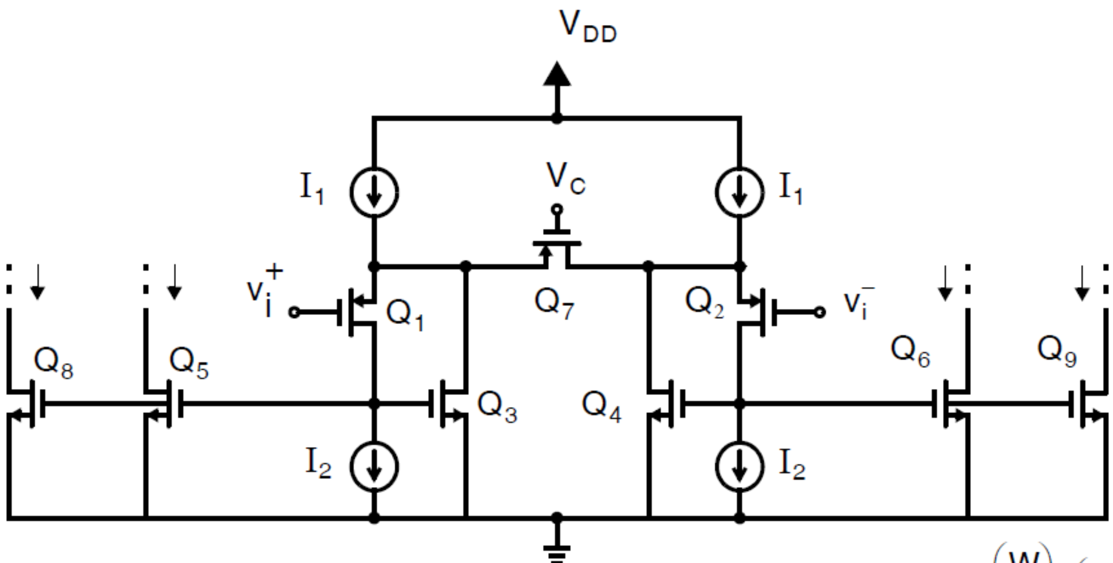


$$G_m = \mu_p C_{ox}\left(\frac{W}{L}\right)_7 (V_{sg7} - |V_{tp}|)$$

[Kwan, 1991]

- Alternative approach with lower complexity and p-channel inputs
  - transconductor has a variable transconductance value that can be adjusted by changing the value of V_{gs9}

