

EE247

Lecture 8

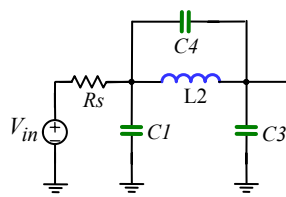
- Summary of various continuous-time filter frequency tuning techniques
- Continuous-time filter design considerations
 - Monolithic highpass filters
 - Active bandpass filter design
 - Lowpass to bandpass transformation
 - Example: 6th order bandpass filter
 - Gm-C bandpass filter using simple diff. pair
 - Various Gm-C filter implementations
- Performance comparison of various continuous-time filter topologies

Summary: Continuous-Time Filter Frequency Tuning

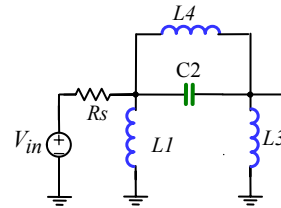
- Trimming
 - Expensive & does not account for temperature and supply etc... variations
- Automatic frequency tuning
 - Continuous tuning
 - Master VCF used in tuning loop, same tuning signal used to tune the slave (main) filter
 - Tuning quite accurate
 - Issue → reference signal feedthrough to the filter output
 - Master VCO used in tuning loop
 - Design of reliable & stable VCO challenging
 - Issue → reference signal feedthrough
 - Single integrator in negative feedback loop forces time-constant to be a function of accurate clock frequency
 - More flexibility in choice of reference frequency → less feedthrough issues
 - DC locking of a replica of the integrator to an external resistor
 - DC offset issues & does not account for integrating capacitor variations
 - Periodic digitally assisted tuning
 - Requires digital capability + minimal additional hardware
 - Advantage of no reference signal feedthrough since tuning performed off-line

RLC Highpass Filters

- Any RLC lowpass 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 $C_r = 1 / (R_r \omega_r)$



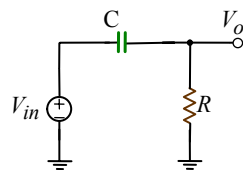
Lowpass



Highpass

Integrator Based High-Pass Filters 1st Order

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



$$\frac{V_o}{V_{in}} = \frac{sRC}{1+sRC}$$

1st Order Integrator Based High-Pass Filter Signal Flowgraph

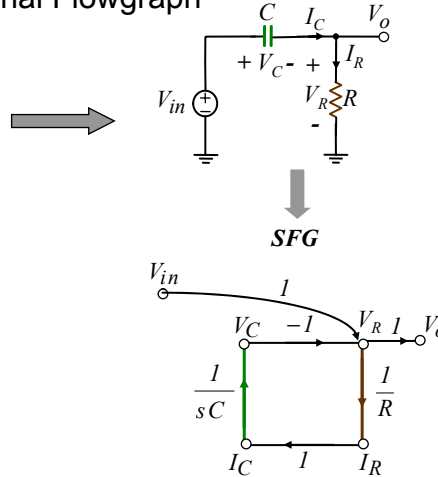
$$V_R = V_{in} - V_C$$

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

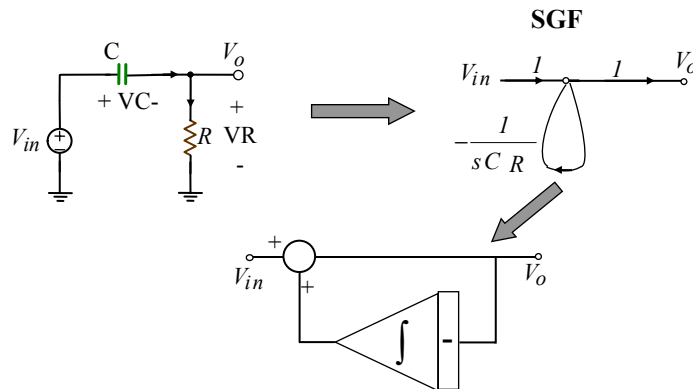
$$V_O = V_R$$

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

$$I_C = I_R$$



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:

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

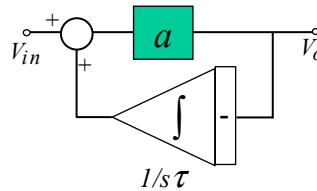
$$\frac{V_O}{V_{in}} = \frac{a}{1+a/s\tau} = \frac{s\tau}{1+s\tau/a}$$

$$\rightarrow \text{zero @ DC} \quad \& \quad \text{pole @ } \omega_{pole} = -\frac{a}{\tau} = -a \times \omega_0^{intg}$$

