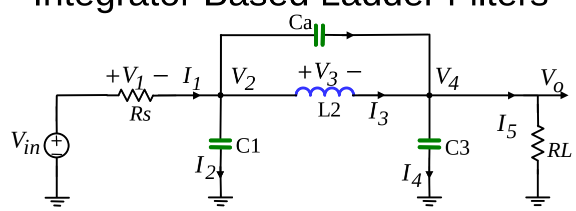
EE247
Lecture 5

- Summary last lecture
- Continuous-time filters
  - Facts about monolithic Rs & Cs and its effect on integrated filter characteristics
  - Opamp MOSFET-C filters
  - Opamp MOSFET-RC filters
  - Gm-C filters
- Frequency tuning for continuous-time filters
  - Trimming
  - Automatic frequency tuning
    - Continuous tuning
    - Periodic tuning

Summary Last Lecture

- High Q high order filters
  - Transmission zero implementation
  - Example
- Various integrator topologies utilized in monolithic filters
  - Resistor + C based filters
  - Transconductance (gm) + C based filters
  - Switched-capacitor filters
- Effect of integrator non-idealities on filter behavior

Summary Last Lecture
Transmission zero Implementation for Integrator Based Ladder Filters



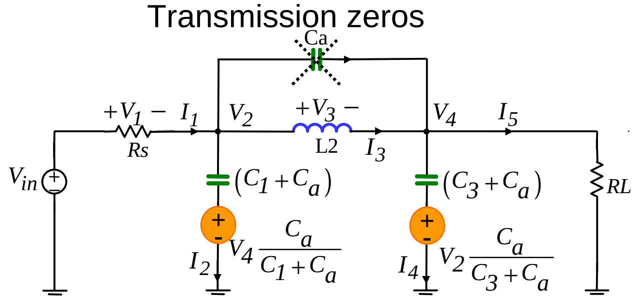
- Use KCL & KVL to derive :

$$V_2 = \frac{I_1 - I_3}{s(C_1 + C_a)} + V_4 \frac{C_a}{C_1 + C_a}$$

Voltage Controlled Voltage Source!

$$V_4 = \frac{I_3 - I_5}{s(C_3 + C_a)} + V_2 \frac{C_a}{C_3 + C_a}$$

Summary Last Lecture
Integrator Based Ladder Filters
Transmission zeros

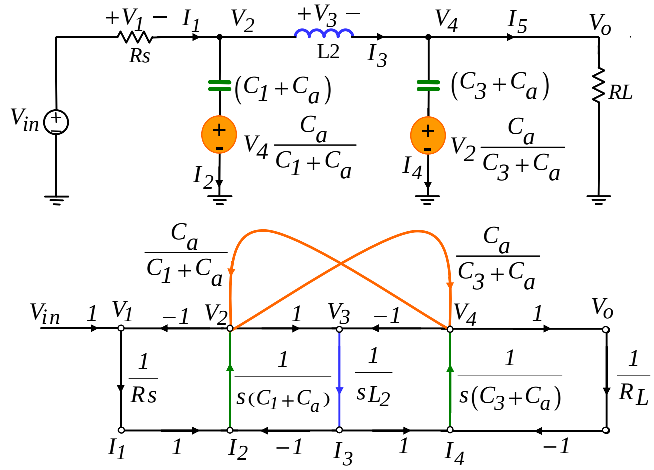


- Replace *shunt capacitor* with *voltage controlled voltage sources*:

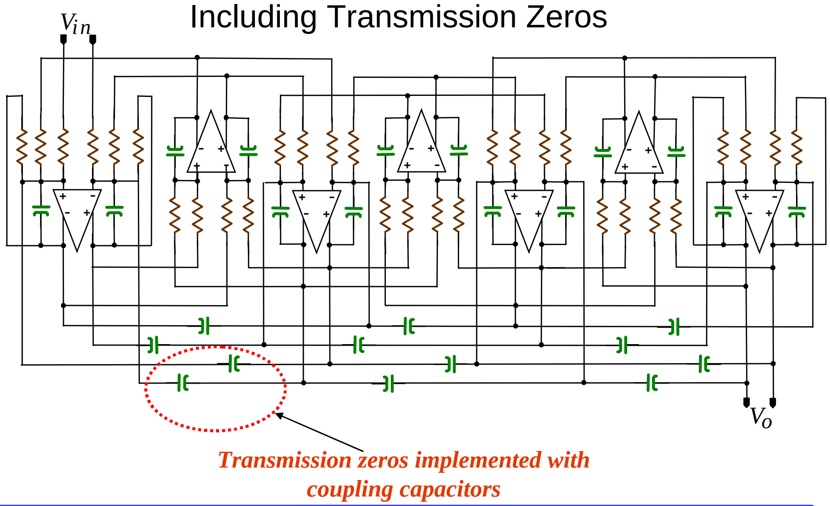
$$V_2 = \frac{I_1 - I_3}{s(C_1 + C_a)} + V_4 \frac{C_a}{C_1 + C_a}$$

$$V_4 = \frac{I_3 - I_5}{s(C_3 + C_a)} + V_2 \frac{C_a}{C_3 + C_a}$$

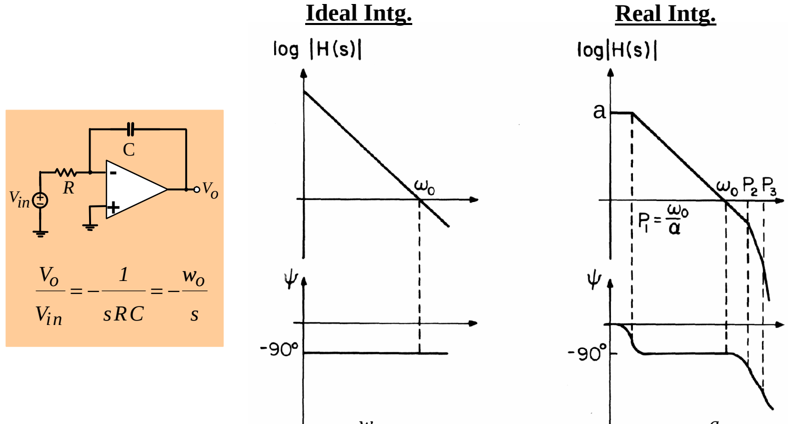
Summary Last Lecture
Integrator Based Ladder Filters
Transmission zeros



