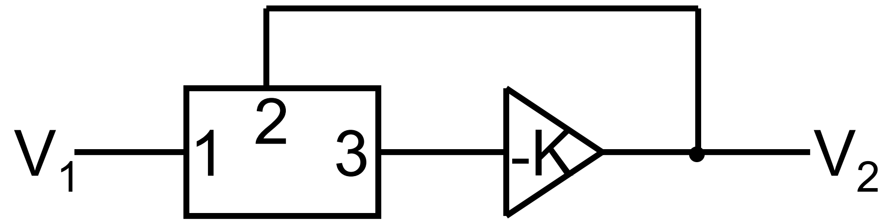
# Building Blocks of Active Filter

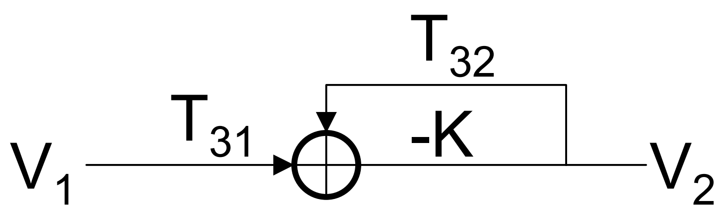
- LC filter
  - high Q, i.e. steep transfer function, is realizable
  - large L for low frequency application.
- RC filter
  - can have complex zeros.
  - poles are restricted to the negative real axis
    => small Q
    => steep slopes are difficult to realize unless very high order.
- Active filter
  - RC + gain element
  - can have complex poles with high Q factors.

# Building Blocks of Active Filter(Cont.)

Example-1



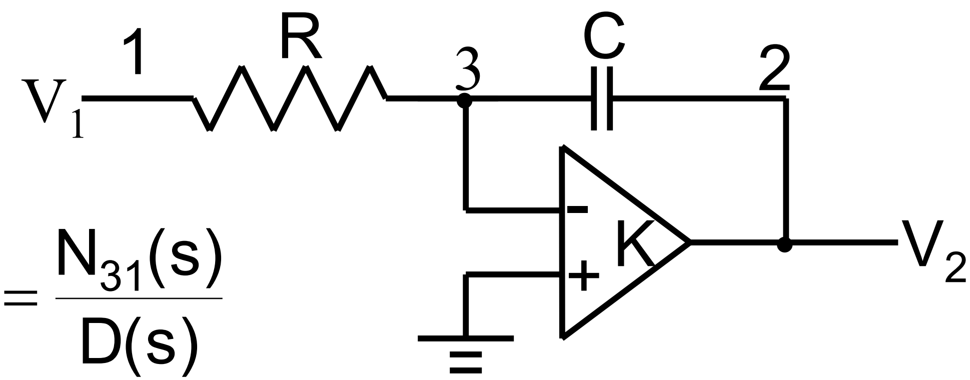

- $T_{3i}(s) = \frac{V_3}{V_i} = \frac{N_{3i}(s)}{D(s)}$
- Signal flow graph




- $H(s) = \frac{V_2}{V_1} = T_{31}(s)\frac{-k}{1+KT_{32}(s)} = -\frac{N_{31}(s)}{N_{32}(s)+\frac{D(s)}{K}}$
- if K>>1, then $H(s) \approx \frac{N_{31}(s)}{N_{32}(S)}$

# Building Blocks of Active Filter(Cont.)

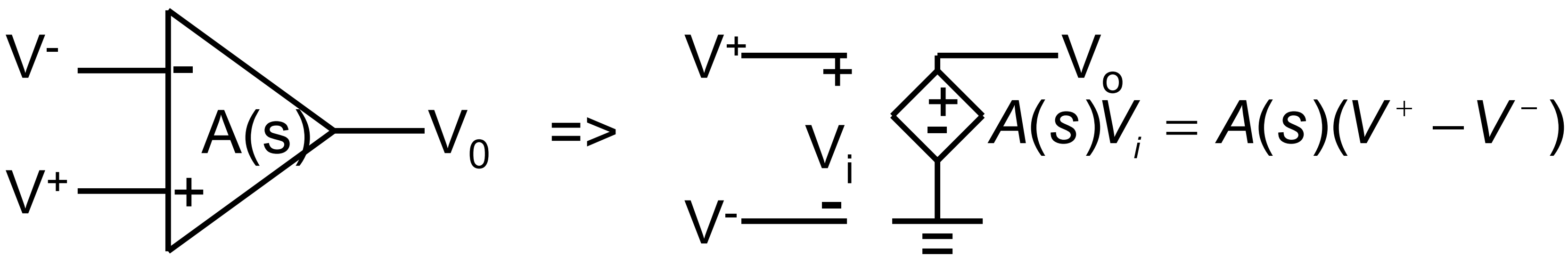
Example-2



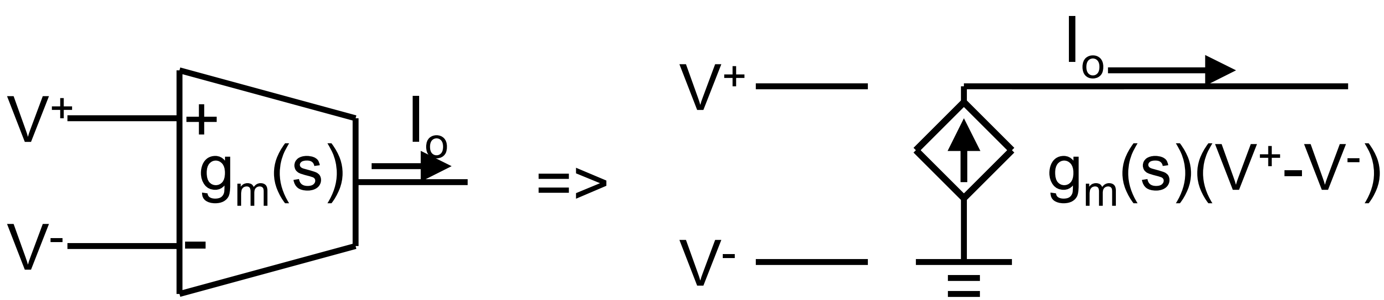

- $T_{31}(s) = \dfrac{\frac{1}{sC}}{R + \frac{1}{sC}} = \dfrac{N_{31}(s)}{D(s)}$
- $T_{32}(s) = \dfrac{R}{R + \frac{1}{sC}} = \dfrac{N_{32}(s)}{D(s)}$
- $H(s) = \dfrac{\frac{1}{sC}}{R + \dfrac{R + \frac{1}{sC}}{K}} = \dfrac{1}{sCR + \dfrac{1 + sCR}{K}}$
- if K>>1, then $H(s) \approx \dfrac{1}{sCR}$

# Building Blocks of Active Filter(Cont.)

- If N31(s) and N32(s) have complex zeros, then
  H(s) can have both complex zeros and poles
- Conventionally, the gain element is an operational amplifier.




- Recent trend, transconductance amplifier is used as active element instead of OPAMP

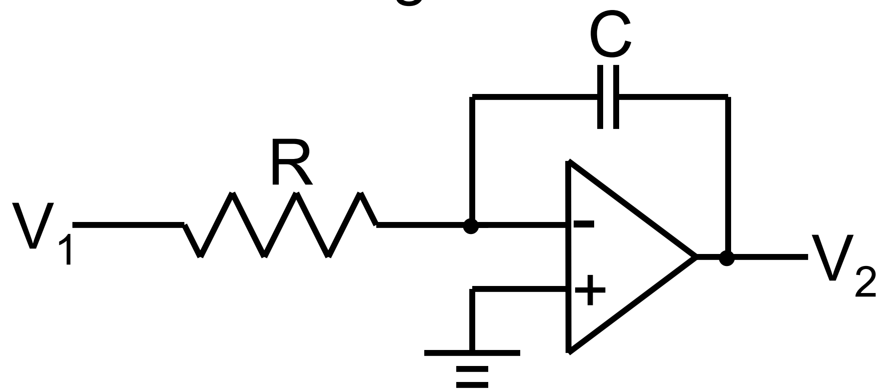


# Active RC Integrator Based on OPAMP

- Ideal OPAMP
  - wideband
  - high gain
  - input virtual ground if negative feedback is applied
  - large linear range
  - many others
- Real OPAMP
  - Limited bandwidth =>limited frequency response
  - limited gain => virtual ground is not necessary
  - limited linear range =>
    1. nonlinearity occurs
    2. harmonic distortion(increase)
    3. noise(increase)
    4. resolution(down)
  - many others

# Active RC Integrator Based on OPAMP(Cont.)

Example: actice RC integrator based on ideal OPAMP




– Time domain

$$V_2(t) = -\frac{1}{RC}\int_{-\infty}^{t} V_1(\lambda)\,d\lambda$$

– Frequency domain

$$V_2(s) = -\frac{1}{sRC}V_1(s)$$

$$\Rightarrow H(s) = -\frac{1}{sRC} \Rightarrow \text{ ideal integrator}$$

$\Rightarrow$ perfect linear, i.e. no error

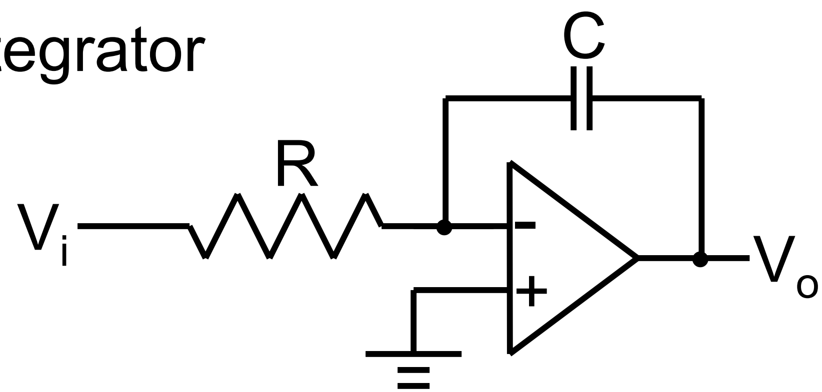
# RC Active Integrator Based on Nonideal OPAMP

- Transfer function of OPAMP (one-pole is assumed)

$$A(s) = \frac{A_o}{1+\frac{s}{s_p}} = \frac{A_o s_p}{s+s_p} = \frac{w_u}{s+s_p}$$ ; where $s_p$ is pole location

$w_u$ is gain-bandwidth product

- Transfer function of real integrator




$$V^-(s) = \frac{V_o(s)}{A(s)}$$

$$A_v(s) = \frac{V_o(s)}{V_i(s)} = \frac{1}{sCR + \frac{1+sCR}{A(s)}} \quad \left(= \frac{N_{31}(s)}{N_{32}(s) + \frac{D(s)}{A(s)}}\right)$$

$$= \frac{1}{\frac{RC}{w_u}[s^2 + s(w_u + s_p + \frac{1}{RC}) + \frac{s_p}{RC}]}$$