Note: For large forward path gain, a , can implement high pass function with high corner frequency

Addition of an integrator in the feedback path \rightarrow zero @ DC + pole @ $a \times \omega_0^{intg}$

This technique used for offset cancellation in systems where the low frequency content is not important and thus disposable

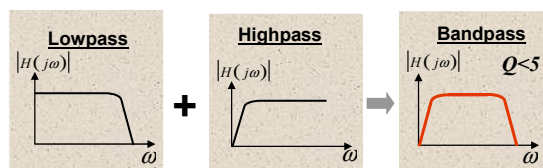


Bandpass Filters

• Bandpass filters \rightarrow two cases:

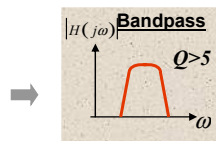
1- Low Q or wideband ($Q < 5$)

\rightarrow Combination of lowpass & highpass



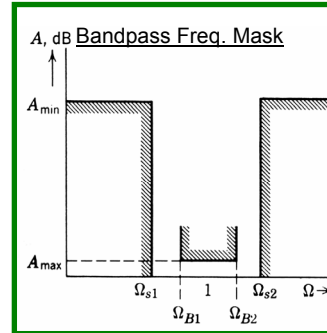
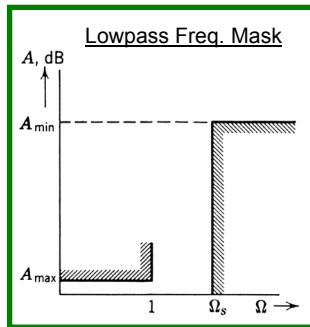
2- High Q or narrow-band ($Q > 5$)

\rightarrow Direct implementation



Narrow-Band Bandpass Filters Direct Implementation

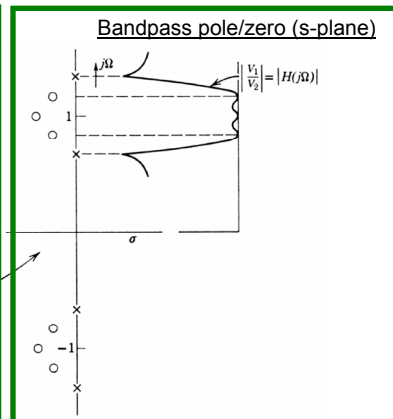
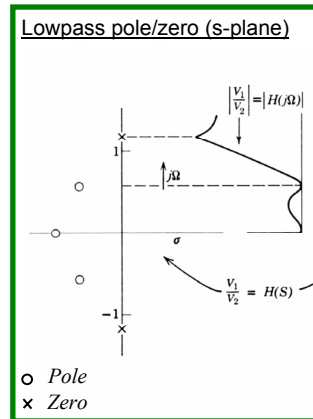
- Narrow-band BP filters → Design based on lowpass prototype
- Same tables used for LPFs are also used for BPFs



$$s \Rightarrow Q \times \left[\frac{s}{\omega_c} + \frac{\omega_c}{s} \right]$$

$$\frac{\Omega_s}{\omega_c} \Rightarrow \frac{\Omega_{s2} - \Omega_{s1}}{\Omega_{B2} - \Omega_{B1}}$$

Lowpass to Bandpass Transformation



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

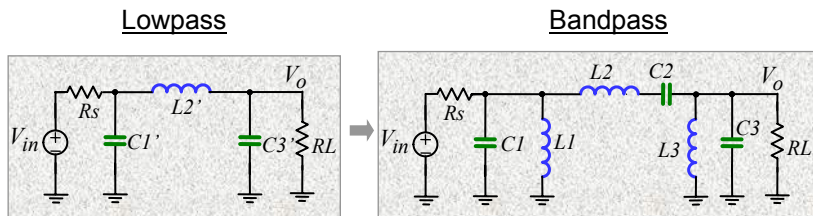
$$Q = Q_{filter}$$

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

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

C' & L' are normalized LP values

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



- 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

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

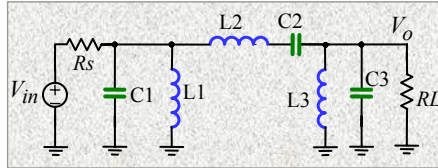
$$L_1 = \frac{1}{QC_1'} \times \frac{R}{\omega_0}$$

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

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

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

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

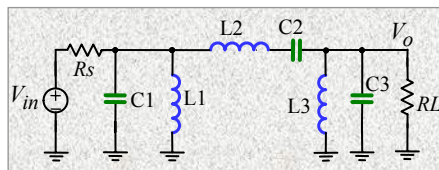


Where:

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

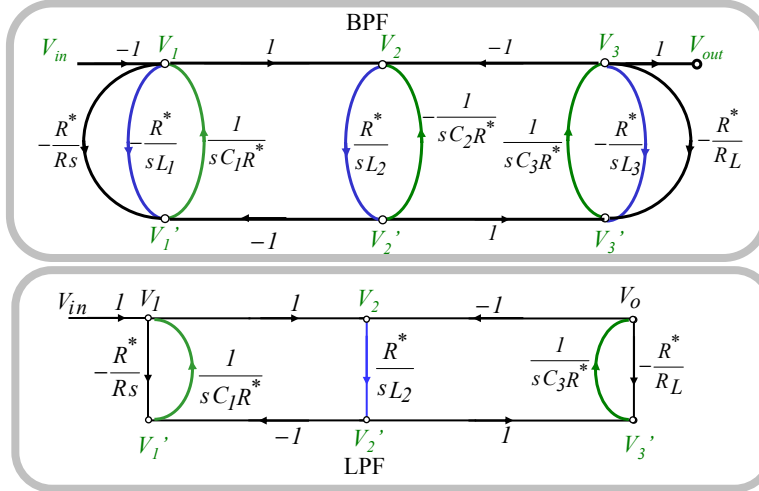
Lowpass to Bandpass Transformation

Signal Flow Graph

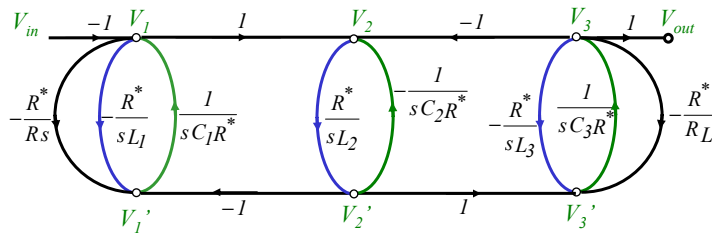


- 1- Voltages & currents named for all components
- 2- Use KCL & KVL to derive state space description
- 3- To have BMFs in the integrator form
 - Cap. voltage expressed as function of its current $V_C = f(I_C)$
 - Ind. current as a function of its voltage $I_L = f(V_L)$
- 4- Use state space description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

Signal Flowgraph 6th Order BPF versus 3rd Order LPF

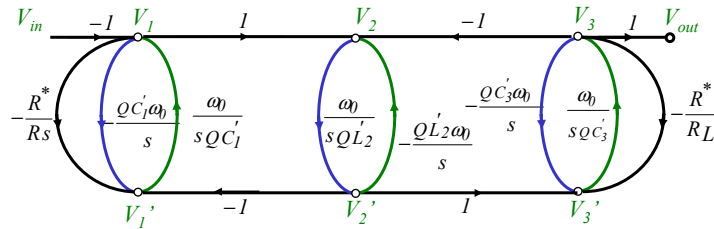


Signal Flowgraph 6th Order Bandpass Filter



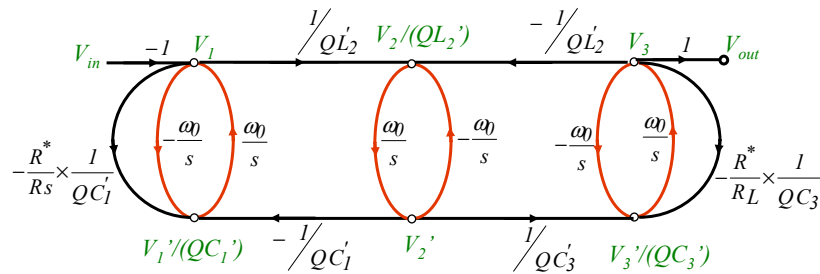
Note: each C & L in the original lowpass prototype \rightarrow replaced by a *resonator*
 Substituting the bandpass LI, CI, \dots by their normalized lowpass equivalent from page 13
 The resulting SFG is:

Signal Flowgraph 6th Order Bandpass Filter



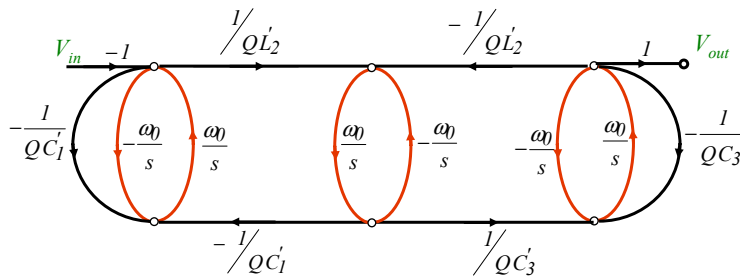
- Note the integrators \rightarrow different time constants
 - Ratio of time constants for two integrator in each resonator $\sim Q^2$
 - \rightarrow Typically, requires high component ratios
 - \rightarrow Poor matching
- Desirable to modify SFG so that all integrators have equal time constants for optimum matching.
 - To obtain equal integrator time constant \rightarrow use node scaling

