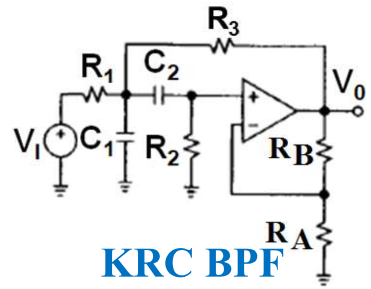
LIC: LECTURE
ACTIVE FILTERS

- **Multiple Feedback Filters**
  - **(i) Concept**
  - **(ii) Multiple Feedback based Band Pass filter**
  - **(iii) Multiple Feedback based Low Pass filter**
  - **(iv) Example**

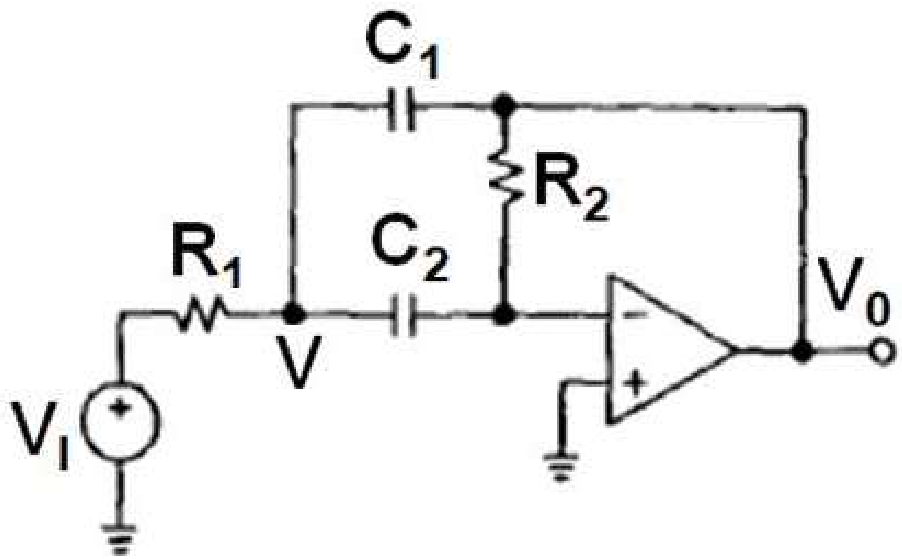
[1] Franco, Sergio. Design with operational amplifiers and analog integrated circuits. Vol. 1988. New York: McGraw-Hill.

Multiple Feedback Filters



KRC BPF

- Multiple Feedback Filters utilizes more than one feedback path.
- Unlike their KRC counterparts, which configure the op-amp for a finite gain K, multiple feedback filters exploit the full open loop gain and are also referred to as infinite gain filters.



Multiple Feedback based Band Pass filter

Multiple Feedback based Band Pass Filter

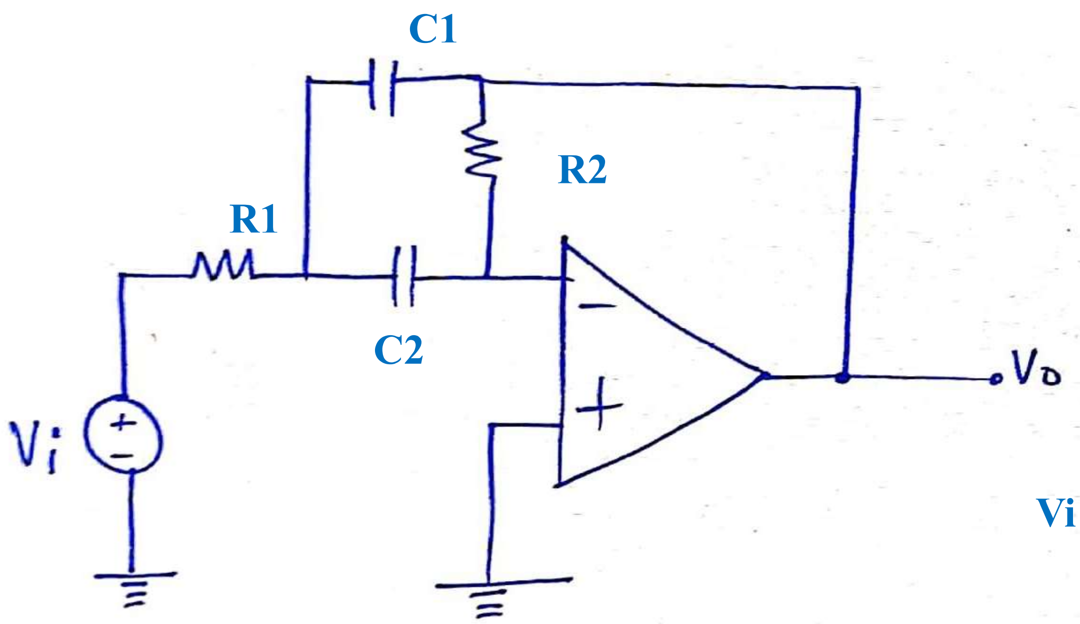
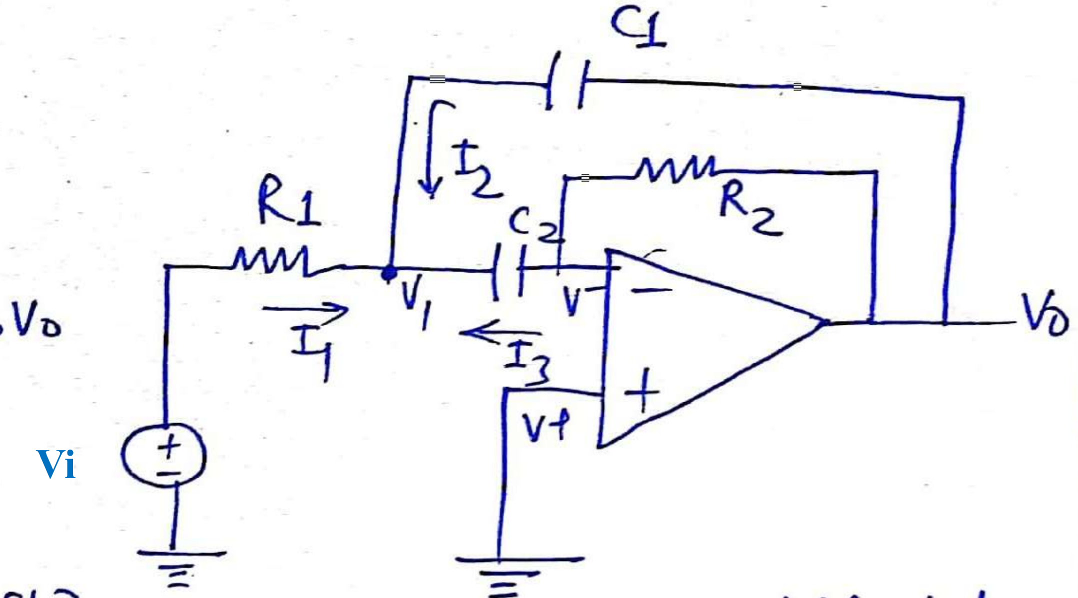


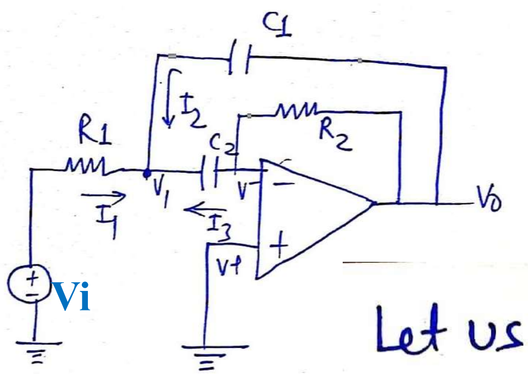
Fig.- Band pass filter (multiple f/B)

The above circuit can be redrawn as;



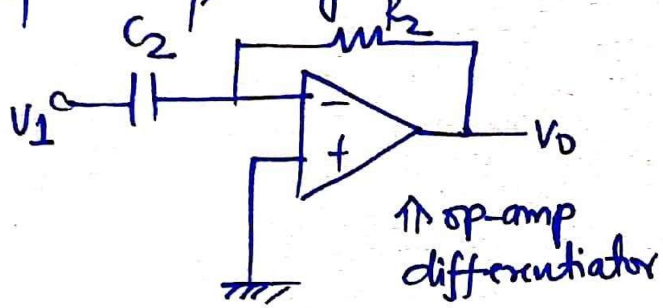
R_2–C_2–opamp: makes differentiator

$\left.\begin{matrix} V^+ = 0 \\ V^- \approx 0 \end{matrix}\right\}$ Virtual short



Multiple Feedback based BPF

Let us analyse C_2-R_2 op-amp separately, we get:



$$\frac{V_0}{V_1} = \frac{-R_2}{X_{C_2}} = \frac{-R_2}{1/SC_2} = -SR_2C_2$$

$$\boxed{V_0 = -SR_2C_2V_1} \quad -(1)$$

KCL at V_1 in fig.1, we get:

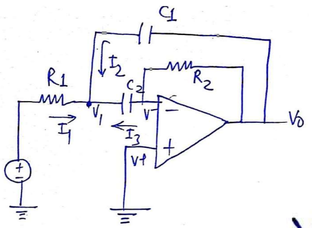
$$I_1 + I_2 + I_3 = 0$$

$$\frac{V_{in} - V_1}{R_1} + \frac{V_0 - V_1}{1/SC_1} + \frac{0 - V_1}{1/SC_2} = 0$$

$$\frac{V_{in}}{R_1} + SC_1V_0 - V_1\left[\frac{1}{R_1} + SC_1 + SC_2\right] = 0$$

from (1); $V_1 = \dfrac{-V_0}{SR_2C_2}$

$$\frac{V_{in}}{R_1} + SC_1V_0 + \frac{V_0}{SR_2C_2}\left[\frac{1 + SC_1R_1 + SC_2R_1}{R_1}\right] = 0$$