# RC Active Integrator Based on Nonideal OPAMP(Cont.)

Assuming $R_i \to \infty$, $R_o = 0$, $A_o >> 1$ and $w_u >> \frac{1}{RC}$

$\Rightarrow w_u = A_o s_p >> s_p$

$$w_u >> \frac{1}{RC}$$

$$w_u >> \frac{1}{A_o RC}$$

$$\Rightarrow w_u + s_p + \frac{1}{RC} \approx w_u \approx w_u + \frac{1}{A_o RC}$$

$$\Rightarrow \frac{V_o(s)}{V_i(s)} = -\frac{1}{\frac{RC}{w_u}[s^2 + s(w_u + \frac{1}{A_o RC}) + \frac{s_p}{RC}]}$$

$$= -\frac{1}{\frac{RC}{w_u}(s + \frac{1}{A_o RC})(s + w_u)}$$

$\Rightarrow$ poles at $-\frac{1}{A_o RC}$ and $-w_u$

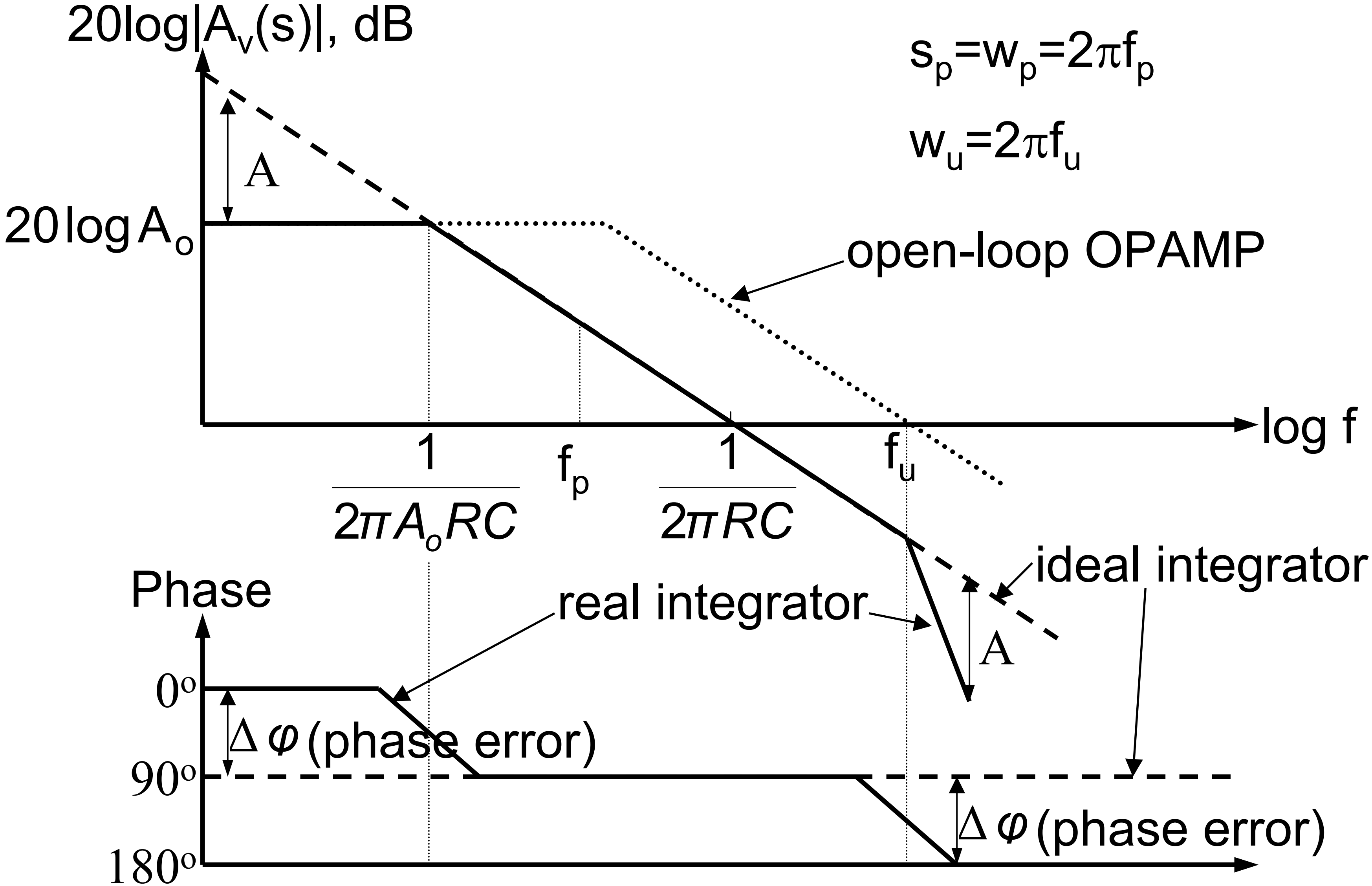
# RC Active Integrator Based on Nonideal OPAMP(Cont.)

– $\Delta A$(gain error): $f < \dfrac{1}{2\pi A_o RC}$ & $f > f_u$ ; where f is signal frequency

– $\Delta \varphi$(phase error): $f < \dfrac{10}{2\pi A_o RC}$ & $f > \dfrac{f_u}{10}$

– For $\dfrac{10}{2\pi A_o RC} < f < \dfrac{fu}{10}$ , the nonideal integrator behaves like an ideal integrator

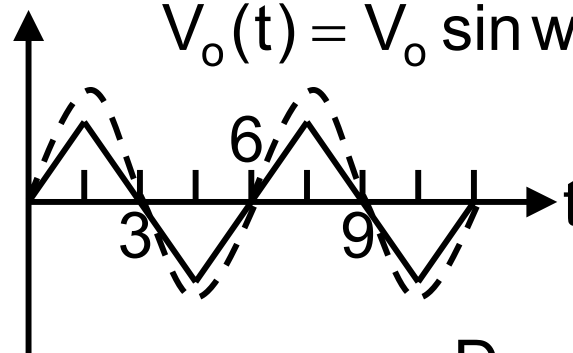
# RC Active Integrator Based on Nonideal OPAMP(Cont.)

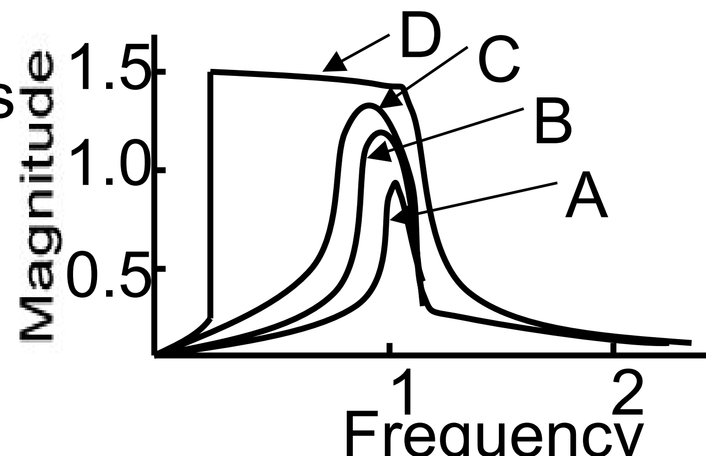


# Nonidealities of OPAMP

- Frequency - dependent gain
  - finite gain
  - finite bandwidth
  - nonideal integrator
- Input and output impedances
- Slew rate limitation => maximum undistorted output signals
  - output signal of OPAMP
  - $\left.\frac{dV_o(t)}{dt}\right|_{max} = V_o \omega_{max} < SR$
  - signal distortion caused by slew-rate limiting
  - shape of a 2nd-order bandpass filter with Q=10 for nominal design (no slewing: A) and for continuously increasing signal amplitudes(B, C, D)

$V_o(t) = V_o \sin wt$

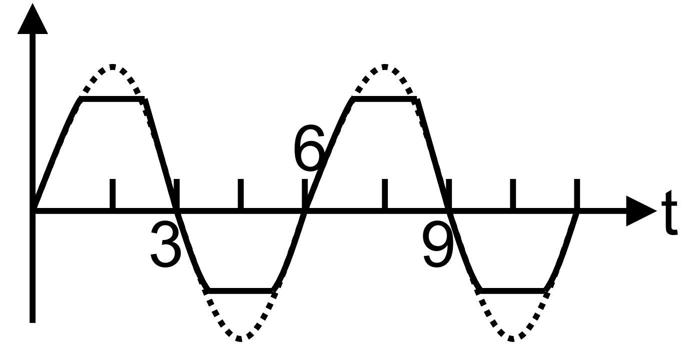


# Nonidealities of OPAMP(Cont.)

- Output voltage swing

$$V_o < \left|V_{power\,supply}\right|$$




- Noise

$$\left|e_{on}\right| = \sqrt{\sum_{i=1}^{m} \left|e_{ni}\right|^2 \left|H_{ni}(jw)\right|^2}$$ where $H_{ni}$ is the noise trnasfer function from the *i*th noise source

$$DR = 20\log\left|\frac{maximum\_undistorted\_rms\_signal}{e_{on}}\right|$$ dB ; typically, 70dB~100dB

(In active RC filters, supply noise and the thermal noise of resistors should be considered)

- Offset voltage, offset current, and DC bias currents
- Common mode rejection

# Transconductance Amplifiers

- Currently, most filters are built with VCVS, i.e. OPAMP
  - frequency-dependent gain of OPAMP

    => serious deviations on filter behavior for high frequency

    => precludes active filter applications significantly above the video range(<50MHz)
- Transconductance amplifiers(TA)
  - generally have significantly higher bandwidth than OPAMPs
  - generally provides simpler circuitry for integration and easy methods for electronic tuning by changing a bias current
  - analog filters built with transconductances usually require fewer components than their OPAMP counterparts

# Transconductance Amplifiers(Cont.)

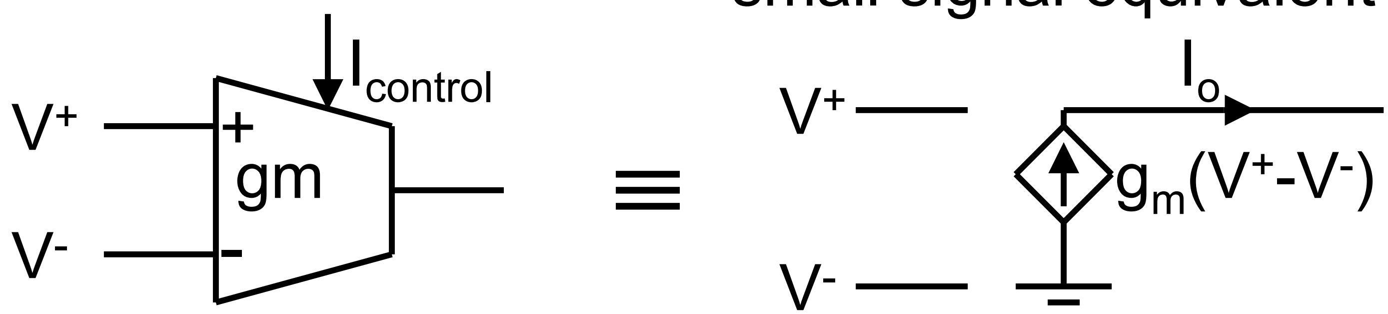
- Ideal operational transconductance amplifier(OTA)
  – a coltage-controlled current source(VCCS)

$$I_o = g_m(V^+ - V^-)$$

  – Input and output impedances are both infinite

(voltage) (current)