Signal Flowgraph 6th Order Bandpass Filter



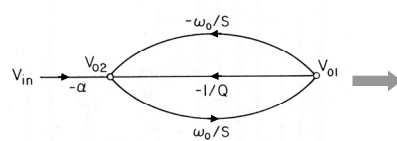
- All integrator time-constants \rightarrow equal
- To simplify implementation \rightarrow choose $RL=Rs=R^*$

Signal Flowgraph 6th 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

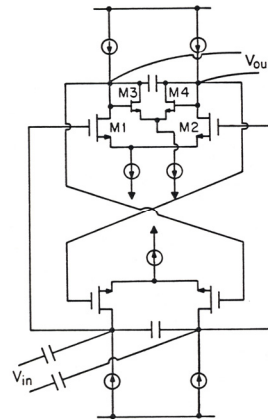


- Center frequency:

$$\omega_0 = \frac{g_m^{M1,2}}{2 \times C_{intg}}$$

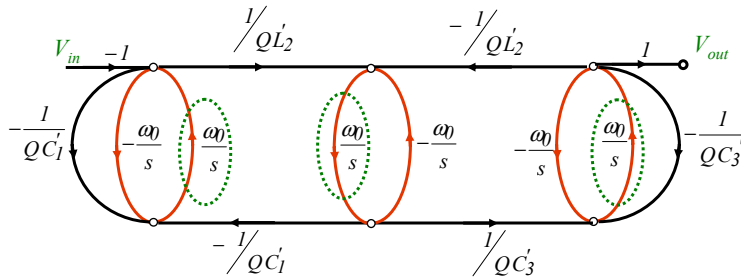
- Q function of:

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



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

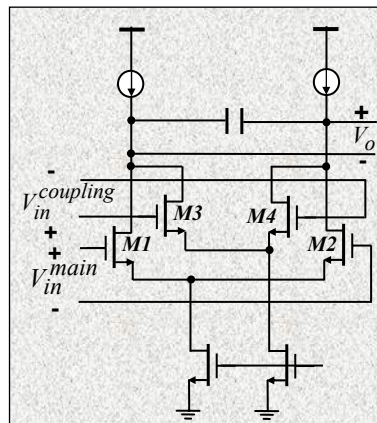
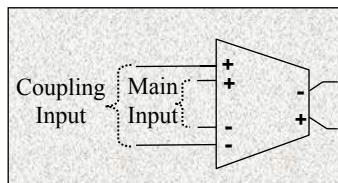
Coupling of the Resonators 1- Additional Set of Input Devices



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

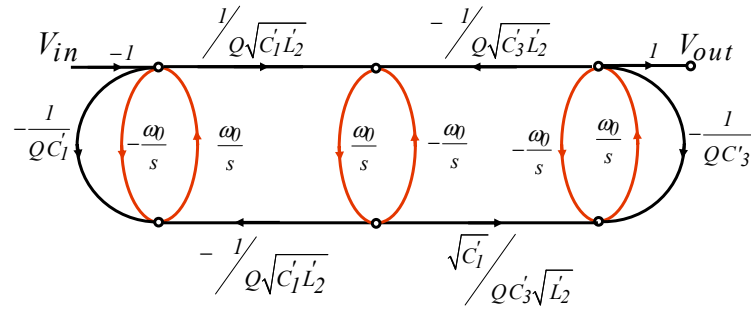
Example: Coupling of the Resonators 1- Additional Set of Input Devices

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



Coupling of the Resonators

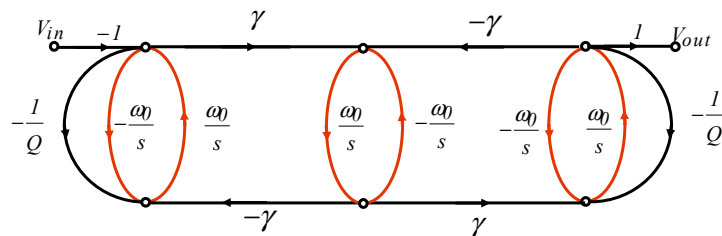
2- Modify SFG → Bidirectional Coupling Paths