Multiple Feedback based BPF

$$\frac{V_{in}}{R_1} + SC_1V_0 + \frac{V_0}{SR_2C_2}\left[\frac{1+SC_1R_1+SC_2R_1}{R_1}\right] = 0$$

$$-\frac{V_{in}}{R_1} = \left[SC_1 + \frac{1}{SC_2R_1R_2}\left(1+SC_1R_1+SC_2R_1\right)\right]V_0$$

$$V_{in} = -R_1\left[\frac{S^2R_1R_2C_1C_2+1+SC_1R_1+SC_2R_1}{SR_2C_2R_1}\right]V_0$$

$$\boxed{H(S) = \frac{V_0(S)}{V_{in}(S)} = \frac{-SR_2C_2}{1+SC_1R_1+SC_2R_1+S^2C_1C_2R_1R_2}} \; * \quad -(2)$$

Let $S = j\omega$, we get:

$$\boxed{H(j\omega) = \frac{-j\omega R_2C_2}{1-\omega^2R_1R_2C_1C_2+j\omega R_1[C_1+C_2]}} \quad -(3)$$

Comparing with standard transfer func. of B.P.F.

$$H(j\omega) = \frac{H_{oBP}\left(j\omega/\omega_0Q\right)}{1-(\omega/\omega_0)^2+\left(j\omega/\omega_0Q\right)} \quad -(4)$$

Compare (3) & (4)

(1) $$\omega^2R_1R_2C_1C_2 = \frac{\omega^2}{\omega_0^2}$$

$$\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}}$$

$$\boxed{f_0 = \frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}} \; *$$

Cut-off Frequ.

(2) $$j\omega R_1[C_1+C_2] = \frac{j\omega}{\omega_0Q}$$

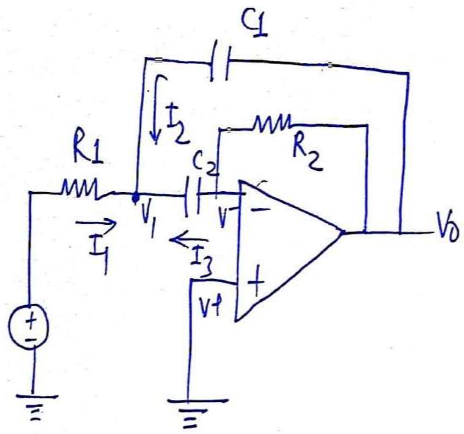
$$Q = \frac{1}{\omega_0R_1[C_1+C_2]}$$

$$Q = \frac{\sqrt{R_1R_2C_1C_2}}{R_1(C_1+C_2)}$$

$$\boxed{Q = \frac{\sqrt{R_2/R_1}}{\sqrt{\frac{C_1}{C_2}}+\sqrt{\frac{C_2}{C_1}}}} \; *$$

(3) Similarly: $$\boxed{H_{oBP} = \frac{-R_2/R_1}{1+C_1/C_2}}$$

Multiple Feedback based BPF



If substitute $C_1 = C_2 = C$ then

$$\boxed{W_0 = \frac{1}{\sqrt{R_1 R_2}\cdot C}}\ *$$

$$\boxed{Q = 0.5\sqrt{\frac{R_2}{R_1}}}\ *$$

$$\boxed{H_{0BP} = -\frac{1}{2}\left(\frac{R_2}{R_1}\right)}$$

or

$$\boxed{H_{0BP} = -2Q^2}$$

DESIGN EQUATIONS:

$$\boxed{R_1 = \frac{1}{2W_0 Q\cdot C}\ ;\ R_2 = \frac{2Q}{W_0 C}}\ *$$

$R_1 = \frac{1}{2w_0 Q\cdot C} \leftarrow$ proof

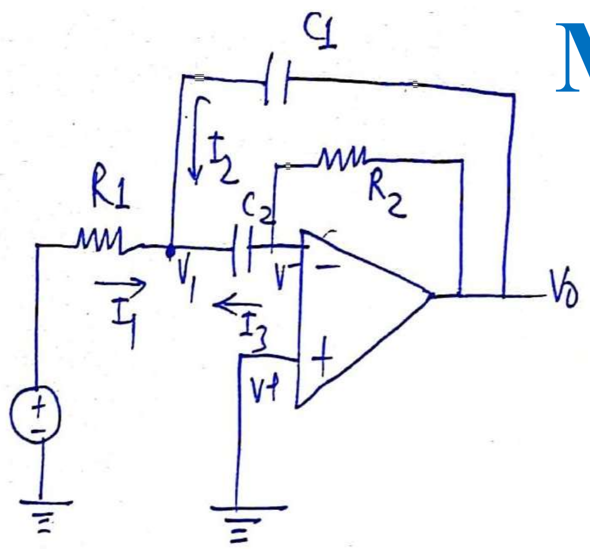
$Q^2 = (0.5)^2 \frac{R_2}{R_1} \rightarrow R_2 = R_1 Q^2 \cdot 4$

$w_0^2 = \frac{1}{R_1 R_2 C^2} \rightarrow \frac{1}{w_0^2 C^2 R_1} = R_1 Q^2 \cdot 4$

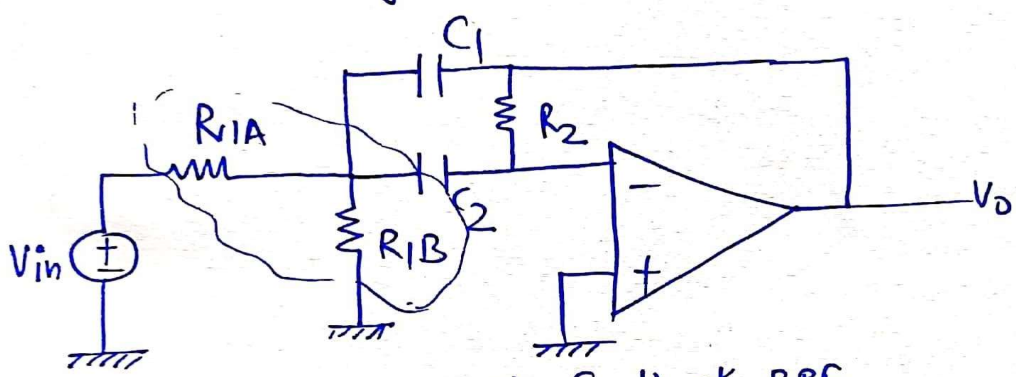
So; $R_1 = \frac{1}{2Q_0 W_0 C}$, Similarly calculate R_2.

$\Rightarrow$ Denoting resonance gain magnitude as $H_0 = |H_{0BP}|$, for simplicity we observe that it increases quadratically with Q. $(H_0 = -2Q^2)$

Multiple Feedback based BPF



DESIGN: If we want $H_o < 2Q^2$, we must replace R_1 with a voltage divider as shown below:



Multiple feedback BPF

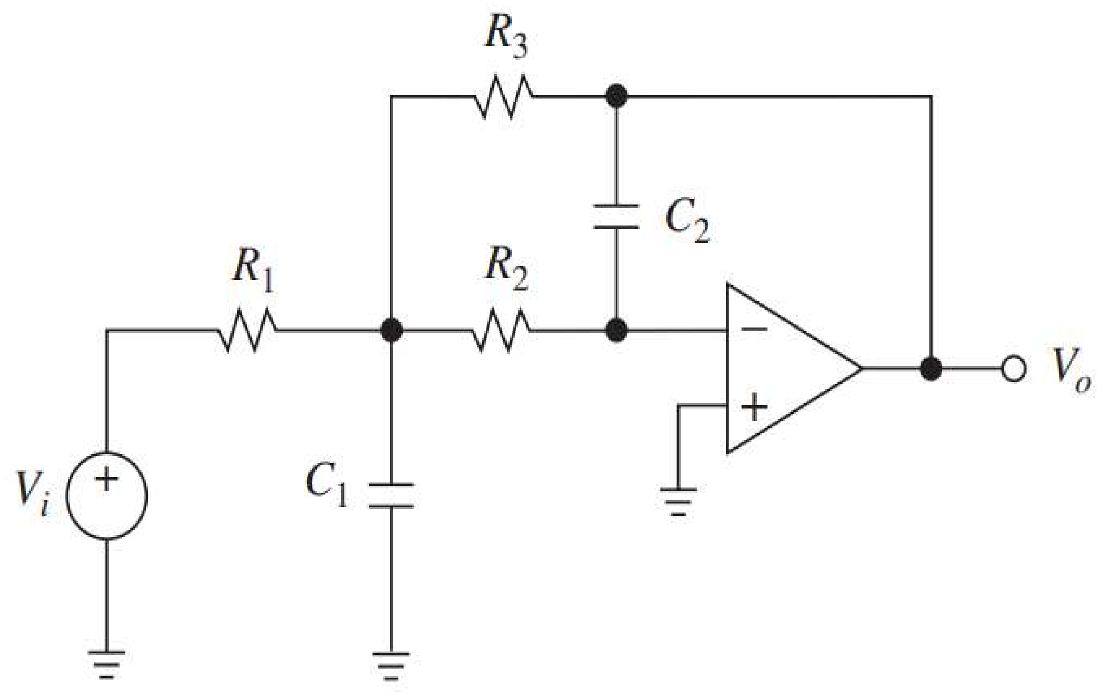
The design equations:

$$\boxed{R_{1A} = \frac{Q}{H_o \omega C}} \qquad \boxed{R_{1B} = \frac{R_{1A}}{\left[\left(\frac{2Q^2}{H_o}\right) - 1\right]}}$$

Problem: Design a multiple feedback band-pass filter with $f_o = 1\ KHz$, $H_{oBP} = 20\ dB$ (resonance gain), $Q = 1.0$

Multiple Feedback based Low Pass Filter

- The circuit consists of the low-pass stage R1-C1 followed by the integrator stage made up of R2, C2, and the op amp.
- So, we anticipate a low-pass response. Moreover, the presence of positive feedback via R3 should allow for Q control.