Summary Last Lecture
Seventh Order Differential Lowpass Filter
Including Transmission Zeros



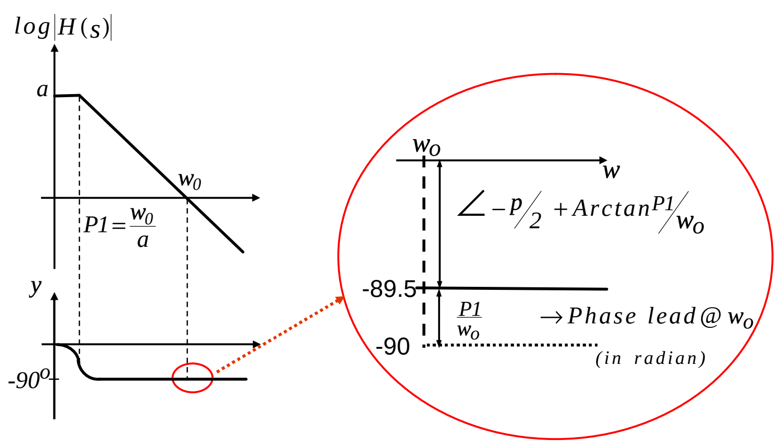
Summary Last Lecture
Effect of Integrator Non-Idealities on Filter Performance



$$H(s) = \frac{-w_o}{s}$$

$$H(s) \approx \frac{-a}{\left(1+s\frac{a}{w_o}\right)\left(1+\frac{s}{p2}\right)\left(1+\frac{s}{p3}\right)\cdots}$$

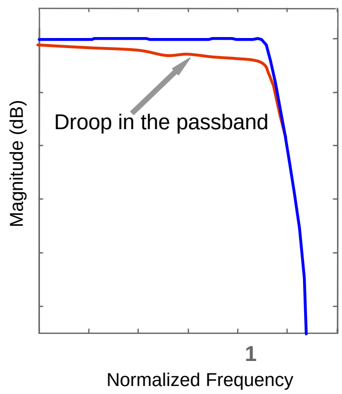
Effect of Integrator Finite DC Gain on Q

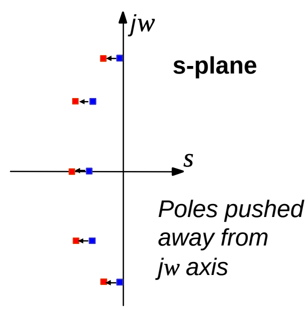


Example: $P1/\ \omega_0 = 1/a = 1/100$

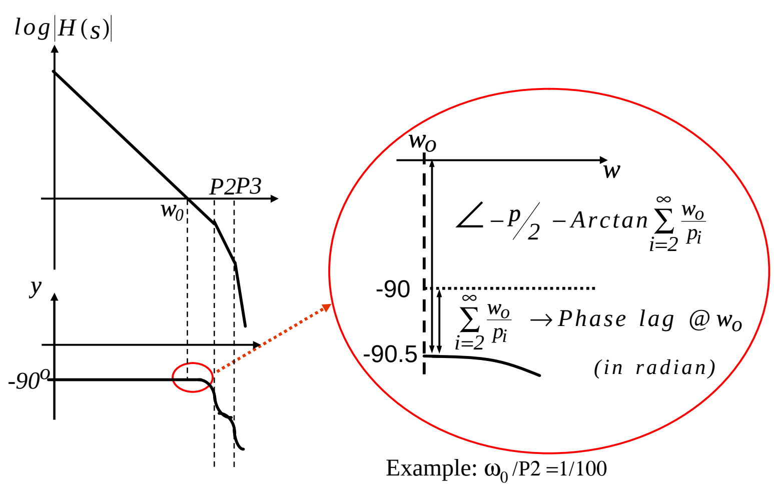
Effect of Integrator Finite DC Gain on Overall Filter Frequency Response

- Phase lead @ ω_0
  → Droop in the passband



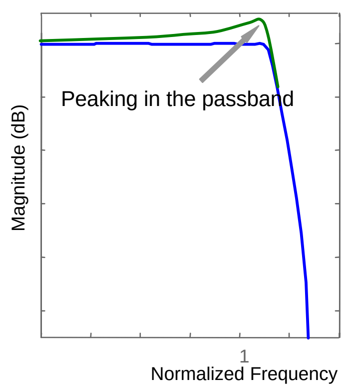


Effect of Integrator Non-Dominant Poles

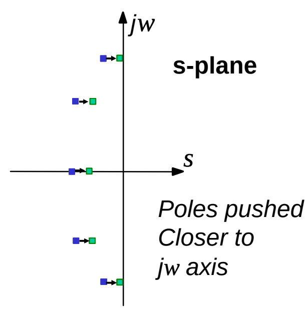


Example: ω_0 /P2 =1/100

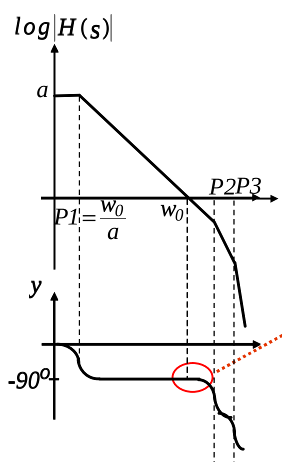
Effect of Integrator Non-Dominant Poles on Overall Filter Frequency Response

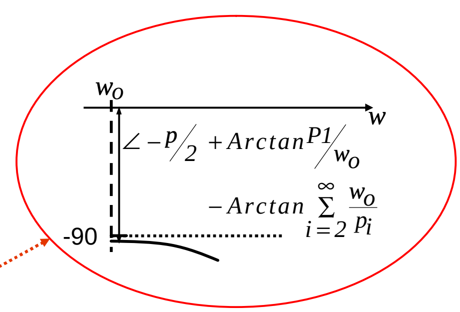


- Phase lag @ ω_0
  → Peaking in the passband
  In extreme cases could result in oscillation!