Modified signal flowgraph to have equal coupling between resonators

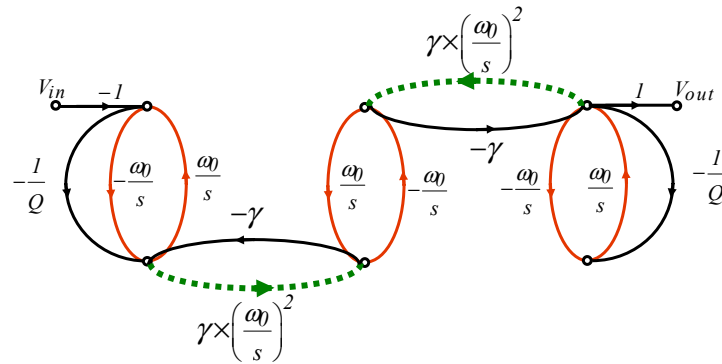
- In most filter cases $C_1' = C_3'$
- Example: For a butterworth lowpass filter $C_1' = C_3' = 1$ & $L_2' = 2$
- Assume desired overall bandpass filter $Q=10$

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

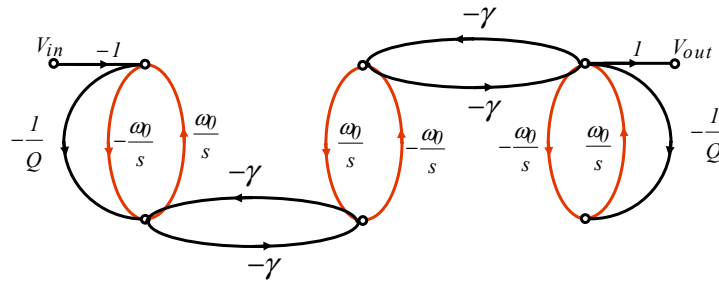
For narrow band filters (high Q) where frequencies within the passband are close to ω_0 , *narrow-band approximation* can be used:

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 \rightarrow no extra power dissipation

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

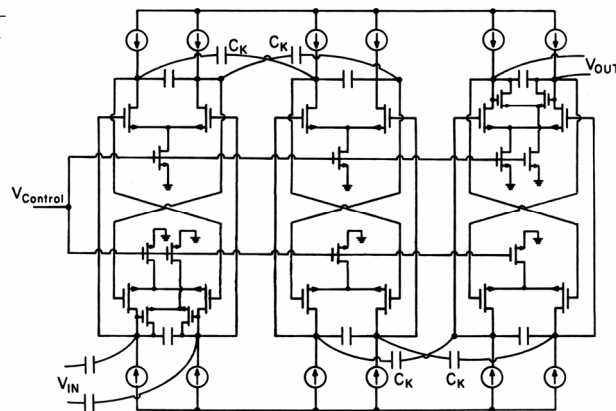
$$\gamma = \frac{C_k}{2 \times C_{int} g + C_k}$$

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

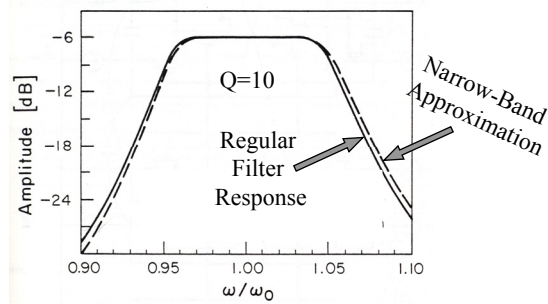
$$\gamma = 1/14$$

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

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

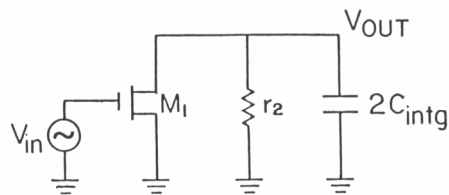


Simplest Form of CMOS Gm-Cell Nonidealities

- DC gain (integrator Q)

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

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



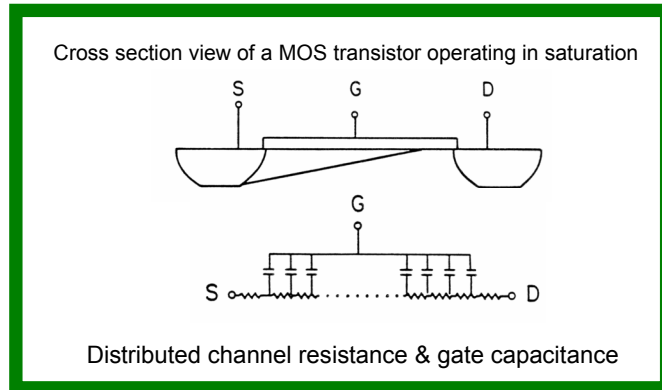
Small Signal Differential Mode Half-Circuit

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

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

- Seems no extra poles!