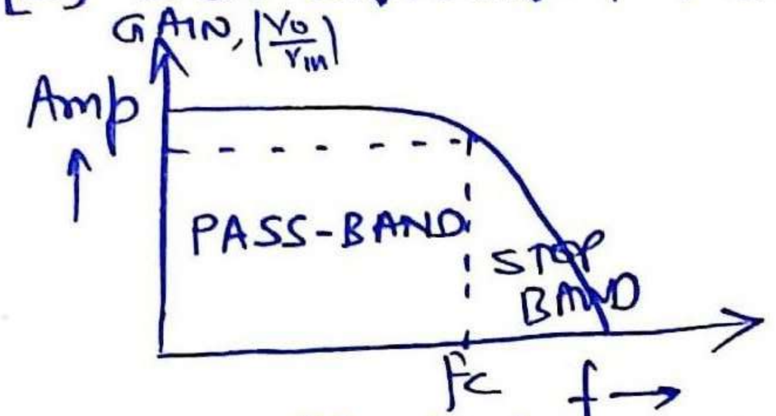
LIC: LECTURE
ACTIVE FILTERS

- **FILTER APPROXIMATIONS:**
  - **(i) Introduction to different type of Filters for Approximation**
  - **(ii) Butterworth Response**
    - **(a) Filter Design Approximation**
    - **(b) Design of First Order BW Filter**
    - **(c) Design of Higher Order BW Filters & Examples**

INTRODUCTION: FITER APPROXIMATIONS

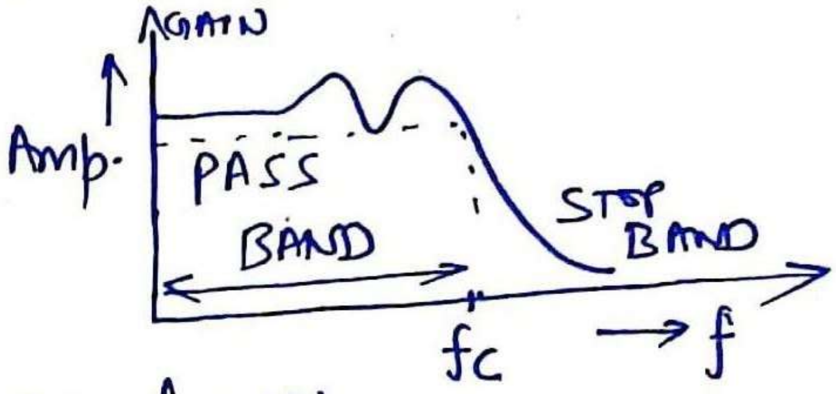
BUTTERWORTH, CHEBYSHEY, ELIPTICAL & BESSEL

[1] BUTTERWORTH FILTER



* Maximally flat approximation
* FLAT BAND! flat response
  No ripples
* Roll-off rate → 20n dB/decade
* Also, no ripple in stop band

[2] CHEBYSHEV FILTER APPROXIMATION

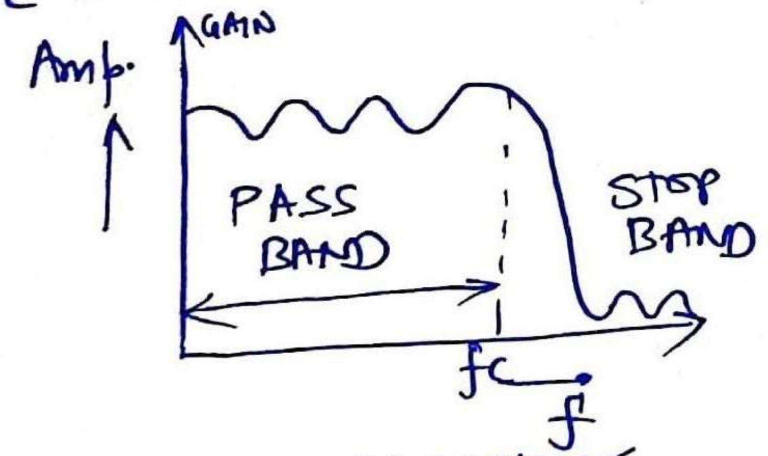


* No of ripple $= \frac{n}{2}$
* Ripple pass band
* Transition very fast

INTRODUCTION: FITER APPROXIMATIONS

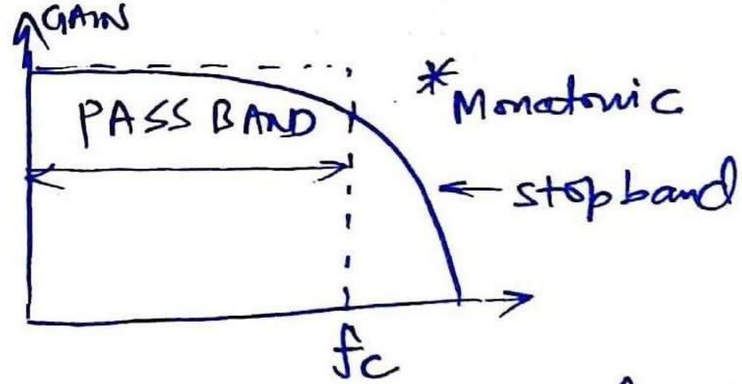
BUTTERWORTH, CHEBYSHEY, ELIPTICAL & BESSEL

[3] ELLIPTICAL FILTER APPROXIMATION



* FASTEST TRANSITION
* Ripple in pass band & stop band

[4] BESSEL FILTER APPROXIMATION



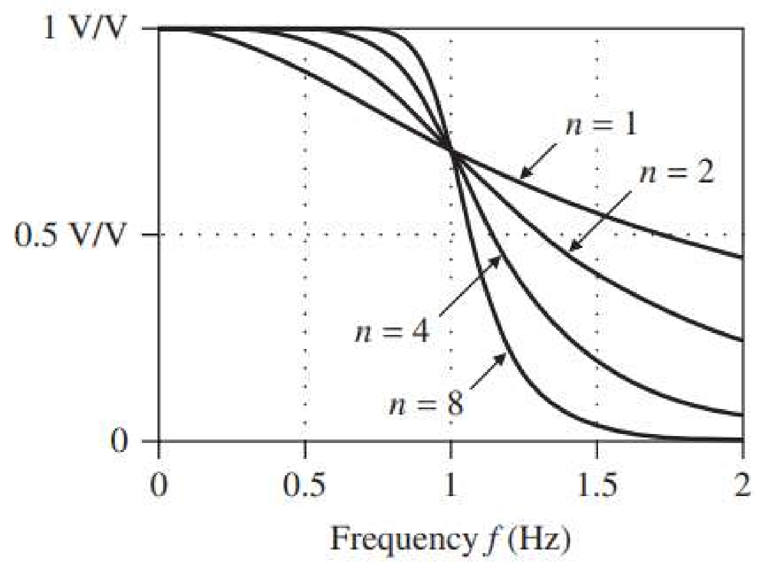
* No ripple in pass band & stop band.
* Phase-freq. relation linear
* Constant delay (slow transition)

Higher order Butterworth Filter design

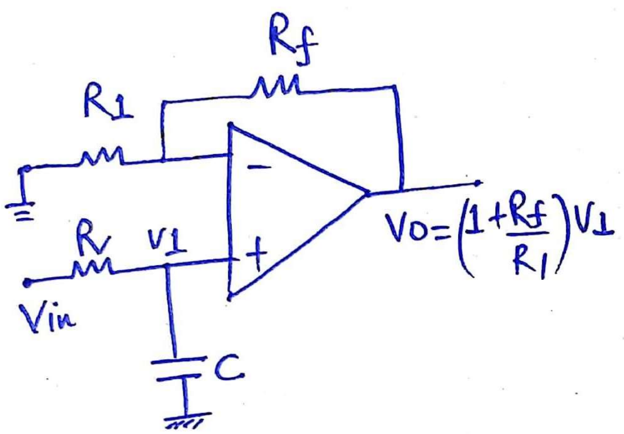
- The gain of the Butterworth approximation is

$$|H(j\omega)| = \frac{1}{\sqrt{1 + \epsilon^2(\omega/\omega_c)^{2n}}}$$

where n is the order of the filter, ω_c is the cutoff frequency, and ϵ is a constant that determines the maximum passband variation as $A_{\max} = A(\omega_c) = 20 \times \log_{10}\sqrt{1+\epsilon^2} = 10\log_{10}(1+\epsilon^2)$. The first $2n-1$ derivatives of $|H(j\omega)|$ are zero at $\omega = 0$, indicating a curve as flat as possible at $\omega = 0$. Aptly referred to as *maximally flat*, a Butterworth curve becomes somewhat rounded near ω_c and rolls off at an ultimate rate of $-20n$ dB/dec in the stopband



FIRST ORDER LOW PASS BUTTERWORTH FILTER APPROXIMATIONS



$$V_O = \left(1 + \frac{R_f}{R_1}\right) V_1$$

$$V_1 = \frac{X_C}{R + X_C} V_{in}$$

$$\boxed{\frac{V_O}{V_{in}} = \frac{A_f}{1 + j\omega/\omega_C}} \quad \text{or} \quad \left|\frac{V_O}{V_{in}}\right| = \frac{A_f}{\sqrt{1 + (\omega/\omega_C)^2}}$$

$$\left.\begin{aligned} A_f &= 1 + R_f/R_1 \\ \omega_C &= \frac{1}{RC} \end{aligned}\right]$$

As per Butterworth approximation; for $\varepsilon = 1$

$$|H(j\omega)| = \frac{1}{\sqrt{1 + (\omega/\omega_C)^{2N}}}$$

for first order $N = 1$

$$|H(j\omega)| = \frac{1}{\sqrt{1 + (\omega/\omega_C)^2}}$$

For $f_C = 1\,KHZ$, $C = 0.01\,\mu f$ & $A_f = 2$
Design Butterworth low pass filter