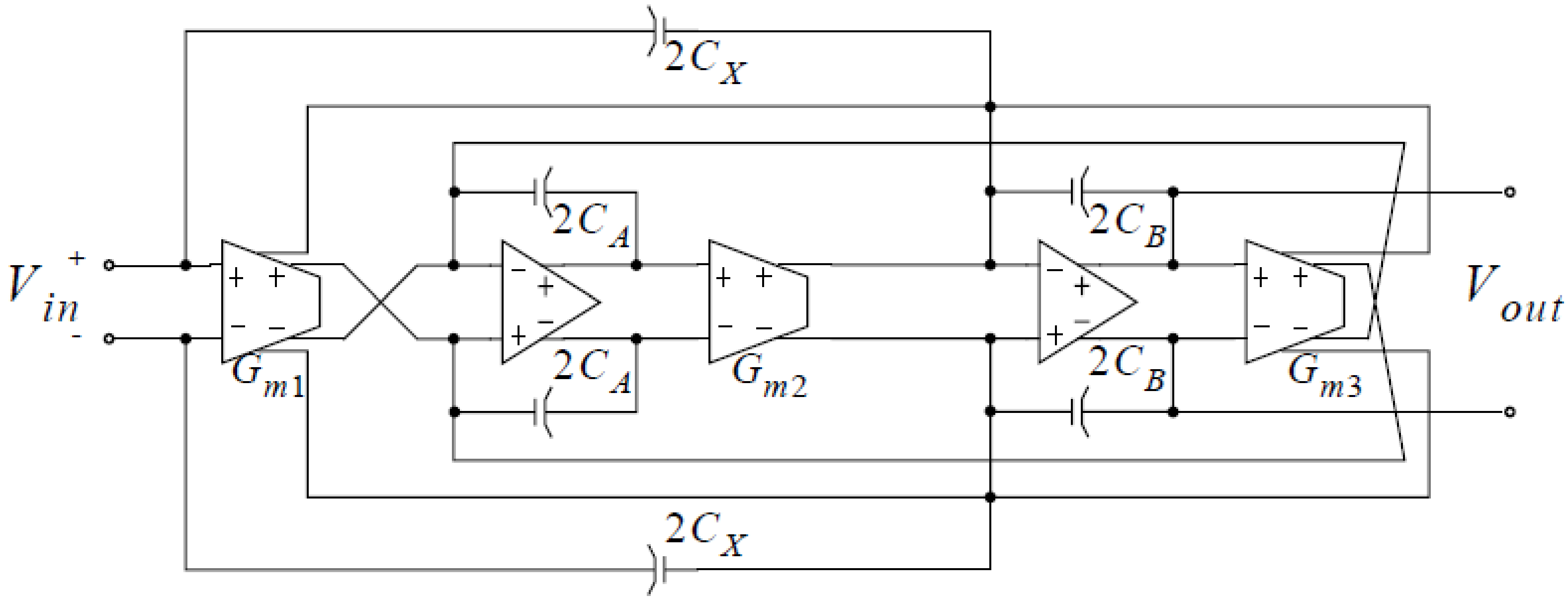
Fixed Bias Triode Tranconductors



$$G_m = \mu_p C_{ox} \left(\frac{W}{L}\right)_7 \left(V_{sg7} - |V_{tp}|\right)$$

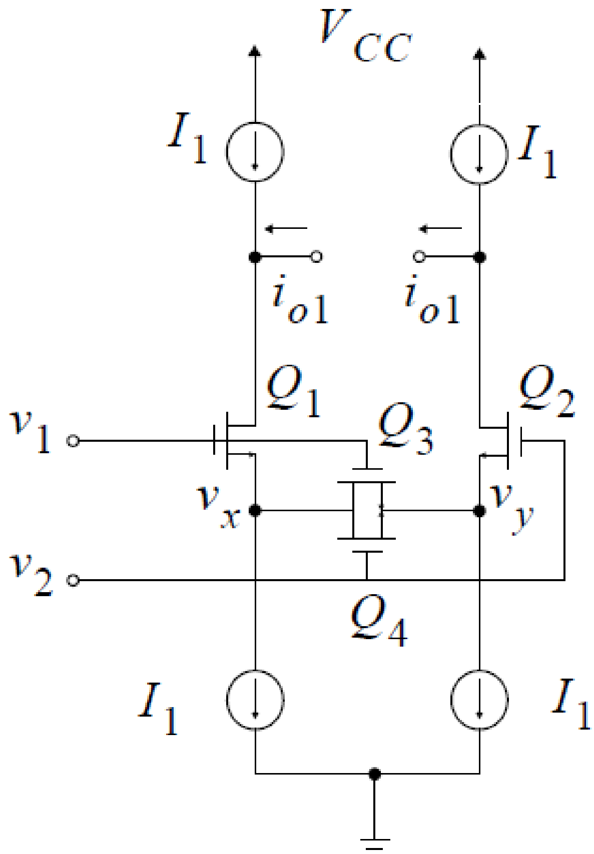
- Circuit can be easily made with multiple scaled output currents
- Multiple outputs allow filters to be realized using fewer transconductors

Biquads Using Multiple Outputs



- Can make use of multiple outputs to build a biquad filter
  - scale extra outputs to desired ratio
- Reduces the number of transconductors
  - saves power and die area
- Above circuit makes use of Miller integrators

Varying-Bias Triode Transconductor



$$G_m = \frac{4k_1k_3\sqrt{I_1}}{(k_1+4k_3)\sqrt{k_1}}$$

[Krummenacher, 1988]

- Linearizes MOSFET differential stage
  - Transistors primarily in triode region

Varying-Bias Triode Transconductor

- gates of Q_3 and Q_4 connected to the differential input (and not to bias voltage)
- Q_3 and Q_4 undergo varying bias conditions to improve linearity
- It can be shown that