CMOS Gm-Cell High-Frequency Poles



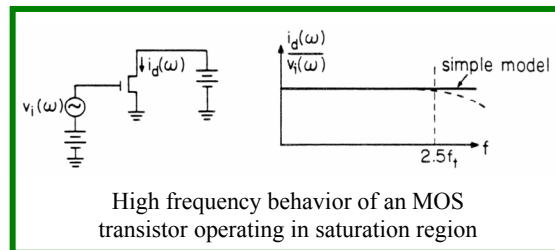
- Distributed nature of gate capacitance & channel resistance results in infinite no. of high-frequency poles

CMOS Gm-Cell High-Frequency Poles

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

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

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



- 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

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

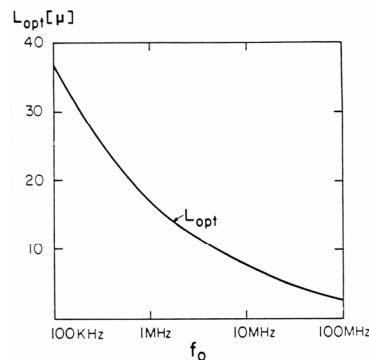
$$Q_{real}^{intg.} \approx \frac{1}{\frac{1}{a} - \omega_b \sum_{i=2}^{\infty} \frac{1}{p_i}}$$

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

- 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_0 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

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



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

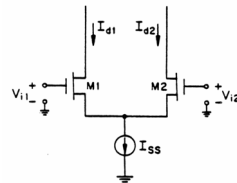
Source-Coupled Pair CMOS Gm-Cell Transconductance

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

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

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

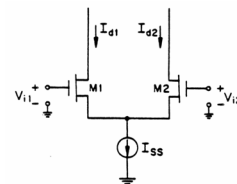
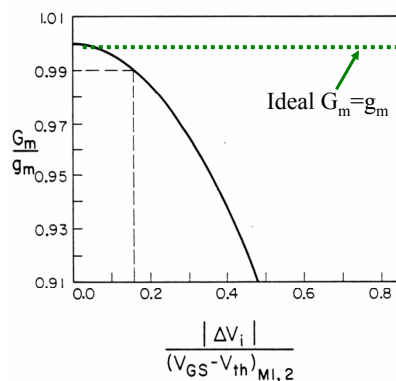
Note: As Δv_i increases $\frac{\Delta I_d}{\Delta v_i}$ or the
effective transconductance decreases



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

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

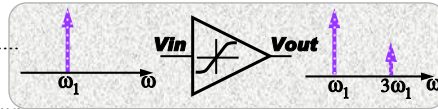
Source-Coupled Pair CMOS Gm-Cell Linearity



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

Measure of Linearity

$$V_{out} = \alpha_1 V_{in} + \alpha_2 V_{in}^2 + \alpha_3 V_{in}^3 + \dots$$

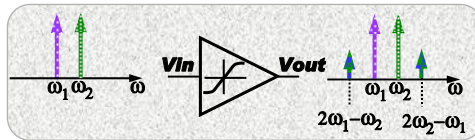


$$HD_3 = \frac{\text{amplitude 3rd harmonic dist. comp.}}{\text{amplitude fundamental}}$$

$$= \frac{1}{4} \frac{\alpha_3}{\alpha_1} V_{in}^2 + \dots$$

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

$$= \frac{3}{4} \frac{\alpha_3}{\alpha_1} V_{in}^2 + \frac{25}{8} \frac{\alpha_5}{\alpha_1} V_{in}^4 + \dots$$



Source-Coupled Pair Gm-Cell Linearity

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

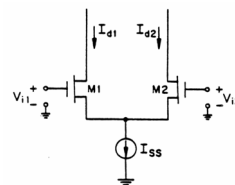
$$\Delta I_d = a_1 \times \Delta v_i + a_2 \times \Delta v_i^2 + a_3 \times \Delta v_i^3 + \dots$$

Series expansion used in (1)

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

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

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

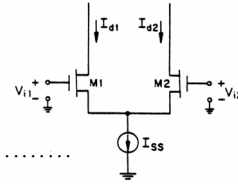


Linearity of the Source-Coupled Pair CMOS Gm-Cell

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

Substituting for a_1, a_3, \dots

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



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

$$IM3 = 1\% \ \& \ (V_{GS} - V_{th}) = 1V \Rightarrow \hat{V}_{in}^{rms} \approx 230mV$$