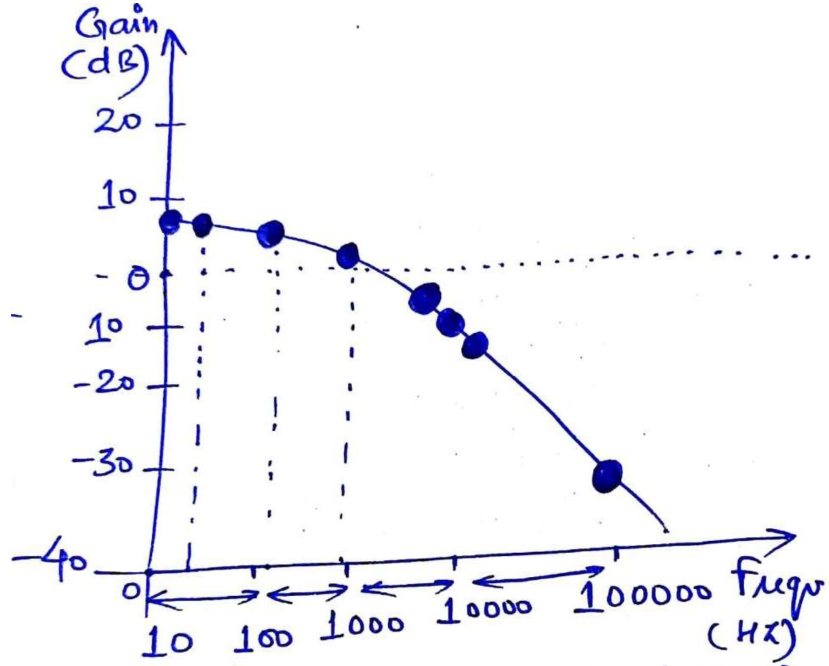
$$A_f = 1 + \frac{R_f}{R_1} = 2$$

Assume - $R_f = R_1 = 10\,K\Omega$

$$R = \frac{1}{2\pi f_C \cdot C} = 15.9\,K\Omega$$

FIRST ORDER LOW PASS BUTTERWORTH FILTER APPROXIMATIONS

F (Hz)	Gain (Vo/Vi)	Gain (dB)
10	2	6.02
20	1.99	5.98
200	1.96	5.85
700	1.64	4.29
1000	1.41	3.01
3000	0.63	-3.98
10000	0.2	-14.02
100000	0.02	-33.98



Decade change ≈ −33.98 − (−14.02)
≃ 19.96
≈ −20dB/decade

HIGHER ORDER LOW PASS BUTTERWORTH FILTER APPROXIMATIONS

- General Low pass equation transfer function

$$H_{LP}(j\omega)=\frac{1}{1-(\omega/\omega_0)^2+(j\omega/\omega_0)/Q}$$

- $H(j\omega)= H_{oLP} * H_{LP}(j\omega)$

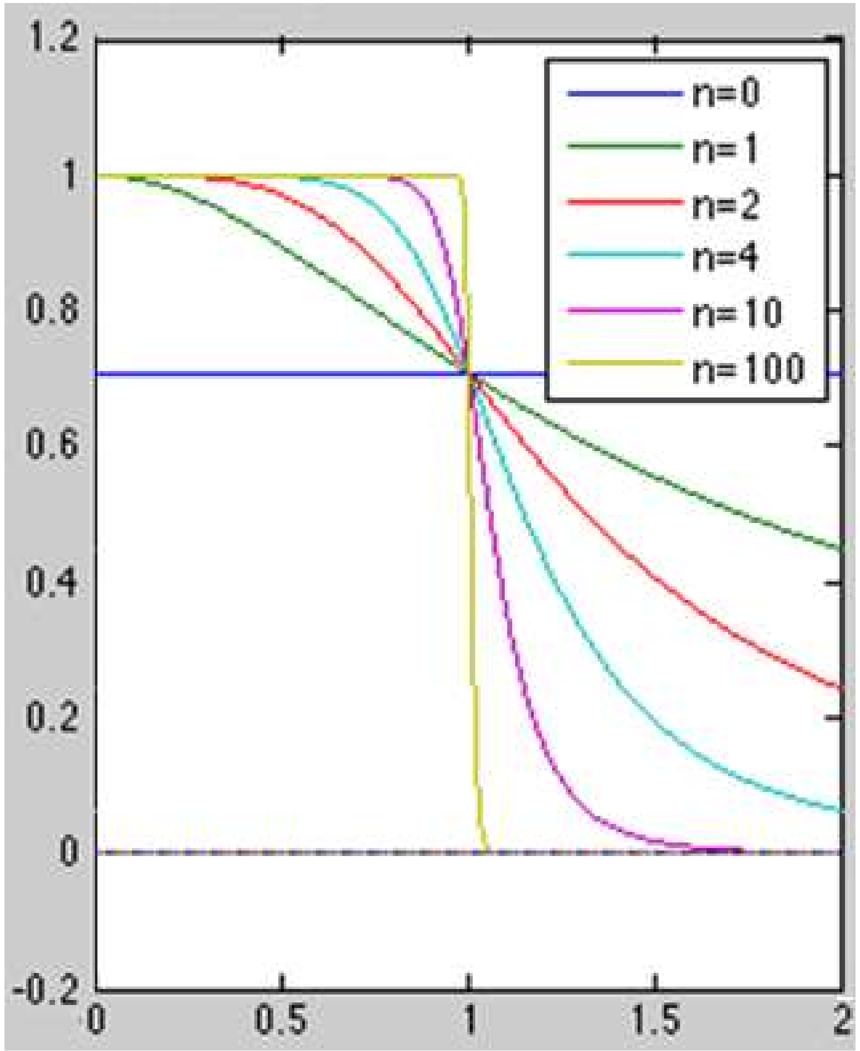
- Butterworth filter Q=0.707 =(1/√2)

- Find out the amplitude ??

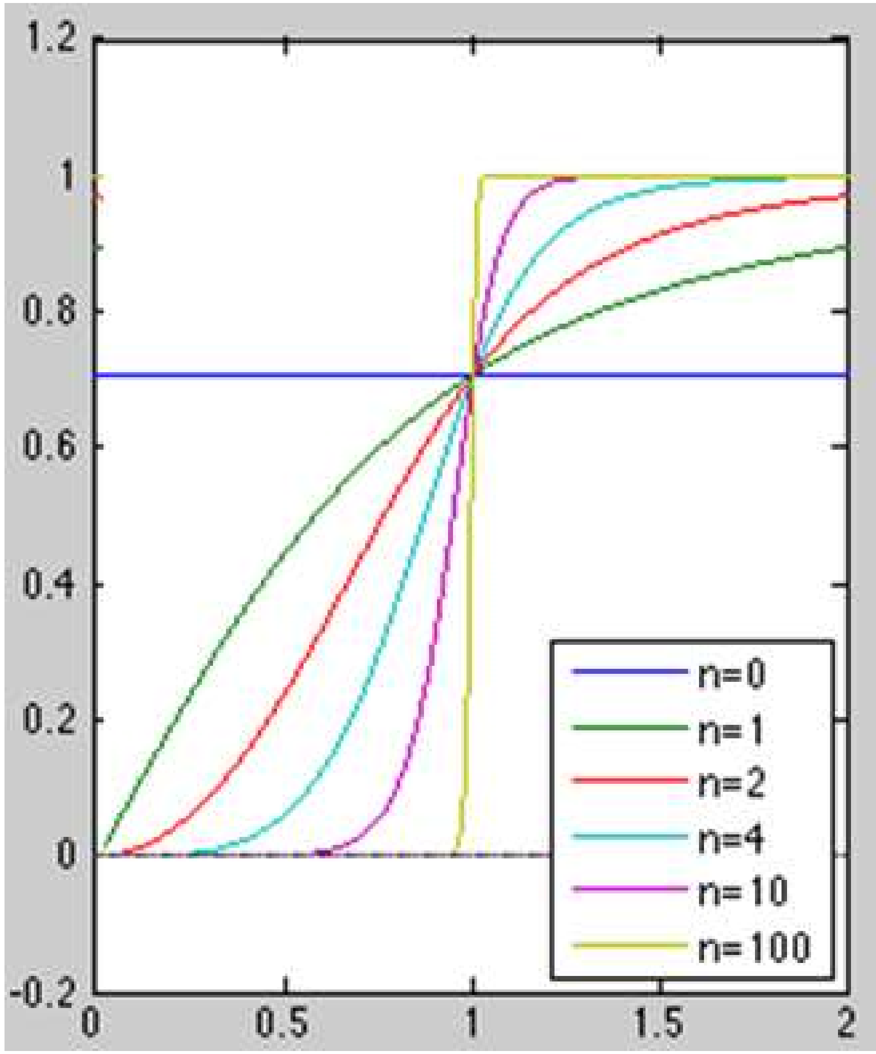
$$H(j\omega)=\frac{K}{\sqrt{1+\left(\frac{\omega}{\omega_0}\right)^{2n}}}$$

n	Normalised Denominator Polynomials in Factored Form (Each coefficient=3-K)				
1	$(1+s)$				
2	$(1+1.414s+s^2)$				
3	$(1+s)$	$(1+s+s^2)$			
4	$(1+0.765s+s^2)$	$(1+1.848s+s^2)$			
5	$(1+s)$	$(1+0.618s+s^2)$	$(1+1.618s+s^2)$		
6	$(1+0.518s+s^2)$	$(1+1.414s+s^2)$	$(1+1.932s+s^2)$		
7	$(1+s)$	$(1+0.445s+s^2)$	$(1+1.247s+s^2)$	$(1+1.802s+s^2)$	
8	$(1+0.390s+s^2)$	$(1+1.111s+s^2)$	$(1+1.663s+s^2)$	$(1+1.962s+s^2)$	
9	$(1+s)$	$(1+0.347s+s^2)$	$(1+s+s^2)$	$(1+1.532s+s^2)$	$(1+1.879s+s^2)$
10	$(1+0.313s+s^2)$	$(1+0.908s+s^2)$	$(1+1.414s+s^2)$	$(1+1.782s+s^2)$	$(1+1.975s+s^2)$