$$G_m = \frac{4k_1k_3\sqrt{I_1}}{(k_1 + 4k_3)\sqrt{k_1}}$$

- With

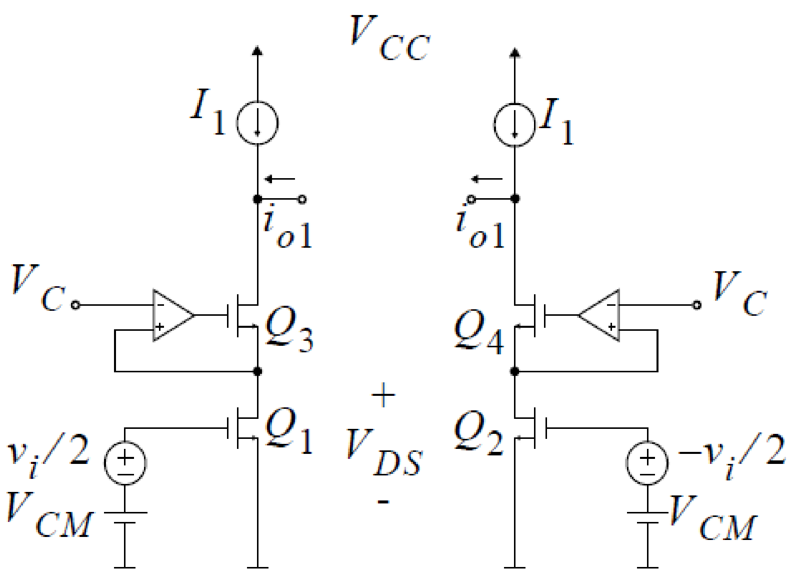
$$k_i = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)_i$$

- Note, G_m is proportional to square-root of as opposed to linear relation for a BJT transconductor
- Transconductance can be tuned by changing bias current I_i

Drain-Source Fixed-Bias Transconductor

$$i_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left((V_{GS} - V_{tn})V_{DS} - \frac{V_{DS}^2}{2}\right)$$

- If v_{DS} is kept constant, then i_D varies linearly with v_{GS}
  - Model is too simple, neglecting second order effects such as velocity saturation, mobility degradation
- Possible implementation using fully differential architecture



$$G_m = \mu_n C_{ox}\left(\frac{W}{L}\right)_1 V_{DS}$$

Drain-Source Fixed-Bias Transconductor

- Can realize around 50 dB linearity (not much better since model is not that accurate)
- Requires a fully-differential structure to cancel even-order terms
- V_C sets v_{DS} voltage
- Requires a non-zero common-mode voltage on input
- Note that the transconductance is proportional to v_{DS}
  - For v_{DS} small the bias current I_1 is also approximately proportional to v_{DS}

Alternative: MOSTFET-C Filters

- Gm-C filters are most commonly used but MOSFET-C have advantages in BiCMOS for low power applications
- MOSFET-C filters similar to active-RC filters but resistors replaced with MOS transistors in triode
- Generally slower than Gm-C filters since opamps capable of driving resistive loads required
- Rely on Miller integrators
- Two main types — 2 transistors or 4 transistors

Alternative: MOSTFET-C Filters

- Gm-C filters are most commonly used but MOSFET-C have advantages in BiCMOS for low power applications
- MOSFET-C filters similar to active-RC filters but resistors replaced with MOS transistors in triode
  - Knowledge and architecture of active RC filters can be transferred
- Generally slower than Gm-C filters since Opamps capable of driving resistive loads required
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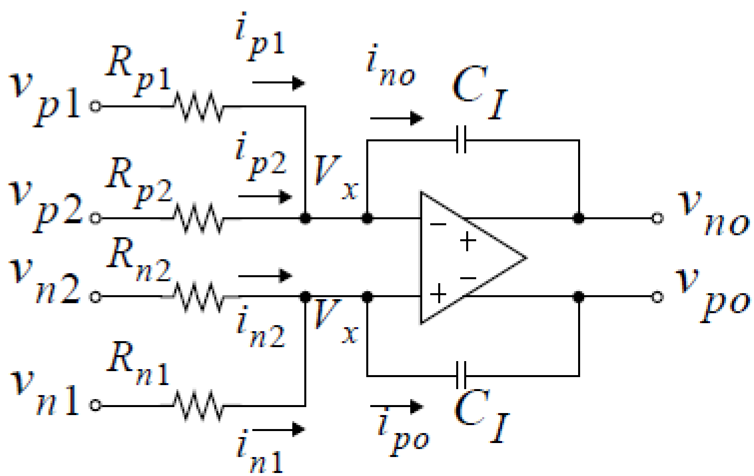
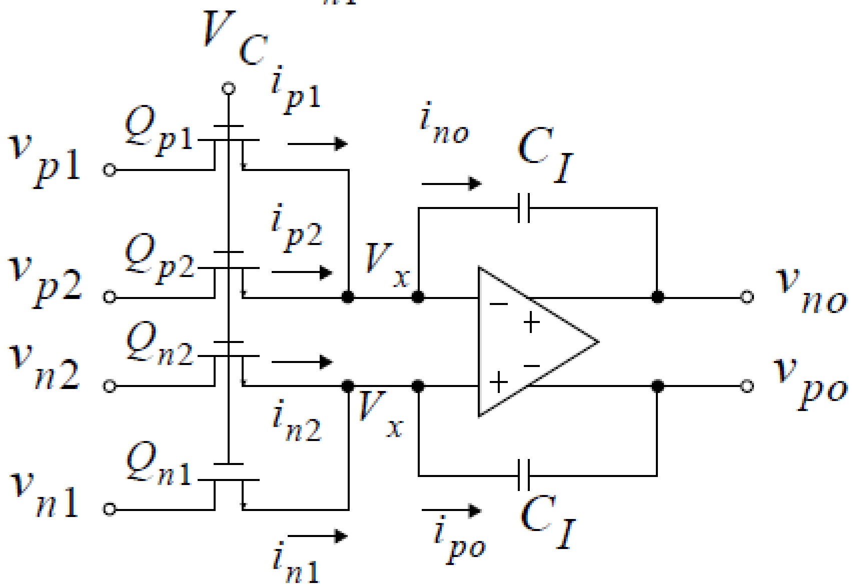
Two Transistor Integrators