Effect of Integrator Non-Dominant Poles & Finite DC Gain on Q





Note that the two terms can cancel each other's effect

Summary

Effect of Integrator Non-Idealities on Q

$$Q_{ideal}^{intg.} = \infty$$

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

Phase lead @ w_0 — *Phase lag @ w_0*

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

Few Facts About Monolithic *Rs & Cs & Gms*

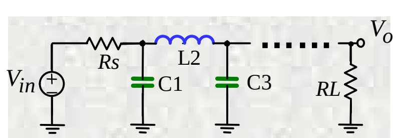
- Monolithic continuous-time filter critical frequency set by *RxC* or *GmxC*
- Absolute value of integrated *Rs & Cs & Gms* are quite variable
  - *Rs* vary due to doping and etching non-uniformities
    - Could vary by as much as ~+-*30 to 40%* due to process & temperature variations
  - *Cs* vary because of oxide thickness variations and etching inaccuracies
    - Could vary ~ +-*10 to15%*
  - *Gms* typically function of mobility, oxide thickness, current, device geometry …
    - Could vary > ~ +- *40%* or more with process & temp. & supply voltage

→Continuous-time filter critical frequency could vary by over +-*50%*

Few Facts About Monolithic Rs & Cs

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

Impact of Process Variations on Filter Characteristics



RLC Filters

Facts about RLC filters

- w_{-3dB} determined by absolute value *Ls & Cs*
- Shape of filter depends on ratios of normalized *Ls & Cs*

$$C_1^{RLC} = C_r \times C_1^{Norm} = \frac{C_1^{Norm}}{R^* \times w_{-3dB}}$$

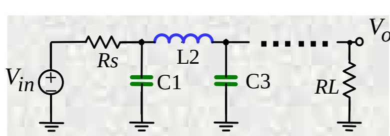
$$L_2^{RLC} = L_r \times L_2^{Norm} = \frac{L_2^{Norm} \times R^*}{w_{-3dB}}$$

Effect of Monolithic R & C Variations on Filter Characteristics

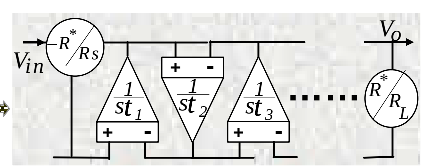
- Filter shape (whether Elliptic with 0.1dB Rpass or Butterworth..etc) is a function of *ratio* of normalized *Ls & Cs* in RLC filters
- Critical frequency (e.g. w_{-3dB}) function of absolute value of *Ls & Cs*
- Absolute value of integrated *Rs & Cs & Gms* are quite variable
- *Ratios* very accurate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

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

Impact of Process Variations on Filter Characteristics



RLC Filters



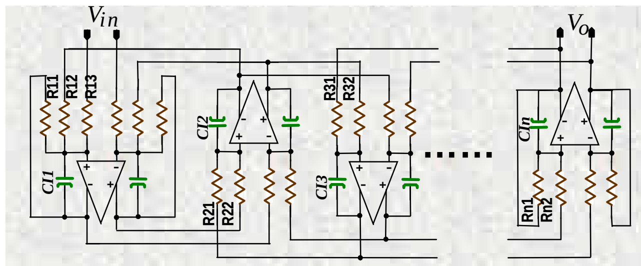
Integrator Based Filters

$$t_1 = C_1^{RLC}.R^* = \frac{C_1^{Norm}}{w_{-3dB}}$$

$$t_2 = \frac{L_2^{RLC}}{R^*} = \frac{L_2^{Norm}}{w_{-3dB}}$$

$$\frac{t_1}{t_2} = \frac{C_1^{Norm}}{L_2^{Norm}}$$

Impact of Process Variations on Filter Characteristics



$$t_1^{int\,g} = C_{I1}.R_1 = \frac{C_1^{Norm}}{w_{-3dB}}$$

$$t_2^{int\,g} = C_{I2}.R_2 = \frac{L_2^{Norm}}{w_{-3dB}}$$

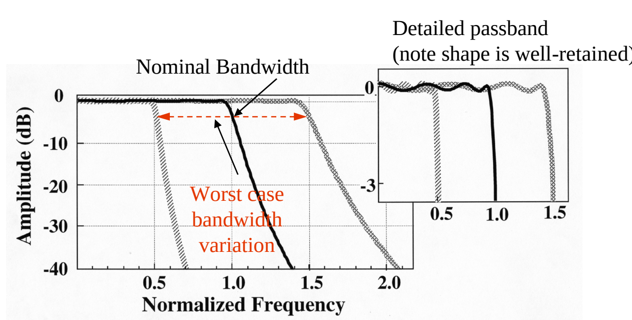
→ Variation in absolute value of integrated *Rs & Cs* causes change in critical freq. (w_{-3dB})

$$\frac{t_1^{int\,g}}{t_2^{int\,g}} = \frac{C_{I1}.R_1}{C_{I2}.R_2} = \frac{C_1^{Norm}}{L_2^{Norm}}$$

→ Since *Ratios* of Rs & Cs very accurate

→ Continuous time monolithic filters fully retain their shape

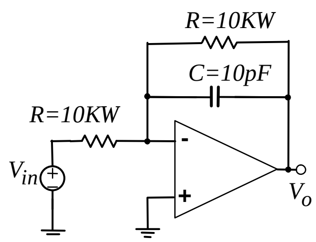
Example: LPF Worst Case Corner Frequency Variations



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

Tunable Opamp-RC Filters

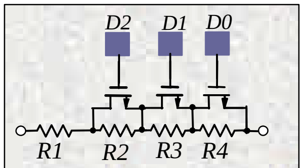
•Example: A 1st order Opamp-RC filter is designed to have a corner frequency of 1.6MHz
•Assuming process variations of:
  •*C* varies by +-10%
  •*R* varies by +-25%
•Build the filter in such a way that the corner frequency can be adjusted post-manufacturing.