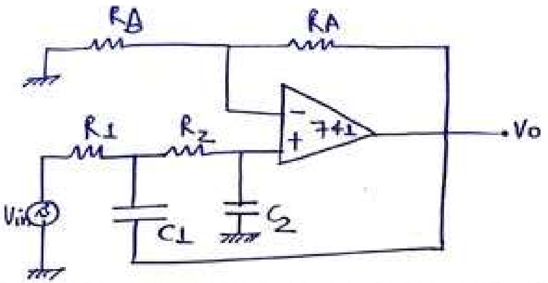
Butterworth LPF



Butterworth HPF



Design a 2nd order Butterworth LPF with f_c = 1kHz



$$H(s) = \frac{K}{1 + R_1 R_2 C_1 C_2 s^2 + [(1-K) R_1 C_1 + R_1 C_2 + R_2 C_2] s}$$

$$\frac{V_0}{V_I} = \frac{K}{(sCR)^2 + sCR(3-K) + 1}$$

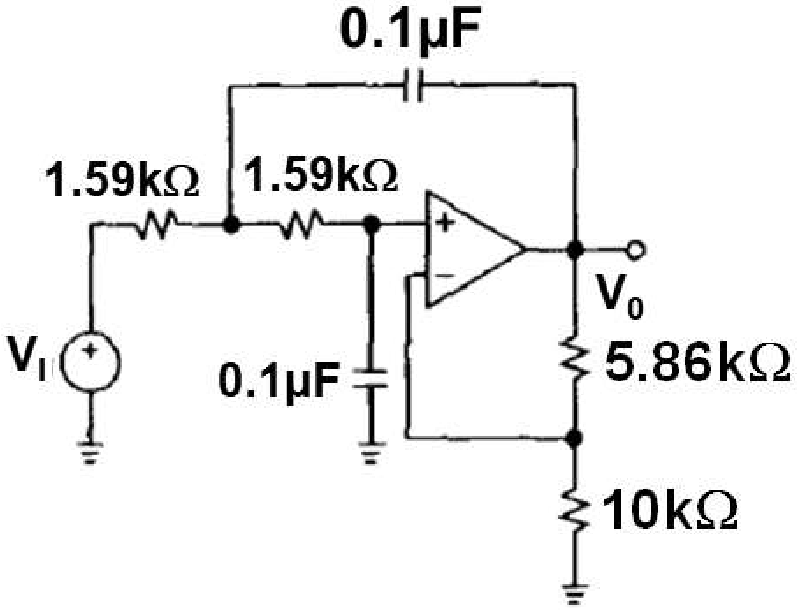
Let C = 0.1µF → $\omega_0 = \frac{1}{CR} \Rightarrow R = \frac{1}{2\pi \times 1 \times 10^3 \times 0.1 \times 10^{-6}} = 1.59k\Omega$

$(1+1.414s+s^2)$ → 3 – K = 1.414 → K = 1.586

$$\Rightarrow \frac{V_0}{V_I} = \frac{1.586}{(sCR)^2 + 1.414sCR + 1}$$

$$K = 1.586 = 1 + \frac{R_B}{R_A} \Rightarrow R_B = R_A(K-1)$$

Let R_A = 10kΩ
→ R_B = 5.86kΩ



Design a 4th order Butterworth LPF with f_c = 1kHz

$(1+0.765s+s^2)$ $(1+1.848s+s^2)$ $\quad \frac{V_0}{V_I} = \frac{K}{(sCR)^2 + sCR(3-K) + 1}$

$(1+0.765s+s^2) \rightarrow 3 - K_1 = 0.765 \rightarrow K_1 = 2.235$

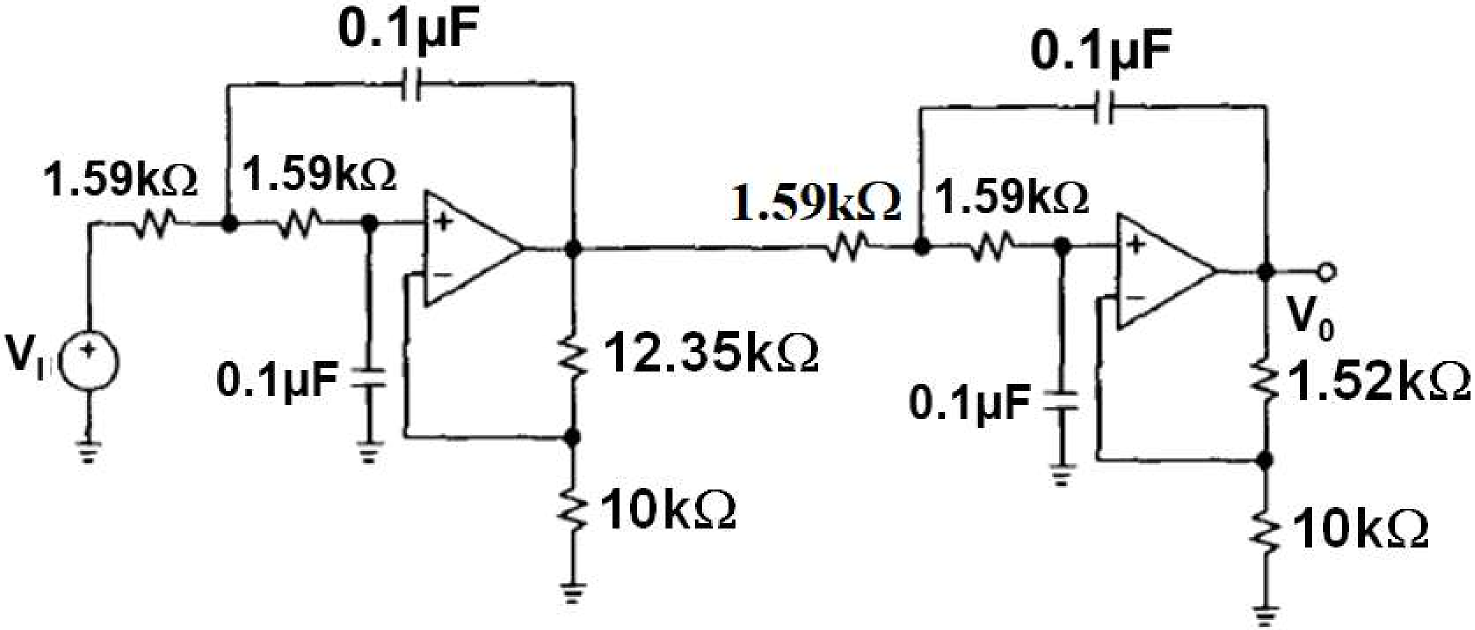
$(1+1.848s+s^2) \rightarrow 3 - K_2 = 1.848 \rightarrow K_2 = 1.152$

Let C = 0.1µF → $\quad \omega_0 = \frac{1}{CR} \Rightarrow R = \frac{1}{2\pi \times 1 \times 10^3 \times 0.1 \times 10^{-6}} = 1.59k\Omega$

$K_1 = 2.235 = 1 + \frac{R_B}{R_A} \Rightarrow R_B = 1.235R_A$ $\quad$ Let R_{A1} = 10kΩ
→ R_{B1} = 12.35kΩ

$K_2 = 1.152 = 1 + \frac{R_B}{R_A} \Rightarrow R_B = 0.152R_A$ $\quad$ Let R_{A2} = 10kΩ
→ R_{B2} = 1.52kΩ

Resulting gain $K = K_1 \times K_2 = 2.575$


0.1μF
0.1μF
1.59kΩ
1.59kΩ
1.59kΩ
1.59kΩ
V_I
0.1μF
12.35kΩ
0.1μF
V_0
1.52kΩ
10kΩ
10kΩ

LIC: LECTURE
Static and Dynamic Op Amp Limitations

- **INPUT OFFSET VOLTAGE:**

  (i) Error Caused by Vos

  (ii) Effective CMRR: Impact of Vos

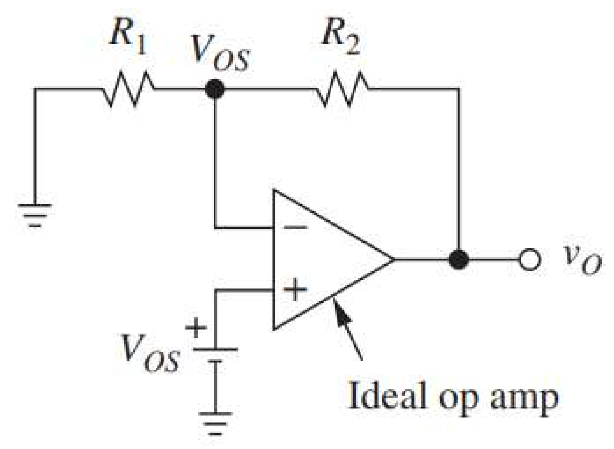
  (iii) Power Supply Rejection Ratio: Impact of Vos

  (iv) Examples

POWER SUPPLY REJECTION RATIO: PSRR

- If we change one of the op amp supply voltages V_S by a given amount ΔV_S the operating points of the internal transistors will be altered, generally causing a small change in V_O. By analogy with the CMRR, we model this phenomenon with a change in the input offset voltage.
- Which we express in terms of the power-supply rejection ratio (PSRR) as $(1/\text{PSRR}) \times \Delta V_S$

$$\frac{1}{\text{PSRR}} = \frac{\partial V_{OS}}{\partial V_S}$$



(a)

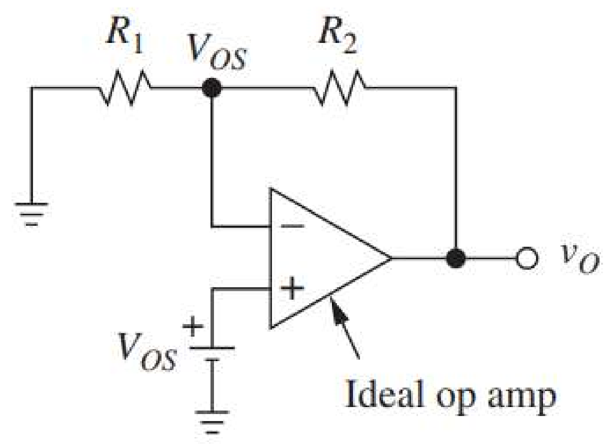
- The parameter 1/PSRR represents the change in V_{OS} brought about by a 1-V change in V_S, and is expressed in microvolts per volt. Like the CMRR, the PSRR deteriorates with frequency

POWER SUPPLY REJECTION RATIO: PSRR

Some data sheets give separate PSRR ratings, one for changes in V_{CC} and the other for changes in V_{EE}. Others specify the PSRR for V_{CC} and V_{EE} changing symmetrically. The PSRR_{dB} ratings of most op amps fall in the range of 80 dB to 120 dB. The devices of superior matching usually offer the highest PSRRs.

the 1/PSRR ratings for the 741C, which are given for symmetric supply changes, are 30 μV/V typical, 150 μV/V maximum. This means that changing, for instance, the supply voltages from ± 15 V to ± 12 V yields $\Delta V_{OS} = (1/\text{PSRR})\Delta V_S = (30\ \mu\text{V})(15 - 12) = \pm 90\ \mu\text{V}$ typical, $\pm 450\ \mu$V maximum. The OP77 op amp has $1/\text{PSRR} = 0.7\ \mu$V/V typical, 3 μV/V maximum.