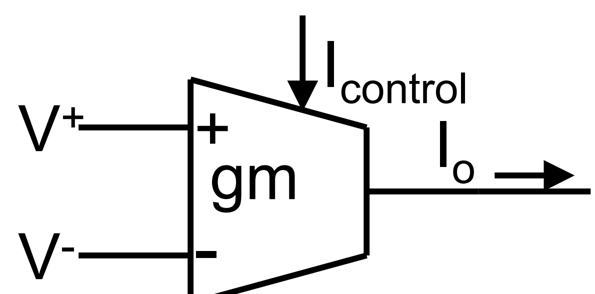
small signal equivalent circuit




- Tunable OTA

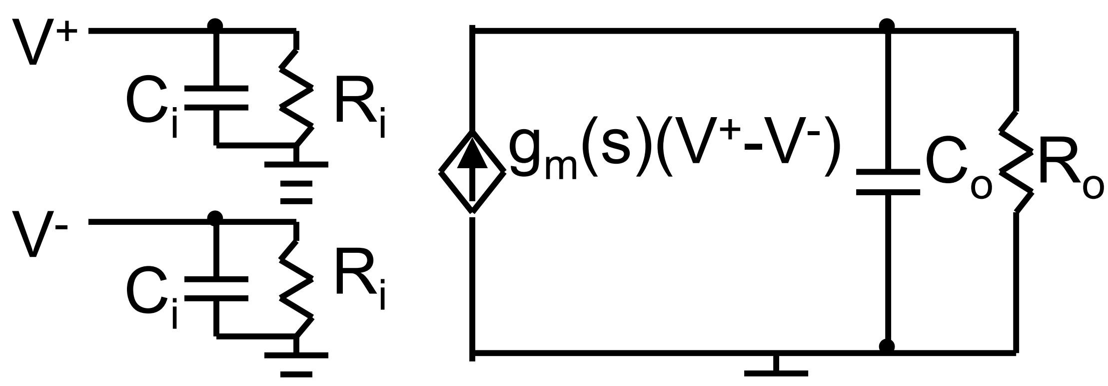
$$g_m = KI_{control}$$

$$\Rightarrow I_o = KI_{control}(V^+ - V^-)$$



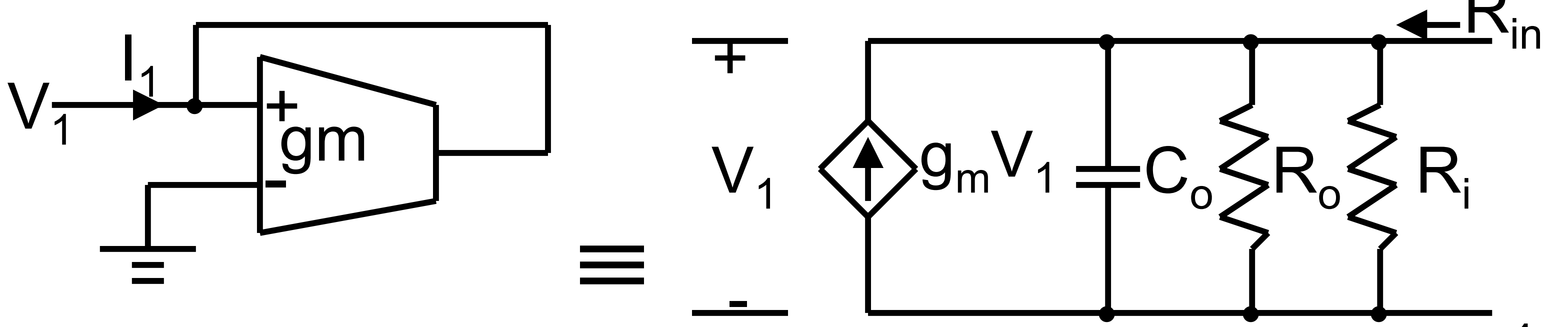
# Transconductance Amplifiers

- Real OTA
  - finite input impedance
  - finite output impedance
  - limited input swing for linear operation(e.g. $\leq 20mV$ )
  - limited output swing
  - input current (e.g. bipolar OTA)
  - frequency dependent $g_m$
  - small signal model with frequency-dependent gm and finite input and output impedances



# Transcondductance Amplifiers(Cont.)

- Simulation of a grounded resistor




$$R_{in} = (g_m)^{-1} // R_o // R_i \approx \frac{1}{g_m}$$

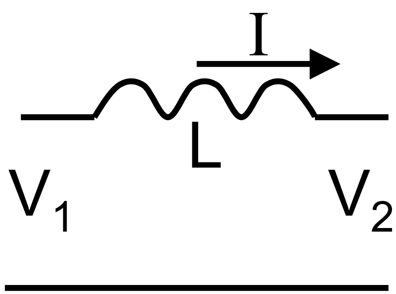
- Example 4-3
- Example 4-4

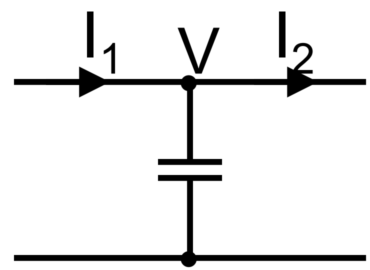
$g_m \uparrow$ Q error$\downarrow$ (To reduce $\frac{G_o}{g_m} \Rightarrow \frac{G_o}{g_m} << 1 \Rightarrow \frac{1}{g_m R_o} << 1$
$\Rightarrow g_m R_o >> 1$)

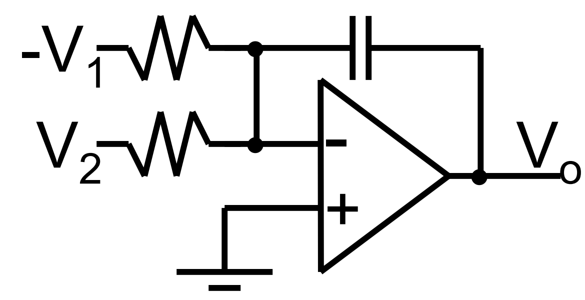
i.e. voltage gain of OTA should be large in this example

# Active Building Blocks using OPAMPs

- OPAMP can perform
  - summation
  - amplification
  - operational simulation of L & C



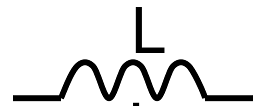

$$I = \frac{1}{sL}(V_1 - V_2)$$

$$V = \frac{1}{sC}(I_1 - I_2)$$

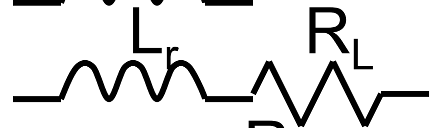
$$V_o = \frac{1}{sCR}(V_1 - V_2)$$

# Active Building Blocks using OPAMPs(Cont.)

- Inductor
  - ideal inductor




  - lossy inductor




$$jwL = jwL_r + R_L = jwL_r(1 - j\frac{R_L}{wL_r}) = jwL_r(1 - j\frac{1}{Q_L(w)})$$

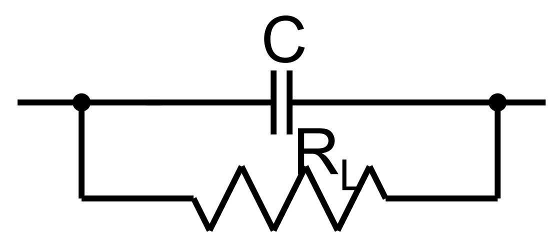
where $Q_L(w) = \frac{wL_r}{R_L}$ is the inductor quality factor evaluated at some appropriate frequencies

  - integration phase error $\Delta\phi_I(w)$

$$\phi_I(w) = -\frac{\pi}{2} + \mathrm{Tan}^{-1}\frac{1}{Q_L(w)} = -\frac{\pi}{2} + \Delta\phi_I(w)$$

$$\Delta\Phi_I(w) = \mathrm{Tan}^{-1}\frac{R_L}{wL_r}$$

- Capacitor




Similarly $Q_c(w) = wCR_L$

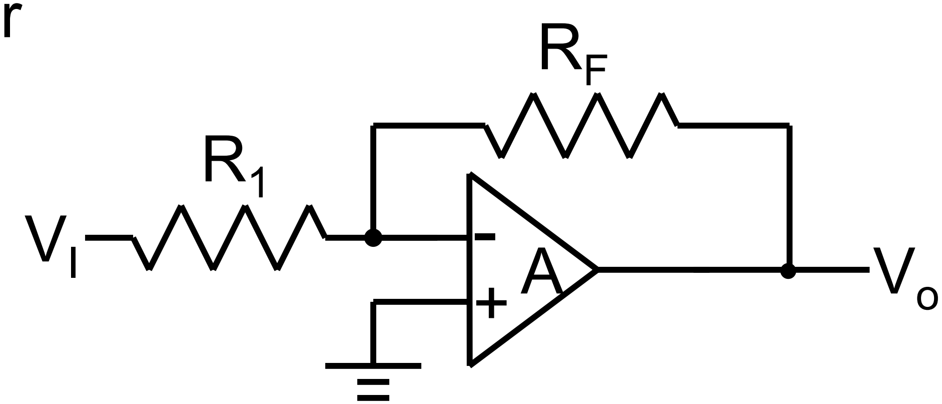
- Quality factor

$$H(s) = \frac{1}{j\mathrm{Im}(w) + \mathrm{Re}(w)} \Rightarrow Q_I = \frac{\mathrm{Im}(w)}{\mathrm{Re}(w)}$$

# Summers and Related Circuits

- Must take into account the noidealities of OPAMPs in high performance system
  - e.g. am inverting amplifier

$$\frac{V_o}{V_I} = -\frac{R_F}{R_1}\frac{1}{1+\dfrac{1}{A(s)(1+\dfrac{R_F}{R_1})}}$$



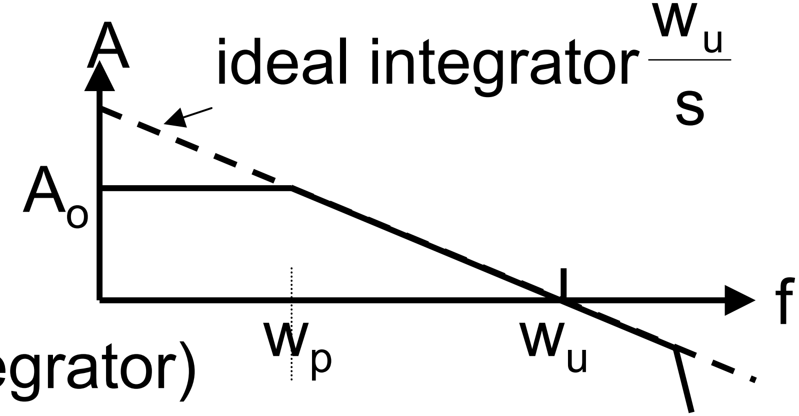

=> gain error and phase error occur if A(s) is nonideal

  - noninverting amplifier
  - summer
  - unity-gain buffer
  - others

# Integrators

- OPAMP can be approximated as an ideal integrator

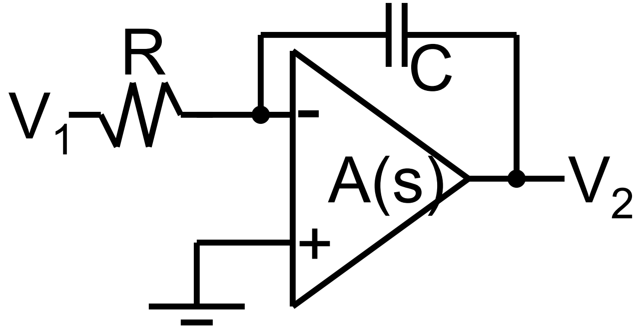
$$\Rightarrow H(s) = \frac{A_o}{1+\frac{s}{s_p}} = \frac{A_o s_p}{s+s_p} = \frac{w_u}{s+s_p}$$




For w>$w_p$ $\Rightarrow H(s) \approx \frac{w_u}{s}$ (ideal integrator)

  – can be used to design “active R” filter
  – example 4-5

- Inverting integrators




$$\frac{V_2}{V_1} = -\frac{1}{sCR + \frac{1+sCR}{A(s)}}$$

$$= -\frac{1}{sCR}\frac{1}{1+\frac{1}{A(s)}(1+\frac{1}{sCR})}$$

# Integrators(Cont.)

– Approximation: $A(s) = \frac{w_u}{s}$ and $w_u CR >> 1$ (i.e. $w_u >> \frac{1}{RC}$)

$$\frac{V_2}{V_1} = -\frac{1}{sCR(1+\frac{2}{w_u}+\frac{1}{w_u CR})} \approx -\frac{1}{sCR}\frac{1}{1+\frac{s}{1+\frac{s}{w_u}}}$$

$$= -\frac{1}{jwCR - \frac{w^2 CR}{w_u}} = -\frac{1}{jwCR - \frac{w^2 CR}{w_u}}$$