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

Simplest Form of CMOS Gm Cell Disadvantages

- Max. signal handling capability function of gate-overdrive

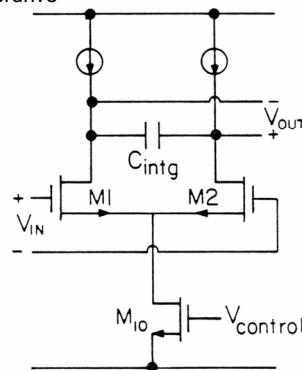
$$IM3 \propto (V_{GS} - V_{th})^{-2}$$

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

$$\omega_0 = \frac{g_m^{M1,2}}{2 \times C_{intg}}$$

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

$$\text{then } \omega_0 \propto (V_{gs} - V_{th})$$

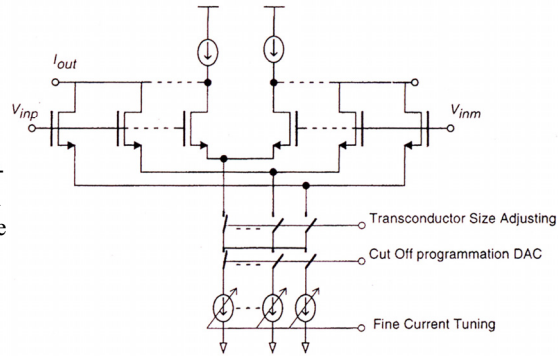


→ Filter tuning affects max. signal handling capability!

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.

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:

$$\sqrt{v_o^2} \approx \sqrt{3Q \frac{kT}{C_{intg}}}$$

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

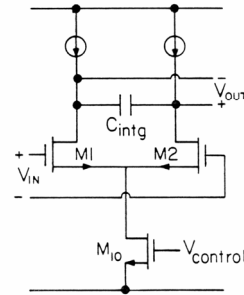
$$v_{noise}^{rms} \approx 160\mu V$$

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

$$Dynamic \ Range = 20 \log \frac{230 \times 10^{-3}}{160 \times 10^{-6}} \approx 63dB$$

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

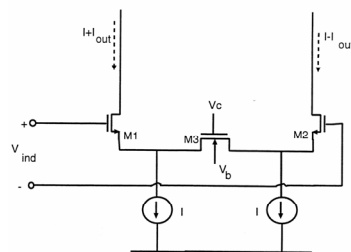
$$I_d = \frac{\mu C_{ox} W}{2L} [2(V_{gs} - V_{th})V_{ds} - V_{ds}^2]$$

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

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

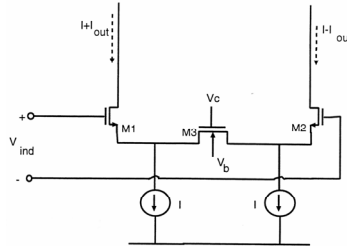
$$\text{for } g_m^{M1,2} \gg g_{ds}^{M3}$$

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



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

Gm-Cell Source-Coupled Pair with Degeneration



- Pros
 - Moderate linearity
 - Continuous tuning provided by varying V_c
 - 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 in triode mode (M1):

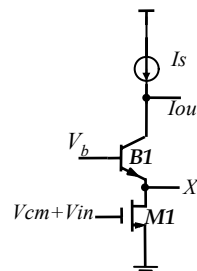
$$I_d = \frac{\mu C_{ox} W}{2L} \left[2(V_{gs} - V_{th})V_{ds} - V_{ds}^2 \right]$$

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

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

$$A_x = \frac{V_x}{V_{in}} = g_m^{M1} / g_m^{B1}$$

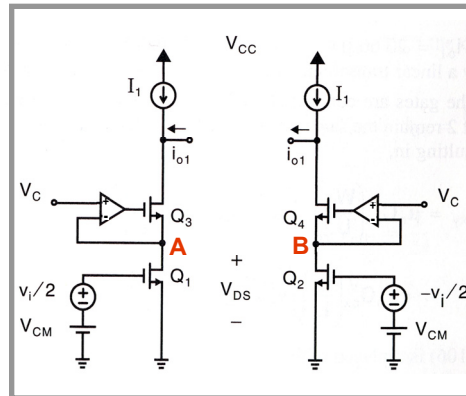
- Since for a given current, g_m of BJT is larger compared to MOS- preferable to use BJT
- Extra pole at node X could limit max. freq.



