EE247
Lecture 8

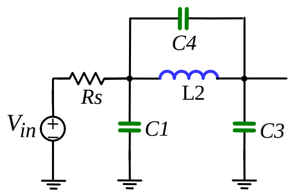
- Continuous-time filter design considerations
  - Monolithic highpass filters
  - Active bandpass filter design
    - Lowpass to bandpass transformation
    - Example: 6^{th} order bandpass filter
    - Gm-C bandpass filter using simple diff. pair
  - Various Gm-C filter implementations
- Performance comparison of various continuous-time filter topologies
- Introduction to switched-capacitor filters

Summary
Lecture 7

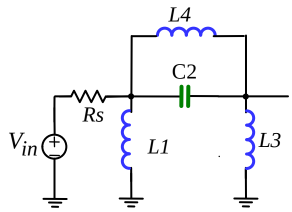
- Automatic on-chip filter tuning (continued from last lecture)
  - Continuous tuning (continued)
    - Replica single integrator in a feedback loop locked to a reference frequency
    - DC tuning of resistive timing element
  - Periodic digitally assisted filter tuning
    - Systems where filter is followed by ADC & DSP, existing hardware can be used to periodically update filter freq. response

RLC Highpass Filters

- Any RLC lowpass and values derived from tables can be converted to highpass by:
  - Replacing all Cs by Ls and $L_{Norm}^{HP} = 1/\ C_{Norm}^{LP}$
  - Replacing all Ls by Cs and $C_{Norm}^{HP} = 1/\ L_{Norm}^{LP}$
  - $L^{HP}=L_r/\ C_{Norm}^{LP}$, $C^{HP}=C_r/\ L_{Norm}^{LP}$ *where* $L_r=R_r/\omega_r$ and $C_r=1/(R_r\omega_r)$



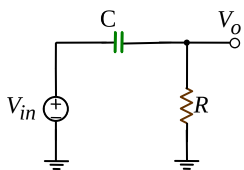
Lowpass



Highpass

Integrator Based High-Pass Filters
1st Order

- Conversion of simple high-pass RC filter to integrator-based type by using signal flowgraph technique



$$\frac{V_o}{V_{in}} = \frac{s\,RC}{1+s\,RC}$$

1st Order Integrator Based High-Pass Filter Signal Flowgraph

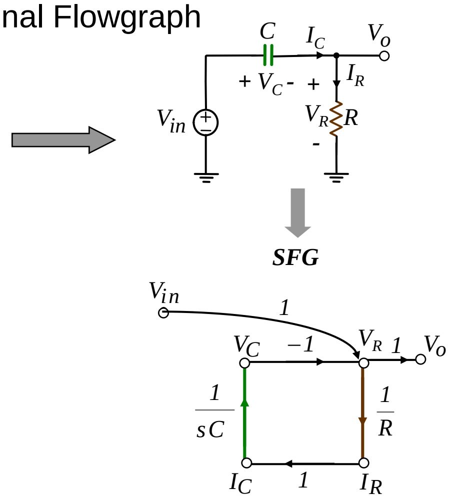
$V_R = V_{in} - V_C$

$V_C = I_C \times \frac{1}{sC}$

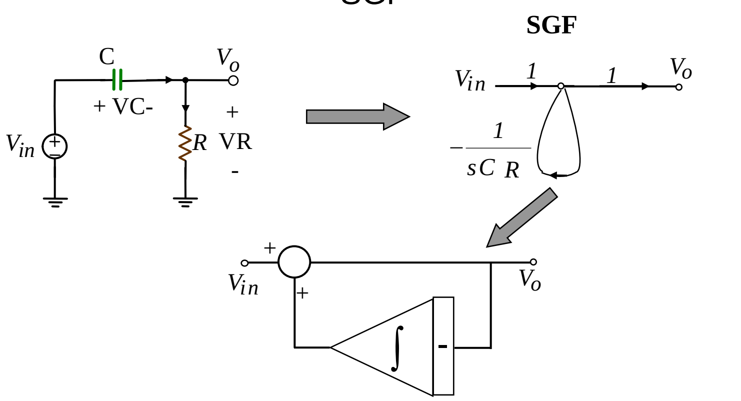
$V_o = V_R$

$I_R = V_R \times \frac{1}{R}$

$I_C = I_R$



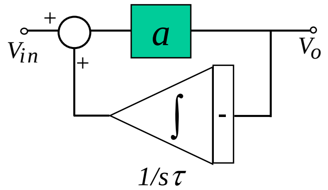
1st Order Integrator Based High-Pass Filter SGF



Note: Addition of an integrator in the feedback path → High pass frequency shaping

Addition of Integrator in Feedback Path

Let us assume flat gain in forward path (a)
Effect of addition of an integrator in the feedback path:



$$\frac{V_o}{V_{in}} = \frac{a}{1+af}$$

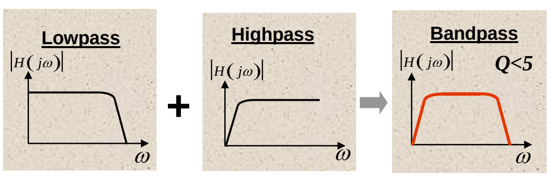
$$\frac{V_o}{V_{in}} = \frac{a}{1+a/s\tau} = \frac{s\tau}{1+s\tau/a}$$

$$\rightarrow zero@DC \quad \& \quad pole\ @ \quad \omega_{pole} = -\frac{a}{\tau} = -a \times \omega_o^{intg}$$

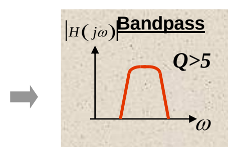
Note: For large forward path gain, a, can implement high pass function with high corner frequency
Addition of an integrator in the feedback path → zero @ DC + pole @ $a x \omega_0^{intg}$
This technique used for offset cancellation in systems where the low frequency content is not important and thus disposable

Bandpass Filters

- Bandpass filters → two cases:
  1- Low Q or wideband ($Q < 5$)
  → Combination of lowpass & highpass



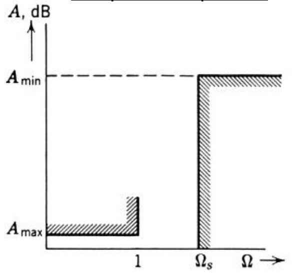
  2- High Q or narrow-band ($Q > 5$)
  → Direct implementation



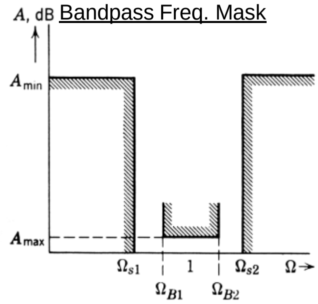
Narrow-Band Bandpass Filters
Direct Implementation

- Narrow-band BP filters → Design based on lowpass prototype
- Same tables used for LPFs are also used for BPFs

Lowpass Freq. Mask



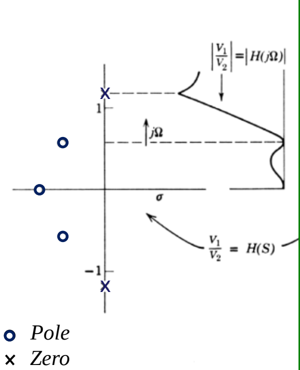
Bandpass Freq. Mask



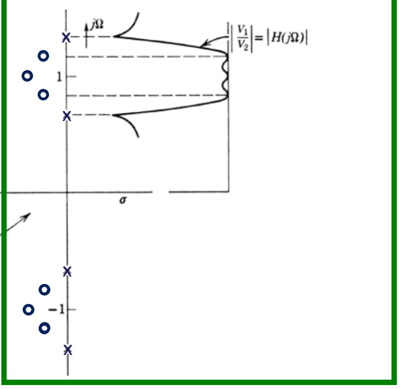
$$s \Rightarrow Q \times \left[\frac{s}{\omega_C} + \frac{\omega_C}{s} \right] \qquad \frac{\Omega_S}{\Omega_C} \Rightarrow \frac{\Omega_{S2} - \Omega_{S1}}{\Omega_{B2} - \Omega_{B1}}$$

Lowpass to Bandpass Transformation
S-plane Comparison

Lowpass pole/zero (s-plane)



Bandpass pole/zero (s-plane)

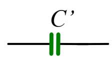
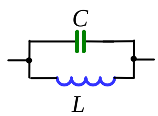
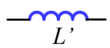
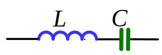


From: Zverev, *Handbook of filter synthesis*, Wiley, 1967- p.156.