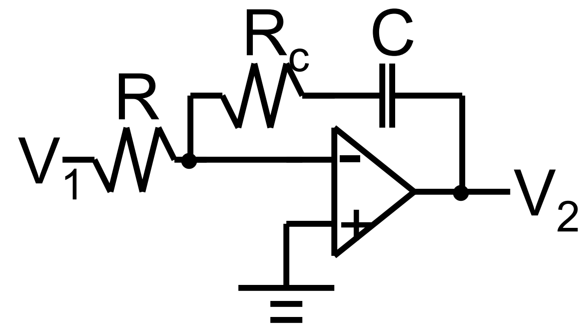
$$\Rightarrow \begin{cases} \text{phase error} & \Delta\varphi = -\mathrm{Tan}^{-1}\frac{w}{w_u} = -\mathrm{Tan}^{-1}\frac{1}{|A(jw)|} \\ \text{quality factor} & Q_I = \frac{\mathrm{Im}(w)}{\mathrm{Re}(w)} \approx -\frac{w_u}{w} = -|A(jw)| \end{cases}$$

(integrator Q is negative and given by OPAMP gain)

# Integrator(Cont.)

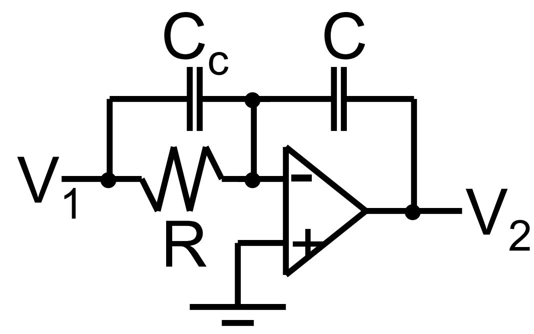
- High-Q integrator
  - passive compensation methods

(a) $\frac{V_2}{V_1} = -\frac{1}{sCR}$ if $R_c = \frac{1}{w_u C}$




(b) $\frac{V_2}{V_1} = -\frac{1}{sCR}$ if $C_c = \frac{1}{w_u C}$

and $w_u >> \frac{1}{RC}$

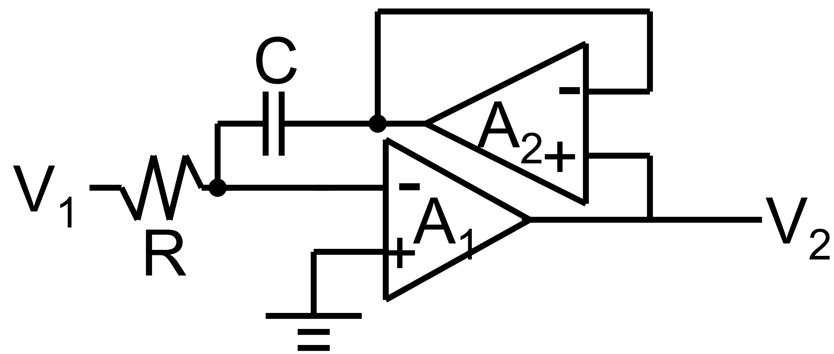



=> $R_c$ or $C_c$ must be very accurate

(In ICs, it is not easy => active compensation for ICs)

# Integrators(Cont.)

– active compensation

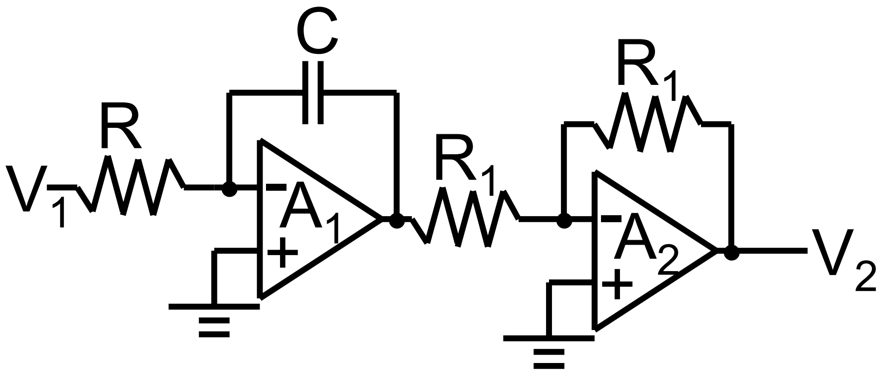



$$Q_I \approx -\left(\frac{w_u}{w}\right)^3 = -\left|A(jw)\right|^3 \quad \text{if } w_{u1} = w_{u2} = w_u$$

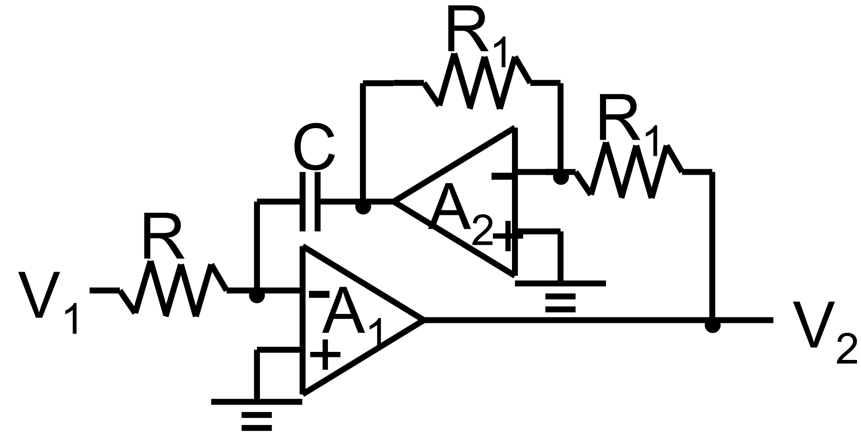
$$Q_I \approx \frac{w_{u2}}{w}\frac{1}{1-\frac{w_{u2}}{w_{u1}}} \approx \frac{A_2(jw)}{1-\frac{w_{u2}}{w_{u1}}} \quad \text{if } w_{u1} \neq w_{u2}$$

# Noninverting Integrators

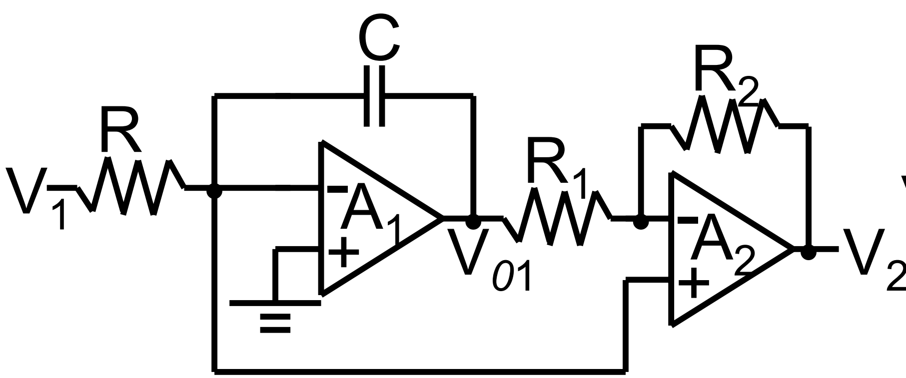
- Miller-inverter cascade $Q_I \approx -|A(jw)|/3$



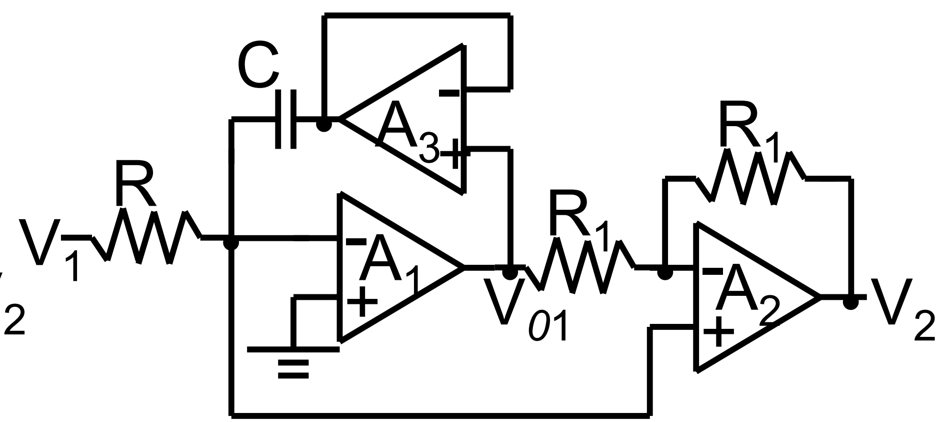

- Positive $Q_I \approx +|A(jw)|$ (phase-lead)




- Modified Miller-inverter cascade, $Q_I = |A(jw)|$

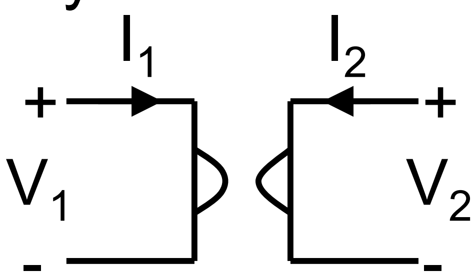


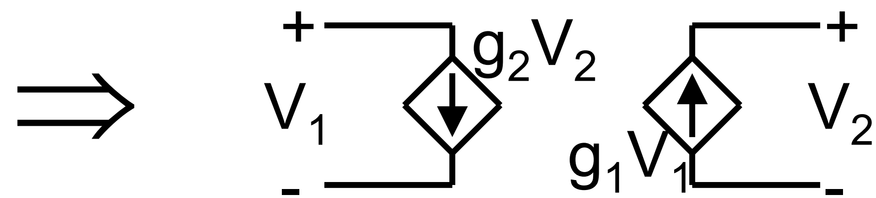

- Modified high-$Q_I$, $Q_I \approx -|A(jw)|^3$



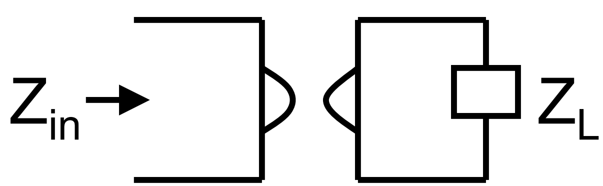
# Gyrator and Immittance Converter

- Inductor simulator
  - Gyrator
  - Immittance converter
- Gyrator
  - Symbol
  - Small-signal equivalent circuit




  - With load



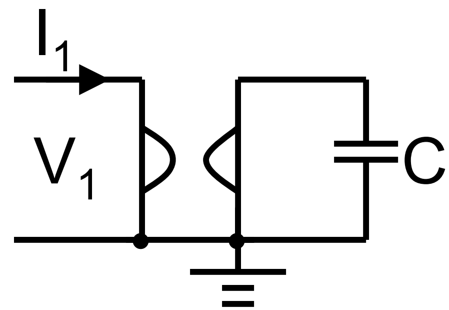

$$Z_{in}(s) = \frac{r^2}{Z_L(S)}; r^2 = \frac{1}{g_1 g_2}$$

=> If $Z_L$ is C, then $Z_{in}$ is L

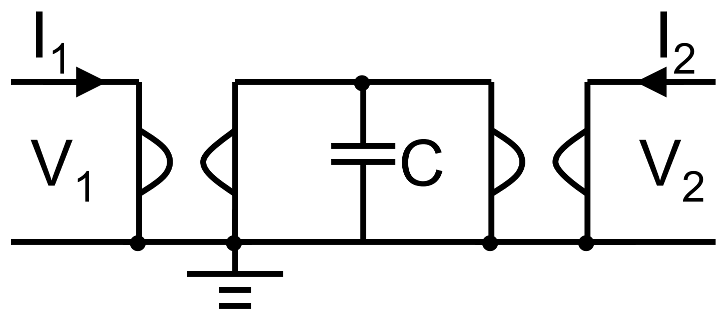
# Gyrator and Immittance Converter(Cont.)