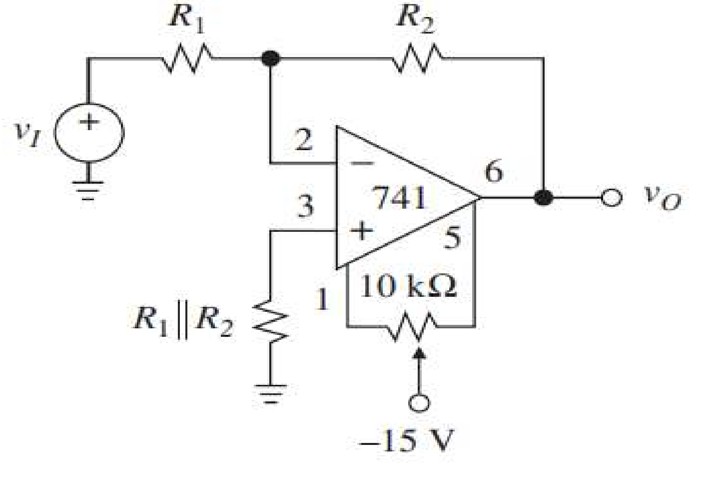
EXAMPLE A 741 op amp is connected as in Fig. 5.13*a* with $R_1 = 100\ \Omega$ and $R_2 = 100\ \text{k}\Omega$. Predict the typical as well as the maximum ripple at the output for a power-supply ripple of 0.1 V (peak-to-peak) at 120 Hz.



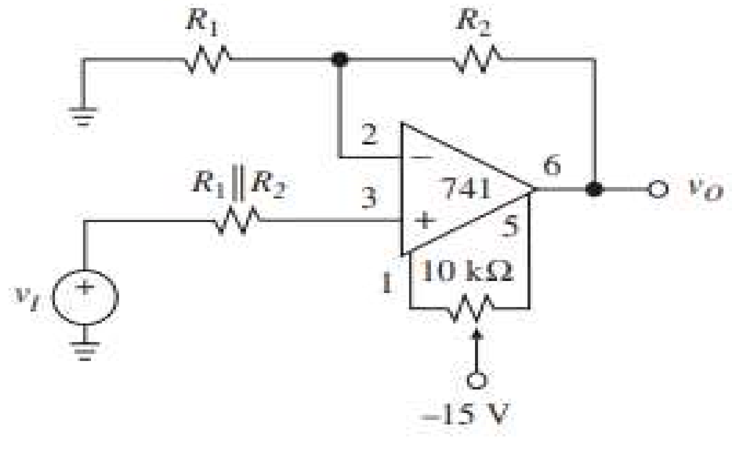
(*a*)

INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES

- Compensation for input offset voltage is required when we want to make it absolutely sure that output voltage is zero for a zero input voltage.
- 741-opamp provide offset compensation pins to nullify the input offset voltage [Pin 1 & 5]



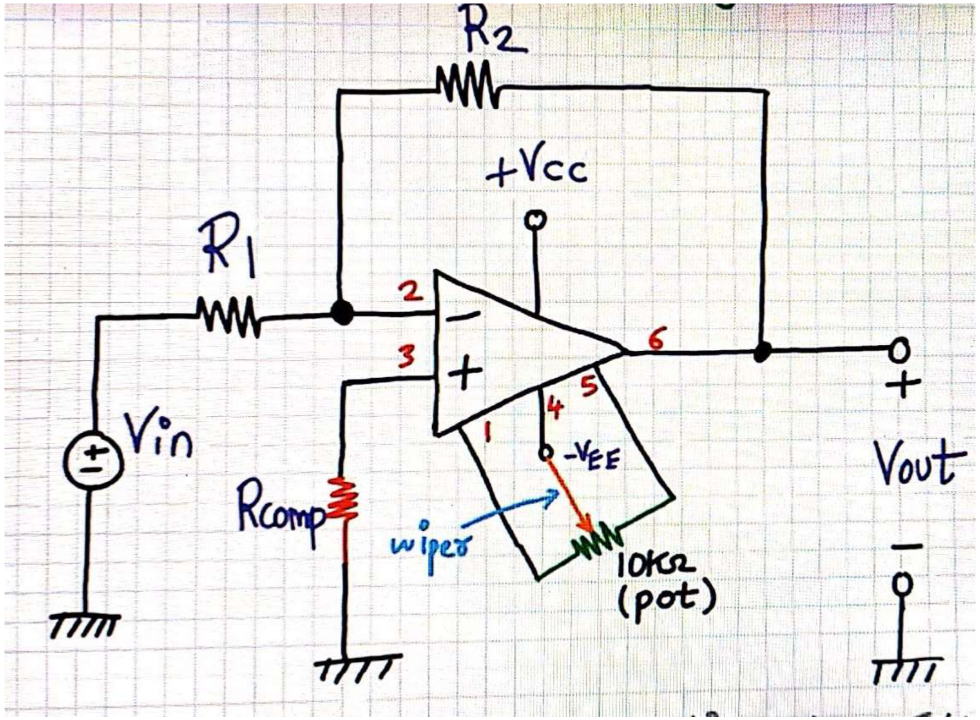
(a)



(b)

INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES

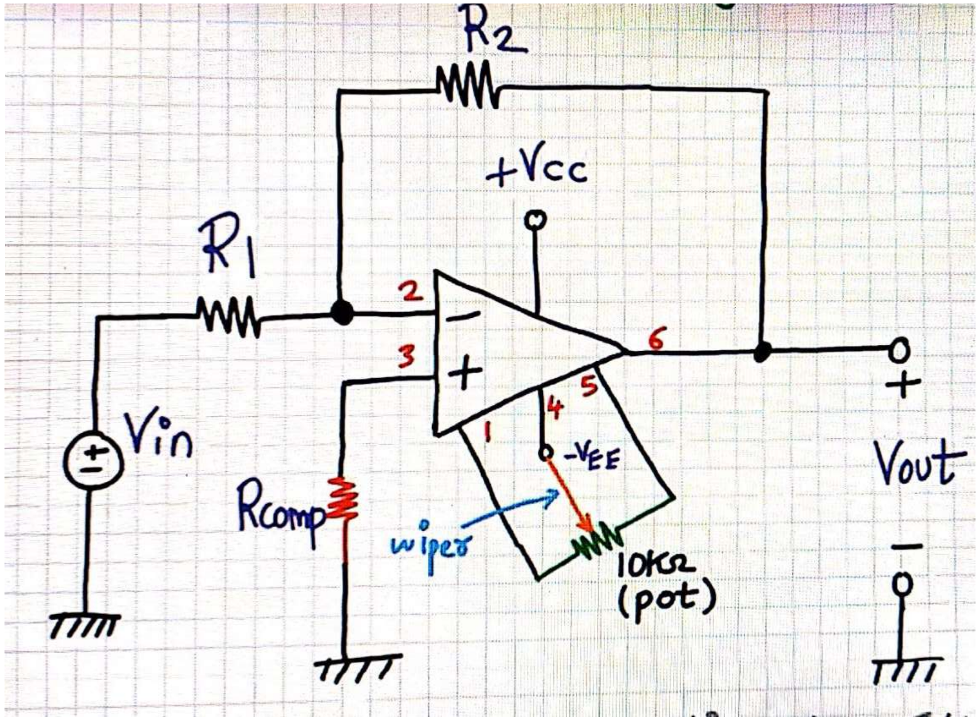
- Here, a 10 KΩ potentiometer is placed across the offset null pins 1&5 and the wiper (variable terminal of pot) is connected to pin no. 4 (i.e. negative supply pin 4).
- The position of the wiper is adjusted to nullify the output offset voltage/output output offset error.



INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES

$$v_O = A_s v_I + E_O$$

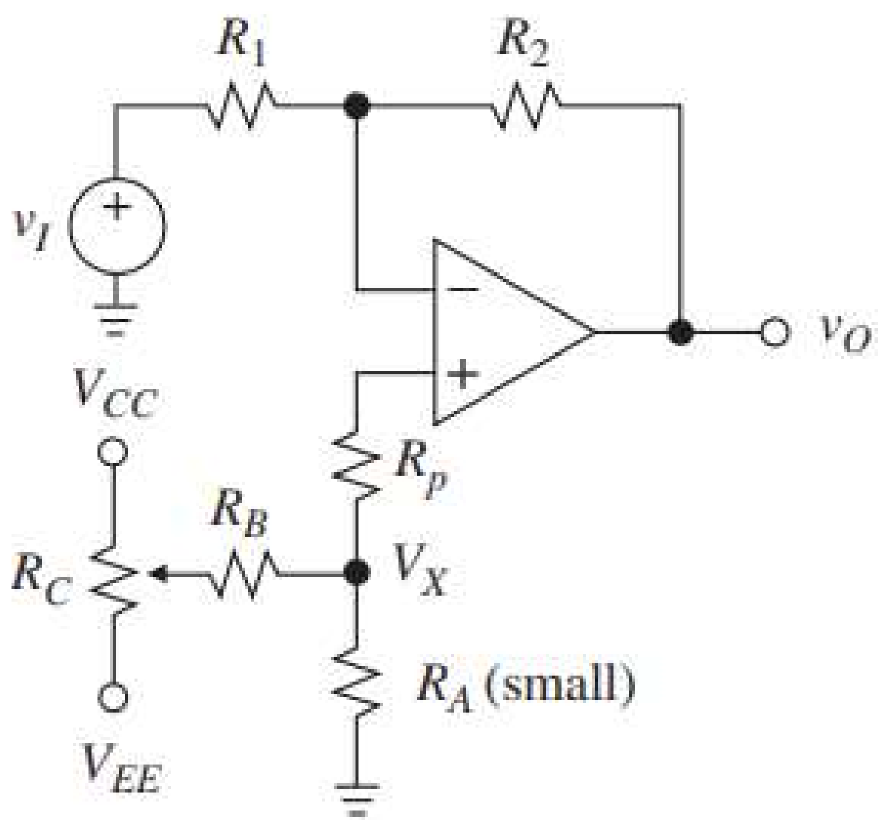
$$E_O = \left(1 + \frac{R_2}{R_1}\right)[V_{OS} - (R_1 \parallel R_2) I_{OS}] = \frac{1}{\beta} E_I$$



Where **As = −R2/R1** for the inverting amplifier, and **As = 1 + R2/R1** for the noninverting one. We call As the signal gain to distinguish it from the **dc noise gain, which is 1/β = 1 + R2/R1** for both circuits. Moreover, **EI = Vos − (R1||R2)Ios is the total offset error** referred to the input, and **Eo the total offset error referred to the output.**

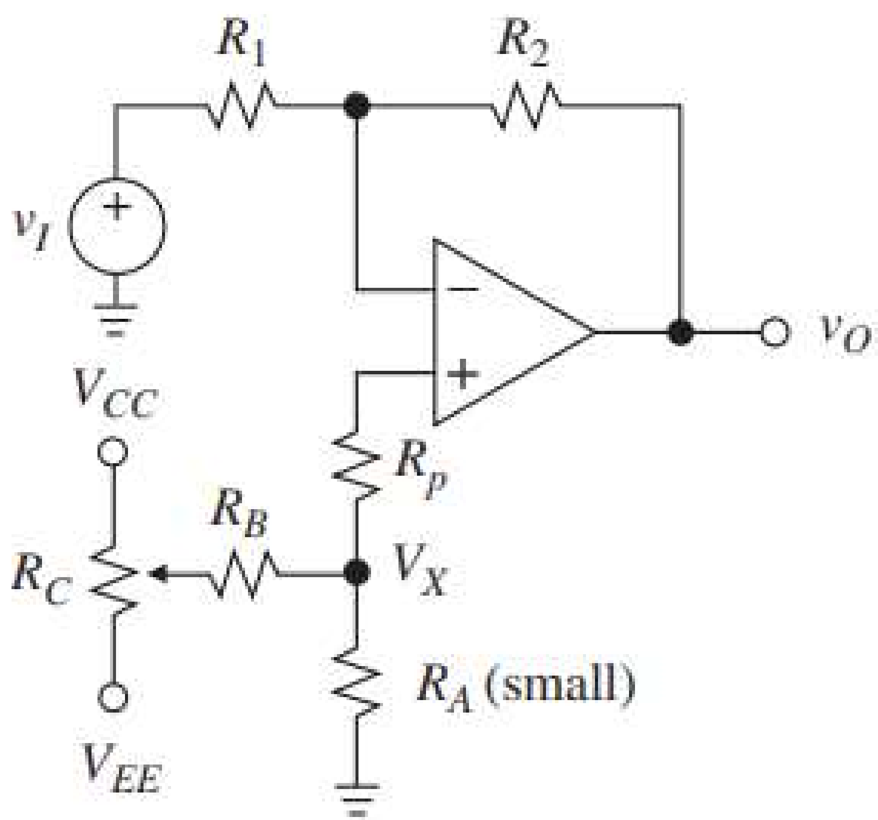
INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES: EXTERNAL NULLING

- External nulling is based on the injection of an adjustable voltage or current into the circuit to compensate for its offset error. This scheme does not introduce any additional imbalances in the input stage, so there is no degradation in drift, CMRR, or PSRR.
- The most convenient point of injection of the correcting signal depends on the particular circuit. For inverting-type configurations like the amplifier, we simply lift Rp off ground and return it to an adjustable voltage Vx .



INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES: EXTERNAL NULLING

- we now have an apparent input error of E_I + V_X, and we can always adjust V_X to neutralize EI
- VX is obtained from a dual reference source, such as the supply voltages if they are adequately regulated and filtered.
- In the circuits shown, we impose RB>>RC to avoid excessive loading at the wiper, and RA<<Rp to avoid perturbing the existing resistance levels.



INPUT OFFSET ERROR AND COMPENSTATION TECHNIQUES: EXTERNAL NULLING

In principle, the foregoing scheme can be applied to any circuit that comes with a dc return to ground. In the circuit of Fig. , R_1 has been lifted off ground and returned to the adjustable voltage V_X. To avoid upsetting the signal gain, we must impose $R_{eq} \ll R_1$, where R_{eq} is the equivalent resistance of the nulling network as seen by R_1 (for $R_A \ll R_B$ we have $R_{eq} \cong R_A$.) Alternatively, we must decrease R_1 to the value $R_1 - R_{eq}$.

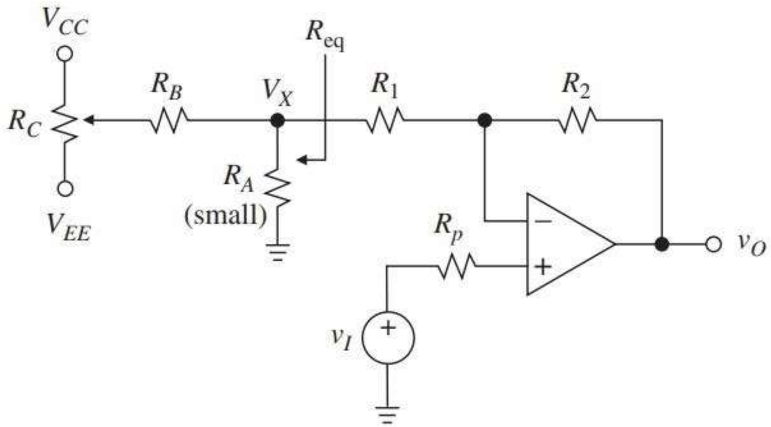


FIGURE
External offset-error nulling for the noninverting amplifier.

LIC: LECTURE
Static and Dynamic Op Amp Limitations

- **DYNAMIC OP-AMP LIMITATIONS: INTRODUCTION**

  (i) Open Loop Response

  (ii) Single Pole Open-loop Gain

  (iii) Transient Response

  (a) Rise Time

  (b) Slew Rate- Definition and Limitations

DYNAMIC OP-AMP LIMITATIONS

- Till now, we have assumed op-amps with extremely high open-loop gains, regardless of frequency.

- A practical op amp provides high gain only from dc up to a given frequency, beyond which gain decreases with frequency and the **output is also delayed with respect to the input**.

- These limitations have a profound impact on the closed-loop characteristics of a circuit: they **affect both its frequency and transient responses**, and also its input and output impedances.

DYNAMIC OP-AMP LIMITATIONS

OPEN LOOP RESPONSE

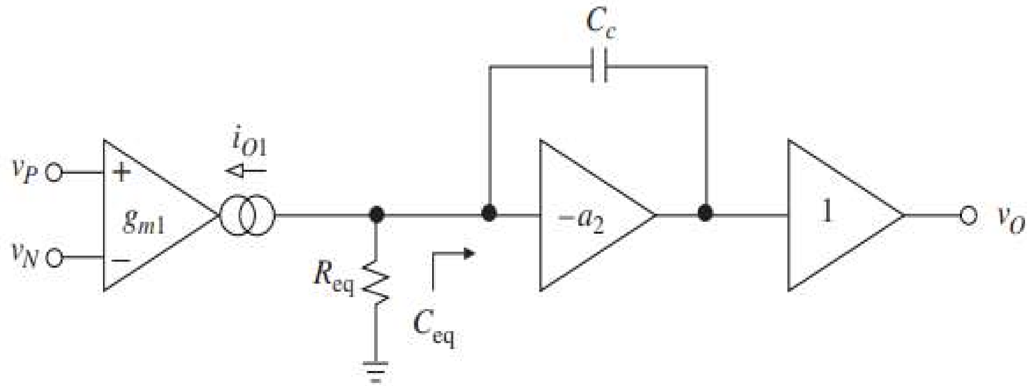
- The most common open-loop response is the dominant-pole response.
- As, its frequency profile is primarily controlled by a single pole.
- At low frequencies, where ***Cc*** acts as an open circuit:

$$V_0 = 1 \times -a_2 \times (-Req.i_{o1})$$

$$V_0 = g_{m1} Req a_2 \times (V_p - V_n)$$

- The low-frequency gain, called the dc gain and denoted as a_0,

$$a_0 = g_{m1} R_{eq} a_2$$



Simplified Op-Amp Block diagram

gm1=**transconductance gain of the first stage**

−a2=**voltage gain of the second stage (inverting stage)**

Req, Ceq= **Net equivalent resistance and capacitance between the node common to the first and second stage, and ground.**

DYNAMIC OP-AMP LIMITATIONS

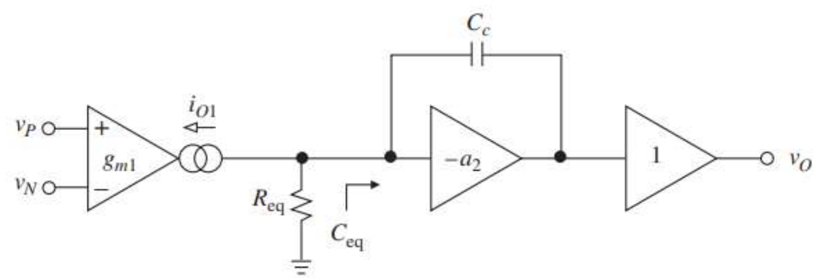
OPEN LOOP RESPONSE

- Increasing the operating frequency will bring the impedance of *Ceq* into the circuit, causing gain to roll off with frequency because of the LPF action provided by *Req* and *Ceq*.
- Gain starts to roll off at the frequency f_b that makes:

$$|Z_{Ceq}| = R_{eq} \quad \text{or} \quad R_{eq} = \frac{1}{2\pi f_b C_{eq}}$$

$$f_b = \frac{1}{2\pi R_{eq} C_{eq}}$$

This frequency (f_b), is called as dominant-pole frequency.