Lowpass to Bandpass Transformation Table

Lowpass RLC filter structures & tables used to derive bandpass filters

$$Q = Q_{filter}$$

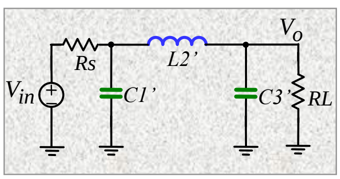
LP	BP	BP Values
		$C = QC' \times \frac{1}{R_r \omega_r}$ $L = \frac{1}{QC'} \times \frac{R_r}{\omega_r}$
		$L = QL' \times \frac{R_r}{\omega_r}$ $C = \frac{1}{QL'} \times \frac{1}{R_r \omega_r}$

C' & L' are normilzed LP values

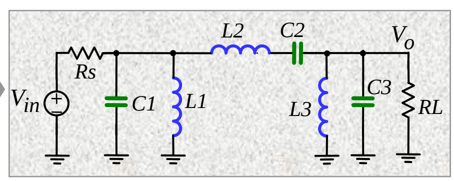
From:
Zverev,
Handbook of filter synthesis,
Wiley, 1967- p.157.

Lowpass to Bandpass Transformation Example: 3rd Order LPF → 6th Order BPF

Lowpass



Bandpass



- Each capacitor replaced by parallel L& C
- Each inductor replaced by series L&C

Lowpass to Bandpass Transformation
Example: 3rd Order LPF → 6th Order BPF

$$C_1 = QC_1' \times \frac{1}{R\omega_0}$$

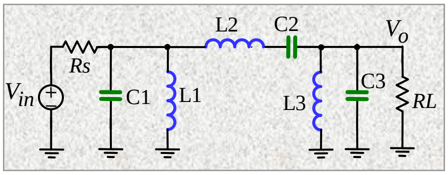
$$L_1 = \frac{1}{QC_1'} \times \frac{R}{\omega_0}$$

$$C_2 = \frac{1}{QL_2'} \times \frac{1}{R\omega_0}$$

$$L_2 = QL_2' \times \frac{R}{\omega_0}$$

$$C_3 = QC_3' \times \frac{1}{R\omega_0}$$

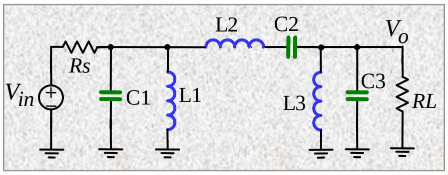
$$L_3 = \frac{1}{QC_3'} \times \frac{R}{\omega_0}$$



Where:

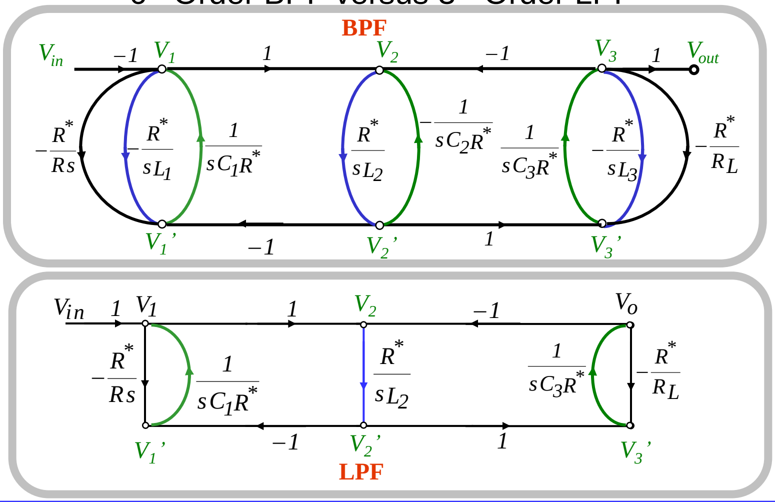
C_1' , L_2' , C_3' → Normalized lowpass values
Q → Bandpass filter quality factor
ω_0 → Filter center frequency

Lowpass to Bandpass Transformation
Signal Flowgraph

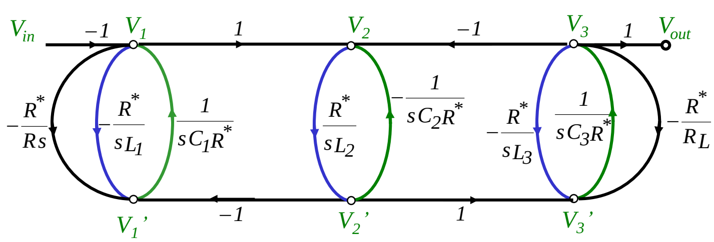


1- Voltages & currents named for all components
2- Use KCL & KVL to derive state space description
3- To have BMFs in the integrator form
   Cap. voltage expressed as function of its current $V_C=f(I_C)$
   Ind. current as a function of its voltage $I_L=f(V_L)$
4- Use state space description to draw SFG
5- Convert all current nodes to voltage

Signal Flowgraph
6^{th} Order BPF versus 3^{rd} Order LPF

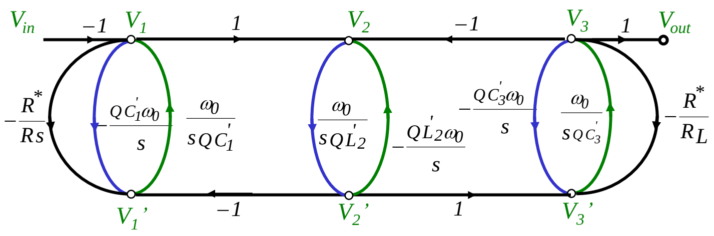


Signal Flowgraph
6^{th} Order Bandpass Filter



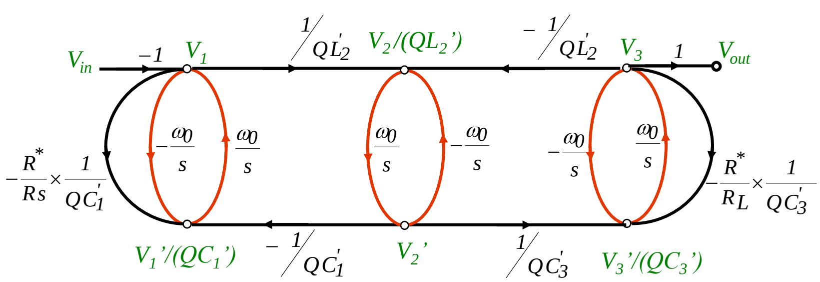
Note: each *C* & *L* in the original lowpass prototype → replaced by a *resonator*
Substituting the bandpass *L1, C1,.....* by their normalized lowpass equivalent from page 13
The resulting SFG is:

Signal Flowgraph
6^{th} Order Bandpass Filter



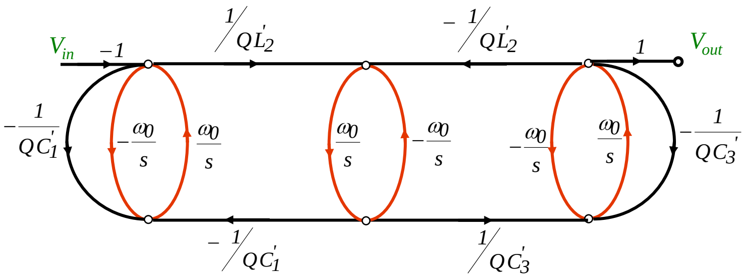
- Note the integrators → different time constants
  - Ratio of time constants for two integrator in each resonator loop~ Q^2
    → Typically, requires high component ratios
    → Poor matching
- Desirable to modify SFG so that <u>all integrators have equal time constants for optimum matching.</u>
  - To obtain equal integrator time constant → use node scaling

Signal Flowgraph
6^{th} Order Bandpass Filter



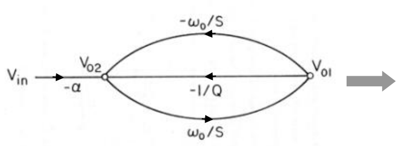
- All integrator time-constants → equal
- To simplify implementation → choose $RL=Rs=R^*$

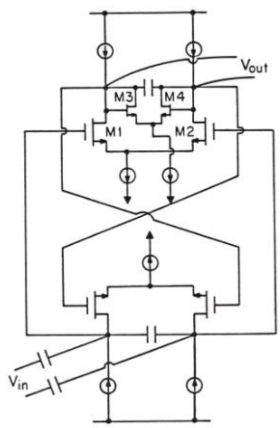
Signal Flowgraph
6^{th} Order Bandpass Filter



Let us try to build this bandpass filter using the simple Gm-C structure

Second Order Gm-C Filter
Using Simple Source-Couple Pair Gm-Cell





- Center frequency:

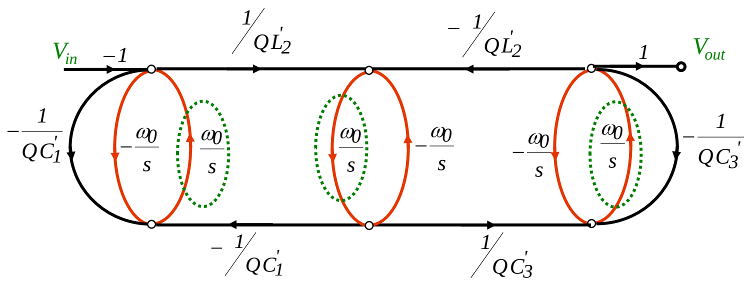
$$\omega_o = \frac{g_m^{M1,2}}{2 \times C_{intg}}$$

- Q function of:

$$Q = \frac{g_m^{M1,2}}{g_m^{M3,4}}$$

Use this structure for the 1^{st} and the 3^{rd} resonator
Use similar structure w/o M3, M4 for the 2^{nd} resonator
How to couple the resonators?

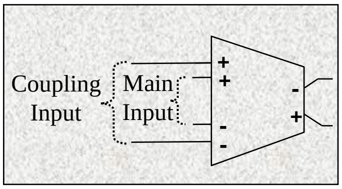
Coupling of the Resonators
1- Additional Set of Input Devices

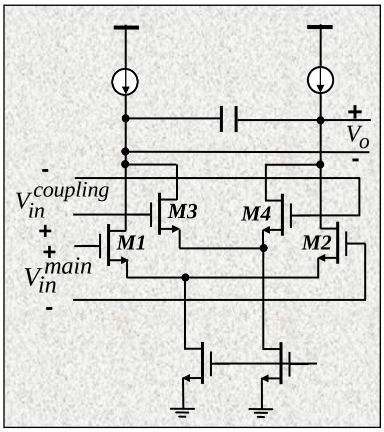


Coupling of resonators:
Use additional input source coupled pairs for the highlighted integrators
For example, the middle integrator requires 3 sets of inputs

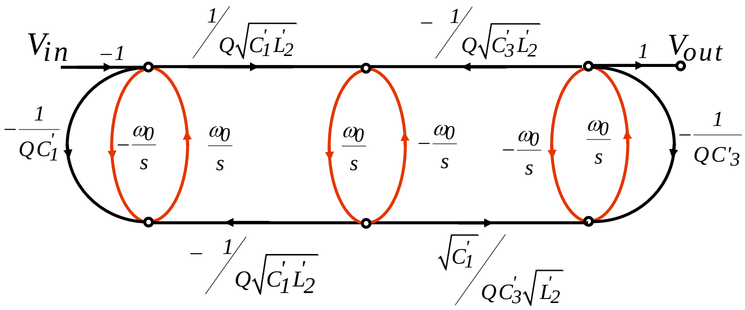
Example: Coupling of the Resonators
1- Additional Set of Input Devices

- Add one source couple pair for each additional input
- Coupling level → ratio of device widths
- Disadvantage → extra power dissipation





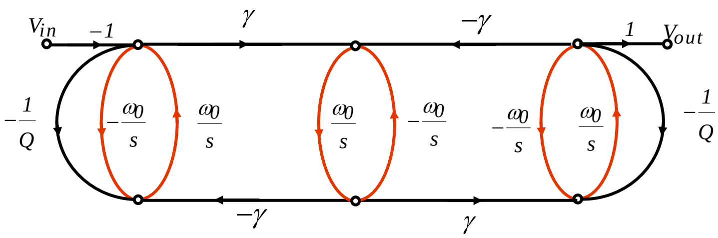
Coupling of the Resonators
2- Modify SFG → Bidirectional Coupling Paths



Modified signal flowgraph to have equal coupling between resonators

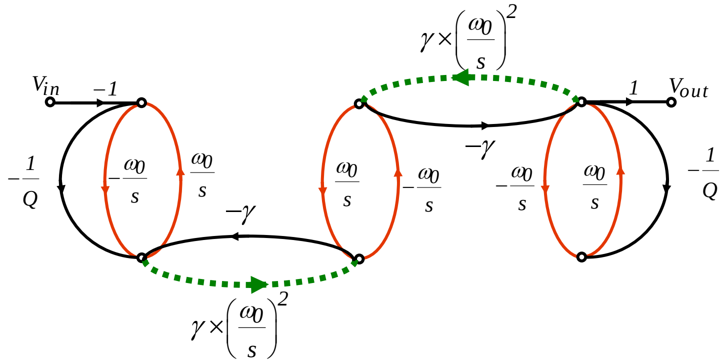
- In most filter cases $C_1' = C_3'$
- Example: For a butterworth lowpass filter $C_1' = C_3' = 1$ & $L_2' = 2$
- Assume desired overall bandpass filter $Q=10$

Sixth Order Bandpass Filter Signal Flowgraph



- Where for a Butterworth shape $\gamma = \frac{1}{Q\sqrt{2}}$
- Since in this example $Q=10$ then: $\gamma \approx \frac{1}{14}$

Sixth Order Bandpass Filter Signal Flowgraph SFG Modification



Sixth Order Bandpass Filter Signal Flowgraph SFG Modification

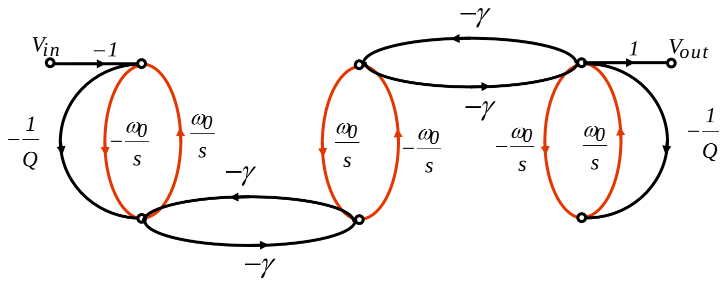
For narrow band filters (high Q) where frequencies within the passband are close to ω_0 *narrow-band approximation* can be used:

Within filter passband: $\left(\frac{\omega_0}{\omega}\right)^2 \approx 1$

$$\gamma \times \left(\frac{\omega_0}{s}\right)^2 = \gamma \times \left(\frac{\omega_0}{j\omega}\right)^2 \approx -\gamma$$

The resulting SFG:

Sixth Order Bandpass Filter Signal Flowgraph SFG Modification



Bidirectional coupling paths, can easily be implemented with coupling capacitors → no extra power dissipation

Sixth Order Gm-C Bandpass Filter Utilizing Simple Source-Coupled Pair Gm-Cell

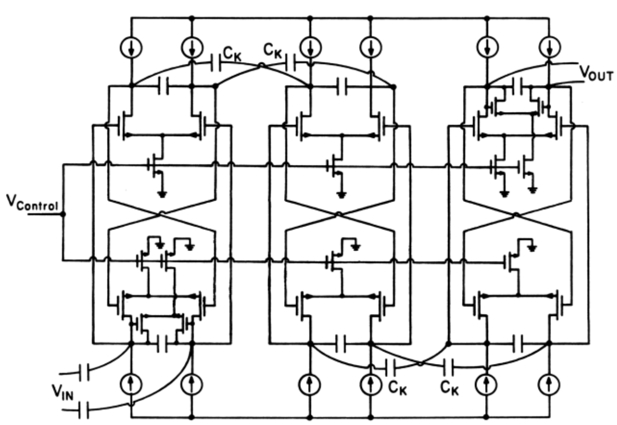
$$\gamma = \frac{C_k}{2\times C_{int\,g} + C_k}$$

$$C_k = \frac{2\times C_{int\,g}}{\frac{1}{\gamma} - 1}$$

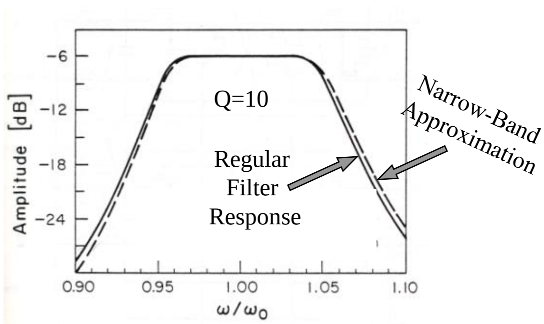
$$\gamma = 1/14$$

$$\rightarrow C_k = \frac{2}{13} C_{int\,g}$$

Parasitic cap. at integrator output, if unaccounted for, will result in inaccuracy in γ



Sixth Order Gm-C Bandpass Filter Narrow-Band versus Exact Frequency Response Simulation

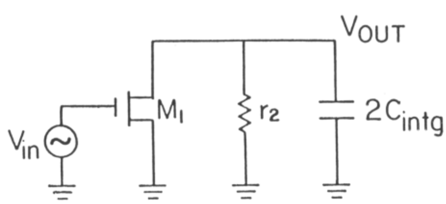


Simplest Form of CMOS Gm-Cell Nonidealities

- DC gain (integrator Q)

$$a = \frac{g_m^{M1,2}}{g_0^{M1,2} + g_{load}}$$

$$a = \frac{2L}{\theta\left(V_{gs} - V_{th}\right)_{M1,2}}$$



Small Signal Differential Mode Half-Circuit

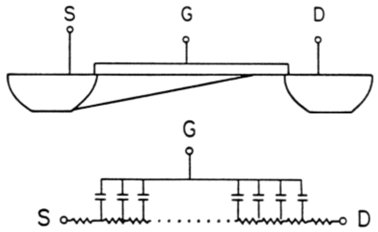
- Where a denotes DC gain & θ is related to channel length modulation (λ)by:

$$\lambda = \frac{\theta}{L}$$

- Seems no extra poles!

CMOS Gm-Cell High-Frequency Poles

Cross section view of a MOS transistor operating in saturation



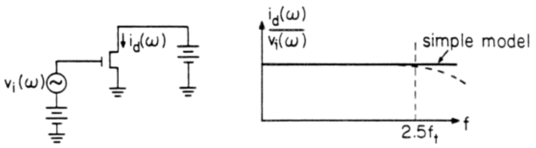
Distributed channel resistance & gate capacitance

- Distributed nature of gate capacitance & channel resistance results in infinite no. of high-frequency poles

CMOS Gm-Cell High-Frequency Poles

$$P_2^{effective} \approx \frac{1}{\sum_{i=2}^{\infty} \frac{1}{P_i}}$$

$$P_2^{effective} \approx 2.5\omega_t^{M1,2}$$



High frequency behavior of an MOS transistor operating in saturation region

$$\omega_t^{M1,2} = \frac{g_m^{M1,2}}{2/3C_{ox}WL} = \frac{3}{2}\frac{\mu\left(V_{gs}-V_{th}\right)_{M1,2}}{L^2}$$

- Distributed nature of gate capacitance & channel resistance results in an effective pole at 2.5 times input device cut-off frequency

Simple Gm-Cell Quality Factor

$$a = \frac{2L}{\theta\left(V_{gs} - V_{th}\right)_{M1,2}} \qquad P_2^{effective} = \frac{15}{4}\frac{\mu\left(V_{gs} - V_{th}\right)_{M1,2}}{L^2}$$

$$Q_{real}^{intg.} \approx \frac{1}{\frac{1}{a} - \omega_o \sum_{i=2}^{\infty} \frac{1}{p_i}}$$

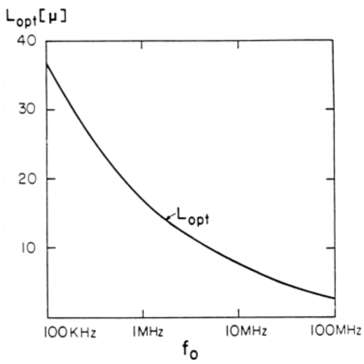
$$\frac{1}{Q^{intg.}} \approx \frac{\theta\left(V_{gs} - V_{th}\right)_{M1,2}}{2L} - \frac{4}{15}\frac{\omega_o L^2}{\mu\left(V_{gs} - V_{th}\right)_{M1,2}}$$

- Note that phase lead associated with DC gain is inversely prop. to L
- Phase lag due to high-freq. poles directly prop. to L

→ For a given ω_o there exists an optimum *L* which cancel the lead/lag phase error resulting in high integrator *Q*

Simple Gm-Cell Channel Length for Optimum Integrator Quality Factor

$$L_{opt.} \approx \left[\frac{15}{4}\frac{\theta\mu\left(V_{gs} - V_{th}\right)^2_{M1,2}}{\omega_o}\right]^{1/3}$$

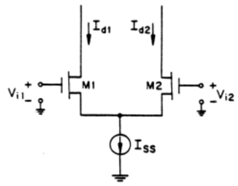


- Optimum channel length computed based on process parameters (could vary from process to process)

Source-Coupled Pair CMOS Gm-Cell Transconductance

For a source-coupled pair the differential output current (ΔI_d) as a function of the input voltage(Δv_i):

$$\Delta I_d = I_{ss}\left[\frac{\Delta v_i}{\left(V_{gs}-V_{th}\right)_{M1,2}}\right]\left\{1-\frac{1}{4}\left[\frac{\Delta v_i}{\left(V_{gs}-V_{th}\right)_{M1,2}}\right]^2\right\}^{1/2}$$



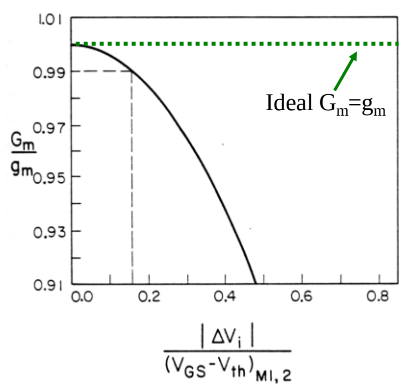
$$Note: For\ small\left[\frac{\Delta v_i}{\left(V_{gs}-V_{th}\right)_{M1,2}}\right]\rightarrow\ \frac{\Delta I_d}{\Delta v_i}=g_m^{M1,M2}$$