– Gyrator simulation of

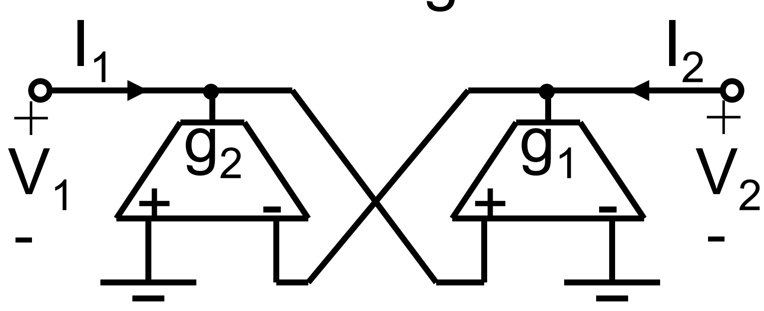
1. Grounded inductor




2. Floating inductor

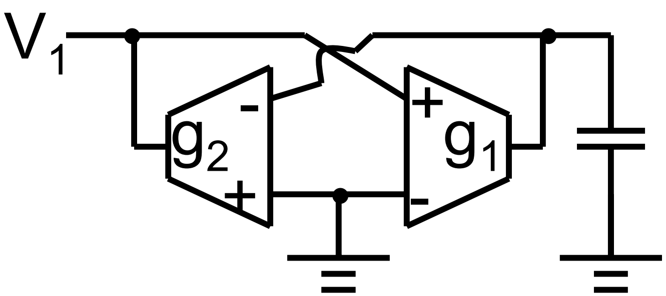



– Gyrator realization using transconductance amplifiers

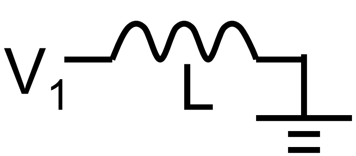


# Inductor Simulation of Gyrator using TA

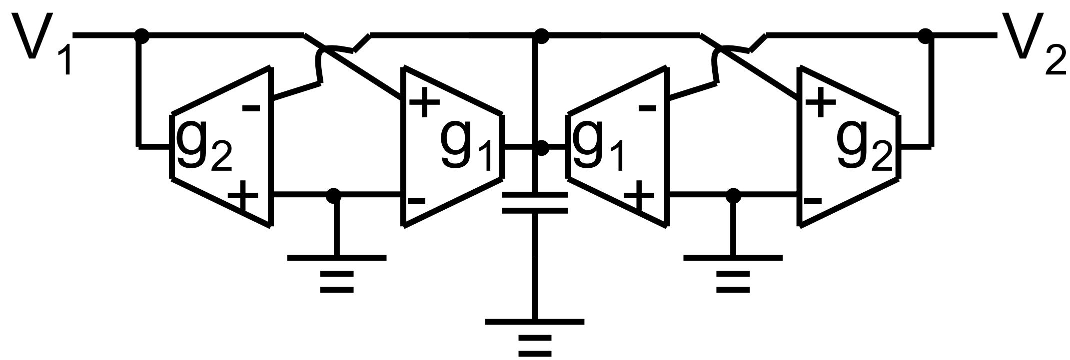
– Grounded inductor



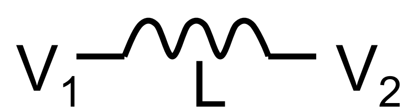

$\equiv$ $L = \dfrac{C}{g_1 g_2}$




– Floating inductor

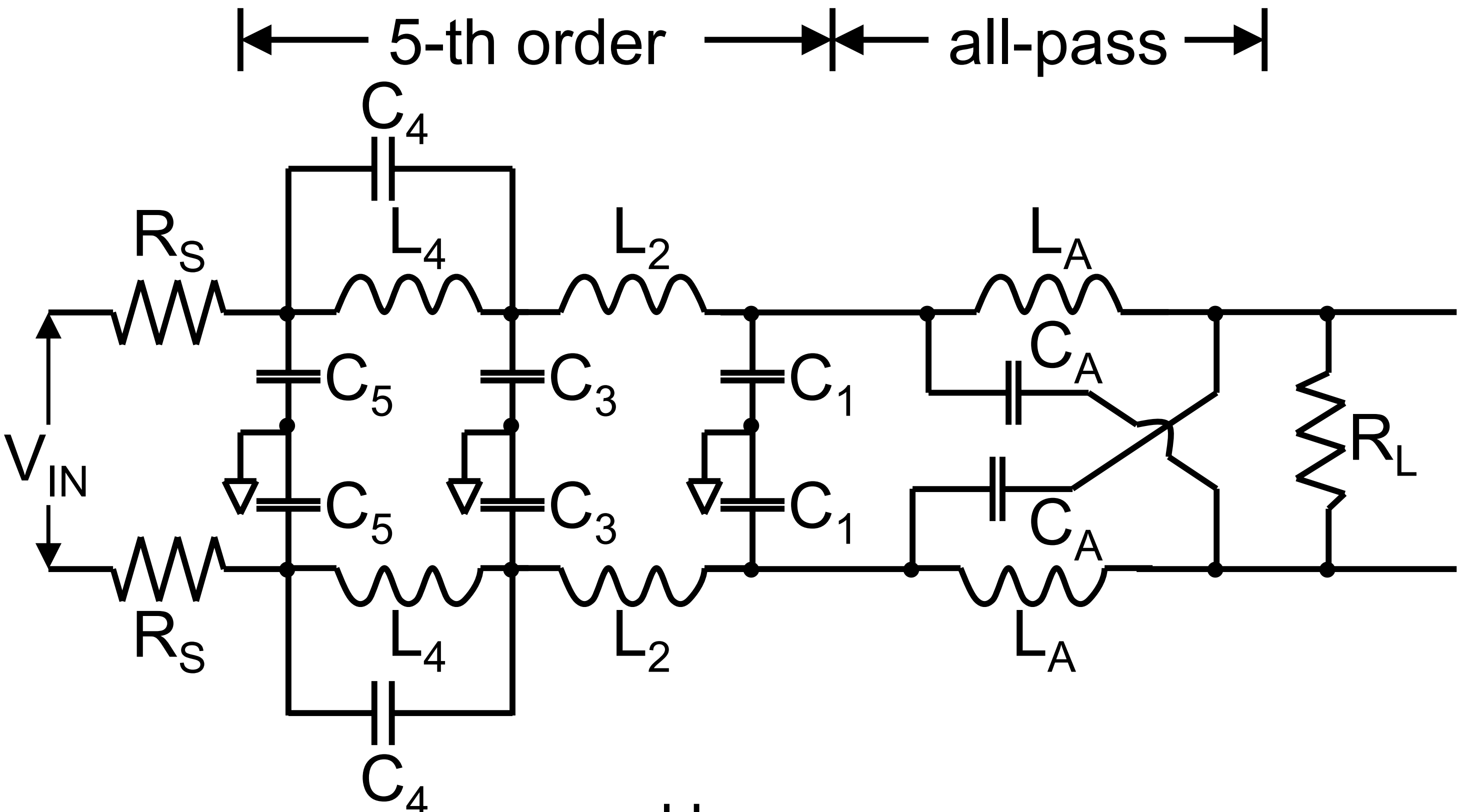



$\equiv$ $L = \dfrac{C}{g_1 g_2}$



# Inductor Simulation of Gyrator using TA(Cont.)

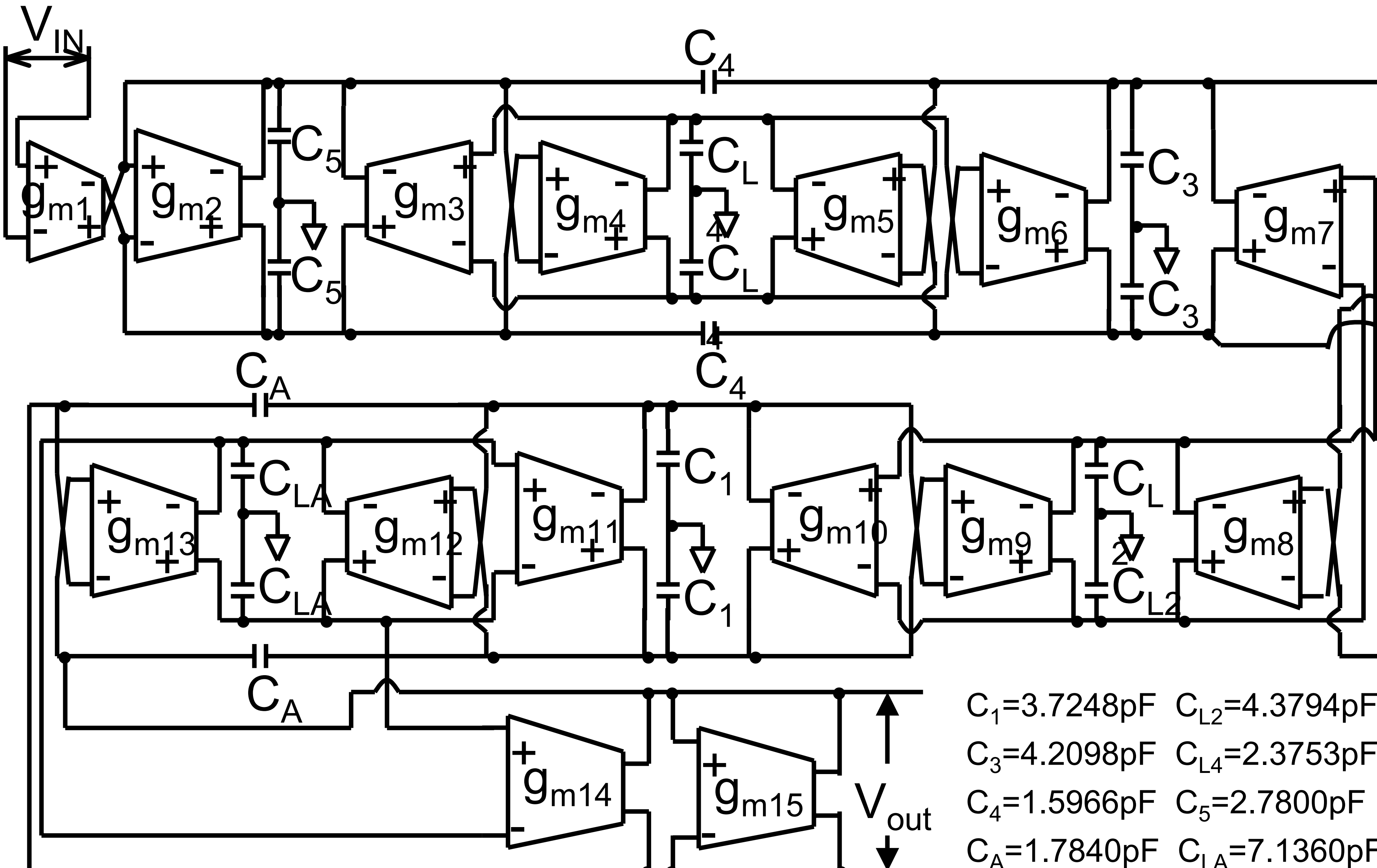
– Example




⇓ Gyrator replacement

next page

# Inductor Simulation of Gyrator using TA(Cont.)




$C_1$=3.7248pF $C_{L2}$=4.3794pF

$C_3$=4.2098pF $C_{L4}$=2.3753pF

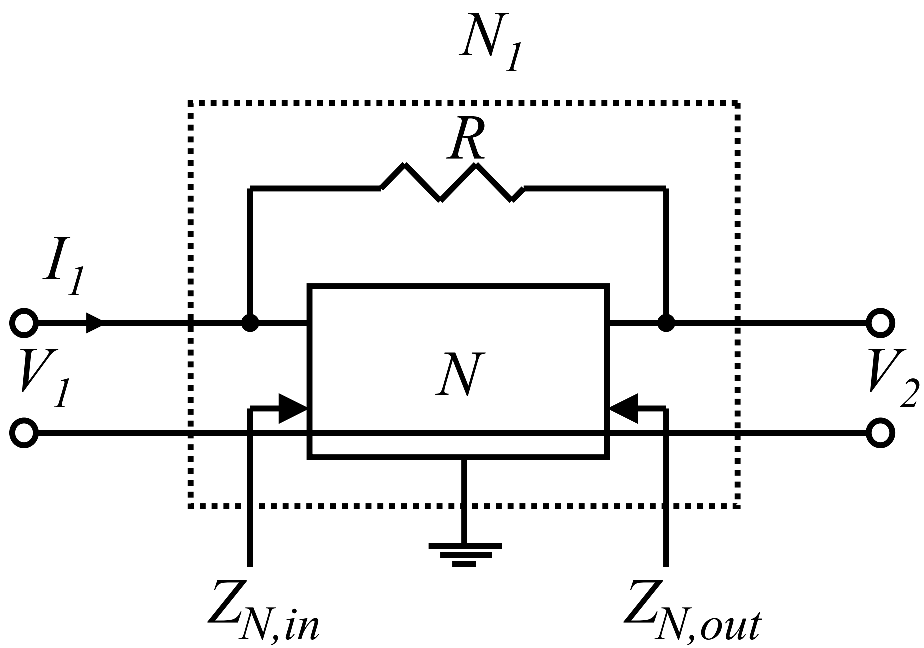
$C_4$=1.5966pF $C_5$=2.7800pF

$C_A$=1.7840pF $C_{LA}$=7.1360pF

gm=200uA/V

# Immittance Converter

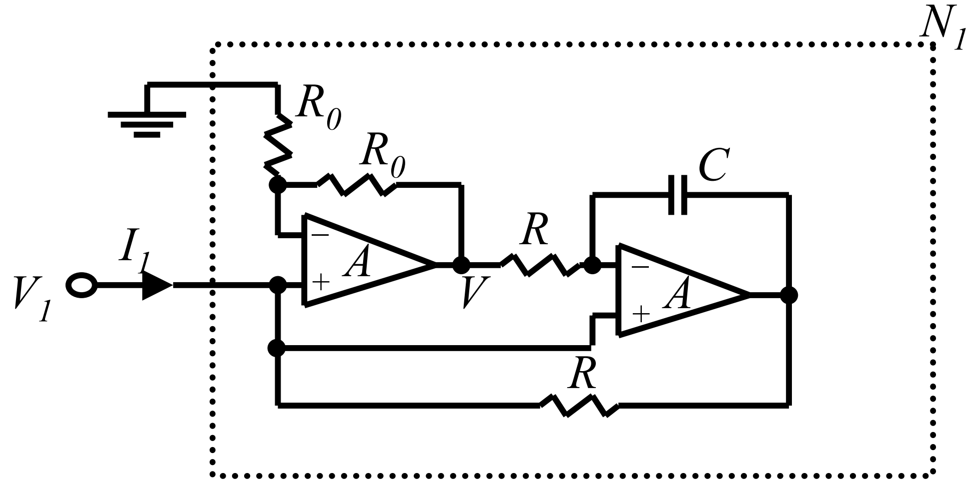
- Generic inductance-simulation circuit



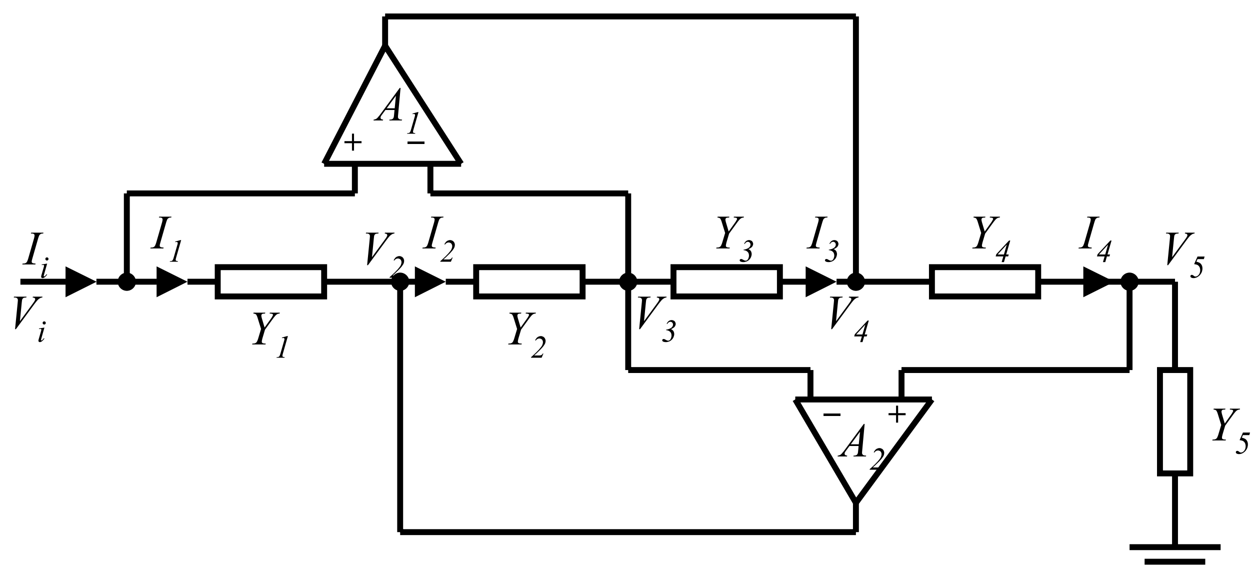

  – A number of such inductance-simulation circuits exist in the literature, all based on the generic configurations as shown above but different realizations of the twopoart N