Nominal R & C values for 1.6MHz corner frequency

Tunable Resistor

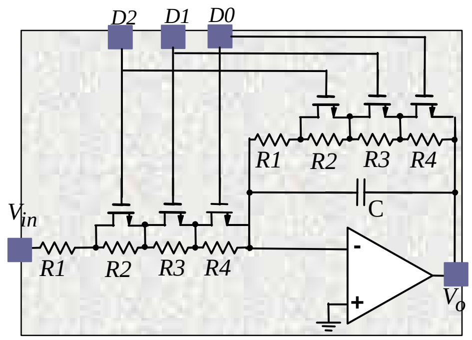
•Make provisions for either R or C to be adjustable (example adjustable R)
•Monolithic Rs can only be made adjustable in discrete steps (not continuous)
•Assuming expected process variations of:
  •Maximum C variations by +-10%→ Cmin=9pF, Cmax=11pF
  •Maximum R variations by +-25%→ Rmin=7.5K, Rmax=12.5K
    →Corner frequency varies by +-35%
•Assuming $n=3$bit (0 or 1) control signal for adjustment
  Rmax=Rnom(1+35%)=13.5KOHM
  Rmin=Rnom(1-35%)=6.5KOHM
  R1=Rmin
  R2 =(Rmax-Rmin)4/7=4K → $(2^{n-1}/(2^n-1))$
  R3 =(Rmax-Rmin)2/7=2K → $(2^{n-2}/(2^n-1))$
  R4 =(Rmax-Rmin)1/7=1K → $(2^{n-3}/(2^n-1))$
  Tuning resolution 10% →(1K/10K)
  If finer resolution needed add more bits & Rs



Variable Resister
MOS xtors act as switches

Tunable Opamp-RC Filter

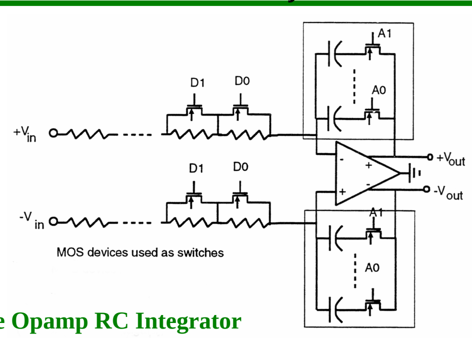
D2	D1	D0	Rtotal
1	1	1	6.5K
1	1	0	7.5K
1	0	1	8.5K
……	……	……	
……	……	……	
……	……	……	
0	0	0	13.5K



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

For higher order filters, all filter integrators tuned simultaneously

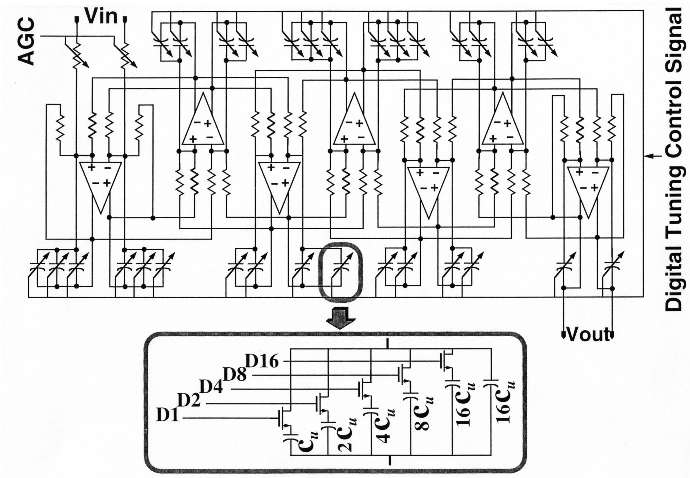
Tunable Opamp-RC Filters Summary



Tunable Opamp RC Integrator

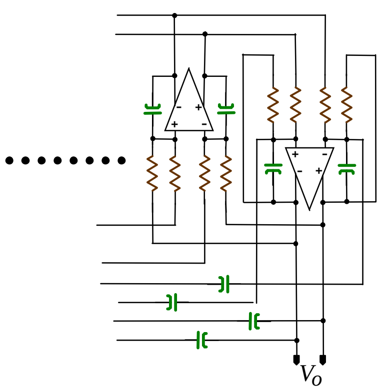
- Program Cs and/or Rs to freq. tune the filter
- All filter integrators tuned simultaneously
- Tuning in discrete steps & not continuous
- Tuning resolution limited
- Switch parasitic C & series R can affect the freq. response of the filter

Example: Tunable Low-Pass Opamp-RC Filter
Adjustable Capacitors

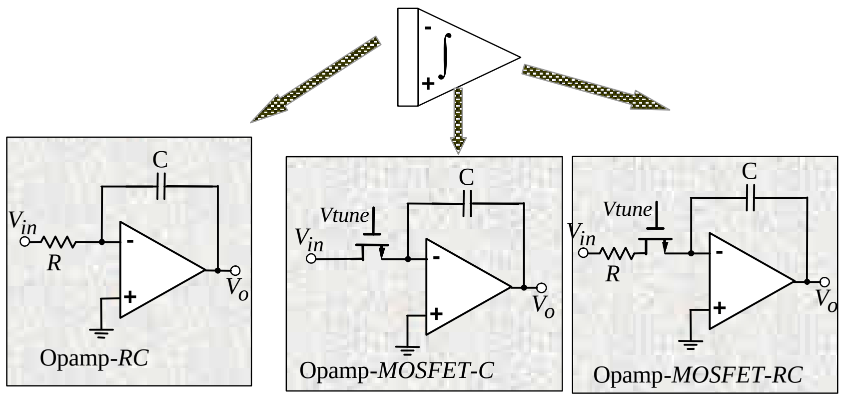


Opamp RC Filters

- Advantages
  - Since resistors are quite linear, linearity only a function of opamp linearity
    - → good linearity
- Disadvantages
  - Opamps have to drive resistive load, low output impedance is required
    - → High power consumption
  - Continuous tuning not possible
  - Tuning requires programmable Rs and/or Cs