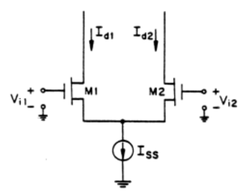
$$Note:\ As\ \ \Delta v_i\ increases\ \frac{\Delta I_d}{\Delta v_i}\ or\ the\ effective\ transconductance\ decreases$$

$$\Delta v_i = V_{i1} - V_{i2}$$

$$\Delta I_d = I_{d1} - I_{d2}$$

Source-Coupled Pair CMOS Gm-Cell Linearity

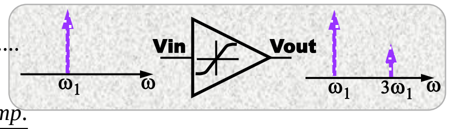




- Large signal G_m drops as input voltage increases
  → Gives rise to nonlinearity

Measure of Linearity

$$Vout = \alpha_1 Vin + \alpha_2 Vin^2 + \alpha_3 Vin^3 + \ldots\ldots\ldots\ldots$$

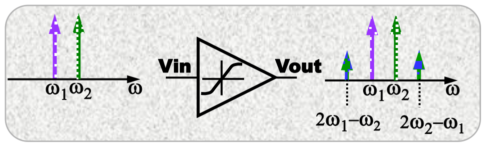


$$HD3 = \frac{amplitude\, 3rd\, harmonic\, dist.\, comp.}{amplitude\ fundamental}$$

$$= \frac{1}{4}\frac{\alpha_3}{\alpha_1} Vin^2 + \ldots\ldots$$

$$IM_3 = \frac{amplitude\, 3rd\, order\ IM\ \ comp.}{amplitude\ fundamental}$$

$$= \frac{3}{4}\frac{\alpha_3}{\alpha_1} Vin^2 + \frac{25}{8}\frac{\alpha_5}{\alpha_1} Vin^4 + \ldots\ldots$$



Source-Coupled Pair Gm-Cell Linearity

$$\Delta I_d = I_{ss}\left[\frac{\Delta v_i}{\left(V_{gs} - V_{th}\right)_{M1,2}}\right]\left\{1 - \frac{1}{4}\left[\frac{\Delta v_i}{\left(V_{gs} - V_{th}\right)_{M1,2}}\right]^2\right\}^{1/2} \qquad (1)$$

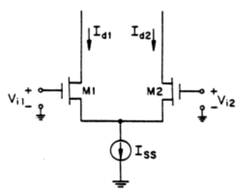
$$\Delta I_d = a_1 \times \Delta v_i + a_2 \times \Delta v_i^2 + a_3 \times \Delta v_i^3 + \ldots\ldots\ldots\ldots$$

Series expansion used in (1)

$$a_1 = \frac{I_{ss}}{\left(V_{gs} - V_{th}\right)_{M1,2}} \qquad \& \qquad a_2 = 0$$

$$a_3 = -\frac{I_{ss}}{8\left(V_{gs} - V_{th}\right)^3_{M1,2}} \qquad \& \qquad a_4 = 0$$

$$a_5 = -\frac{I_{ss}}{128\left(V_{gs} - V_{th}\right)^5_{M1,2}} \qquad \& \qquad a_6 = 0$$

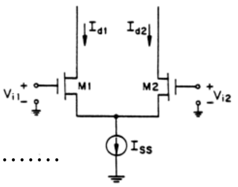


Linearity of the Source-Coupled Pair CMOS Gm-Cell

$$IM3 \approx \frac{3a_3}{4a_1}\hat{v}_i^2 + \frac{25a_5}{8a_1}\hat{v}_i^4 \;...........$$

$$Substituting\ for\ a_1, a_3, ...$$

$$IM3 \approx \frac{3}{32}\left(\frac{\hat{v}_i}{(V_{GS}-V_{th})}\right)^2 + \frac{25}{1024}\left(\frac{\hat{v}_i}{(V_{GS}-V_{th})}\right)^4 \;...........$$



$$\hat{v}_{i\,max} \approx 4(V_{GS}-V_{th}) \times \sqrt{\frac{2}{3} \times IM3}$$

$$IM_3 = 1\% \ \& \ (V_{GS}-V_{th}) = 1V \ \Rightarrow \hat{V}_{in}^{rms} \approx 230mV$$

- Note that max. signal handling capability function of gate-overdrive voltage

Dynamic Range for Source-Coupled Pair Based Filter

$$IM_3 = 1\% \ \& \ (V_{GS}-V_{th}) = 1V \ \Rightarrow V_{in}^{rms} \approx 230mV$$

- Minimum detectable signal determined by total noise voltage
- It can be shown for the *6th* order Butterworth bandpass filter fundamental noise contribution is given by:

$$\sqrt{\overline{v_o^2}} \approx \sqrt{3\,Q\frac{k\,T}{C_{intg}}}$$

$$Assuming \quad Q = 10 \quad C_{intg} = 5pF$$

$$v_{noise}^{rms} \approx 160\mu V$$

$$since \ v_{max}^{rms} = 230mV$$

$$Dynamic\ Range = 20\log\frac{230x10^{-3}}{160x10^{-6}} \approx 63dB$$

Simplest Form of CMOS Gm Cell Disadvantages

•Max. signal handling capability function of gate-overdrive

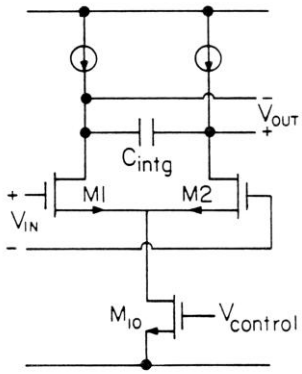
$$IM_3 \propto (V_{GS} - V_{th})^{-2}$$

•Critical freq. is also a function of gate-overdrive

$$\omega_o = \frac{g_m^{M1,2}}{2 \times C_{intg}}$$

since $g_m = \mu C_{ox} \frac{W}{L} (V_{gs} - V_{th})$

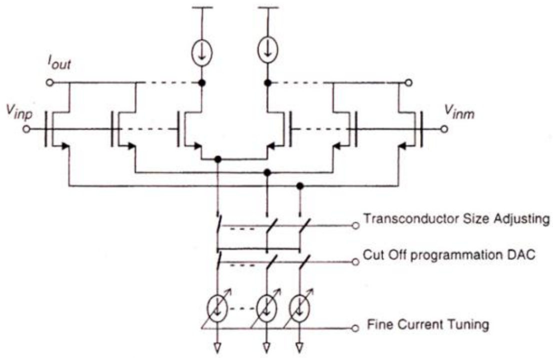
then $\omega_o \propto (V_{gs} - V_{th})$



→ Filter tuning affects max. signal handling capability!

Simplest Form of CMOS Gm Cell Removing Dependence of Maximum Signal Handling Capability on Tuning

- Can overcome problem of max. signal handling capability being a function of tuning by providing tuning through :
  - Coarse tuning via switching in/out binary-weighted cross-coupled pairs→ Try to keep gate overdrive voltage constant
  - Fine tuning through varying current sources

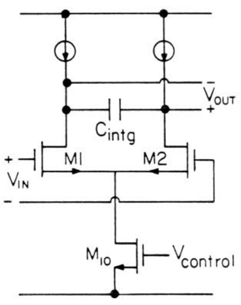


→ Dynamic range dependence on tuning removed (to 1st order)

Ref: R.Castello ,I.Bietti, F. Svelto , "High-Frequency Analog Filters in Deep Submicron Technology , "*International Solid State Circuits Conference*, pp 74-75, 1999.