Banu 1983

Two Transistor Integrators

- For resistor integrator can be shown

$$v_{\text{diff}} = \frac{1}{sR_1C_I}(v_{p1} - v_{n1}) + \frac{1}{sR_2C_I}(v_{p2} - v_{n2})$$

  – If negative integration is required — cross-couple wires

- For MOSFET-C integrator, assuming transistors are biased in triode region, the small-signal resistance is given by:

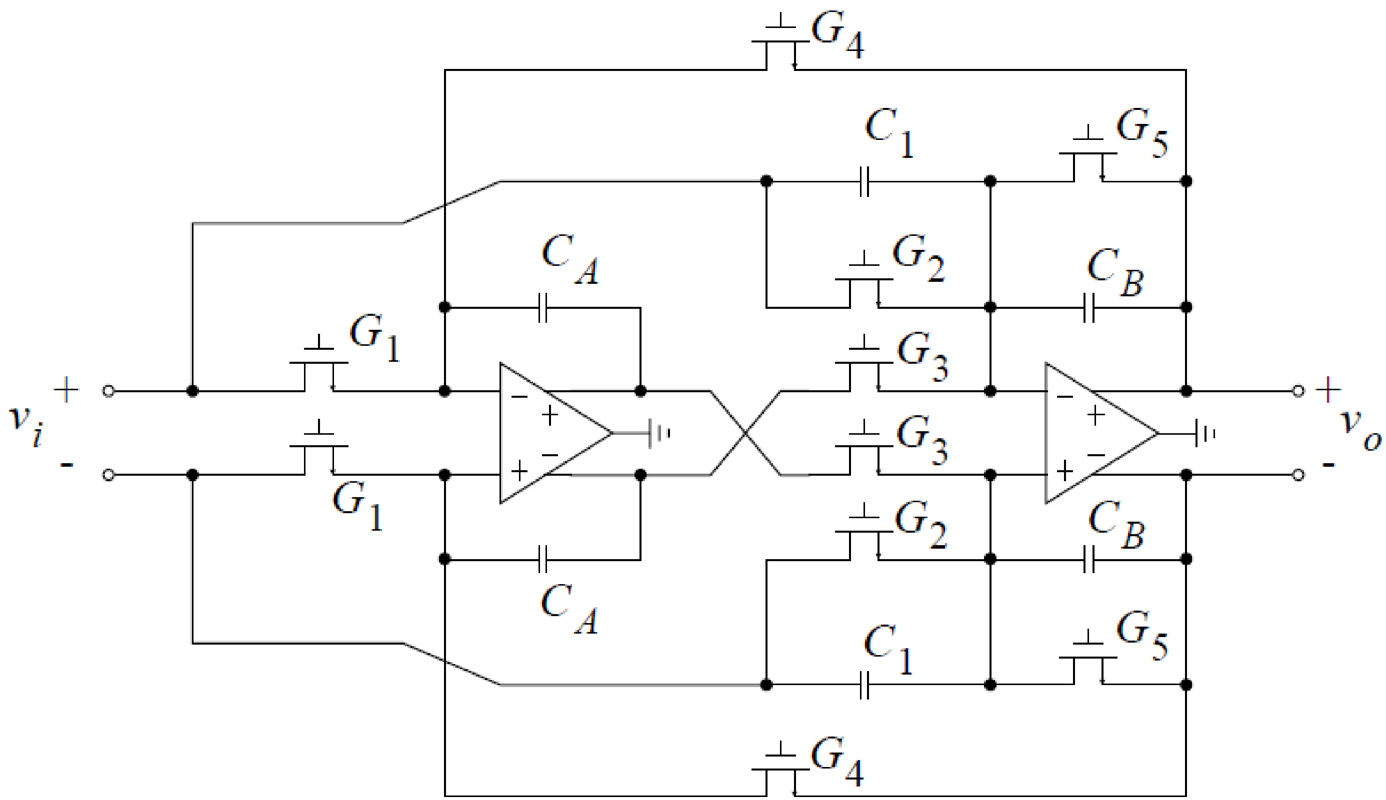
$$r_{DS} = \left(\mu_n C_{ox}\left(\frac{W}{L}\right)(v_{GS} - V_{tn})\right)^{-1}$$