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

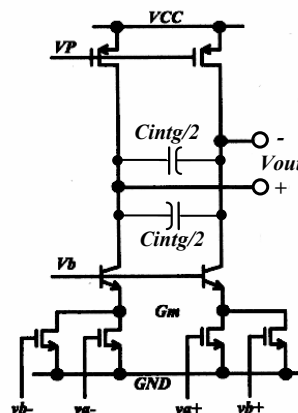
Alternative Fully CMOS Gm-Cell Example

- 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

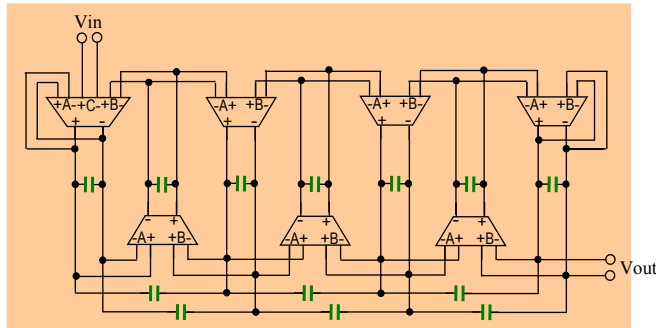


BiCMOS Gm-C Integrator

- Differential- needs common-mode feedback ckt
- Freq. tuned by varying V_b
- 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 \rightarrow could drive ckt into non-linear region
 - Small devices \rightarrow high $1/f$ noise



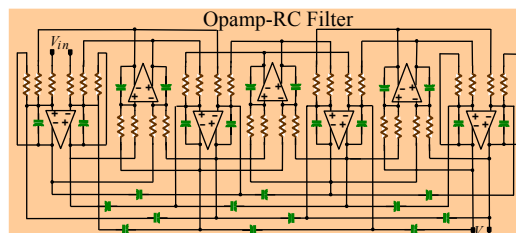
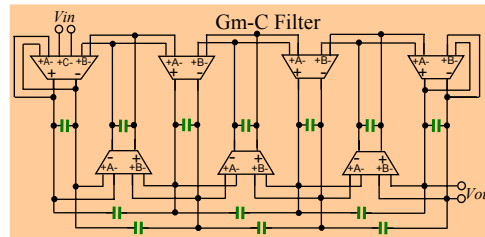
7th Order Elliptic Gm-C LPF For CDMA RX Baseband Application



- 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

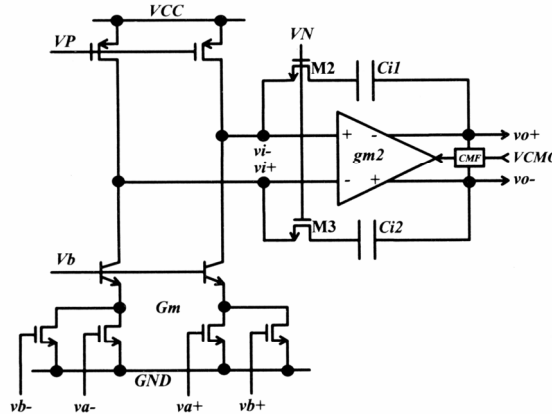
Comparison of 7th Order Gm-C versus Opamp-RC LPF

- Gm-C filter requires 4 times less intg. cap. area compared to Opamp-RC
 - For low-noise applications where filter area is dominated by C_s , 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



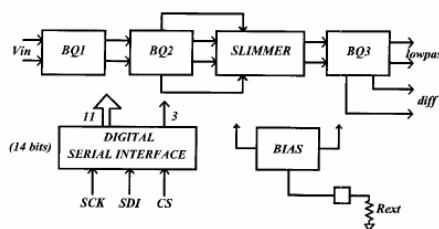
BiCMOS Gm-OTA-C Integrator

- 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 to compensate for phase lag (provides phase lead)



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.

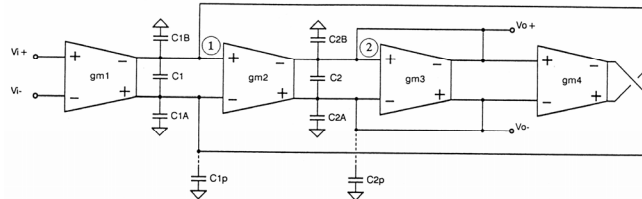
6th Order BiCMOS Continuous-time Filter & Second Order Equalizer for Disk Drive Read Channels



- 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
- 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.

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. Baschiroto, 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 ($<100MHz$)
 - Dynamic range not as high as Opamp RC but better than Opamp MOSFET-C ($40-70dB$)