Simplest Form of CMOS Gm-Cell

- Pros
  - Capable of very high frequency performance (highest?)
  - Simple design
- Cons
  - Tuning affects max. signal handling capability (can overcome)
  - Limited linearity (possible to improve)
  - Tuning affects power dissipation



Ref: H. Khorramabadi and P.R. Gray, "High Frequency CMOS continuous-time filters," *IEEE Journal of Solid-State Circuits*, Vol.-SC-19, No. 6, pp.939-948, Dec. 1984.

Gm-Cell
Source-Coupled Pair with Degeneration

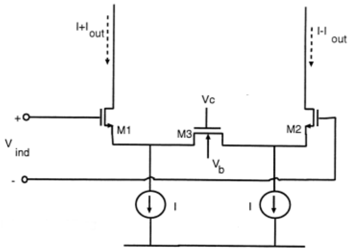
$$I_d = \frac{\mu C_{ox}}{2}\frac{W}{L}\left[2\left(V_{gs}-V_{th}\right)V_{ds}-V_{ds}^2\right]$$

$$g_{ds} = \frac{\partial I_d}{\partial V_{ds}} \approx \mu C_{ox}\frac{W}{L}\left(V_{gs}-V_{th}\right)\Bigg|_{V_{ds}\ small}$$

$$g_{eff} = \frac{1}{\dfrac{1}{g_{ds}^{M3}}+\dfrac{2}{g_m^{M1,2}}}$$

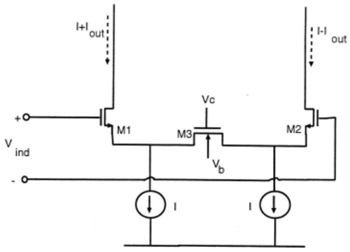
$$for\ g_m^{M1,2} >> g_{ds}^{M3}$$

$$g_{eff} \approx g_{ds}^{M3}$$



M3 operating in triode mode → source degeneration→ determines overall gm
Provides tuning through varying Vc (DC voltage source)

Gm-Cell
Source-Coupled Pair with Degeneration



- Pros
  - Moderate linearity
  - Continuous tuning provided by varying Vc
  - Tuning does not affect power dissipation
- Cons
  - Extra poles associated with the source of M1,2,3 → Low frequency applications only

Ref: Y. Tsividis, Z. Czarnul and S.C. Fang, "MOS transconductors and integrators with high linearity," *Electronics Letters*, vol. 22, pp. 245-246, Feb. 27, 1986

BiCMOS Gm-Cell
Example

- MOSFET operating in triode mode (M1):

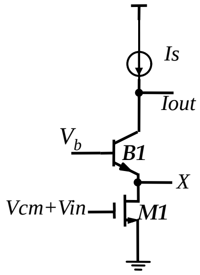
$$I_d = \frac{\mu C_{ox}}{2}\frac{W}{L}\left[2\left(V_{gs}-V_{th}\right)V_{ds}-V_{ds}^2\right]$$

$$g_m^{M1} = \frac{\partial I_d}{\partial V_{gs}} = \mu C_{ox}\frac{W}{L}V_{ds}$$

- Note that if V_{ds} is kept constant → g_m stays constant
- Linearity performance → keep gm constant as Vin varies→ function of how constant V_{ds}^{M1} can be held
  - Need to minimize gain @ node X

$$A_X = \frac{V_X}{V_{in}} = g_m^{M1}/g_m^{B1}$$

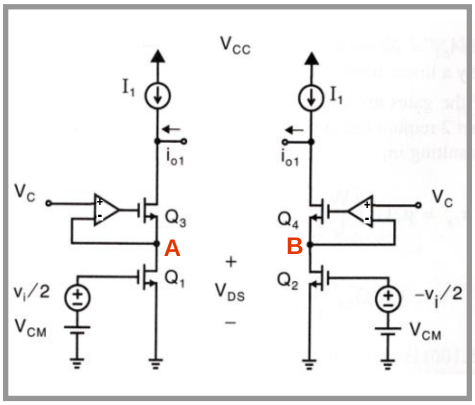
- Since for a given current, g_m of BJT is larger compared to MOS- preferable to use BJT
- Extra pole at node X could limit max. freq.



Varying V_b changes V_{ds}^{M1}
→ Changes g_m^{M1}
→ adjustable overall stage g_m

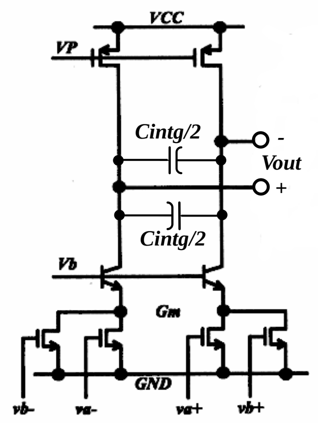
Alternative Fully CMOS Gm-Cell Example

- BJT replaced by a MOS transistor with boosted g_m
- Lower frequency of operation compared to the BiCMOS version due to more parasitic capacitance at nodes A & B

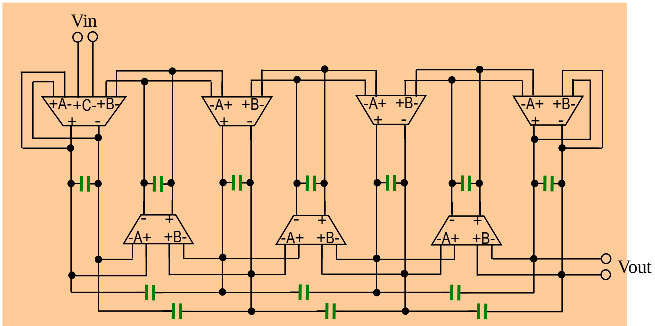


BiCMOS Gm-C Integrator

- Differential- needs common-mode feedback circuit
- Frequency tuned by varying Vb
- Design tradeoffs:
  - Extra poles at the input device drain junctions
  - Input devices have to be small to minimize parasitic poles
    - Results in high input-referred offset voltage → could drive circuit into non-linear region
    - Small devices → high 1/f noise



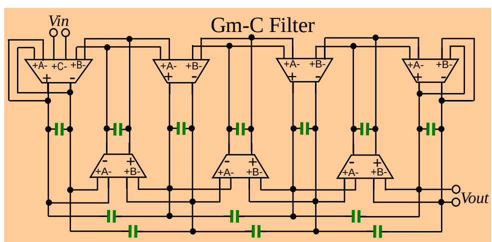
7^th Order Elliptic Gm-C LPF
For CDMA RX Baseband Application

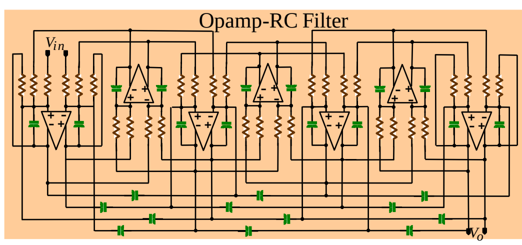


- Gm-Cell in previous page used to build a 7th order elliptic filter for CDMA baseband applications (650kHz corner frequency)
- In-band dynamic range of <50dB achieved

Comparison of 7^th Order Gm-C versus Opamp-RC LPF

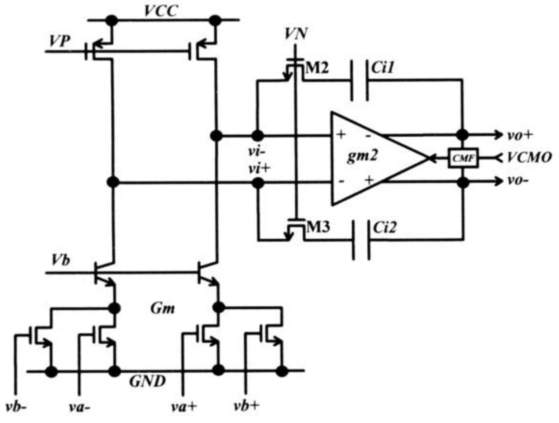
- Gm-C filter requires 4 times less intg. cap. area compared to Opamp-RC
  → For low-noise applications where filter area is dominated by Cs, could make a significant difference in the total area
- Opamp-RC linearity superior compared to Gm-C
- Power dissipation tends to be lower for Gm-C since OTA load is C and thus no need for buffering