# Immittance Converter(cont.)

- Riordan inductor-simulation circuit



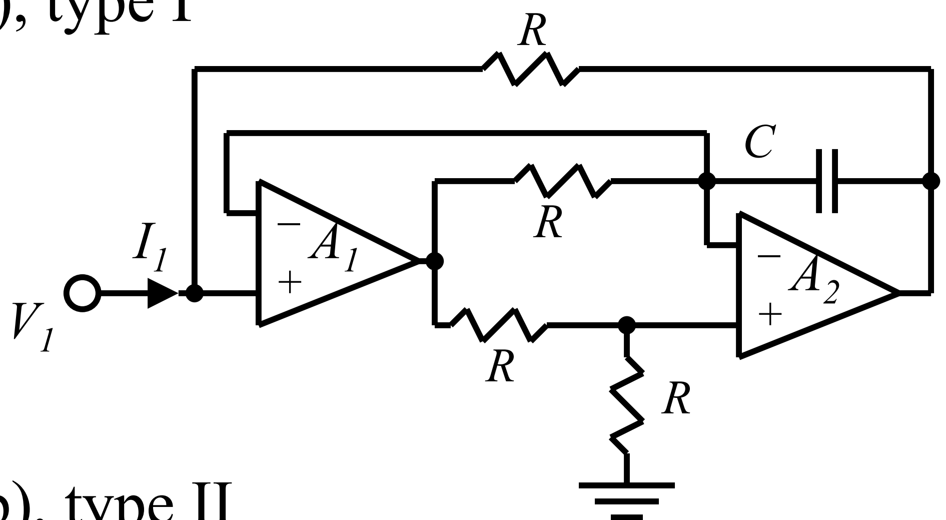

- Antoniou's general immittance converter(GIC)



# Immittance Converter(cont.)

– Two types to realize $L=CR^2$

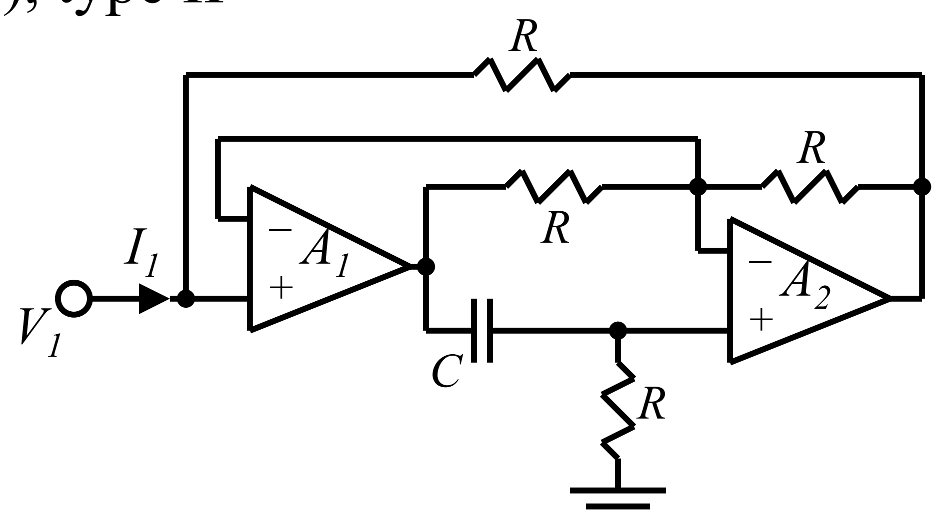
(a), type I




△ Can be used for one of the best-performing biquad filter sections

△ Disadvantage: requires matched OPAMP

(b), type II




△ Preferred device for inductance simulation

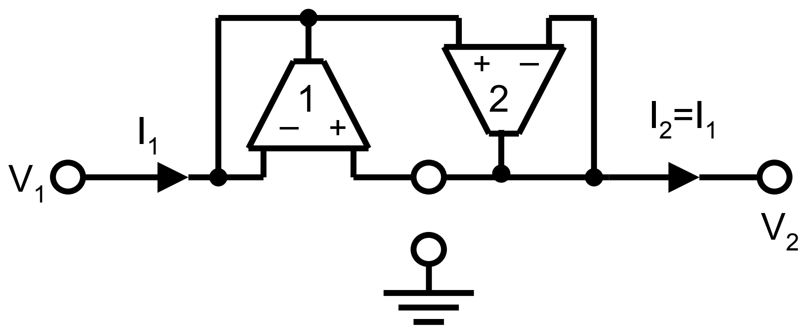
# Active Building Blocks Using OTAs

- OPAMP
  - widely available
  - Highly developed
  - conventionally, it’s the primary active component in the design of (discrete) active filter
  - disadvantages : mentioned previously
- OTA
  - Advantage : mentioned previously
- Active gm-C filter
  - gm & C are sufficient
  - simpler to integrate

# Active Building Blocks Using OTAs(cont.)

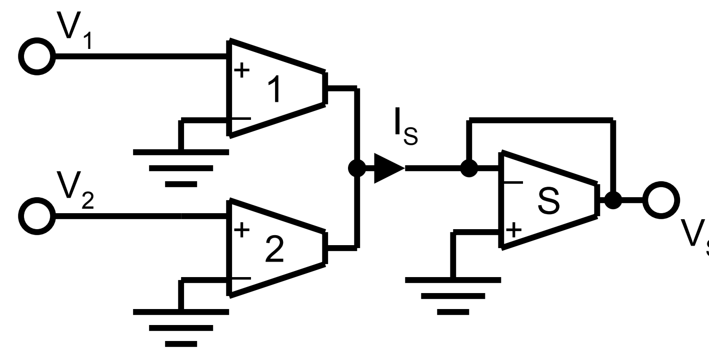
- Basic building blocks

(1) Floating resistor



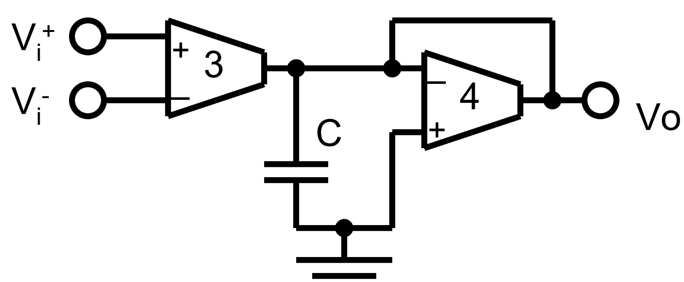

$I_1=g_{m1}(V_1-V_2)$ and $I_2=g_{m2}(V_1-V_2)$
For $g_{m1}=g_{m2}=g_m$, $R=1/g_m$