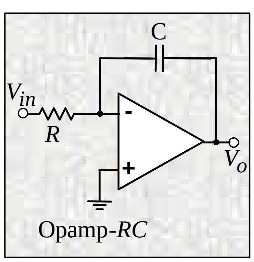


Integrator Implementation
Opamp-RC & Opamp-MOSFET-C & Opamp-MOSFET-RC

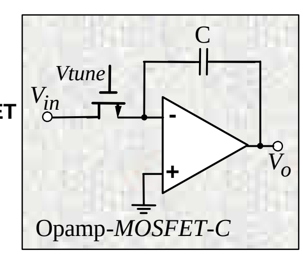


$$\frac{V_o}{V_{in}} = \frac{-w_o}{s} \quad \text{where} \quad w_o = \frac{1}{R_{eq}C}$$

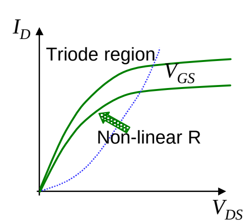
Use of MOSFETs as Resistors



R replaced by MOSFET
→Continuously variable resistor:



MOSFET IV characteristic:

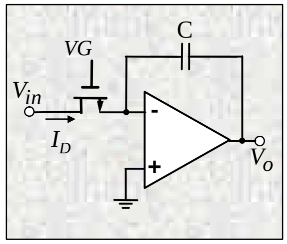


Use of MOSFETs as Resistors
Single-Ended Integrator

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

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

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



→**Tunable by varying *VG*:**

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

Use of MOSFETs as Resistors
Differential Integrator

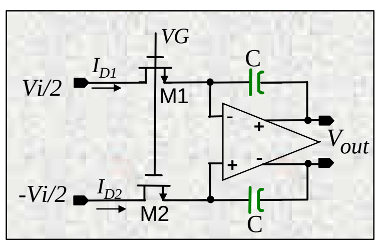
$$I_D = \mu C_{ox} \frac{W}{L}\left(V_{gs} - V_{th} - \frac{V_{ds}}{2}\right)V_{ds}$$

$$I_{D1} = \mu C_{ox} \frac{W}{L}\left(V_{gs} - V_{th} - \frac{V_i}{4}\right)\frac{V_i}{2}$$

$$I_{D2} = -\mu C_{ox} \frac{W}{L}\left(V_{gs} - V_{th} + \frac{V_i}{4}\right)\frac{V_i}{2}$$

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

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



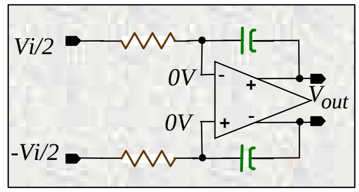
Opamp-*MOSFET-C*

- **Non-linear term cancelled!**
- **Admittance independent of Vi**

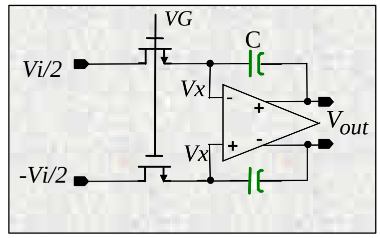
Problem: Threshold voltage dependence

MOSFET-C Integrator

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



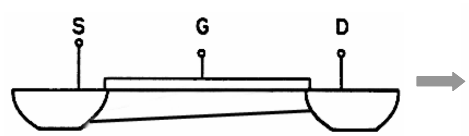
•For the MOSFET-C integrator, opamp input stays at the voltage Vx which is a function of 2nd order MOSFET non-linearities



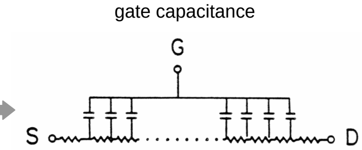
→ **Common-mode voltage sensitivity**

Use of MOSFET as Resistor Issues

MOS xtor operating in triode region
Cross section view



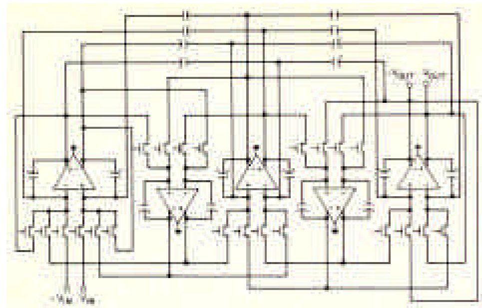
Distributed channel resistance & gate capacitance



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

Example: Opamp MOSFET-C Filter

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



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

Ref: Y. Tsividis, M.Banu, and J. Khoury, "Continuous-Time MOSFET-C Filters in VLSI", *IEEE Journal of Solid State Circuits* Vol. SC-21, No.1 Feb. 1986, pp. 15-30

Improved MOSFET-C Integrator

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

$$I_{D1} = \mu C_{ox}\frac{W}{L}\left(V_{gs1} - V_{th} - \frac{V_i}{4}\right)\frac{V_i}{2}$$

$$I_{D3} = -\mu C_{ox}\frac{W}{L}\left(V_{gs2} - V_{th} + \frac{V_i}{4}\right)\frac{V_i}{2}$$