- Therefore, the differential output of the MOSFET-C integrator is:

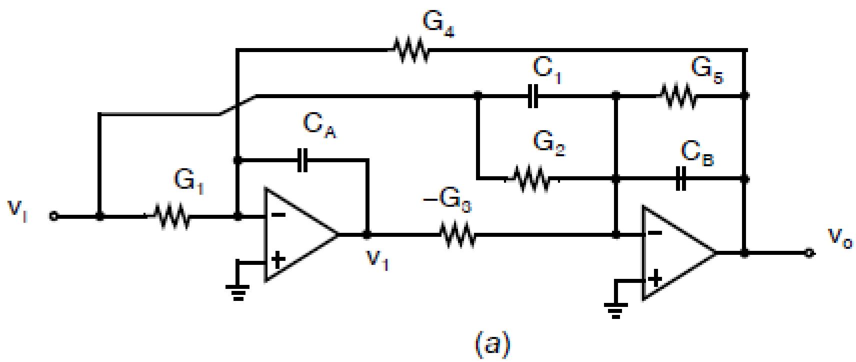
$$v_{\text{diff}} = \frac{1}{sr_{DS1}C_I}(v_{p1} - v_{n1}) + \frac{1}{sr_{DS2}C_I}(v_{p2} - v_{n2})$$

With: $$r_{DSi} = \left(\mu_n C_{ox}\left(\frac{W}{L}\right)_i(V_C - V_x - V_{tn})\right)^{-1}$$