(2) Summer



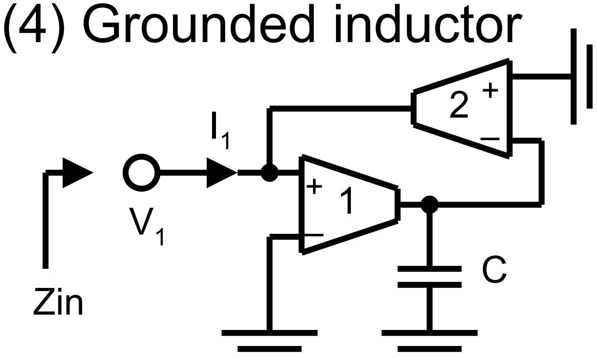

$Vs=g_{m1}V_1/g_{ms}+g_{m2}V_2/g_{ms}$

(3) Losy differential gm-C integrator




$$Vo=\frac{g_{m3}}{SC+g_{m4}}(Vi^+-Vi^-)$$

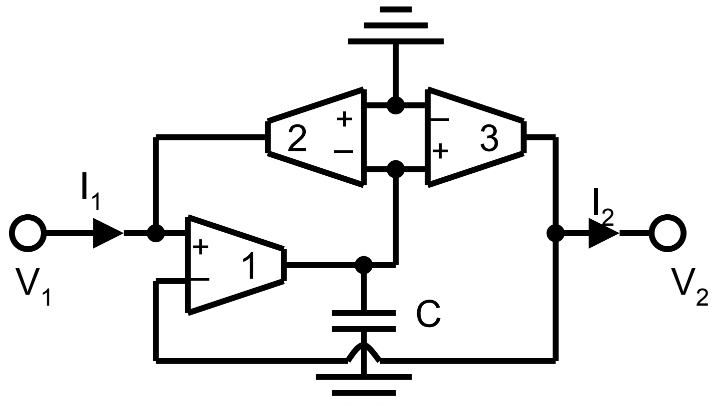
(4) Grounded inductor




$$Zin=\frac{V_1}{I_1}=S\frac{C}{g_{m1}g_{m2}}$$

# Active Building Blocks Using OTAs(cont.)

(5) Floating inductor




$$\textbf{For } g_{m2} = g_{m3} = g_m, \quad L = \frac{C}{g_m g_{m1}}$$

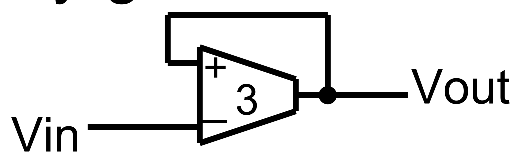
- The above simulations are based on the assumption that OTAs are ideal
  - If OTAs are not ideal, i.e., real OTA

    Example 4-6 shows that the gm-C simulated indrector is far from ideal if real OTAs are used in the design
  - The real transconductances are not ideal current source and their Ro $\neq \infty$

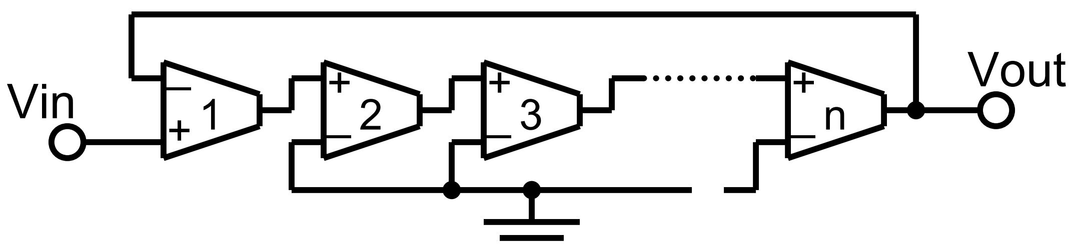
# Active Building Blocks Using OTAs(cont.)

- ∆ Transconductances circuits generally should be designed such that the building blocks drive high-impedance nodes, such as input of other OTAs.
- ∆ If large loads must be driven, the OTA circuit must be buffered.

- OTA buffer circuit
  (1) unity-gain OPAMP




  (2) Transconductance buffer circuit




Assuming for simplicity that all OTAs with output conductance $y_p$ are identical and that $V_{out}$ is loaded by $Y_L$, we calculate

# Active Building Blocks Using OTAs(cont.)

$$\mathbf{Vout} = \frac{g_{m1}}{y_P}\frac{g_{m2}}{y_P}\frac{g_{m3}}{y_P}\cdots\cdots\frac{g_{mn}}{(Y_L + y_P)}(\mathbf{Vin} - \mathbf{Vout})$$

i.e.,
$$\frac{\mathbf{Vout}}{\mathbf{Vin}} = \frac{1}{1+(y_P/g_m)^n(1+Y_L/y_P)} \approx 1 \text{ where } \left|\frac{y_P}{g_m}\right|^n << 1$$

Similarly, the output impedance $Z_{out}(s)$ is calculated as

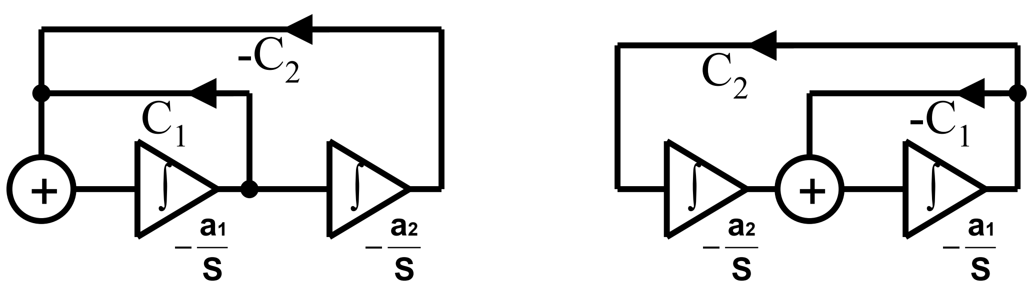
$$Z_{out}(s) = \frac{1/y_P}{1+(g_m/y_P)^n} \approx \frac{1}{y_P}\left(\frac{y_P}{g_m}\right)^n \text{ where } \left|\frac{y_P}{g_m}\right|^n << 1$$

Because it is easy to design the transconductance such that $\mathbf{gm} >> |y_P|$, clearly $|V_{out}/V_{in}| \approx 1$ and $|Z_{out}| << |1/y_P|$ for sufficiently large values of n. Usually, n=2 or 3 will be satisfactory.

# Transconductance Biquads

- Biquad design using TAs
  - convenient realization

    only gm and C are required
  - Equality and ratio may be difficult to maintain in descrete circuits without tuning; They are easily implemented in integrated circuits
- Example: two-integrator biquad
  - signal flow graph




(a) (b)

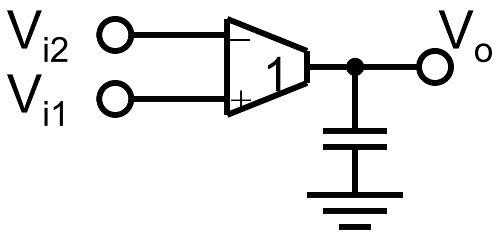
System poles are given by the polynomial

$S_2+Sa_1c_1+a_1a_2c_2=0$

# Transconductance Biquads(cont.)

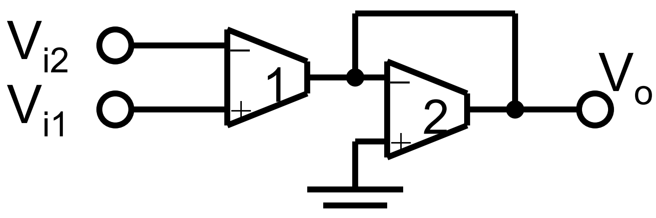
– Basic building blocks

(i) integrating



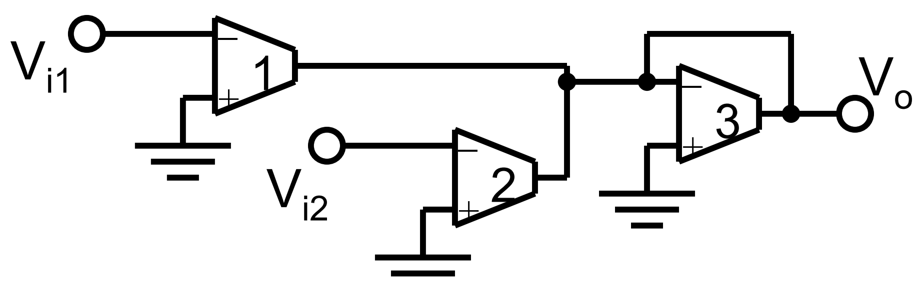

$$V_O = \frac{g_{m1}}{sC}(V_{i1} - V_{i2})$$

(ii) scaling




$$V_O = \frac{g_{m1}}{g_{m2}}(V_{i1} - V_{i2})$$

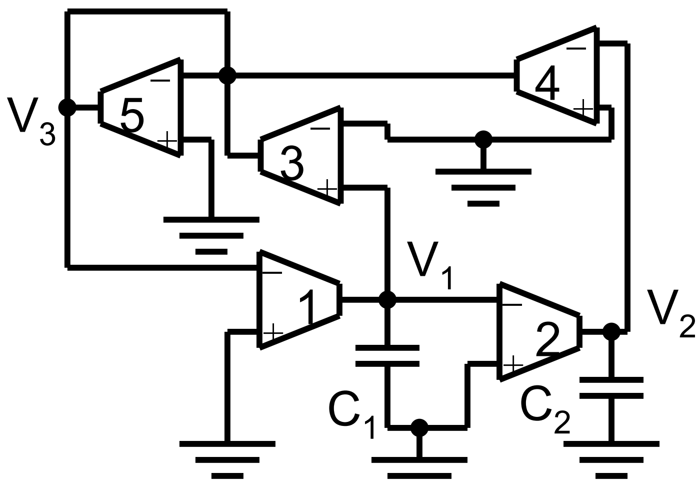
(iii) summing



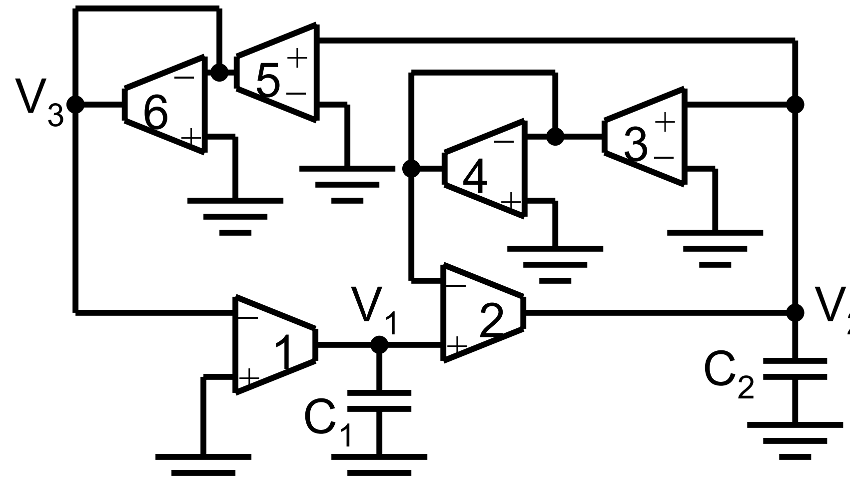

$$V_O = -\frac{g_{m1}}{g_{m3}}V_{i1} + \frac{g_{m2}}{g_{m3}}V_{i2}$$

# Transconductance Biquads(Cont.)

– OTA implementation




(a)




(b)

System polynomials

For Fig(a), $$S^2 + S\frac{w_0}{Q} + w_0^2 = S^2 + S\frac{1}{C_1}\frac{g_{m1}g_{m3}}{g_{m5}} + \frac{g_{m1}g_{m2}g_{m4}}{C_1C_2g_{m5}}$$

For Fig(b), $$S^2 + S\frac{w_0}{Q} + w_0^2 = S^2 + S\frac{1}{C_2}\frac{g_{m2}g_{m3}}{g_{m4}} + \frac{g_{m1}g_{m2}g_{m5}}{C_1C_2g_{m6}}$$

# Transconductance Biquads(Cont.)

It is probably preferable, especially for IC implementations, to choose all transconductances identical. In that case, with $g_{mi}=g_m$ for all i.