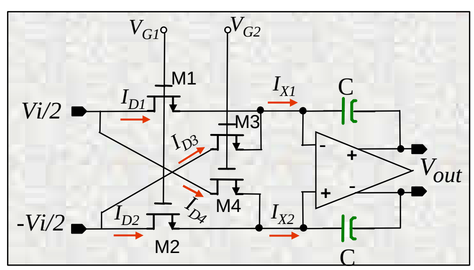
$$I_{X1} = I_{D1} + I_{D3}$$

$$= \mu C_{ox}\frac{W}{L}\left(V_{gs1} - V_{gs2} - \frac{V_i}{2}\right)\frac{V_i}{2}$$

$$I_{X2} = \mu C_{ox}\frac{W}{L}\left(V_{gs2} - V_{gs1} - \frac{V_i}{2}\right)\frac{V_i}{2}$$

$$I_{X1} - I_{X2} = \mu C_{ox}\frac{W}{L}\left(V_{gs1} - V_{gs2}\right)V_i$$

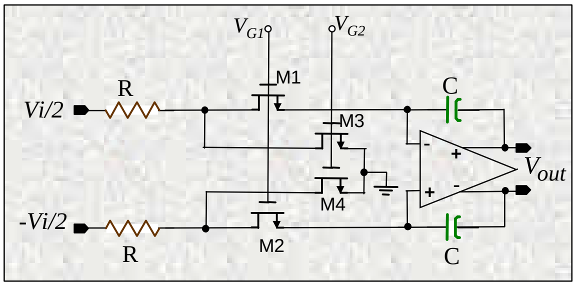
$$G = \frac{\partial\left(I_{D1} - I_{D2}\right)}{\partial V_i} = \mu C_{ox}\frac{W}{L}\left(V_{gs1} - V_{gs2}\right)$$



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

Ref: Z. Czarnul, "Modification of the Banu-Tsividis Continuous-Time Integrator Structure," *IEEE Transactions on Circuits and Systems,* Vol. CAS-33, No. 7, pp. 714-716, July 1986.

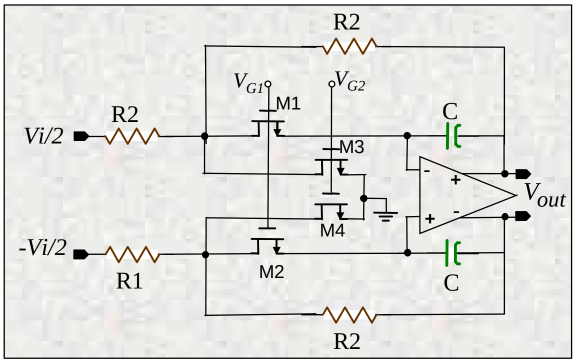
R-MOSFET-C Integrator



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

Ref: U-K Moon, and B-S Song, "Design of a Low-Distortion 22-kHz Fifth Order Bessel Filter," *IEEE Journal of Solid State Circuits*, Vol. 28, No. 12, pp. 1254-1264, Dec. 1993.

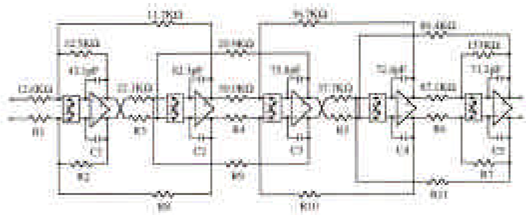
R-MOSFET-C Lossy Integrator



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

Ref: U-K Moon, and B-S Song, "Design of a Low-Distortion 22-kHz Fifth Order Bessel Filter," *IEEE Journal of Solid State Circuits*, Vol. 28, No. 12, pp. 1254-1264, Dec. 1993.

Example:
Opamp MOSFET-RC Filter



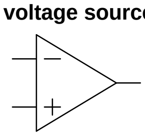
5th Order Bessel MOSFET-RC LPF -22kHz bandwidth
THD → -90dB for 4Vp-p 2kHz input signal

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

Ref: U-K Moon, and B-S Song, "Design of a Low-Distortion 22-kHz Fifth Order Bessel Filter," *IEEE Journal of Solid State Circuits*, Vol. 28, No. 12, pp. 1254-1264, Dec. 1993.

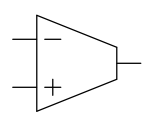
Operational Amplifiers (Opamps) versus Operational Transconductance Amplifiers (OTA)

Opamp
Voltage controlled voltage source



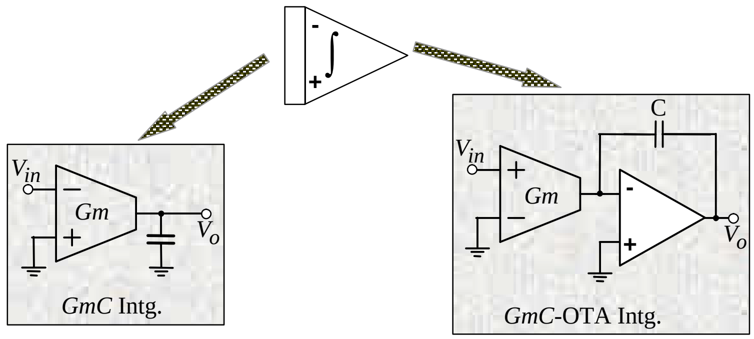
- Low output impedance
- Output in the form of voltage
- Can drive R-loads
- Good for RC filters, OK for SC filters
- Extra buffer adds complexity, power dissipation

OTA
Voltage controlled current source



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

Integrator Implementation
Gm-C & Opamp-Gm-C



$$\frac{V_o}{V_{in}} = \frac{-w_o}{s} \quad \text{where} \quad w_o = \frac{G_m}{C}$$