General MOSFET-C Biquad Filter



Equivalent active RC half circuit

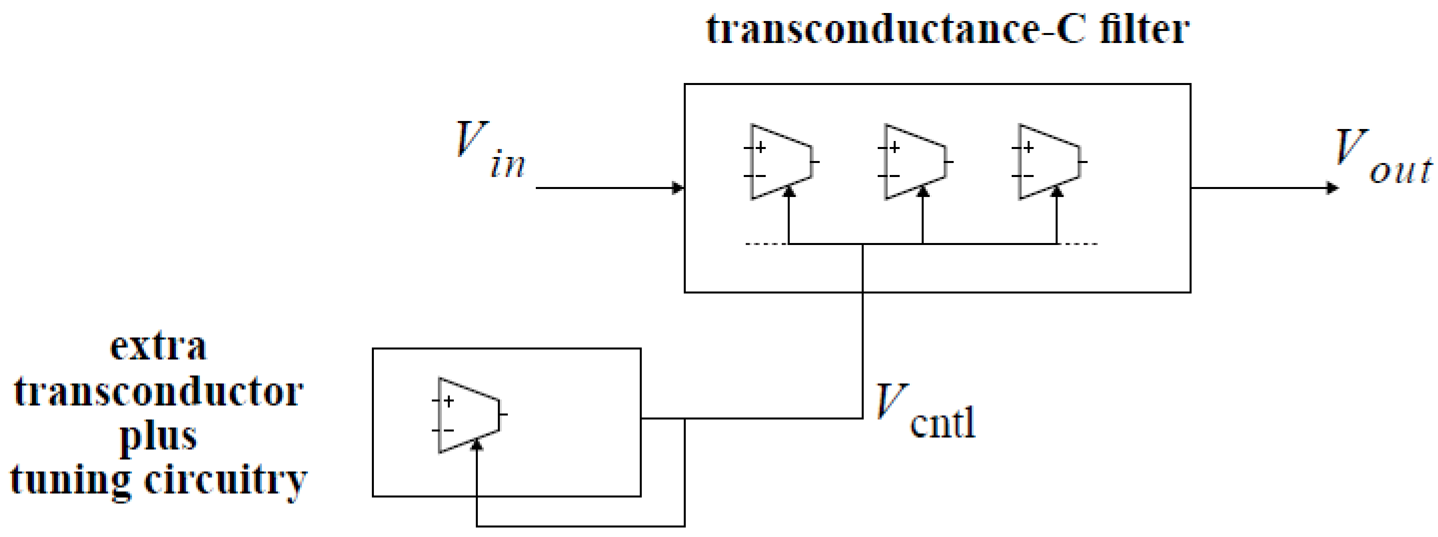


$$\frac{v_{0(s)}}{v_{i(s)}} = \frac{\frac{C_1}{C_B}s^2 + \frac{G_2}{C_B}s + \frac{G_1 G_3}{C_A\,C_B}}{s^2 + \frac{G_5}{C_B}s + \frac{G_3 G_4}{C_A C_B}}$$

Tuning Circuitry

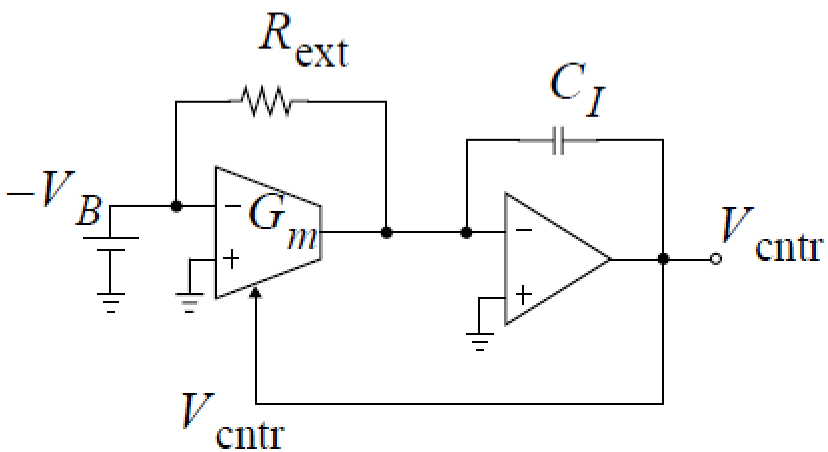
- Tuning can often be the MOST difficult part of a continuous-time integrated filter design
- Tuning required for continuous-time integrated filters to account for capacitance and transconductance variations — 30 percent time-constant variations
- Must account for process, temperature, aging, etc.
- While absolute tolerances are high, ratio of two like components can be matched to under 1 percent
- Note that SC filters do not need tuning as their transfer-function accuracy set by ratio of capacitors and a clock-frequency

Indirect Tuning

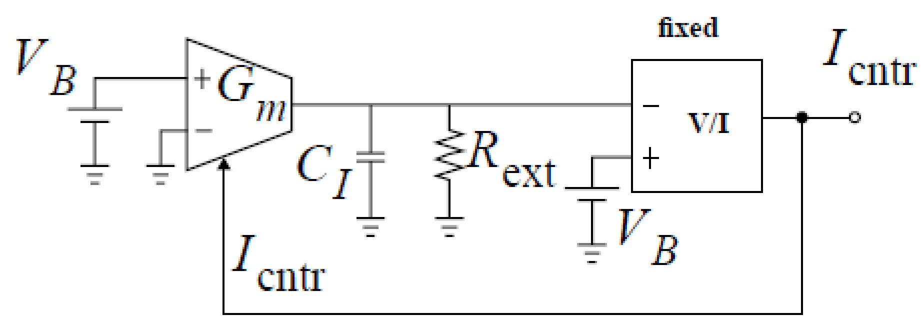


- Most common method — build an extra transconductor and tune it
- Same control signal is sent to filter’s transconductors which are scaled versions of tuned extra
- Indirect since actual filter’s output is not measured

Constant Transconductance



$$G_m = \frac{1}{R_{ext}}$$



- Can tune Gm to off-chip resistance and rely on capacitor absolute tolerance to be around 10 percent