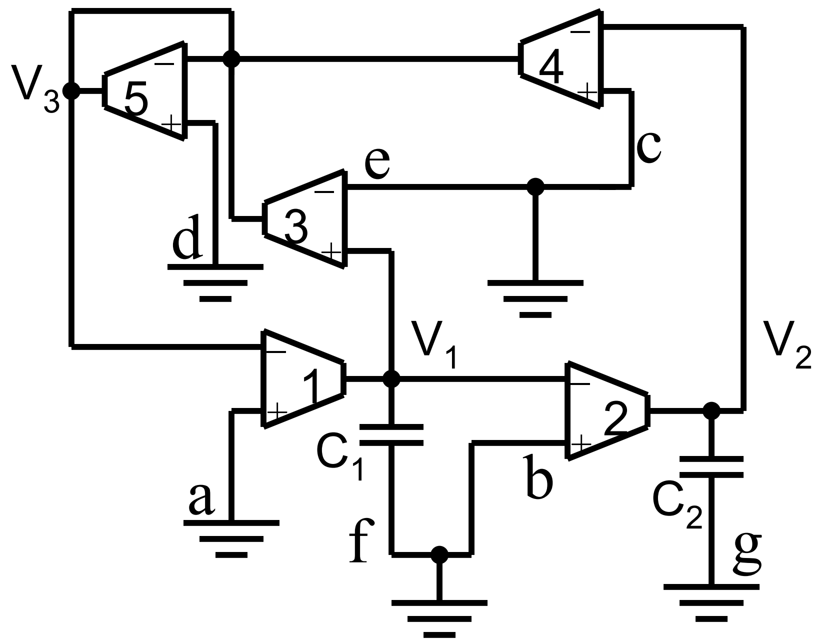
$$w_0 = \frac{g_m}{\sqrt{C_1 C_2}} \quad \text{and} \quad Q = \sqrt{\frac{C_1}{C_2}}$$

Note that all sensitivities are low, i.e., equal to 1/2 in magnitude, but that the capcitor ratio equals $Q^2$. Obviously, for a circuit with identical $g_m$ values, the $gm_3$-$gm_4$ and the $gm_5$-$gm_6$ combinations in (b) can be replaced by short circuits for a saving of four OTAs.

# General OTA Biquads

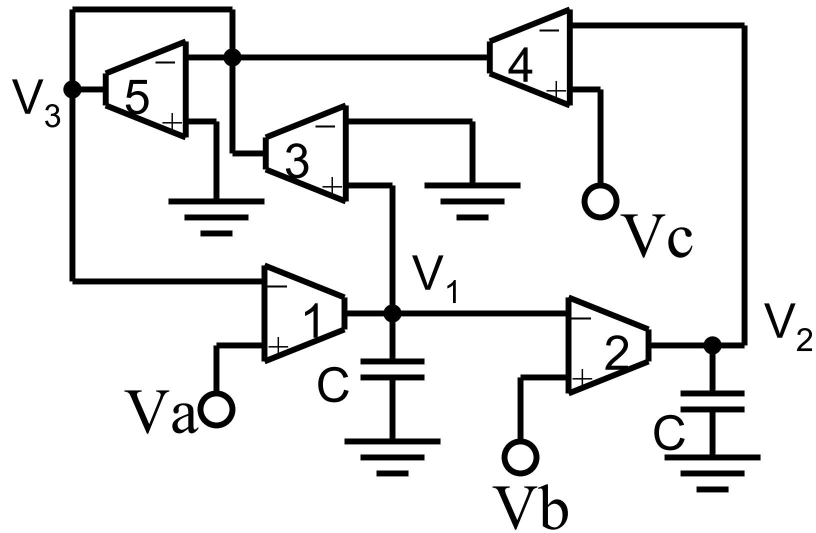
- To generate transmission zeros for Fig (a) by connecting input voltages to grounded terminal
  - seven ground nodes in Fig (a), nodes a,b,c,d,e,f,g.

    => there are many possibilities for zero generation



# General OTA Biquads(cont.)

- Example: choose 3 grounded terminals, Va, Vb, and Vc. (Grounded capacitors are retained. )




$$V_1 = \frac{sC_2 g_{m1}(g_{m5}V_a - g_{m4}V_c) + g_{m1}g_{m2}g_{m4}V_b}{D(s)}$$

$$V_2 = \frac{(sC_1 g_{m2}g_{m5} + g_{m1}g_{m2}g_{m3})V_b + g_{m1}g_{m2}(g_{m4}V_c - g_{m5}V_a)}{D(s)}$$

$$V_3 = \frac{s^2C_1C_2g_{m4}V_c + s(C_2g_{m1}g_{m3}V_a - C1g_{m2}g_{m4}V_b) + g_{m1}g_{m2}g_{m4}V_a}{D(s)}$$

# General OTA Biquads(cont.)

where

$$D(s) = C_1 C_2 g_{m5}\left(s^2 + s\frac{1}{C_1}\frac{g_{m1}g_{m3}}{g_{m5}} + \frac{g_{m1}g_{m2}g_{m4}}{C_1 C_2 g_{m5}}\right)$$

Choosing, for example, $V_a=V_b=0$ and $V_c=V_i$ result in

$$\frac{V_1}{V_i} = H_{BP}(s) = -\frac{sC_2 g_{m1}g_{m4}}{D(s)}$$

$$\frac{V_2}{V_i} = H_{LP}(s) = \frac{g_{m1}g_{m2}g_{m4}}{D(s)}$$

$$\frac{V_3}{V_i} = H_{HP}(s) = \frac{s^2 C_1 C_2 g_{m4}}{D(s)}$$

A general biquadratic transfer function can be obtained by setting $V_a=V_b=V_c=V_i$

$$\frac{V3}{Vi} = \frac{s^2 C_1 C_2 g_{m4} + s(C_2 g_{m1} g_{m3} - C_1 g_{m2} g_{m4}) + g_{m1} g_{m2} g_{m4}}{D(s)}$$

# General OTA Biquads(cont.)

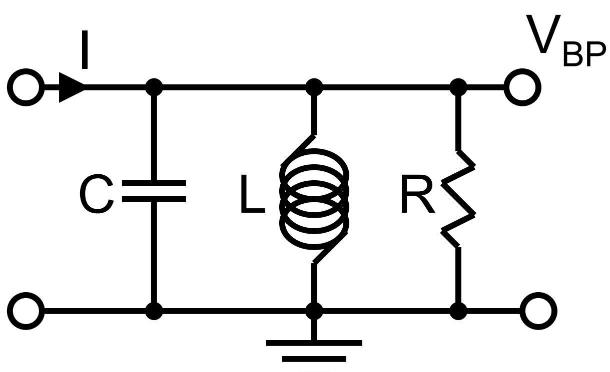
Clearly, $C_2 g_{m1} g_{m3} = C_1 g_{m2} g_{m4}$ results in a band-rejection function, and setting $C_1 g_{m2} g_{m4} = 2C_2 g_{m1} g_{m3}$ and $g_{m4} = g_{m5}$ yields an allpass filter.

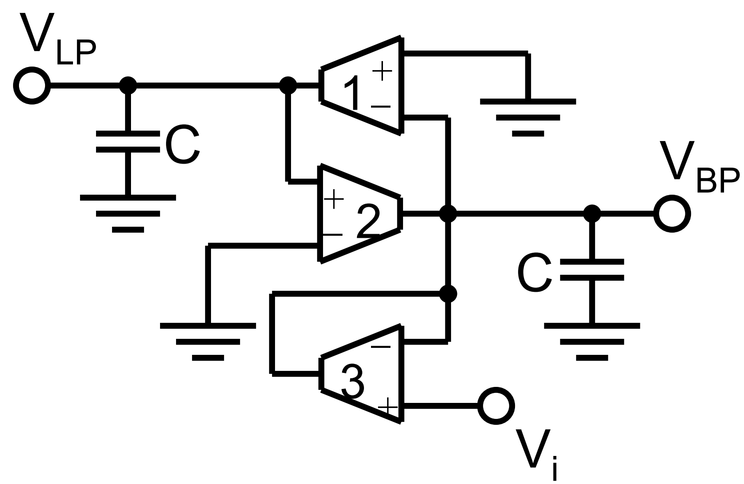
– The indicated equality conditions may be difficult to maintain in descrete circuits without tuning; they are however, easily implemented in integrated form.
– Using different inputs or the topology of Fig (b) results in different ways for realizing second-order transfer characteristics which may have advantages or disadvantages in number of components, element spread, ease of tuning, or sensitivity to parasitics.
– Note that the voltage transfer functions given in the above equations are ideal, and that parasitic input and, especially, output impedances cause errors in filter performance that should be investigated form case to case.
– At high operating frequencies for which OTA designs are intended, the deviations of these poles may well not be negligible. For filters in the audio-frequency range, however, errors caused by parasitic components and also by the finite bandwidth of OTAs are usually negligible.

# Active OTA-C Simulation of Passive RLC Resonance Circuit

- No need of voltage scaling
- Reduce parasitic effect
- Simple : Only three OTAs and 2Cs are required

A final very simple lowpass-bandpass configuration that uses OTAs and avoids many of the parasitic problems is shown below.




It is derived from a passive RLC resonance circuit by making use of $R_i=1/g_{m3}$ and $L=C/(g_{m1}g_{m2})$. The input current I is obtained from $I=g_{m3}V_i$. Thus, setting

# Active OTA-C Simulation of Passive RLC Resonance Circuit (cont.)

$g_{m1}=g_{m2}=g_m$, we obtain in complete analogy to the passive circuit given by

$$\frac{V_{BP}}{I}=\frac{1}{sC+1/(sL)+1/R}=\frac{sL}{s^2LC+sL/R+1}$$

the following active realization:

i.e., $$\frac{V_{BP}}{g_{m3}V_i}=\frac{1}{sC+g_m{}^2/(sC)+g_{m3}}\ ;\ \text{where } g_{m1}=g_{m2}=g_m$$

$$\frac{V_{BP}}{Vi}=\frac{sg_{m3}C}{s^2C^2+sg_{m3}C+gm^2}$$

and $$\frac{V_{LP}}{V_i}=-\frac{g_mg_{m3}}{s^2C^2+sg_{m3}C+g_m{}^2}$$

All parasitic OTA capacitors are in parallel with the main circuit capacitors; thus they do not increase the order of the transfer function and can be absorbed by predistortion. Also, the output conductances of OTAs 2 and 3 are in parallel

# Active OTA-C Simulation of Passive RLC Resonance Circuit (cont.)

with the "conductance" $g_{m3}$ and can be absorbed by predistortion. Thus, the only "separate" parasitic is $g_0$ of OTA 1; its effect is to lower Q and increase the pole frequency. Evidently, the circuit realizes nominally

$$w_0 = \frac{g_m}{C} \quad \text{and} \quad Q = \frac{g_m}{g_{m3}}$$

and has the additional advantage that, at $w=w_0$, where the maximum output voltages occur, both voltage gains are equal to unity:

$$\left|\frac{V_{BP}}{V_i}(jw_0)\right| = 1 \qquad \left|\frac{V_{LP}}{V_i}(jw_0)\right| = \frac{g_m}{w_0 C} = 1$$

The maximum internal voltage seen by the OTAs is, therefore, given by $V_i$, which can be controlled easily without scaling; this latter fact is convenient in practice because all available OTA designs suffer from a limited linear signal range.