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

- MOSFET in saturation region:

$$I_d = \frac{mC_{ox}}{2}\frac{W}{L}\left(V_{gs} - V_{th}\right)^2$$

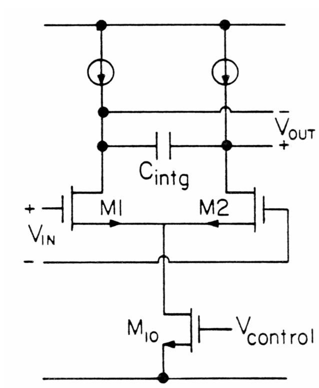
- Gm is given by:

$$g_m = \frac{\partial I_d}{\partial V_{gs}} = mC_{ox}\frac{W}{L}\left(V_{gs} - V_{th}\right)$$

$$= 2\frac{I_d}{\left(V_{gs} - V_{th}\right)}$$

$$= 2\left(\frac{1}{2} mC_{ox}\frac{W}{L} I_d\right)^{1/2}$$

Id varied via Vcontrol
→ *gm* tunable via *Vcontrol*



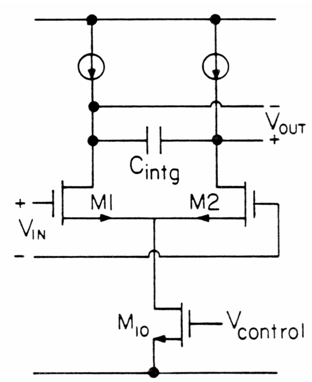
Gm-C Filters
Simplest Form of CMOS Gm

- Integrator behavior:

$$\frac{V_o}{V_{in}} = \frac{-w_o}{s}$$

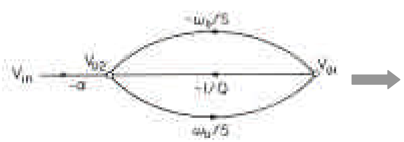
$$\text{where} \quad w_o = \frac{g_m^{M1,2}}{2 \times C_{intg}}$$

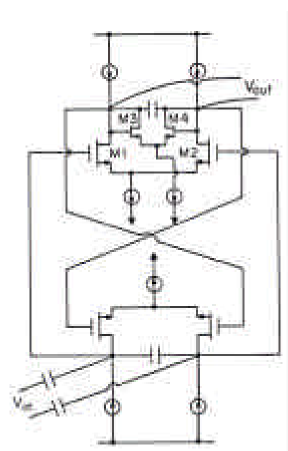
- Critical frequency continuously tunable via $V_{control}$



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.

Second Order Gm-C Filter





- Simple design
- Tunable
- Q function of device ratios:

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

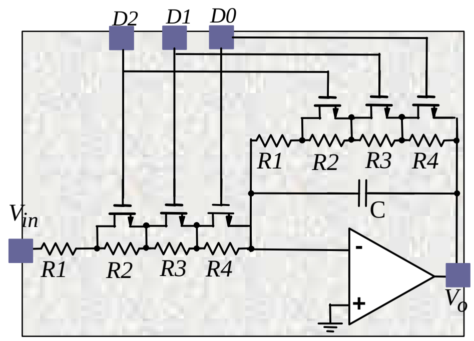
Filter Frequency Tuning Techniques

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

Example: Tunable Opamp-RC Filter

Post manufacturing:
•Usually at wafer-sort tuning performed

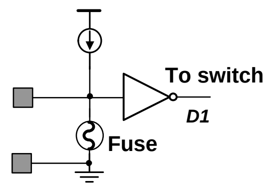
•Measure -3dB frequency
  •If frequency too high decrement D to D-1
  •If frequency too low increment D to D+1
  •If frequency within 10% of the desired corner freq. stop



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

Trimming

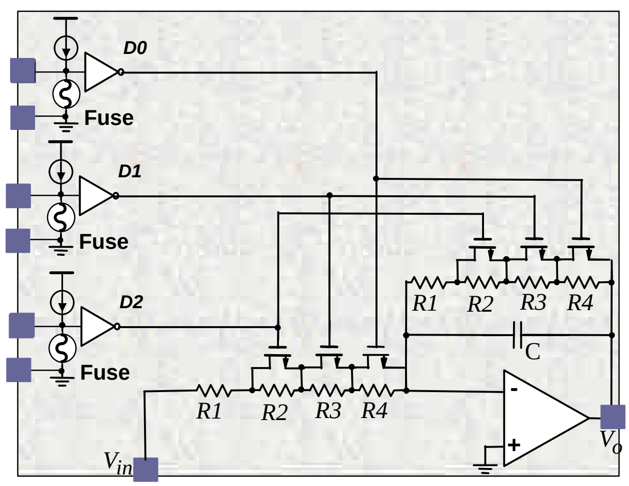
- Component trimming
  - Build fuses on-chip,
    - Based on measurements @ wafer-sort blow fuses by applying high current to the fuse
      - Expensive
      - Fuse regrowth problems!
      - Does not account for temp. variations & aging
  - Laser trimming
    - Trim components or cut fuses by laser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



Fuse not blown → D1=1
Fuse blown → D1=0

Example:Tunable/Trimmable Opamp-RC Filter

D2	D1	D0	Rtotal
1	1	1	6.5K
1	1	0	7.5K
1	0	1	8.5K
…	…	…	
…	…	…	
…	…	…	
0	0	0	13.5K



Automatic Frequency Tuning

- By adding additional circuitry to the main filter circuit
  - Have the filter critical frequency automatically tuned
    - → Expensive trimming avoided
    - → Accounts for critical frequency variations due to temp. and voltage changes

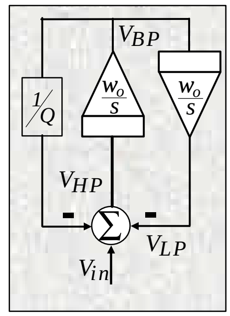
Master-Slave Automatic Frequency Tuning

- Following facts used in this scheme:
  - Use a replica (master) of the main filter (called the slave) in the tuning circuitry
  - Place the replica in close proximity of the main filter
  - Use the tuning signal generated to tune the replica, to also tune the main filter
  - In the literature, this scheme is called master-slave tuning!