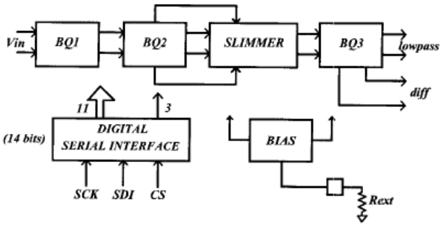
BiCMOS Gm-OTA-C Integrator

- Used to build filter for disk-drive applications
- Since high frequency of operation, time-constant sensitivity to parasitic caps significant.
  → Opamp used
- M2 & M3 added → provides phase lead to compensate for phase lag due to amp extra poles



Ref: C. Laber and P.Gray, "A 20MHz 6th Order BiCMOS Parasitic Insensitive Continuous-time Filter & Second Order Equalizer Optimized for Disk Drive Read Channels," *IEEE Journal of Solid State Circuits*, Vol. 28, pp. 462-470, April 1993.

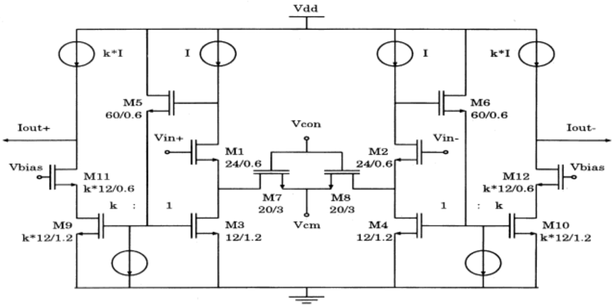
6th Order BiCMOS Continuous-time Filter & Second Order Equalizer for Disk Drive Read Channels



- Gm-C-opamp of the previous page used to build a 6th order filter for Disk Drive
- Filter consists of cascade of 3 biquads with max. Q of 2 each
- Tuning → DC tuning of gm-cells (Lect. 7 page 32) + trimming of Cs
- Performance in the order of 40dB SNDR achieved for up to 20MHz corner frequency

Ref: C. Laber and P.Gray, "A 20MHz 6th Order BiCMOS Parasitic Insensitive Continuous-time Filter & Second Order Equalizer Optimized for Disk Drive Read Channels," *IEEE Journal of Solid State Circuits*, Vol. 28, pp. 462-470, April 1993.

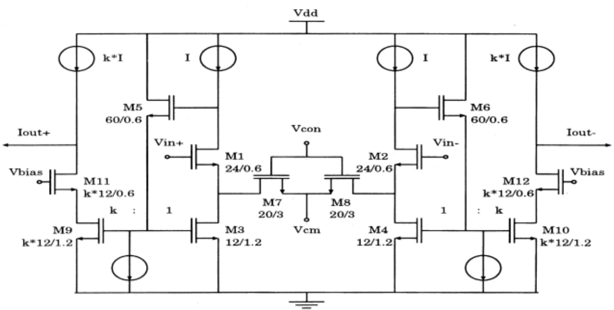
Gm-Cell
Source-Coupled Pair with Degeneration



- Gm-cell intended for low Q disk drive filter
- *M7,8* operating in triode mode provide source degeneration for *M1,2*
  → determine the overall g_m of the cell

Ref: I.Mehr and D.R.Welland, "A CMOS Continuous-Time Gm-C Filter for PRML Read Channel Applications at 150 Mb/s and Beyond", IEEE Journal of Solid-State Circuits, April 1997, Vol.32, No.4, pp. 499-513.

Gm-Cell
Source-Coupled Pair with Degeneration



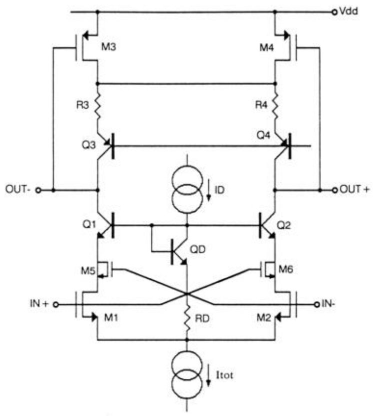
– Feedback provided by M5,6 maintains the gate-source voltage of M1,2 constant by forcing their current to be constant→ helps deliver Vin across M7,8 with good linearity

– Current mirrored to the output via M9,10 with a factor of k → overall gm scaled by k

– Performance level of about *50dB* SNDR at f_{corner} of *25MHz* achieved

BiCMOS Gm-C Integrator

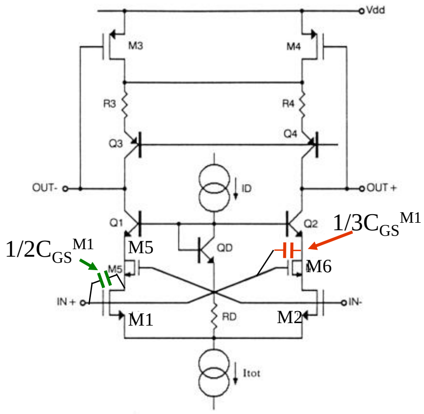
- Needs higher supply voltage compared to the previous design since quite a few devices are stacked vertically
- M1,2 → triode mode
- Q1,2 → hold V_{ds} of M1,2 constant
- Current ID used to tune filter critical frequency by varying V_{ds} of M1,2 and thus controlling *gm* of M1,2
- M3, M4 operate in triode mode and added to provide common-mode feedback



Ref: R. Alini, A. Baschirotto, and R. Castello, "Tunable BiCMOS Continuous-Time Filter for High-Frequency Applications," *IEEE Journal of Solid State Circuits,* Vol. 27, No. 12, pp. 1905-1915, Dec. 1992.

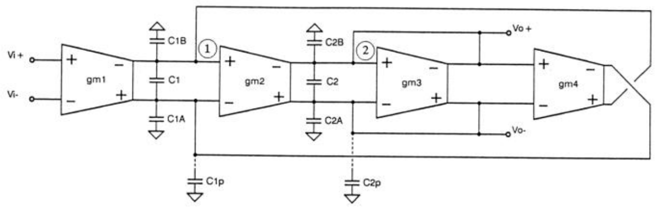
BiCMOS Gm-C Integrator

- M5 & M6 configured as capacitors- added to compensate for RHP zero due to Cgd of M1,2 (moves it to LHP) size of M5,6 → 1/3 of M1,2



Ref: R. Alini, A. Baschirotto, and R. Castello, "Tunable BiCMOS Continuous-Time Filter for High-Frequency Applications," *IEEE Journal of Solid State Circuits,* Vol. 27, No. 12, pp. 1905-1915, Dec. 1992.

BiCMOS Gm-C Filter For Disk-Drive Application



- Using the integrators shown in the previous page
- Biquad filter for disk drives
- *gm1=gm2=gm4=2gm3*
- *Q=2*
- Tunable from 8MHz to 32MHz

Ref: R. Alini, A. Baschirotto, and R. Castello, "Tunable BiCMOS Continuous-Time Filter for High-Frequency Applications," *IEEE Journal of Solid State Circuits,* Vol. 27, No. 12, pp. 1905-1915, Dec. 1992.