Simplified Op-Amp Block diagram

gm1**=transconductance gain of the first stage**

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DYNAMIC OP-AMP LIMITATIONS

Single Pole Open Loop gain

- Assume that the open-loop gain a(s) possesses just a single pole; this, both to facilitate our mathematical manipulation and to help us develop a basic feel for the effect of the gain roll-off on the closed-loop parameters. Gain can be written as:

$$a(s) = \frac{a_0}{1 + {}^{s}/_{w_b}}$$

Where, S is the complex frequency
a_0 =Open loop DC gain
w_b = s-plane pole location

or

In terms of frequency:

$$a(if) = \frac{a_0}{1 + {}^{if}/_{f_b}}$$

where j is the imaginary unit ($j^2 = -1$), and $f_b = \omega_b/(2\pi)$ is the *open-loop –3-dB frequency*, also called the *open-loop bandwidth*. We calculate gain magnitude and phase as

$$|a(jf)| = \text{mag}\, a(jf) = \frac{a_0}{\sqrt{1 + (f/f_b)^2}}$$

$$\measuredangle a(jf) = \text{ph}\, a(jf) = -\tan^{-1}(f/f_b)$$

DYNAMIC OP-AMP LIMITATIONS

Single Pole Open Loop gain

- The gain is high and approximately constant only from dc up to f_b. Past f_b it rolls off at the rate of −20 dB/dec, until it drops to 0 dB (or 1 V/V) at $f=f_t$.
- **This frequency f_t** is called the **unity-gain frequency**, or also the transition frequency.
- Because it marks the transition from amplification (positive decibels) to attenuation (negative decibels).

$$1 = \frac{a_0}{\sqrt{1+(ft/f_b)^2}}$$ and $f_t >> f_b$

So,

$$1 = \frac{a_0}{ft/f_b}$$ We get, $$\boldsymbol{f_t = a_0 f_b}$$

- Over the frequency region $f >> f_b$ the op amp behaves as an integrator, and that its gain-bandwidth product, defined as ***GBP* = |*a*(*j f*)|×*f* , is constant.**

So, $GBP = f_t$ For; $f >> f_b$

- For this reason, op amps with dominant-pole compensation are also referred to as constant-GBP op amps.
- **Increasing (or decreasing) *f* by a given amount in the region of integrator behavior will decrease (or increase) |*a*| by the same amount.**

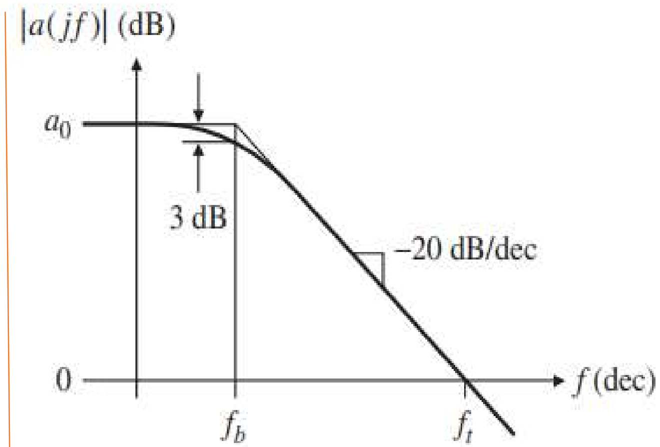


Fig. Single-pole open-loop gain

TRANSIENT RESPONSE

- That is, the response to an input step as a function of time.
- This response, like its frequency domain counterpart, varies with the amount of feedback applied.

Calculation of Rise Time t_R : Small signal B.W. of voltage follower is ft. The freq. response can be written as:

$$A(jt) = \frac{1}{1 + {jf}/{f_t}}$$

Where, S=-$2\pi f t$ $(pole)$

- Subjecting the voltage follower of Fig. to an input voltage step of sufficiently small amplitude ***Vm*** will result in the well-known exponential response

$$\boldsymbol{V_0(t) = V_m(1 - e^{-t/\tau})}$$

$$where, \tau = \frac{1}{2\pi f_t}$$

- The time t_R it takes for '**Vo**' to swing from 10% to 90% of '**Vm**' is called the rise time.
- It provides an indication of how rapid the exponential swing is. So,

t_R = τ(ln 0.9 – ln 0.1), or

$$t_R = \frac{0.35}{f_t}$$

- **This provides a link between the frequency-domain parameter f_t and the time domain parameter t_R.**

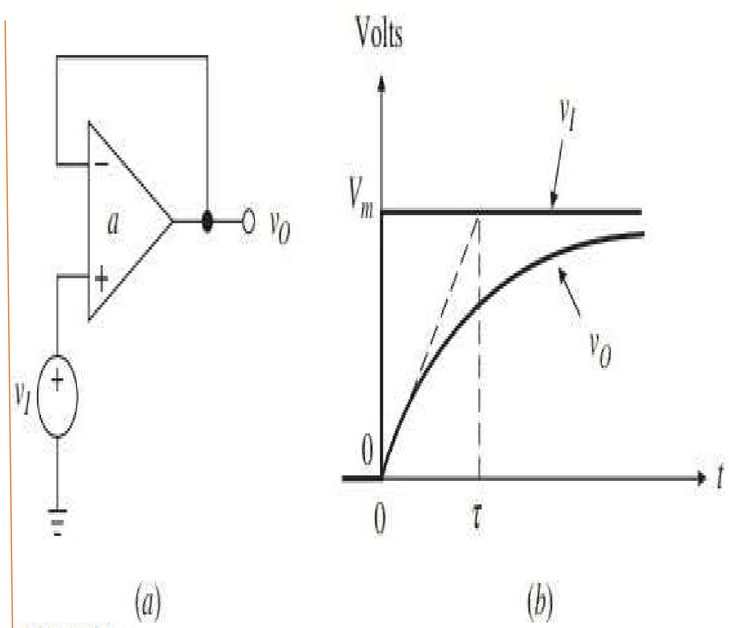
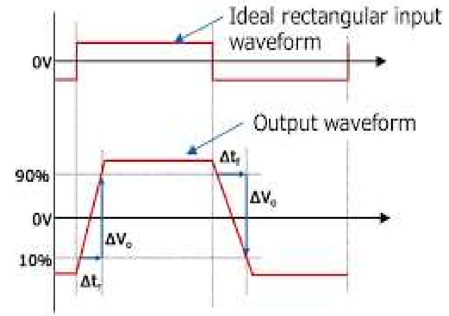


FIGURE
Voltage follower and its small-signal step response.

TRANSIENT RESPONSE



Calculation of Slew Rate (SR=ΔV / Δ t) : Max. rate of change of an op-amps output per unit time (change of voltage or current per unit time)

SLEW RATE LIMITATIONS: The rate at which ***Vo*** changes with time is highest at the beginning of the exponential transition.

$$V_0(t) = V_m(1 - e^{-t/\tau}) \qquad \frac{dV_0}{dt} = \frac{Vm}{\tau} \quad \textit{(at t=0)}$$

- If we increase V_m, the rate at which the output slews will have to increase accordingly in order to complete the 10%-to-90% transition within the time t_R.
- In practice it is observed that above a certain step amplitude the output slope saturates at a constant value called the slew rate (SR).The output waveform, rather than an exponential curve, is now a ramp.
- ***SR*** is a nonlinear large-signal parameter, while t_R is a linear small signal parameter.

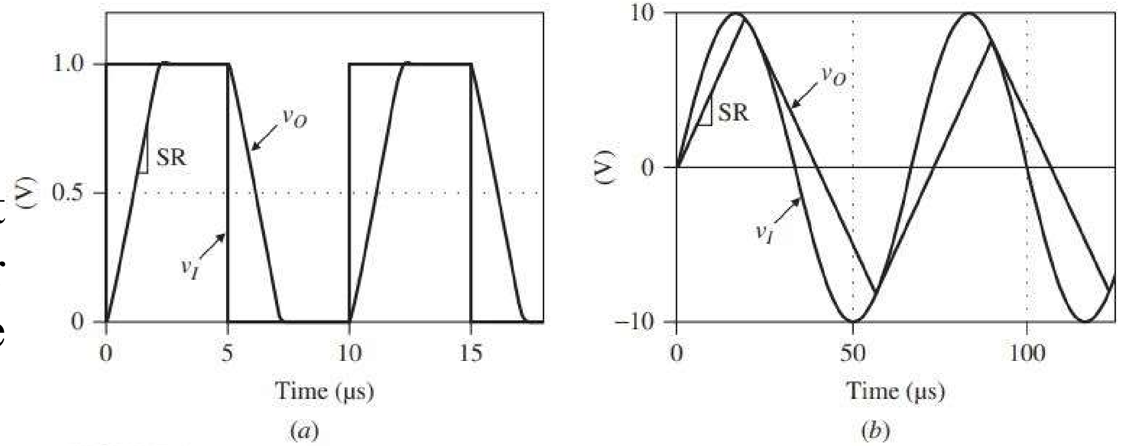


FIGURE
Slew-rate limited responses of the 741 follower (*a*) a pulse and (*b*) a sinusoid.

- The output-step magnitude corresponding to the onset of slew-rate limiting is such that Vom(crit)/τ = SR.

$$V_{om(\text{crit})} = \frac{SR}{2\pi f_t}$$