Master-Slave Frequency Tuning Reference Filter (VCF)

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

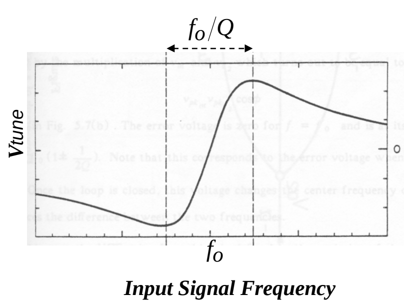
$$\frac{V_{LP}}{V_{in}} = \frac{1}{\frac{s^2}{w_o^2} + \frac{s}{Qw_o} + 1} \quad @ \; w = w_o \;\; f = -90^o$$

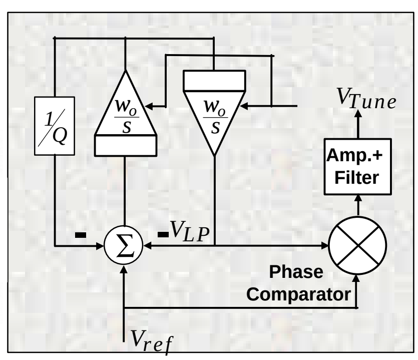


- Apply a sinusoid at the desired f_o
- Compare the LP output phase to the input
- Based on the phase difference
  - Increase or decrease filter critical freq.

Master-Slave Frequency Tuning Reference Filter (VCF)

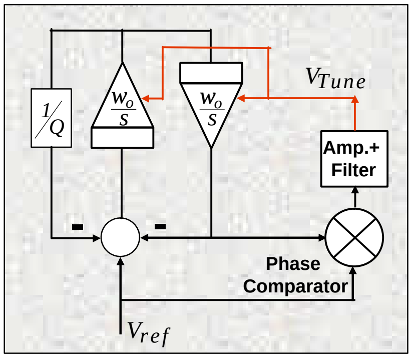
$$V_{tune} \approx -K \times V_{ref}^{rms} \times V_{LP}^{rms} \times \cos f$$



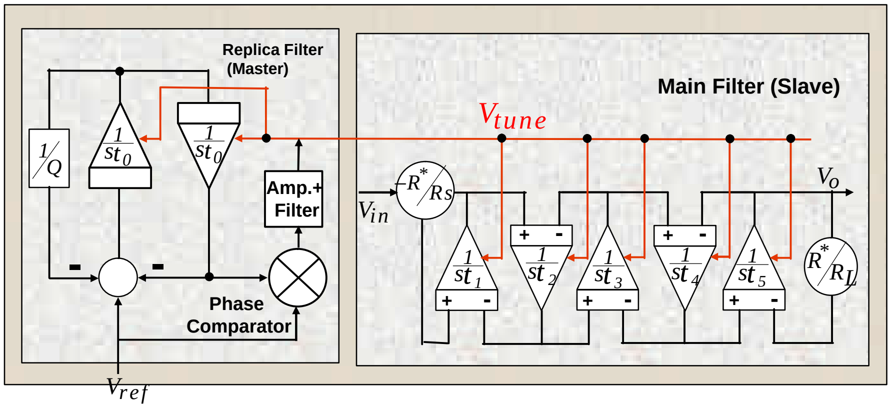


Master-Slave Frequency Tuning Reference Filter (VCF)

- By closing the loop, feedback tends to drive the error voltage to zero.
  → Locks *fo*, the critical frequency of the filter to the accurate reference frequency
- Typically the reference frequency is provided by a crystal oscillator with accuracies in the order of few ppm



Master-Slave Frequency Tuning Reference Filter (VCF)



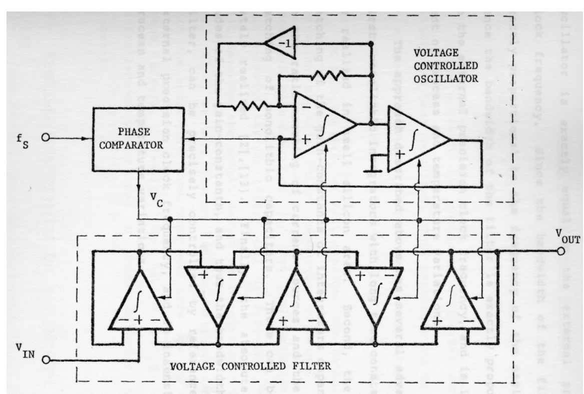
Ref: H. Khorramabadi and P.R. Gray, "High Frequency CMOS continuous-time filters," IEEE Journal of Solid-State Circuits, Vol.-SC-19, No. 6, pp.939-948, Dec. 1984.

Master-Slave Frequency Tuning Reference Filter (VCF)

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

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

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

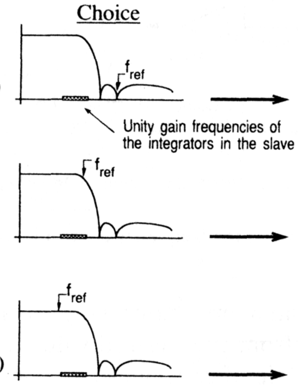


Ref: K.S. Tan and P.R. Gray, "Fully integrated analog filters using bipolar FET technology," IEEE, J. Solid-State Circuits, vol. SC-13, no.6, pp. 814-821, December 1978..

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

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

Master-Slave Frequency Tuning
Choice of Ref. Frequency wrt Feedthrough Immunity



Choice	Matching	Immunity to reference freq. feedthrough
(a)	Worst	Best
(b)	Best	Worst
(c)	Medium	Medium

Ref: V. Gopinathan, et. al, "Design Considerations for High-Frequency Continuous-Time Filters and Implementation of an Antialiasing Filter for Digital Video," *IEEE JSSC*, Vol. SC-25, no. 6 pp. 1368-1378, Dec. 1990.