Summary
Continuous-Time Filters

- Opamp RC filters
  - Good linearity → High dynamic range (*60-90dB*)
  - Only discrete tuning possible
  - Medium usable signal bandwidth (<*10MHz*)
- Opamp MOSFET-C
  - Linearity compromised (typical dynamic range *40-60dB*)
  - Continuous tuning possible
  - Low usable signal bandwidth *(<5MHz*)
- Opamp MOSFET-RC
  - Improved linearity compared to Opamp MOSFET-C (D.R. *50-90dB*)
  - Continuous tuning possible
  - Low usable signal bandwidth *(<5MHz*)
- Gm-C
  - Highest frequency performance -at least an order of magnitude higher compared to other integrator-based active filters *(<100MHz*)
  - Typically, dynamic range not as high as Opamp RC but better than Opamp MOSFET-C (*40-70dB*)

Switched-Capacitor Filters

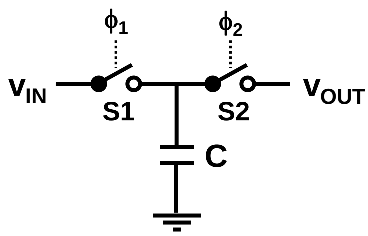
- S.C. filters are sampled-data type circuits operating with continuous signal amplitude & quantized time
- First product including switched-capacitor filters
  - Intel 2912 voice-band CODEC
- Stand-alone filter IC: LMF100 from National Semi.
  - Dual S.C. biquad with LP, HP, BP outputs
- Other than filters, S.C. circuits are used in oversampled data converters
- Pioneering work on S.C. filter technology was mostly performed at UC Berkeley

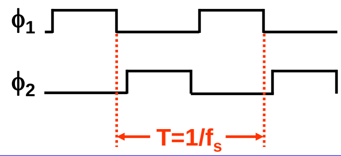
Switched-Capacitor Filters

- Emulating resistor via switched-capacitor network
- Switched-capacitor 1st order filter
- Switch-capacitor filter considerations:
  - Issue of aliasing and how to prevent aliasing
  - Tradeoffs in choice of sampling rate
  - Effect of sample and hold
  - Switched-capacitor filter electronic noise

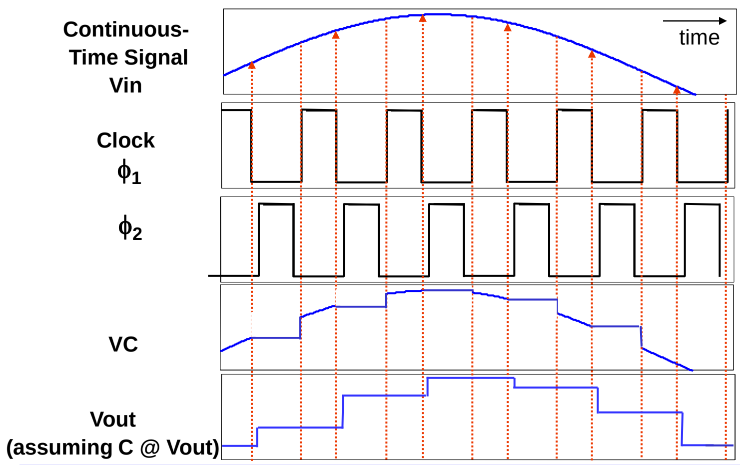
Switched-Capacitor Resistor

- Capacitor C is the “switched capacitor”
- Non-overlapping clocks ϕ_1 and ϕ_2 control switches S1 and S2, respectively
- v_{IN} is sampled at the falling edge of ϕ_1
  - Sampling frequency f_S
- Next, ϕ_2 rises and the voltage across C is transferred to v_{OUT}





Switched-Capacitor Resistor
Waveforms



Switched-Capacitor Resistors

- Why does this behave as a resistor?
- Charge transferred from v_{IN} to v_{OUT} during each clock cycle is:

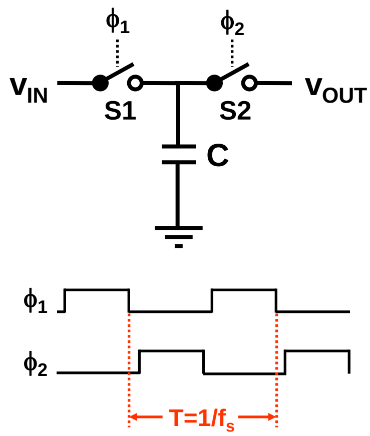
$$Q = C(v_{IN} - v_{OUT})$$

- Average current flowing from v_{IN} to v_{OUT} is:

$$i = Q/t = Q \cdot f_s$$

Substituting for Q:

$$i = f_S C(v_{IN} - v_{OUT})$$



Switched-Capacitor Resistors

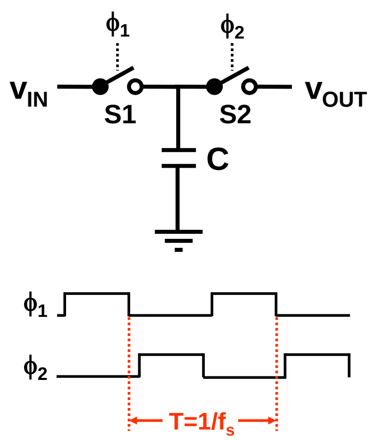
$$i = f_S C(v_{IN} - v_{OUT})$$

With the current through the switched-capacitor resistor proportional to the voltage across it, the equivalent "switched capacitor resistance" is:

$$R_{eq} = \frac{V_{IN} - V_{OUT}}{i} = \frac{1}{f_S C}$$

Example:
$f_S = 100KHz, C = 0.1pF$
$\rightarrow R_{eq} = 100Mega\Omega$

Note: Can build large time-constant in small area

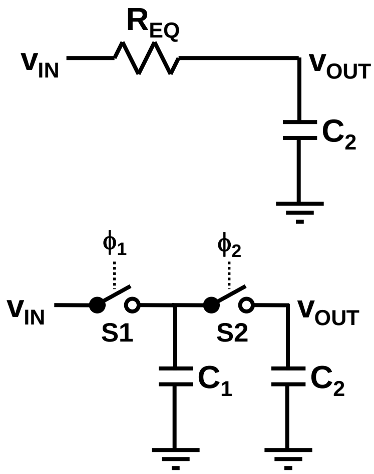


Switched-Capacitor Filter

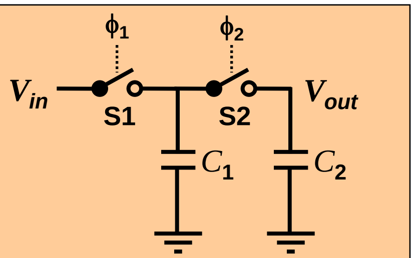
- Let's build a "switched- capacitor " filter …
- Start with a simple RC LPF
- Replace the physical resistor by an equivalent switched-capacitor resistor
- 3-dB bandwidth:

$$\omega_{-3dB} = \frac{1}{R_{eq}C_2} = f_s \times \frac{C_1}{C_2}$$

$$f_{-3dB} = \frac{1}{2\pi} f_s \times \frac{C_1}{C_2}$$

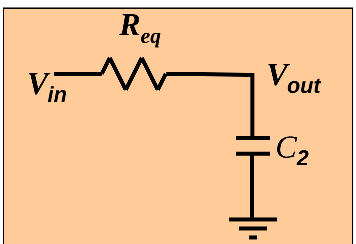


Switched-Capacitor Filter Advantage versus Continuous-Time Filter



$$f_{-3dB} = \frac{1}{2\pi} f_s \times \frac{C_1}{C_2}$$

- Corner freq. proportional to:
  System clock (accurate to few ppm)
  C ratio accurate → < 0.1%



$$f_{-3dB} = \frac{1}{2\pi} \times \frac{1}{R_{eq}C_2}$$

- Corner freq. proportional to:
  Absolute value of Rs & Cs
  Poor accuracy → 20 to 50%

Main advantage of SC filters → inherent critical frequency accuracy